Technical Digest

12th Optoelectronics and Communications Conference

16th International Conference on Integrated Optics and Optical Fiber Communication

July 9-13, 2007
Pacifico Yokohama
Yokohama
Kanagawa, Japan

Co-sponsored by
IEICE Communications Society
IEICE Electronics Society

Technically Co-sponsored by
IEEE Communications Society
IEEE Lasers and Electro-Optics Engineering
Optical Society of America
SPIE
Dear OECC/IOOC2007 Attendees:

Welcome to Yokohama and the 12th Optoelectronics and Communications Conference and the 16th International Conference on Integrated Optics and Optical Fiber Communication – OECC/IOOC2007! We hope that you will enjoy both the excellent technical content and recreational parts of the meeting. OECC, which is held annually in the Asian Pacific region, will provide an international forum to present and discuss significant progress in research, development, and application of optical communication systems and optoelectronics. Especially, this year, the OECC will be held with the IOOC, which has a history of 30 years in this field circulating in Asian Pacific, Europe and North America.

In OECC/IOOC2007 a total of 381 papers are to be presented in both oral and poster sessions which will stimulate active discussions among the participants. The Plenary Session of the conference begins with three excellent invited Plenary Talks, Dr. Tatsuo Izawa, NTT Electronics Corp., and former President of IEICE, on “Fiber Technology in the Early days and the Future”; by Dr. Nim K. Cheung, Telcordia Technologies, and President of IEEE Communications Society, on “The Challenge of the Last 100 Meters – Access Technologies in 2007” and by Prof. Jintong Lin, President of Beijing University of Post and Telecommunications, on “Telecommunications in China”.

Following the Plenary Session, OECC/IOOC2007 will continue with many exciting contributed technical talks and posters. The main technical program consists of 204 Regular Oral, 107 Poster, and 70 Invited papers arranged in 66 technical sessions. The Technical Program Committee, Co-chaired by M. Shikada and Y. Yoshikuni has done an excellent job arranging the program.

Related Workshops will be held at the same conference site on Monday, July 9. The themes of the workshops will cover hot topics as follows: 1) 3rd International Workshop on the Future of Optical Networking, 2) Recent Progress in Integrated Optics and their Applications and 3) IEEE/LEOS Workshop for Advanced Optics.

Finally we are very happy to acknowledge all of the kind support and contributions from many individuals who spent their time and resources to make the OECC/IOOC2007 a great success. We also acknowledge the financial contributions of many foundations, and all attendees for both technical and financial contributions without whose support, this conference could not be held. On behalf of the Organizing Committee, we hope you enjoy both technical part of OECC/IOOC2007 and wonderful Yokohama.

Dr. Hideo Kuwahara, Prof. Fumio Koyama, Prof. Tetsuya Miki
Co-Chairs, Organizing Committee of OECC / IOOC 2007
Committee Members

ORGANIZING COMMITTEE

Organizing Committee
Co-Chairs: H. Kuwahara (Fujitsu Labs.), F. Koyama (Tokyo Inst. of Technol.), T. Miki (Univ. of Electro-Communs.)

General Affairs Committee
Chair: K. Oguchi (Seikei Univ.)
Members: S. Watanabe (Fujitsu Labs.), N. Kishi (Univ. of Electro-Communs.), T. Sugawara (Hitachi), K. Himeno (Fujikura)

Treasury Committee
Chair: H. Ogoshi (Furukawa Electr.)
Members: S. Shikii (Optohub), M. Yoshida (Kinki Univ.), M. Fuse (Matsushita Electr.)

Publication & Registration Committee
Chair: M. Koga (Oita Univ.)
Members: K. Sasaki (Oki Electr.), K. Tanaka (KDDI R&D Labs.), H. Hasegawa (Nagoya Univ.)

Local Arrangement Committee
Chair: M. Nishimura (Sumitomo Electr.)
Members: H. Yusa (Sumitomo Electr.), T. Ito (NEC), K. Ishida (Mitsubishi Electr.)

IOOC Special Program Committee
Chair: T. Miki (Univ. of Electro-Communs.)
Members: K. Iga (JSPS), T. Izawa (NEL), T. Aoyama (Keio Univ.), K. Kobayashi (Tokyo Inst. of Technol.), J. Yoshida (Chitose Inst. of Sci. & Technol.), S. Akiba (KDDI R&D Labs.), K. Kikuchi (The Univ. of Tokyo), N. Chinone (Aichi Univ. of Technol.)

TECHNICAL PROGRAM COMMITTEE

Co-Chairs: M. Shikada (NEC), Y. Yoshikuni (Kitasato Univ.), T. Mizuochi (Mitsubishi Electr.), S. Matsuo (NTT)

1: Optical Networks and Broadband Access
Chair: H. Morikawa (The Univ. of Tokyo)
Co-Chair: S. Hanatani (Hitachi)
Members: M. Jinno (NTT), M. Tsurusawa (KDDI R&D Labs.), N. Wada (NICT), S. Oshiba (Kyoto Inst. of Technol.), K. Fukuchi (NEC), Y. Ji (Beijing Univ. of Post and Telecommuns.), H. Kim (Samsung Electr.), R. Ranganathan (CIENA), F.-J. Westphal (GmbH Business Unit Technologiezentrum)

2: Transmission Systems and Switching Technologies
Chair: T. Morioka (NICT)
Co-Chair: M. Hanawa (Univ. of Yamanashi)
Members: M. Matsumoto (Osaka Univ.), J. Maeda (Tokyo Univ. of Sci.), H. Murai (Oki Electr.), T. Terahara (Fujitsu Labs.), H. Taga (Natl. Sun Yat-sen Univ.), P. K. A. Wai (Hong Kong Polytechnic Univ.), C.-S.Park (Gwangju Inst. of Sci. & Technol.), C. R. Menyuk (Univ. of Maryland Baltimore Country), K. Petermann (TU Berlin)

3: Optical Fibers, Cables and Fiber Devices
Chair: K. Kokura (Furukawa Electr.)
Members: M. Ohashi (Osaka Prefecture Univ.), A. Mori (NTT), M. Onishi (Sumitomo Electr.), S. Namiki (AIST) S. Matsuo (Fujikura), C. Martijn de Sterke (Univ. of Sydney), P. Shum (Nanyang Technol. Univ.), C. Lin (Chinese Univ. of Hong Kong), J. Knight (Univ. of Bath), T. Hanson (Corning)
4: Optical Active Devices and Modules
Chair: Y. Nakano (The Univ. of Tokyo)
Members: M. Aoki (Hitachi), H. Ishii (NTT), T. Katsuyama (Sumitomo Electr.),
K. Morito (Fujitsu Labs.), Y. H. Lee (KAIST), J. Coleman (Univ. of Illinois),
M. C. Amann (Technical Univ. of Munich)

5: Optical Passive Devices and Modules
Chair: A. Sugita (NTT)
Members: T. Suhara (Osaka Univ.), T. Mizumoto (Tokyo Inst. of Technol.),
H. Uetsuka (Hitachi Cable), O. Mikami (Tokai Univ.), H. Yamada (Tohoku Univ.),
S. Kobayashi (Chitose Inst. of Sci. & Technol.),
P. S. Chung (City Univ. of Hong Kong), T.H. Rhee (POINTek)

INTERNATIONAL ADVISORY COMMITTEE
Members: A. Ahmad (Telekom Malaysia), S. Akiba (KDDI R&D Lab.), T. Aoyama (Keio Univ.),
S. Arai (Tokyo Inst. of Technol.), Y.-X. Chen (Shanghai Jiao Tong Univ.), N. Chinone (Aichi Univ. of Technol.),
C. T. Chong (Data Storage Inst.), P. S. Chung (City Univ. of Hong Kong), Z. Deng (China Inst. of Communs.),
A. K. Ghatak (Indian Inst. of Technol.), J. Harvey (Univ. of Auckland), T. R. HSING (SPIE), K. Iga (JSPS),
S. S. Jian (Chitose Inst. of Sci. & Technol.), K. Kikuchi (The Univ. of Tokyo), C. M. Kim (Univ. Of Seoul),
B. Y. Kim (KAIST), K. Kobayashi (Tokyo Inst. of Technol.), J. D. Love (Australian Natl. Univ.),
Y.-F. Lu (Univ. of Nebraska Lincoln), T. Miki (Univ. of Electro-Communs.),
M. Nakamura (Hitachi), K. Oguchi (Seikei Univ.), M. Saruwatari (Natl. Defense Acad.),
P.T.C. Shih (Natl. Sci. Council), W. Surakamponthorn (King Mongkut's Inst. of Technol. Landkrabang),
R. S. Tucker (Univ. of Melbourne), J. Yoshida (CIST)

ADVISORY COMMITTEE
Chair: K. Kikuchi (The Univ. of Tokyo)
Advisors: T. Aoyama (Keio Univ.), S. Akiba (KDDI R&D Lab.), K. Iga (JSPS), T. Izawa (NEL),
N. Chinone (Aichi Univ. of Technol.), K. Kobayashi (Tokyo Inst. of Technol.), Y. Fukunaga (Hitachi),
T. Miki (Univ. of Electro-Communs.), K. Murano (Fujitsu), J. Yoshiida (CIST)
Members: Y. Inoue (IEICE/CS), H. Kuwahara (IEICE/OCS), T. Suhara (IEICE/LOE), K. Hotate (IEICE/ES),
M. Nishimura (IEICE/PE), Y. Yoshikuni (IEEE LEOs), N. Ohta (IEEE-COMSOC), R. Yamauchi (Fujikura),
K. Onaka (Matsushita Electr.), H. Kan (Hamamatsu Photonics), K. Oyamada (NHK),
M. Suzuki (KDDI R&D Labs.), Y. Tawara (MIC), K. Aoki (SCAT), Y. Mochida (Fujitsu Labs.),
K. G. Ravikumar (Corning International), Y. Miyajima (Sumitomo Elec.), K. Hagimoto (NTT),
H. Yanagawa (Furukawa Electr.), K. Kyuma (Mitsubishi Electr.), M. Kobayashi (Hitachi Cable),
H. Sugimoto (Oki Elec.), S. Nishimura (Hitachi), Y. Toriumi (Sony), R. Ito (OITDA),
S. Niki (Advantest Labs.), M. Watanabe (AIST), S. Suyama (NEC)

*NOTE
NICT: National Institute of Information and Communications Technology
AIST: Advanced Industrial Science and Technology
KAIST: Korea Advanced Institute of Science and Technology
JSPS: Japan Society for the Promotion of Science
CIST: Chitose Institute of Science and Technology
MIC: Ministry of Internal Affairs and Communications
OITDA: Optoelectronic Industry and Technology Development Association
Acknowledgments

OECC/IOOC 2007 is cosponsored by The Institute of Electronics, Information and Communication Engineers (IEICE) Communications Society and Electronics Society and technically cosponsored by IEEE/COMSOC, IEEE/LEOS, OSA and SPIE in cooperation with the following academic societies:

- The Japan Society of Applied Physics
- The Institute of Electrical Engineers of Japan
- The Institute of Image Information and Television Engineers
- Information Processing Society of Japan
- The Laser Society of Japan
- The Society of Instrument and Control Engineers
- IEEE COMSOC Japan Chapter
- IEEE LEOS Japan Chapter

and on support by Optoelectronics Industry and Technology Development Association (OITDA), Communications and Information Network Association of Japan (CIAJ) and Photonic Internet Forum (PIF).

OECC/IOOC2007 is financially supported by the following foundations:

- The Telecommunications Advancement Foundation
- The City of Yokohama
- Support Center for Advanced Telecommunications Technology Research, Foundation (SCAT)
- International Communications Foundation
- The Ogasawara Foundation for The Promotion of Science & Engineering
- The Murata Science Foundation
- Nippon Sheet Glass Foundation for Materials and Engineering
Tuesday, July 10

Room P  
**Plenary Session (9:45-12:00)**

9:45-10:30  Plenary Talk-1  
Fiber Technology in the Early days and the Future  
T. Izawa  
NTT Electronics Corp, and former President of the Institute of Electronics, Information and Communication Engineers (IEICE) of Japan ................................................................. 2

10:30-11:15  Plenary Talk-2  
The Challenge of the Last 100 Meters – Access Technologies in 2007  
N. Cheung  
Telcordia Technologies, and President of IEEE Communications Society, USA ....................... 4

11:15-12:00  Plenary Talk-3  
Telecommunications in China  
J. Lin  
President of Beijing University of Post and Telecommunications, China ................................. 6

Room A  
**10A1: PON Architectures (14:00-15:30)**  
Chair: Y. C. Chung

14:00-14:30  10A1-1 (Invited)  
Flexible Access Technologies to Upgrade PONs  
M. Tsubokawa  
NTT, Japan .................................................................................................................................. 8

14:30-14:45  10A1-2  
Improvement of TDM-PON Upstream Bandwidth Utilization Adopting Two Wavelengths  
J. Chung(1), J. Park(1), Y. Park(1), B.-W. Kim(2)  
(1) Kookmin Univ., Korea, (2) Electr. and Telecommuns. Res. Inst., Korea .................................... 10

14:45-15:00  10A1-3  
Limited Bandwidth Allocation with Sliced Credit in Gigabit Passive Optical Network  
Y. Shi, H. Yoshiuchi  
Hitachi R&D, China ....................................................................................................................... 12

15:00-15:15  10A1-4  
Impact of Polarization Mode Dispersion on Analog Video Transmission over PON System  
Y. Akasaka, P. Palacharla, M. Bouda, T. Naito  
Fujitsu Labs. of America, USA ......................................................................................................... 14

15:15-15:30  10A1-5  
Properties of Optical Modulator Cascade for Multi-Channel Transmission Systems  
K. Kikushima, T. Fujiwara  
NTT, Japan .................................................................................................................................... 16

10A2: WDM-PON (16:00-17:30)  
Chair: TBD

16:00-16:30  10A2-1 (Invited)  
A Review on WDM PON Technologies  
Y. C. Chung  
Korea Advanced Inst. of Sci. and Technol., Korea ........................................................................... 18

16:30-17:00  10A2-2  
Performance of Wavelength Shared Hybrid PON  
P. Palacharla, M. Bouda, Y. Akasaka, A. Umnov, T. Naito  
Fujitsu Labs. of America, USA ........................................................................................................ 20

16:45-17:00  10A2-3  
Virtual Private Group Formation in a WDM Passive Optical Network  
Q. Zhao, C.-K. Chan  
The Chinese Univ. of Hong Kong, China ......................................................................................... 22

17:00-17:15  10A2-4  
A Novel Wavelength Router Architecture for Optical VPNs Using Multistage AWGs in a WDM-PON  
K. Okada, S. Terada, K. Oguchi  
Seikei Univ., Japan .......................................................................................................................... 24

17:15-17:30  10A2-5  
Demonstration of an 80-Channel DWDM-PON Based on Wavelengthlocked Fabry-Perot Laser Diodes  
Korea Advanced Inst. of Sci. and Technol., Korea ........................................................................... 26
Room B

10B1: High Capacity Transmission Systems (14:00-15:30)
Chair: T. Morioka

14:00-14:30 10B1-1 (Invited) Ultra High Capacity WDM Transmission Using CSRZ-DQPSK Format and Extended L-Band Optical Amplifiers
A. Sano, H. Masuda, E. Yoshida, Y. Miyamoto
NTT, Japan ................................................................. 28

14:30-14:45 10B1-2 Performance Improvement of DPSK Signal Transmission by a Phase-Preserving Amplitude Limiter
T. Kamio, K. Sanuki, M. Matsumoto
Osaka Univ., Japan ........................................................ 30

14:45-15:00 10B1-3 Suppression of BER Fluctuations by Electrical Equalizer in the Ultra-Long 43Gb/s Transmission Experiments
T. Ito
NEC, Japan ................................................................. 32

15:00-15:30 10B1-4 (Invited) Statistics of PMD Outages
M. Brodsky
AT&T Labs. Research, USA ........................................... 34

10B2: Novel Transmission Systems (16:00-17:30)
Chair: H. Murai

16:00-16:30 10B2-1 (Invited) Ultrahigh-Speed Signal Transmission / Processing Technologies
T. Hirooka, M. Nakazawa
Tohoku Univ., Japan ....................................................... 36

16:30-16:45 10B2-2 Low Power-Penalty 160 Gbit/s DPSK Transmission over 600 km
K. Osawa, M. Okazaki, T. Hirooka, M. Nakazawa
Tohoku Univ., Japan ....................................................... 38

16:45-17:00 10B2-3 100 Gb/s Ethernet over Multimode Fiber Based on MIMO with Spatial Pre-Coding
M. Greenberg(1), M. Nazarathy(2), M. Orenstein(2)
(1)Student Member IEEE, (2)Senior Member IEEE, Member OSA .................................................. 40

17:00-17:15 10B2-4 Increased Transmission Bandwidth of Multimode Fiber by Using Mode-Field Matched Center Launching Technique
D. H. Sim, Y. Takushima, Y. C. Chung
Korea Advanced Inst. of Sci. and Technol., Korea .................................................. 42

17:15-17:30 10B2-5 10 Gb/s-2 km Photonic Crystal Fiber Transmission at 850 nm with a Directly Modulated Single-Mode VCSEL
H. Hasegawa, Y. Oikawa, M. Nakazawa
Tohoku Univ., Japan ....................................................... 44

Room C

10C1: Highly Nonlinear Fibers and Applications (14:00-15:30)
Chair: A. Maruta, S. Ramachandran

14:00-14:30 10C1-1 (Invited) Recent Progress on Silica-Based Highly Nonlinear Fiber
M. Hirano, T. Nakanishi, T. Okuno, M. Onishi
Sumitomo Electric Inds., Japan ......................................... 46

14:30-15:00 10C1-2 (Invited) Nonlinearity and Dispersion Characteristics of Bismuth-Based Highly Nonlinear Fiber
N. Sugimoto, T. Nagashima, T. Hasegawa, S. Ohara
Asahi Glass, Japan .......................................................... 48

15:00-15:30 10C1-3 (Invited) Highly Nonlinear Non-Silica Fibers for Efficient Slow Light Devices
K. S. Abedin, G.-W. Lu, T. Miyazaki
NICT, Japan ................................................................. 50

10C2: Photonic Crystal Fibers (I) (16:00-17:15)
Chair: P. Nouchi

16:00-16:30 10C2-1 (Invited) Resonant Coupling in Photonic Crystal Fibers
K. Saitoh, S. K. Varshney, M. Koshiba
Hokkaido Univ., Japan ..................................................... 52
16:30-16:45  10C2-2 Periodic Refocusing in Microstructured Fiber for Efficient Quasi-Phase Matched Optical Nonlinearity
M. Hisatomi(1), M. C. Parker(1), S. D. Walker(2)
(1)Fujitsu Labs. of Europe, UK, (2)Univ. of Essex, UK ................................................ 54

16:45-17:00  10C2-3 Square Photonic Crystal Fiber (SPCF) with Flattened Chromatic Dispersion and High Nonlinearity
Univ. of the Ryukyus, Japan .................................................................................. 56

17:00-17:15  10C2-4 Square-Lattice Photonic Crystal Fiber with High Negative Dispersion and Low Confinement Loss for Dispersion Compensating Applications
Univ. of the Ryukyus, Japan .................................................................................. 58

10C2-5 Withdrawn

Room D

10D1: Materials for Long Wavelength Semiconductor Lasers (Tutorial + Oral Presentation) (14:00-15:30)
Chair: H. Ishii

14:00-15:00  10D1-1 (Tutorial) Materials and Concepts for Semiconductor Lasers in the Near Infrared (λ ≥ 2 μ m)
J. Wagner
Fraunhofer IAF, Germany ..................................................................................... 62

15:00-15:30  10D1-2 (Invited) Integrated Hybrid Silicon Evanescent Racetrack Laser and Photodetector
A. W. Fang, R. Jones, H. Park, O. Cohen, O. Raday, M. Paniccia, J. E. Bowers
(1)Univ. of California, Santa Barbara, USA, (2)Intel, USA, (3)Intel, Israel .................. 64

10D2: Optical Signal Processing (I) (16:00-17:30)
Chair: K. Morito

16:00-16:30  10D2-1 (Invited) High Speed All-Optical Packet Switching
(1)Eindhoven Univ. of Technol., The Netherlands, (2)Univ. Politécnica de Valencia, Spain, (3)CIP, UK, (4)Technical Univ. of Denmark, Denmark ................................. 66

16:30-16:45  10D2-2 Double Wavelength Conversion with Multi-Resonant Saturable Absorber-Based Vertical-Cavity Semiconductor Gate
C. Porzi(1), L. Poti(2), A. Bogoni(2), L. Orsila(3), M. Guina(3)
(1)Scuola Superiore Sant’Anna, Italy, (2)CNIT, Italy, (3)Tampere Univ. of Technol., Finland ................................................................. 68

16:45-17:00  10D2-3 Frequency-Dependent Electric DC Power Consumption in Ultrafast All-Optical Semiconductor Gates
J. Sakaguchi, F. Salleras, Y. Ueno
Univ. of Electro-Communs., Japan ........................................................................ 70

17:00-17:15  10D2-4 Wavelength Conversion Characteristics of SOA-MZI Based All-Optical NRZ-OOK/RZ-BPSK Modulation Format Converter
S. M. Nissanka(1), K. Mishina(1), A. Maruta(1), S. Mitani(2), K. Ishida(2), K. Shimizu(2), T. Hatta(1,2,3), K. Kitayama(1)
(1)Osaka Univ., Japan, (2)Mitsubishi Electric, Japan, (3)OITDA, Japan .................... 72

17:15-17:30  10D2-5 Modeling of Optical Nonlinear-Effect Compensator with Vertical-Cavity Saturable Absorber
S. Suda, F. Koyama
Tokyo Inst. of Technol., Japan ............................................................................. 74

Room E

10E1: Functional Passive Devices (I) (14:00-15:30)
Chair: P. V. Daele

14:00-14:30  10E1-1 (Invited) Photonic Device Breakthrough for New Generation Optical Communications
K. Kobayashi
Tokyo Inst. of Technol., Japan ............................................................................. 76
14:30-14:45 10E1-2  Monolithically Integrated Tandem Waveguide-Type Acoustooptic Frequency Shifter Driven by Surface Acoustic Waves
S. Kakio(1), M. Kitamura(1), Y. Nakagawa(1), T. Hara(2), H. Ito(2), T. Kobayashi(3), M. Watanabe(3)
(1)Univ. of Yamanashi, Japan, (2)Tohoku Univ., Japan, (3)Optoquest, Japan ......................... 78

14:45-15:00 10E1-3  Temperature Dependence of Long Term Modulation Applied to Electro-Optic Polymer Waveguide
S. Nakamura, T. Kikuchi, R. Thapliya
Fuji Xerox, Japan ........................................................................................................................ 80

15:00-15:15 10E1-4  All-Optical Tunable Wavelength Filter Using Photorefractive Material
S. Honma(1), M. Komatsu(1), H. Ito(1), S. Muto(1), T. Ito(2), A. Okamoto(2)
(1)Yamanashi Univ., Japan, (2)Hokkaido Univ., Japan ................................................................. 82

15:15-15:30 10E1-5  Integrated OCDM En/Decoder Module for Practical Deployment
S. Kobayashi, K. Sasaki, T. Ushikubo
Oki Electric Industry, Japan ...................................................................................................... 84

10E2: Optical Interconnection Devices (16:00-17:30)
Chair: L. Eldada

16:00-16:30 10E2-1 (Invited)  Optical Interconnections Integrated in Printed Circuit Boards
P. Van Daele(1), G. Van Steenberge(1), N. Hendrickx(1), E. Bosman(1), P. Karioja(2), C. Debaes(3)
(1)IMEC-Ghent Univ., Belgium, (2)VTT, Finland, (3)Vrije Univ. Brussels, Belgium .................. 86

16:30-16:45 10E2-2  An Efficient Optical Coupling Method for Multilayer Optical Printed Circuit Boards
Y. Matsuoka(1), T. Ban(1), T. Shibata(2), A. Takahashi(2), M. Shishikura(1)
(1)Hitachi, Japan, (2)Hitachi Chemical, Japan ............................................................................. 88

16:45-17:00 10E2-3  Replication of Multimode Polymer Optical Waveguides from Flexible Film Stamps
S. Yagi(1), Y. Hatakeyama(2), N. Kawakami(2), J. Kobayashi(1)
(1)NTT, Japan, (2)NTT AT, Japan ............................................................................................... 90

17:00-17:15 10E2-4  Self-Written Optical Funnel for Optical Interconnect by Microlens-Transfer Method
H. Kubo, M. Kanda, O. Mikami
Tokai Univ., Japan ....................................................................................................................... 92

17:15-17:30 10E2-5  Polymeric Waveguide Optical Switch Using Rotary Drive Mechanism
T. Uesugi, S. Zaizen, A. Sugitatsu, T. Hatta
Mitsubishi Electric, Japan ........................................................................................................... 94

Wednesday, July 11

Room A

11A1: Optical Access with Optical Amplifiers (9:00-10:30)
Chair: S. Oshiba

Y. Nakanishi, T. Nakanishi, Y. Fukada, K. Suzuki, K. Iwatsuki
NTT, Japan ................................................................................................................................ 96

9:15-9:30 11A1-2  Power Stable Fiber Loop Buffers with EDFA Followed by SOA
K. Takano(1), Y. Takahashi(1), D. Morita(1), K. Nakagawa(1), H. Ito(2)
(1)Yamagata Univ., Japan, (2)Tohoku Univ., Japan ..................................................................... 98

9:30-9:45 11A1-3  Self-Seeded Reflective Semiconductor Optical Amplifier for Upstream Access and Local Area Networking in Passive Optical Networks
N. Nadarajah, K. L. Lee, A. Nirmalathas
Natl. ICT Australia, Australia ..................................................................................................... 100

9:45-10:00 11A1-4  Reflection Tolerance of RSOA-Based WDM PON
(1)Korea Advanced Inst. of Sci. and Technol., Korea, (2)KDDI R&D Labs., Japan .................. 102

10:00-10:15 11A1-5  Wavelength Monitoring of Remote Node by Using Self-Locked Reflective Semiconductor Optical Amplifier for Bidirectional WDM-PON
S. H. Han, T.-Y. Kim, C.-S. Park
GIST, Korea .................................................................................................................................. 104
10:15-10:30  11A1-6  MZI-SOA Based All-Optical Router Implementation with OCDMA Header Recognition
P. Teixeira(1), L. Oliveira(1), T. Silveira(1), P. André(1), R. Nogueira(1), M. Lima(1), A. Teixeira(1)
(1)Inst. de Telecommuns., Portugal, (2)Siemens Networks, Portugal……………………………………106

11A2: Access Network Design (11:00-12:15)
Chair: K. Fukuchi
11:00-11:15  11A2-1  Crosstalk in Downlink Carrier Reused WDM-PONs Based on Subcarrier Modulation
(1)A*STAR, Singapore, (2)Nanyang Technological Univ., Singapore, (3)Natl. ICT Australia, Australia, (4)Hong Kong Polytechnic Univ., China……………………………………108

11:15-11:30  11A2-2  Power Penalty Dependency on Sideband Suppression Ratio in Optical SSB Signal Transmission
T. Fujiwara, K. Kikushima
NTT, Japan …………………………………………………………………………………………………………………110

11:30-11:45  11A2-3  Simple Multi-Wavelength Stabilization Technique Using a Periodic Optical Filter for WDM Access Networks
NTT, Japan …………………………………………………………………………………………………………………112

11:45-12:00  11A2-4  Local Area Network Emulation in Passive Optical Networks by Wavelength Switching the Distributed Feedback Laser
N. Nadarajah, C.-J. Chae
Natl. ICT Australia, Australia …………………………………………………………………………………………………………………………………………………114

12:00-12:15  11A2-5  Remote Nodes for Wavelength Shared Hybrid PON Supporting Video Overlay
M. Bouda, P. Palacharla, Y. Akasaka, A. Umno, T. Naito
Fujitsu Labs. of America, USA ……………………………………………………………………………………………………………………………………………………………116

11A3: OCDMA Technology (14:00-16:00)
Chair: N. Wada
14:00-14:30  11A3-1 (Invited)  Optical-Label and Code Empowered Systems for Next Generation Photonic Networks
S. J. B. Yoo
Univ. of California, Davis, USA……………………………………………………………………………………………………118

14:30-15:00  11A3-2 (Invited)  Enabling Techniques for Multi-User Asynchronous OCDMA System
X. Wang(1), N. Wada(1), G. Cincotti(2), K. Kitayama(3)
(1)NICT, Japan, (2)Univ. of Roma Tre, Italy, (3)Electr. and Information Systems, Japan……120

15:00-15:15  11A3-3  Optical Code Label Processing Using Multi-Port Optical Spectrum Synthesizer and Frequency Comb Generator
M. Mieno(1), F. Moritureka(1), Y. Komai(1), S. Anzai(1), K. Kodate(1), N. Wada(2), T. Sakamoto(2), T. Kawanishi(2), M. Izutsu(2)
(1)Japan Women's Univ., Japan, (2)NICT, Japan, Japan………………………………………………………………………………………………………………………………………………122

15:15-15:30  11A3-4  Transmission Characteristics of Loss and Fiber Nonlinearity Tolerance on Coherent Time-Spreading OCDM with Optical Time-Gatings
N. Minato, S. Kutsuzawa, S. Kobayashi, K. Sasaki
Oki Electric Industry, Japan ……………………………………………………………………………………………………………………………………………………………124

G. Cincotti(1), G. Manzacca(1), V. Sacchieri(1), N. Wada(2), X. Wang(2), K. Kitayama(3)
(1)Univ. Roma Tre, Italy, (2)NICT, Japan, (3)Osaka Univ., Japan…………………126

Room B
11B1: Ultrafast Transmission Technologies (Tutorial) (9:00-10:00)
Chair: T. Morioka
9:00-10:00  11B1-1 (Tutorial)  Ultrafast Transmission Technology
R. Ludwig
FhG Heinrich-Hertz-Inst., Germany……………………………………………………………………………………………………128
**11B2: Microwave Photonics (11:00-12:30)**
Chair: J. Maeda

<table>
<thead>
<tr>
<th>Time</th>
<th>Session</th>
<th>Title</th>
<th>Authors</th>
</tr>
</thead>
<tbody>
<tr>
<td>11:00-11:30</td>
<td>11B2-1 (Invited)</td>
<td>Optical/Wireless Physical Layer Integration – Radio-over-Fiber Systems</td>
<td>C. Lim(1), M. Attygalle(2), A. Nirmalathas(1,2), K. Lee(1), D. Novak(1,3), R. Waterhouse(1,3)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>*(1)CUBIN, Australia, *(2)Natl. ICT Australia, Australia, <em>(3)Pharad, USA</em></td>
</tr>
<tr>
<td>11:30-11:45</td>
<td>11B2-2</td>
<td>20-GHZ 16-QAM Radio-Over-Fiber Data Transmission Using a Directly Modulated Laser</td>
<td>H. Kim(1), H. C. Ji(2)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>*(1)Samsung Electr., Korea, <em>(2)Osaka Univ., Japan</em></td>
</tr>
<tr>
<td>11:45-12:00</td>
<td>11B2-3</td>
<td>UWB Monocyte Pulse Generation by Optical Polarization Time Delay Method</td>
<td>M. Chen, H. Chen, J. Zhang, S. Xie</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>Tsinghua Univ., China</em></td>
</tr>
<tr>
<td>12:00-12:15</td>
<td>11B2-4</td>
<td>Multi-Channel Ultrawideband Monocyte Pulse Generation via Cross Phase Modulation and Spectral Filtering</td>
<td>B. P. P. Kuo, P. C. Chui, K. K. Y. Wong</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>The Univ. of Hong Kong, China</em></td>
</tr>
<tr>
<td>12:15-12:30</td>
<td>11B2-5</td>
<td>High-Quality Electrical Gaussian-Monocyte Pulse Generation by Electrical-Optical Hybrid Signal Processing</td>
<td>M. Hanawa(1,3), K. Nakamura(1), K. Nonaka(2)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>*(1,3)Univ. of Yamanashi, Japan, <em>(2)Kochi Univ. of Technol., Japan</em></td>
</tr>
</tbody>
</table>

**11B3: Novel Optical Modulation/Demodulation Technologies (14:00-16:00)**
Chair: H. Taga

<table>
<thead>
<tr>
<th>Time</th>
<th>Session</th>
<th>Title</th>
<th>Authors</th>
</tr>
</thead>
<tbody>
<tr>
<td>14:00-14:30</td>
<td>11B3-1 (Invited)</td>
<td>Coherent Detection of Multi-Level Coded Optical Signals</td>
<td>K. Kikuchi</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>The Univ. of Tokyo, Japan</em></td>
</tr>
<tr>
<td>14:30-15:00</td>
<td>11B3-2 (Invited)</td>
<td>Impairment Mitigation in High Speed Optical Communications Using Digital Signal Processing</td>
<td>R. I. Killey(1), P. M. Watts(1), S. J. Savory(1), R. Waegemans(1), Y. Benlachtar(1), V. Mikhailov(1), M. Glick(1), P. Bayvel(1)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>*(1)Univ. College London, UK, <em>(2)Intel Res., USA</em></td>
</tr>
<tr>
<td>15:00-15:30</td>
<td>11B3-3 (Invited)</td>
<td>Advanced Modulation/Demodulation Technologies for High-Speed Optical Transmission Systems</td>
<td>I. Morita, S. L. Jansen</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>KDDI R&amp;D Labs., Japan</em></td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>Alcatel-Lucent Res. and Innovation, Germany</em></td>
</tr>
</tbody>
</table>

**11C1: Photonic Crystal Fibers (II) (9:00-10:30)**
Chair: K. Saitoh

<table>
<thead>
<tr>
<th>Time</th>
<th>Session</th>
<th>Title</th>
<th>Authors</th>
</tr>
</thead>
<tbody>
<tr>
<td>9:00-9:30</td>
<td>11C1-1 (Invited)</td>
<td>Solid Photonic Crystal Fibres</td>
<td>R. Goto, S. Matsuo, K. Himeno</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>Fujikura, Japan</em></td>
</tr>
<tr>
<td>9:30-9:45</td>
<td>11C1-2</td>
<td>Designs and Fabrications of Photonic Crystal Fiber Couplers with Air Hole Collapsed Taper Regions</td>
<td>H. Yokota, H. Kawashiri, H. Yashima, Y. Sasaki</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>Ibaraki Univ., Japan</em></td>
</tr>
<tr>
<td>9:45-10:00</td>
<td>11C1-3</td>
<td>Photonic Crystal Fiber Interferometer Composed of a Long-Period Grating and One Point Collapsing</td>
<td>H. Y. Choi, M. J. Kim, S. H. Kim, B. H. Lee</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>GIST, Korea</em></td>
</tr>
<tr>
<td>10:00-10:15</td>
<td>11C1-4</td>
<td>A Simple and Practical Design Approach to Realize Band-Pass Photonic Crystal Fiber Filters</td>
<td>S. K. Varshney, K. Saitoh, N. J. Florous, M. Koshiba</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td><em>Hokkaido Univ., Japan</em></td>
</tr>
</tbody>
</table>
10:15-10:30 11C1-5  Wavelength-Dependent Core-Mode Blocker Using Micro Air Bubble in the Hollow Optical Fiber
J. Jeon\textsuperscript{1}, J. Kim\textsuperscript{2}, Y. Jung\textsuperscript{2}, W. Shin\textsuperscript{2}, D. Ko\textsuperscript{2}, J. Lee\textsuperscript{2}, K. Oh\textsuperscript{1}
\textsuperscript{1}Yonsei Univ., Korea, \textsuperscript{2}GIST, Korea
\vspace{0.5cm}

11C2: Ultra-High-Speed Fiber-Optic Technologies (Tutorial) (11:00-12:00)
Chair: K. Kokura, K. Mukasa
11:00-12:00 11C2-1 (Tutorial)  Advanced Optical Fiber and Fiber Device Technologies for Ultrahigh-Speed Optical Transmission
M. Nakazawa
Tohoku Univ., Japan
\vspace{0.5cm}

11C3: Fiber Optic Cables and Connection Technologies for FTTH (14:00-16:00)
Chair: Y. Miyajima, H. Yokosuka
14:00-14:30 11C3-1 (Invited)  The Optical Access Technologies and Applications in FTTH
Y. Ji, Y. Sun, R. Liao
Beijing Univ. of Posts and Telecomms., China
\vspace{0.5cm}
14:30-15:00 11C3-2 (Invited)  Technologies for FTTH System in Korea
E. D. Park, K. Han, W. J. Jung
LS cable, Korea
\vspace{0.5cm}
15:00-15:30 11C3-3 (Invited)  Optical Fibres and Cables for the Access Networks; Standardization & FTTH Development in Europe
G. Kuyt, P. Matthijssen
Draak Comteq Optical Fibre, The Netherlands
\vspace{0.5cm}
15:30-16:00 11C3-4 (Invited)  Development of Optical Fiber Cable and Wiring Techniques in Full-Scale FTTH Era
M. Kama, K. Shiraki, H. Aoyama, O. Inoue, S. Matsui
NTT, Japan
\vspace{0.5cm}

Room D
11D1: Optical Switch (9:00-10:30)
Chair: H. J. S. Dorren
9:00-9:30 11D1-1 (Invited)  InGaAs/AlAs/AlAsSb Intersubband Transition All-Optical Switch for Ultrafast All-Optical Signal Processing
AIST, Japan
\vspace{0.5cm}
9:30-9:45 11D1-2  Wavelength Switching Using GainAs/InP MQW Variable Index Arrayed Waveguides by Thermo-Optic Effect
Y. Shimizu, M. Mogi, T. Yoshioka, K. Shimomura
Sophia Univ., Japan
\vspace{0.5cm}
9:45-10:00 11D1-3  Integrated Bandpass Filter for DWDM Systems with Gain
B. Stelzig, U. Barabas
Univ. der Bundeswehr München, Germany
\vspace{0.5cm}
10:00-10:15 11D1-4  Analysis and Design of Wavelength Selective Switches Based on MMI Assisted Microring Resonators
T. T. Le, L. W. Cahill
La Trobe Univ., Australia
\vspace{0.5cm}
10:15-10:30 11D1-5  All-Optical Transistor Operation Based on Bistability Principle in Nonlinear DFB GainAsP-InP Waveguide: A Transient Perspective
Yosia\textsuperscript{1}, Y. Akano\textsuperscript{2}, K. Tamura\textsuperscript{2}, T. Mizumoto\textsuperscript{2}, S. Ping\textsuperscript{1}
\textsuperscript{1}Nanyang Technological Univ., Singapore, \textsuperscript{2}Tokyo Inst. of Technol., Japan
\vspace{0.5cm}

11D2: Novel Devices (11:00-12:30)
Chair: H. Ishikawa
11:00-11:30 11D2-1 (Invited)  Nano- and Micro-Photonic Devices for Inter- and Intra-Board Level Optical Interconnects
R. T. Chen
The Univ. of Texas, USA


12:00-12:15  11D2-4 Anti-Shake Mechanism for Mobile Optical Wireless Communication M. Hattori, T. Yazaki, H. Tanaka KDDI R&D Labs., Japan ........................................................................................................... 184


11D3: Photo Detectors and Modulator (14:00-15:15)
Chair: T. Aoyagi

14:00-14:15  11D3-1 Very Low Noise AlInAs Avalanche Photodiodes with Gain-Bandwidth Product of 140 GHz M. Achouche, A. Rouvié, J. Decobert, N. Lagay, F. Pommereau, D. Carpentier Alcatel Thales III-V Lab, France ........................................................................................................ 188


14:30-14:45  11D3-3 Characterization of Metal/Semiconductor/Metal Photo Diode for Optical Interface of Single-Flux-Quantum Circuit S. Shinada, H. Terai, N. Wada, Z. Wang, T. Miyazaki NICT, Japan ......................................................................................................................................... 192

14:45-15:00  11D3-4 Novel Optical Interconnection for Silicon-on-Insulator Waveguide Tap Monitor S.-H. Hsu Natl. Taiwan Univ. of Sci. and Technol., Taiwan ........................................................................................................ 194

15:00-15:15  11D3-5 High-Speed (>55 GHz) Electro-Absorption Modulator Based on Two-Step Undercut Active Region Waveguide T. H. Wu, F. Z. Lin, D. R. Lee, Y. J. Chiu NSYSU, Taiwan ......................................................................................................................................... 196

Room E

11E1: Functional Passive Devices (II) (9:00-10:30)
Chair: K. Kobayashi

9:00-9:30   11E1-1 (Invited) Polymeric Integrated Optical Devices for ROADM Systems L. Eldada DuPont Photonics Technol., USA ........................................................................................................ 198

9:30-10:00  11E1-2 (Invited) Polymer-Based Waveguide Grating Devices K. S. Chiang City Univ. of Hong Kong, China ......................................................................................................... 200

10:00-10:15 11E1-3 A Planar-Waveguide 8-Channel Arrayed Wavelength Division Multiplexing Coupler Using Directional Coupler with Higher-Order Mode Cutting Filter for Waveguide Amplifier T. Sugimoto(1), Y. Dekii(1), T. Takeuchi(1), S. Takaesu(1), Y. Urino(1) (1)NEC, Japan, (2)OITDA, Japan ........................................................................................................ 202


11E2: Photonic Crystal Devices (Tutorial + Oral Presentation) (11:00-12:30)
Chair: M. Notomi

11:00-12:00 11E2-1 (Tutorial) Photonic Crystal Devices J. O'Brien Univ. of Southern California, USA ........................................................................................................... 206
12:00-12:15  11E2-2  40-Gbit/s Operation of Dispersion Compensator Based on Multiple 1D Coupled-Defect-Type Photonic Crystals
M. Sagawa(1,2,3), S. Goto(1,2,3), K. Hosomi(1,2,3), T. Sugawara(2,3), T. Katsuyama(1,3), Y. Arakawa(1)
(1) The Univ. of Tokyo, Japan, (2) Hitachi, Japan, (3) OITDA, Japan ................................. 208

12:15-12:30  11E2-3  A Study on Photonic Crystal Waveguide Based on Absolute Photonic Band Gap
Y. Morita, Y. Tsuji, K. Hirayama
Kitami Inst. of Technol., Japan .................................................................................................. 210

11E3: Functional Photonic Crystal Component and Devices (14:00-16:00)
Chair: J. O'Brien

14:00-14:30  11E3-1 (Invited)  Bonded Photonic Crystal Components and Circuits: Toward 2.5-D Micro-Nano-Photonics
INL, France .......................................................................................................................... 212

14:30-15:00  11E3-2 (Invited)  Functional Nanocavities Based on Photonic Crystals
M. Notomi, E. Kuramochi, T. Tanabe, A. Shinya, H. Taniyama
NTT, Japan ................................................................................................................................ 214

15:00-15:30  11E3-3 (Invited)  Prospect of Optical Devices with One-Dimensional Photonic Crystal Structure
T. Katsuyama(1,2), K. Hosomi(1,2,3), M. Sagawa(1,2,3), Y. Arakawa(1)
(1) The Univ. of Tokyo, Japan, (2) OITDA, Japan, (3) Hitachi, Japan ................................. 216

15:30-15:45  11E3-4  Add-Drop Multiplexing in WDM Signal Transmission Link Using Silicon Photonic Crystal R-OADM Devices
S. Nakamura(1,2), T. Chu(1), A. Gomyo(1), J. Ushida(1), H. Yamada(2), S. Ishida(3), Y. Arakawa(3)
(1) NEC, Japan, (2) Tohoku Univ., Japan, (3) The Univ. of Tokyo, Japan ............................. 218

15:45-16:00  11E3-5  Development of Directional Coupler Switch with Ultra-Short Switching Length Based on Flat-Band Photonic Crystal Structure
J. Sugisaka(1,2), N. Yamamoto(1), M. Okano(1), K. Komori(1), M. Itoh(2), T. Yatagai(2)
(1) AIST Photonics, Japan, (2) Univ. of Tsukuba, Japan ....................................................... 220

4F Lobby
11P: Poster Session (12:30-14:00)

11P-1  RF Output Power Improvement Using Heterodyne Technique for Radio-on-Fiber Down-Link Transmission System Eliminating Electric Power Supply at Base Station
M. Matsuura, T. Okabe, N. Kishi, T. Miki
Univ. of Electro-Communs., Japan .......................................................................................... 222

11P-2  A Study of Interference Reduction in Radio over Fiber System Applying MIMO Technique
Y. Kanaoka(1), I. Yamashita(1), S. Kashimura(2), K. Asano(3), S. Shimizu(2)
(1) Kansai Electric Power, Japan, (2) Oki Electric Industry, Japan ......................................... 224

11P-3  Cycle Attack by MAI Noise Propagation and Its Online Detection in OCDM-Based Transparent Optical Networks
S. Huang, K. Baba, M. Murata, K. Kitayama
Osaka Univ., Japan .................................................................................................................. 226

11P-4  Optical CDMA Networks Using Extended Hamming Code for Interference Elimination
C.-C. Yang, L.-Y. Yeh
Kun Shan Univ., Taiwan ............................................................................................................ 228

11P-5  Comparison of Physical Network Configurations for Next Generation Home Network
Y. Kakishima, K. Okada, K. Oguchi
Seikei Univ., Japan .................................................................................................................. 230

11P-6  QoS Management in a Next Generation Home Network
S. Yamakawa, K. Tojo, S. Terada, K. Oguchi
Seikei Univ., Japan .................................................................................................................. 232

11P-7  G-WAPS Internet Access Network Based on an Active Optical Network
T. Nakata, T. Kaminogou, H. Matsuda, T. Yasui
Toyama Prefectural Univ., Japan .............................................................................................. 234

xvi
11P-8 40 Gb/s-2 km Photonic Crystal Fiber Transmission at 850 nm with a Single-Mode VCSEL
H. Hasegawa, Y. Oikawa, T. Hirooka, M. Nakazawa
Tohoku Univ., Japan ................................................................. 236

11P-9 Sampled FBG Based All Optical Wavelet Analyzer for High Speed Optical Communication Signals
M. Hanawa
Univ. of Yamanashi, Japan .......................................................... 238

11P-10 Resolution Improvement of All-Optical ADC Using SPM-Induced Spectral Compression
T. Nishitani, T. Konishi, K. Itoh
Osaka Univ., Japan .................................................................. 240

11P-11 Effects of Frequency Allocations and Zero Dispersion Frequencies on FDM Lightwave Transmission Systems
J. Onishi, S. Kojima, T. Numai
Ritsumeikan Univ., Japan .......................................................... 242

11P-12 Influence of Modulation Formats on FWM Noises in FDM Optical Fiber Transmission Systems
Y. Ito, J. Onishi, S. Kojima, T. Numai
Ritsumeikan Univ., Japan .......................................................... 244

11P-13 Effects of Frequency Allocations and Polarization Allocations on FDM Lightwave Transmission Systems
J. Onishi, S. Kojima, T. Numai
Ritsumeikan Univ., Japan .......................................................... 246

11P-14 Scalability of Code Matching for Optical Time-Series WDM Encoded Label Using Collinear Acoustooptic Devices
N. Goto(1), Y. Miyazaki(2)
(1) Toyohashi Univ. of Technol., Japan, (2) Aichi Univ. of Technol., Japan .......................................................... 248

11P-15 Crossover Reducing Integrated Nested Rings Optical Switch
N. Xie, K. Utaka
Waseda Univ., Japan ................................................................ 250

11P-16 Novel Type BOTDA for Measuring Distributed Strain and Temperature with cm Spatial Resolution in km-long Fiber
Y. Koyamada, S. Sotoyama
Ibaraki Univ., Japan .................................................................. 252

11P-17 All-Optical Variable Optical Attenuator Based on Nonlinear Optical Fiber with Fiber Bragg Grating
S. Ju(1), P. R. Watekar(2), Y. H. Kim(3), T. Kim(1), W.-T. Han(2)
(1) OptoNest, Korea, (2) GIST, Korea, (3) LG Chem., Korea .......................................................... 254

11P-18 Multi-Wavelength Fiber Laser Sources with an Elliptical Core Side-Hole Fiber Sagnac Loop Filter
D. S. Moon(1), B. H. Kim(1), G. Sun(1), Y.-G. Han(2), W.-T. Han(1), Y. Chung(1)
(1) GIST, Korea, (2) Hanyang Univ., Korea .......................................................... 256

11P-19 Investigations on Sensitivity of Long Period Gratings Modified by Gold Nanoparticles
(1) Natl. Chung Cheng Univ., Taiwan, (2) Natl. Yunlin Univ. of Sci. and Technol., Taiwan, (3) Chung Shan Inst. of Sci. and Technol., Taiwan .......................................................... 258

11P-20 Simple Pure Apodization Method for Fiber Bragg Gratings by Sequential UV Writing
K.-C. Hsu, Y. Lai
Natl. Chiao Tung Univ., Taiwan .......................................................... 260

11P-21 The Refractive Index Sensing of Different Types of Long-Period Gratings
(1) Natl. Chung Cheng Univ., Taiwan, (2) Natl. Yunlin Univ. of Sci. and Technol., Taiwan, (3) Chung Shan Inst. of Sci. and Technol., Taiwan .......................................................... 262

11P-22 Cross Phase Modulation by Pump Pulses Emitted from High-Power YAG Laser on Fiber Grating Couplers
F. Abrishamian, K. Narumi, S. Sato, M. Imai
Muroran Inst. of Technol., Japan .......................................................... 264
11P-23  Effective Automatic Gain Spectrum Adjustment of Bi-Directionally Pumped Broadband Raman Amplifiers  
Z. Tong, J. Zhang, D. Sun, S. Jian  
Beijing Jiaotong Univ., China ...........................................................................................................266

11P-24  Configuration Comparison of C/L-Band EDFA for DWDM Systems  
T.-C. Liang, Y.-L. Liu, H.-S. Huang  
Natl. Kaohsiung First Univ. of Sci. and Technol., Taiwan ................................................................268

K. S. Chang, Y. M. Song, Y. T. Lee  
GIST, Korea .....................................................................................................................................270

11P-26  Investigation of Oxide Mode in 1.3 μm InGaAsN Vertical Cavity Surface Emitting Lasers  
H. P. D. Yang(4), T. D. Lee(5)  
(1)Nat. Cheng Kung Univ., Taiwan, (2)Yuan Ze Univ., Taiwan, (3)Nat. Chiao Tung Univ., Taiwan,  
(4)Industrial Technol. Res. Inst., Taiwan, (5)Nat. Yunlin Univ. of Sci. and Technol., Taiwan .........272

11P-27  Etching Characteristics of Low-k Polymer for Wide Modulation Bandwidth VCSELs  
A. Matsutani, N. Jogan, F. Koyama, K. Kobayashi  
Tokyo Inst. of Technol., Japan ...........................................................................................................274

11P-28  A Quantitative Postweld Shift Compensation Technique in Butterfly Laser Module Packages  
M.-T. Sheen(1), Y.-C. Hsu(2), W.-H. Cheng(3)  
(1)Yung Ta Inst. of Technol. and Commerce, Taiwan, (2)Pingtung Univ. of Sci. and Technol., Taiwan,  
(3)Natl. Sun Yat-sen Univ., Taiwan .................................................................................................276

11P-29  Growth and Characterisation of InSb Films on GaAs Substrate Grown Using Molecular Beam Epitaxy  
L. Li(1,2), G.-J. Liu, Z. Li, M. Li, X.-H. Wang  
(1)Changchun Univ. of Sci. and Technol., China, (2)The Univ. of Tokyo, Japan .........................278

11P-31  Reconfigurable Photonic Microwave Bandpass Filter Based on Polarization Modulation  
C. K. Oh, T.-Y. Kim, C.-S. Park  
GIST, Korea .....................................................................................................................................280

11P-32  All-Optical Interconnection for Two-Dimensional Data Using Photorefractive Organic Film  
S. Honma(1), S. Muto(1), H. Kobayashi(1), M. Bunsen(2), A. Okamoto(3)  
(1)Yamanashi Univ., Japan, (2)Fukuoka Univ., Japan, (3)Hokkaido Univ., Japan .......................282

11P-33  Self-Written Micro Optical Pin for Optical Interconnect  
T. Tokuhara, O. Mikami  
Tokai Univ., Japan ............................................................................................................................284

11P-34  Optical Parallel Interconnects Using Organic Photorefractive Polymer  
K. Shimayabu(1), K. Harasaka(1), A. Okamoto(1), S. Rokutanda(2)  
(1)Hokkaido Univ., Japan, (2)Nitto Denko Technical, Japan ..............................................................286

11P-35  Measurements of Lattice Constant and Electrooptic Constant in Reverse-Proton-Exchanged LiNbO3 Waveguides  
S. Kakio, M. Maeda, H. Watanabe, Y. Nakagawa  
Univ. of Yamanashi, Japan ..................................................................................................................288

11P-36  Low Birefringence Silicon-on-Insulator Waveguide and Its Optical Interconnection Using High Numerical Aperture Fiber  
S.-H. Hsu  
Natl. Taiwan Univ. of Sci. and Technol., Taiwan ..............................................................................290

11P-37  Comparison of Transmission Media for Next Generation Home Network  
S. Terada(1), Y. Kakishima(1), S. Yamakawa(1), K. Tojo(1), T. Taniguchi(2),  
T. Kawakami(2), K. Oguchi(1)  
(1)Seikei Univ., Japan (2)Sekisui Chemical, Japan ........................................................................292
Thursday, July 12

Room A
12A1: Control Plane Technology: Challenge and Issues (9:00-10:30)
Chair: S. Hanatani

9:00-9:30 12A1-1 (Invited) Challenges and Requirements for ASON/GMPLS Photonic Networks
W. Imajuku(1), Y. Sameshima(1,2)
(1)NTT, Japan, (2)NICT, Japan ................................................................. 294

9:30-10:00 12A1-2 (Invited) Optical Grid Networking Supports for GT4
Y. -M. Lu, Y. -F. Ji, H. -X. Wang
Beijing Univ. of Posts and Telecommunications, China ................................................. 296

10:00-10:15 12A1-3 Fault Management by Prioritized Alarm Correlation in Multi-Layer GMPLS Networks
M. Miyazawa, K. Ogaki, T. Otani
KDDI R&D Labs., Japan ................................................................. 298

10:15-10:30 12A1-4 A Study on Optimized Assignment of Dispersion Compensation Capability for Dynamic Optical Paths
S. Seno(1), Y. Baba(1), T. Mizuochi(1), T. Sugihara(1), K. Motoshima(1), T. Ideguchi(2)
(1)Mitsubishi Electric, Japan, (2)Aichi Prefectural Univ., Japan ............................................ 300

12A2: Optical Network (11:00-12:30)
Chair: Y. Lu

11:00-11:30 12A2-1 (Invited) Verizon Optical Network — Strategic Vision
W. C. Uliasz
VTO, USA ............................................................................................................. 302

11:30-11:45 12A2-2 A Field Trial of On-Demand Optical Grid Lightpath Network Services
S. Xu, H. Harai
NICT, Japan ........................................................................................................ 304

11:45-12:00 12A2-3 Demonstration of a GMPLS-Controlled Transparent Optical Network with Wavelength Continuity Constraint
H. Guo, T. Tsutsumi, N. Yoshikane, T. Otani
KDDI R&D Labs., Japan ................................................................................................. 306

12:00-12:15 12A2-4 An Effective Light-Path Setup Scheme for Dynamic Traffic in Multi-Ring Wavelength-Routed Networks
M. Park(1), Y. Baba(1), S. Seno(1), N. Okazaki(2)
(1)Mitsubishi Electric, Japan, (2)Univ. of Miyazaki, Japan .............................................. 308

12:15-12:30 12A2-5 The First Optically-Virtual-Concatenated Lambdas over Multiple Domains in Chicago Metro Area Network Achieved Through Interworking of Network Resource Managers
(1)NTT, Japan, (2)Univ. of Illinois at Chicago, USA ................................................ 310

12A3: Home & In-building Optical Networks (14:00-15:30)
Chair: H. Imaizumi

14:00-14:30 12A3-1 (Invited) Flexible Optical Access and In-Building Networks
A. M. J. Koonen, P. J. Urban, H. Yang, M. G. Larrode, H. de Waardt
Eindhoven Univ. of Technol., The Netherlands ..................................................................... 312

14:30-14:45 12A3-2 Effectiveness of Mode-Selective Spatial Filtering in Mode Group Diversity Multiplexing Links
C. P. Tsekrekos, A. M. J. Koonen
Eindhoven Univ. of Technol., The Netherlands ..................................................................... 314

14:45-15:00 12A3-3 1-Gbps Link System for Home Network with Plastic Optical Fiber Cable
T. Kawakami(1), H. Sueyoshi(1), T. Taniguchi(1), K. Oguchi(2)
(1)Sekisui Chemical, Japan, (2)Seikei Univ., Japan ..................................................... 316

15:00-15:15 12A3-4 Low-Power Digital Fiber Optic Sensing Network with Microprocessor-Controlled Sensor Terminals
C. Rothe, T. Nishimura, Y. Tanaka, T. Shioda, T. Kurokawa
Tokyo Univ. of Agriculture and Technol., Japan ..................................................................... 318
15:15-15:30 12A3-5 Reliable Sensor Data Transmission Method for Optical Home Network
K. Tojo(1), T. Murooka(2), S. Terada(1), S. Yamakawa(1), K. Oguchi(1)
(1)Seikei Univ., Japan, (2)NTT, Japan .................................................................320

12A4: Optical Ethernet (Tutorial) (16:00-17:00)
Chair: A. Kirstadter
16:00-17:00 12A4-1 (Tutorial) Ethernet Optical Interfaces: Today and Tomorrow
O. Ishida
NTT, Japan ........................................................................................................322

Room B
12B1: Coherent Systems (I) (9:00-10:30)
Chair: K. Takano
9:00-9:30 12B1-1 (Invited) Incoherent Optical Field Detection for High-Speed Binary and Multilevel Optical Communication
N. Kikuchi
Hitachi, Japan .......................................................................................................324
9:30-9:45 12B1-2 Linewidth-Tolerant 8PSK by Pilot-Carrier Added Homodyne
M. Nakamura, Y. Kamio, T. Miyazaki
NICT, Japan ........................................................................................................326
9:45-10:00 12B1-3 Carrier Synchronization in 43 Gbit/s Coherent QPSK Receiver
L. Li(1), Z. Tao(1), T. Hoshida(2), J. C. Rasmussen(2)
(1)Fujitsu R&D Center, China, (2)Fujitsu Labs., Japan ........................................328
10:00-10:15 12B1-4 Phase Noise-Tolerance of Optical M-QAM Signals in Self-Homodyne Detection with Polarization-Multiplexed Pilot Carrier
Y. Kamio, M. Nakamura, T. Miyazaki
NICT, Japan .........................................................................................................330
10:15-10:30 12B1-5 Numerical Study of APSK Format for Long-Haul Transmission and Its Performance Improvement by Zero-Nulling Method
H. Taga, J.-Y. Wu, W.-T. Shih, S.-S. Shu
Natl. Sun Yat-Sen Univ., Taiwan ..........................................................................332

12B2: Advanced Modulation Techniques (Tutorial) (11:00-12:00)
Chair: M. Shikada
11:00-12:00 12B2-1 (Tutorial) Advanced Optical Data Modulation Techniques for High Speed Transmission
Y.-J. Wen
A*STAR, Singapore ............................................................................................334

12B3: Coherent Systems (II) (14:00-15:30)
Chair: J. Rasmussen
14:00-14:30 12B3-1 (Invited) High-Speed Lithium Niobate Modulator and Its Application to Huge Capacity Transmission System
T. Kawanishi
NICT, Japan .........................................................................................................336
14:00-14:45 12B3-2 A Novel RZ-DQPSK Transmitter Setup and Comparison Regarding their Tolerance in Spectral Efficient 8 x 40 Gb/s WDM Systems
M. Haris(1), J. Yu(1,2), G.-K. Chang(1)
(1)Georgia Inst. of Technol., USA, (2)NEC Labs. America, USA .....................338
14:45-15:00 12B3-3 Experimental Measurement and Numerical Estimation of Optical Phase Jitter in RZ-DPSK Systems
D. Boivin(1), M. Haris(1), M. Hanna(2), J. Yu(3), G.-K. Chang(1), J. R. Barry(1)
(1)Georgia Inst. of Technol., USA, (2)Inst. d’Optique, France, (3)NEC Labs. America, USA ..................................................340
15:00-15:15 12B3-4 Demonstration of Optical Quadrature Amplitude Modulation by Using a High-Speed Optical DQPSK Modulator and Precise Voltage-Control Technique
A. Chiba, T. Sakamoto, T. Kawanishi, M. Iizutsu
NICT, Japan .........................................................................................................342
15:15-15:30 12B3-5 Tolerance of Fiber-Optic Systems to CD and PMD
M. Secondini, E. Forestieri, G. Prati
Scuola Superiore Sant’Anna, Italy ..........................................................................344
**12B4: Performance Monitoring (16:00-17:30)**

Chair: H. Takara

**16:00-16:30 12B4-1 (Invited) Polarization-Based Monitoring in WDM Systems**
M. Sköld(1), B.-E. Olsson(2), M. Karlsson(1), P. A. Andrekson(1)
(1)Chalmers Univ. of Technol., Sweden, (2)Chalmers Industrietechnik, Sweden

16:30-16:45 12B4-2 Performance Evaluation of the Improved Polarization-Nulling Technique for the OSNR Monitoring in Dynamic Optical Networks
(1)Korea Advanced Inst. of Sci. and Technol, Korea, (2)TeraLink Commun., Korea

16:45-17:00 12B4-3 Optical Performance Monitoring in Amplitude Sampling Receivers
F. N. Hauske(1), M. Kuschnerov(1), K. Piyawanno(1), B. Lank(1), E.-D. Schmidt(2)
(1)Univ. of the Federal Armed Forces, Germany, (2)Siemens Networks, Germany

17:00-17:15 12B4-4 In-Service Monitoring of WDM Passive Optical Network Using Novel Optical Reflectometry Based on Correlation Detection
Y. Takushima, Y. C. Chung
Korea Advanced Inst. of Sci. and Technol, Korea

17:15-17:30 12B4-5 In-Band OSNR Monitoring System Based on Link-by-Link Estimation for Reconfigurable Transparent Optical Networks
J. H. Lee, N. Yoshikane, T. Tsuritani, T. Otani
KDDI R&D Labs., Japan

**Room C**

**12C1: Transmission Fibers and Systems (9:00-10:15)**

Chair: M. Tateba

9:00-9:30 12C1-1 (Invited) New Fiber Designs and Fabrications for Data Transmission and Novel Fiber Devices
P. Nouchi, L. Gasca, P. Sillard, L. A. de Montmorillon, G. Melin, A. Pastouret
Draka Comteq, France

9:30-9:45 12C1-2 Ultra-High-Density Optical Fiber Cable and Its Application as Pre-Connectorized Cable with Adjustable Excess Length
Y. Yamada, K. Toge, K. Hogari
NTT, Japan

9:45-10:00 12C1-3 10 Gb/s WDM Transmission at 1064 and 1550 nm over 24 km PCF with Negative Power Penalties
K. Kurokawa, K. Tsujikawa, K. Tajima, K. Nakajima, I. Sankawa
NTT, Japan

10:00-10:15 12C1-4 Splice Characteristics of Trench-Assisted Bend-Insensitive Fiber Having Equivalent Dispersion Characteristics to SMF
T. Nunome, T. Yoshida, J. Takahashi, S. Matsuo
Fujikura, Japan

12C2: Fiber Devices and Lasers (11:00-12:30)

Chair: K. Oh

11:00-11:30 12C2-1 (Invited) Novel Fibers for Dispersion and Nonlinearity Management
S. Ramachandran
OFS Labs., USA

11:30-11:45 12C2-2 A Ring-Type Polarization-Maintaining \(\lambda/4\)-Shifted Distributed Feedback Fiber Laser Pumped by a 0.98 \(\mu\)m Laser Diode
Q. A. Ho Thi, A. Suzuki, M. Nakazawa
Tohoku Univ., Japan

11:45-12:00 12C2-3 Frequency Locking of a Single-Frequency Fiber Laser with Dual-Wavelength External Frequency-Stabilized Light Source
K. Mori, M. Matsuura, N. Kishi
Univ. of Electro-Communs., Japan

12:00-12:15 12C2-4 Active Birefringent Optical Loop Filter Using an SOA as a Phase Shifter
N. Onodera, T. Mansyo, K. Tsuji, T. Yamaguchi, J. Kim, M. Saruwatari
Natl. Defense Acad., Japan

12:15-12:30 12C2-5 High Birefringence Ring Filter with a Reflector: Application in Single-Frequency Fiber Lasers
G. Sun, D. S. Moon, D. Hwang, Y. Chung
GIST, Korea
12C3: Novel Amplifier Glass Material (Tutorial + Oral Presentation) (14:00-15:30)
Chair: M. Yamada, Y. Kondo

14:00-15:00  12C3-1 (Tutorial)  Glass Material Technology for Optical Amplification Media: Past, Present and Future
Y. Ohishi
Toyota Technological Inst., Japan ................................................................. 374

15:00-15:15  12C3-2  11 dB Gain in Tm-Doped Silica Glass Fiber
P. R. Watekar, S. Ju, W.-T. Han
GIST, Korea ......................................................................................... 376

15:15-15:30  12C3-3  Energy Transfer Analysis for Tb3+-Yb3+ Codoped Fluorophosphates Glasses
T. Yamashita(1,2), S. Horiguchi(1), Y. Arai(1), T. Suzuki(1), Y. Ohishi(1)
(1)Toyota Technological Inst., Japan, (2)Toyota Central R&D Labs., Japan .................. 378

12C4: Novel Fiber Devices (16:00-17:15)
Chair: K. Shiraishi, N. Shibata

16:00-16:30  12C4-1 (Invited)  Novel Bandwidth Control Mechanism in Fiber Based Tunable Filters
K. Oh(1), Y. Jung(2), W. Shin(2)
(1)Yonsei Univ., Korea, (2)GIST, Korea ......................................................... 380

16:30-16:45  12C4-2  Embedded Optical Microfiber Coil Resonator
F. Xu, G. Brambilla
Univ. of Southampton, UK ........................................................................ 382

16:45-17:00  12C4-3  A New Scheme of Hyperbolic-End Microlens Using Fusion Technology
I-SHOU Univ., Taiwan ............................................................................. 384

17:00-17:15  12C4-4  Plasmon Emission from Photoexcited Gold Nanoparticles Embedded in Germano-Silicate Fiber
A. Lin, H. J. Cho, W.-T. Han
GIST, Korea .......................................................................................... 386

Room D
12D1: Silicon Active Devices (9:00-10:00)
Chair: M. Aoki

9:00-9:30  12D1-2 (Invited)  Hetero III-V-N alloy/Si Technologies toward Monolithic OEIC Chips Including High-Dense Light Emitting Devices
H. Yonezu, Y. Furukawa, A. Wakahara
Toyohashi Univ. of Technol., Japan ......................................................... 388

9:30-9:45  12D1-3  GaInAsP/InP Membrane DFB Lasers Directly Boded on SOI Substrate with Rib-Waveguide Structure
T. Maruyama(1,2), T. Okumura(1), M. Kanemaru(1), S. Sakamoto(1), S. Tamura(1),
S. Arai(2)
(1)Tokyo Inst. of Technol., Japan, (2)JST-CREST, Japan .......................... 390

9:45-10:00  12D1-4  Fabrication of Silicon-on-Insulator Arrayed Waveguide Grating and Monolithic Power Monitor Array
P. S. Chan(1), D. X. Dai(2), A. W. Poon(3), H. K. Tsang(1)
(1)The Chinese Univ. of Hong Kong, China, (2)Zhejiang Univ., China, (3)Hong Kong Univ. of Sci. & Technol., China ..................................................... 392

12D2: Organic Optical Devices (11:00-12:00)
Chair: T. Hattori

11:00-11:15  12D2-1  CW Operation of Optical Amplification in Organic Dye-Doped Polymeric Channel Waveguide
K. Yamashita(1), K. Hase(1), H. Yanagi(2), K. Oe(1)
(1)Kyoto Inst. of Technol., Japan, (2)Nara Inst. of Sci. and Technol., Japan ................ 394

11:15-11:30  12D2-2  Ultrafast Optical Gain Switching for Wavelength Conversion from Silica to Polymer Optical Fibre Communication Wavelengths
R. Xia(1), C. Cheung(1), D. D. C. Bradley(1), A. Ruseckas(2), D. Amarasinghe(2),
I. D. W. Samuel(2)
(1)Imperial College London, UK, (2)Univ. of St Andrews, UK ........................ 396
12D3: Advanced High-speed Lasers (14:00-15:30)
Chair: Y. Nakano
14:00-14:30 12D3-1 (Invited) 1.3-μm-Wavelength Quantum-Dot Lasers for Temperature-Stable High-Speed Direct Modulation
(1)Fujitsu Labs., Japan, (2)Fujitsu, Japan, (3)OITDA, Japan, (4)Collaborative Inst. for Nano Quantum Information Electr., Japan, (5)The Univ. of Tokyo, Japan, (6)QD Laser, Japan .............................................................................................................................. 402

14:30-15:00 12D3-2 (Invited) High-Performance GaInNAs Lasers for Optical Communications
T. Kitatani, J. Kasai, K. Nakahara, K. Adachi, M. Aoki
Hitachi, Japan ................................................................................................................. 404

15:00-15:30 12D3-3 (Invited) Recent Progress of 40 Gbit/s Directly Modulated Lasers
U. Troppenz, J. Kreissl, W. Rehbein, C. Bornholdt
Heinrich-Hertz-Inst., Germany ........................................................................................ 406

12D4: VCSEL and Related Devices (16:00-17:30)
Chair: N. Yokouchi
16:00-16:30 12D4-1 (Invited) High-Speed 1.55 μm VCSELs
M.-C. Amann, W. Hofmann
Technical Univ. of Munich, Germany ............................................................................. 408

16:30-16:45 12D4-2 Tapered Hollow Waveguide Multiplexer for Multi-Wavelength VCSEL Array
N. Kitabayashi, A. Imamura, A. Matsutani, F. Koyama
Tokyo Inst. of Technol., Japan ........................................................................................ 410

16:45-17:00 12D4-3 Transverse-Mode and Polarization Characteristics of 1.55 μm Micromachined VCSELs
K. Hasebe(1), W. Janto(1), N. Nishiyama(1,2), C. Caneau(2), T. Sakaguchi(1), A. Matsutani(1), F. Koyama(1), C. E. Zah(2)
(1)Tokyo Inst. of Technol., Japan, (2)Corning, USA ..................................................... 412

17:00-17:15 12D4-4 10 Gbps/ch 1x10 VCSEL Array at 850-nm Wavelength
K. Takeda, T. Kondo, M. Yamamoto, R. Ishii, N. Ueki
Fuji Xerox, Japan ........................................................................................................... 414

17:15-17:30 12D4-5 Modeling of Slow Light Modulator with Tilt Coupling Scheme
K. Kuroki, G. Hirano, F. Koyama
Tokyo Inst. of Technol., Japan ........................................................................................ 416

Room E
12E1: Functional Passive Devices (III) (9:00-10:30)
Chair: S. C. Kin
9:00-9:30 12E1-1 (Invited) PLZT Photonic Functional Devices for Future Network Systems
H. Tsuda
Keio Univ., Japan ...................................................................................................... 418

9:30-9:45 12E1-2 Detection of Multiple Hydrocarbon Gases by Broadband Difference Frequency Generation Using Apodized \( \chi^{(2)} \) Grating
T. Umeki, M. Asobe, Y. Nishida, O. Tadanaga, K. Magari, T. Yanagawa, H. Suzuki
NTT, Japan ..................................................................................................................... 420

9:45-10:00 12E1-3 Novel Wavelength Filter in a Periodically Poled Ti:LiNbO\textsubscript{3} Channel Waveguide
GIST, Korea .................................................................................................................. 422

10:00-10:15 12E1-4 40 Gbit/s Switchable OR/XOR Logic Gates Using a PPLN Waveguide
J. Wang, J. Sun, Q. Sun
Huazhong Univ. of Sci. and Technol., China ................................................................... 424
10:15-10:30  12E1-5  Evaluation of Phase Change by a HfO\textsubscript{2}-Waveguide Mach-Zehnder Interferometer with a Ferro-Electric Liquid Crystal Cladding  
H. Sato, K. Nakatsuhara, T. Nakagami  
Kanagawa Inst. of Technol., Japan ................................................................. 426

12E2: Novel Waveguide Devices (I) (11:00-12:30)  
Chair: T. Katsuyama

11:00-11:30  12E2-1 (Invited)  Micromachined Tunable 1.55 \( \mu \)m Vertical Cavity, Multiple Air-Gap Filters and Lasers: Fabrication, Characterization, Scaling Potential and Applications  
H. Hillmer\(^{(1)}\), T. Kusserow\(^{(1)}\), N. Dharmarasu\(^{(1)}\), S. Irmer\(^{(1)}\), J. Daleiden\(^{(1)}\), A. Hasse\(^{(1)}\), M. Bartels\(^{(1)}\), B. Vengatesan\(^{(2)}\), T. Hayakawa\(^{(2)}\)  
\(^{(1)}\) Univ. of Kassel, Germany, \(^{(2)}\) Canare Electric, Japan ................................................................. 428

11:30-11:45  12E2-2  Silicon Photonic Wire Filter Using Asymmetric Sidewall Long-Period Gratings in a Two-Mode Waveguide  
Korea Advanced Inst. of Sci. and Technol., Korea .............................................. 430

11:45-12:00  12E2-3  Cross Absorption Modulation Enhancement in Silicon Waveguides  
Y. Liu, H. K. Tsang  
The Chinese Univ. of Hong Kong, China .......................................................... 432

12:00-12:15  12E2-4  Fabrication of Si Wire Optical Waveguides by Clad Formation by Selective Oxidation of Si  
K. Iiyama, S. Asai, M. Wakashima  
Kanazawa Univ., Japan ................................................................. 434

12:15-12:30  12E2-5  Apodization of Photorefractive Transmission Grating for Wavelength Filter  
H. Yoshida, A. Okamoto, K. Harasaka  
Hokkaido Univ. Japan ................................................................. 436

12E3: Planar Waveguide Circuits (I) (Tutorial + Oral Presentation) (14:00-15:30)  
Chair: M. Itoh

14:00-15:00  12E3-1 (Tutorial)  Recent Progress on Waveguide Device Design and Its Applications  
K. Okamoto  
Univ. of California, Davis, USA ................................................................. 438

15:00-15:15  12E3-2  Multi-Wavelength Channel Selective Switch by Cascading TO-tunable Quadruple Series-Coupled Microrings  
Y. Goebuchi\(^{(1)}\), M. Hisada\(^{(1)}\), T. Kato\(^{(2)}\), Y. Kokubun\(^{(1)}\)  
\(^{(1)}\) Yokohama Natl. Univ., Japan, \(^{(2)}\) Tokyo Inst. of Technol., Japan ................................................................. 440

15:15-15:30  12E3-3  Fabrication and Characterization of Tunable Chromatic Dispersion Compensator Based on Hollow Waveguide  
T. Takeishi, T. Sakaguchi, F. Koyama  
Tokyo Inst. of Technol., Japan ................................................................. 442

12E4: Planar Waveguide Circuits (II) (16:00-17:30)  
Chair: H. Tsuda

16:00-16:30  12E4-1 (Invited)  Compact ROADM Devices Based on PLC Technology  
M. Itoh  
NTT, Japan ................................................................. 444

16:30-17:00  12E4-2  Polarization Independent Microring Resonator Filter Using Internal Stress and Temperature Control  
N. Kobayashi, N. Zaizen, Y. Kokubun  
Yokohama Natl. Univ., Japan ................................................................. 446

16:45-17:00  12E4-3  Super-High-\( \Delta \) Silica-Based Flat-Passband Filter Using AWG and Cascaded Mach-Zehnder Interferometers  
K. Maru\(^{(1,2,3)}\), T. Mizumoto\(^{(1)}\), H. Uetsuka\(^{(2,3)}\)  
\(^{(1)}\) Tokyo Inst. of Technol., Japan, \(^{(2)}\) Hitachi Cable, Japan, \(^{(3)}\) OITDA, Japan ................................................................. 448

17:00-17:15  12E4-4  Fabrication and Evaluation of Tunable Band-Selection Interleaver Switch on Silicon Waveguide  
Waseda Univ., Japan ................................................................. 450

17:15-17:30  12E4-5  Roughness Reduction of Si Waveguides by KrF Excimer Laser Reformation  
E.-Z. Liang\(^{(1)}\), S.-C. Hung\(^{(2)}\), C.-F. Lin\(^{(2)}\)  
\(^{(1)}\) Diwan College of Management, Taiwan, \(^{(2)}\) Natl. Taiwan Univ., Taiwan ................................................................. 452
### 12P: Poster Session (12:30-14:00)

12P-1  
**A Wavelength and Converter Assignment Scheme for Decreasing Blocking Probability in Wavelength-Routed Networks**  
Y. Fukushima, T. Ooishi, T. Yokohira  
Okayama Univ., Japan  

12P-2  
**Experimental Analysis of Bidirectional WDM Networks Using Two Identical Sets of Wavelengths on a Single Fiber**  
H. Obara, M. Sakata  
Akita Univ., Japan  

12P-3  
**A Novel Star-Ring Optical Regional Network Architecture with Dynamic Wavelength/Waveband Broadcast and Select**  
Y. Cai, M. Matsuura, N. Kishi, T. Miki  
The Univ. of Electro-Communs., Japan  

12P-4  
**Application of Two Upstream Wavelength in Fairness and Priority Environments**  
J. Park, B. Choi, Y. Park  
Kookmin Univ., Korea  

12P-5  
**Traffic Grooming & Wavelength Assignment System over WWW for Regional WDM Optical IP Networks**  
A. Ueno, S. Kawase, M. Taniue, O. Koyama, Y. Katsuyama  
Osaka Prefecture Univ., Japan  

12P-6  
**Design Tool for Wavelength Routing Functions**  
G. Mouri, K. Okada, K. Oguchi  
Seikei Univ., Japan  

12P-7  
**Filterless Optical Networks: A Unique and Novel Passive WAN Network Solution**  
C. Tremblay(1), F. Gagnon(1), B. Châtelain(1), É. Bernier(2), M. P. Bélanger(2)  
(1)Université du Québec – ÉTS, Canada, (2)Nortel, Canada  

12P-8  
**Wavelength Monitoring of NRZ Signal Using NRZ-to-PRZ Conversion by π-Phase Shifted FBG**  
T.-Y. Kim(1), M. Hanawa(2), C. K. Oh(1), C.-S. Park(1)  
(1)GIST, Korea, (2)Univ. of Yamanashi, Japan  

12P-9  
**A Spatial BER Contour for Indoor Visible Light LAN and Mobile Networking**  
KOPTI, Korea  

12P-10  
**Simulation Results on Transmission of 10Gb/s Optical MSK System with Narrowband Frequency Discrimination Receiver over 560 km Standard Single Mode Fiber without Dispersion Compensation**  
T. L. Huynh, T. Sivahumaran, L. N. Binh, K. K. Pang  
Monash Univ., Australia  

12P-11  
**Performance Analysis of New Switch Architecture for Variable Packet Size**  
M. S. Salleh(1,2), K. Dimyati(2)  
(1)UPM-MTDC, Malaysia, (2)Univ. Malaya, Malaysia  

12P-12  
**Signal Quality Monitoring of 40 Gb/s Optical Signal Using a Low Bit-Rate Reference Channel**  
M. N. Petersen, T. Tokle  
Technical Univ. of Denmark, Denmark  

12P-13  
**Investigation of Performance Degradations of DPSK RZ/NRZ Signals Due to Non-Optimal Driving Voltage and MZDI Delay**  
M. Haris(1), J. Yu(1,2), G.-K. Chang(1)  
(1)Georgia Inst. of Technol., USA, (2)NEC Labs. America, USA  

12P-14  
**Selectable Multicast Using Raman-Assisted Four-Wave Mixing in Dispersion Shifted Fiber**  
K. Lau(1), L. Xu(1,2), S. H. Wang(1), L. F. K. Lu(1), P. K. A. Wai(1), C. Lu(1), H. Y. Tam(1)  
(1)The Hong Kong Polytechnic Univ., China, (2)Univ. of Sci. and Technol. of China, China  

12P-15  
**Fabrication and Characteristics of Broadband Cr-Doped Fibers by Drawing Tower**  
(1)Natl. Sun Yat-sen Univ., Taiwan, (2)Natl. Taiwan Univ., Taiwan, (3)LeHigh Univ., USA  

---

**xxv**
12P-16  Gain and Noise Figure Enhancement through C-Band Signal Injection in a Double-Pass L-Band EDFA  
K.-M. Feng, P.-J. Tsai  
Natl. Tsing Hua Univ., Taiwan ................................................................. 484

12P-17  Large Effective Area Photonic Crystal Fibers with Negative Dispersion and Ultra-Low Splicing Loss  
Univ. of the Ryukyus, Japan ........................................................................... 486

12P-18  A Novel Distributed Fiber-Optic Sensor Based on a Mach-Zehnder Interferometer  
Q. Sun, D. Liu, J. Wang  
Huazhong Univ. of Sci. and Technol., China .................................................. 488

12P-19  Asymmetrical Raman Resonator for Multiwavelength Raman Fiber Laser  
Y.-E. Im(1-2), S. Hann(2), D.-H. Kim(2), C.-S. Park(1)  
(1)GIST, Korea, (2)KOPTI, Korea ................................................................. 490

12P-20  Design of Dispersionless Fiber Bragg Grating Filters by Using Lagrange Multiplier Constrained Optimization Algorithm  
C.-L. Lee(1), R.-K. Lee(2), S.-K. Liaw(3)  
(1)Natl. United Univ., Taiwan, (2)Natl. Tsing-Hua Univ., Taiwan, (3)Natl. Taiwan Univ. of Sci. and Technol., Taiwan ........................................ 492

12P-21  Effect of EDFA Saturation Cross Talk in Direct-Detection Schemes  
P. S. Chan, H. K. Tsang  
The Chinese Univ. of Hong Kong, China ...................................................... 494

12P-22  Experimental Demonstration of Wavelength-Selective In-Fiber Optical Intensity Modulation  
M. Rajabvand, F. Behnia, M. T. Fatehi  
Sharif Univ. of Technol., Iran ........................................................................ 496

12P-23  Defect-Core Hexagonal-Lattice Photonic Crystal Fibers with Flattened Chromatic Dispersion and Low Confinement Loss  
Univ. of the Ryukyus, Japan ........................................................................... 498

12P-24  A Simple Optical Power Limiter for 40 GHz Pulses Based on SOA Saturation  
G. Contestabile, M. Presi, R. Proietti, N. Calabretta, E. Ciaramella  
Scuola Superiore Sant’Anna, Italy ................................................................. 500

12P-25  Low-Cost Undercut-Etching-Active-Region Method to Fabricate Laterally Tapered Active Waveguide for Electroabsorption Modulator and Spot-Size Converters Integration  
F. Z. Lin, S. A. Tsai, T. H. Wu, J. S. You, Y. J. Chiu  
NSYSU, Taiwan ............................................................................................ 502

12P-26  Power Equalization for SOA-Based All-Optic Switch by Optical Control Pulse Optimization  
S. Fu(1), P. Shum(1), Y. D. Gong(2), X. Li(1), H. Q. Lam(1)  
(1)Nanyang Technological Univ., Singapore, (2)Inst. for InfoComm, Singapore ................................................................. 504

12P-27  A Multiple-Operation-State Optoelectronic Switch  
D.-F. Guo(1), J.-H. Tsai(2), M.-Y. Fu(1)  
(1)Air Force Acad., Taiwan, (2)Natl. Kaohsiung Normal Univ., Taiwan ................................................................. 506

12P-28  Electrorefractive Effect in InGaAsP Five Layer Asymmetric Coupled Quantum Well (FACQW) for High Performance Optical Modulators  
M. Fukuoka(1), T. Toy(1), T. Arakawa(1), K. Tada(2)  
(1)Yokohama Natl. Univ., Japan, (2)Kanazawa Inst. of Technol., Japan ................................................................. 508

12P-29  Wavelength Converter Using an Optical MEMS Switch  
J. Lee(1), J.-H. Bang(2)  
(1)Korea Polytechnic Univ., Korea, (2)Electr. and Communs. Res. Inst., Korea ................................................................. 510

12P-30  Expanding the Bandwidth of Slow and Fast Pulse Propagation in Coupled Micro-Resonators  
D. D. Smith(1,2), H. Chang(2)  
(1)NASA Marshall Space Flight Center, USA, (2)Univ. of Alabama in Huntsville, USA ................................................................. 512

12P-31  Novel 2x2 Photonic Switch Based on Multimode Interference Effect  
Y. Narita, M. Yasumoto, H. Tsuda  
Keio Univ., Japan ........................................................................................ 514
12P-32 Micro-Spherical Refractive Index Variation Sensor Using High Refractive Index Layer
A. Kobayashi, J. Osawa, H. Okayama, H. Nakajima
Waseda Univ., Japan ................................................................. 516

12P-33 Formation of MgO:LiNbO₃ Domain-inverted Grating for Quasi-phase Matching by Voltage Application with Insulation Layer Cladding
T. Tsubouchi, N. Horikawa, M. Fujimura, T. Suhara
Osaka Univ., Japan ................................................................. 518

12P-34 DMD Evaluation of Rectangular Optical Waveguide for 10Gb/s Transmission with Numerical Analysis
AIST, Japan ................................................................. 520

12P-35 Cesaro Means of Fourier Series for Designing Interleaver Filters with Planar Lightwave Circuit-Type Lattice Structure
J. Zhang
Shanghai Univ., China ................................................................. 522

12P-36 High Extinction Ratio of Switched Packets by Two-Stage Four Wave Mixing in Highly Nonlinear Dispersion-Shifted Fiber
(1)The Univ. of Hong Kong, China, (2)The Univ. of Cambridge, UK ................................................................. 524

Friday, July 13

Room A

13A1: Large Capacity Optical Network Platform (9:00-10:30)
Chair: M. Tsurusawa

9:00-9:30 13A1-1 (Invited) 100G Ethernet for Packet Transport Networks
A. Kirstädter
Nokia Siemens Networks, Germany ................................................................. 526

B. Kozicki, H. Takara
NTT, Japan ................................................................. 528

9:45-10:00 13A1-3 40 Gb/s All-Optical Multi-Wavelength Conversion Via a Single SOA-MZI for WDM Wavelength Multicast
N. Yan(1), T. Silveira(2,3), A. Teixeira(2), A. Ferreira(2), E. Tangdiongga(1), P. Monteiro(2,3), A. M. J. Koonen(1)
(1)Eindhoven Univ. of Technol., The Netherlands, (2)Inst. de Telecomms., Portugal, (3)Siemens Networks, Portugal ................................................................. 530

10:00-10:15 13A1-4 Generation-Free-Platform Architecture with Flexible Physical and Logical Links Using Optical Technology
S. Yanagimachi, Y. Hidaka, J. Suzuki, H. Higuchi, T. Yoshikawa, A. Iwata
NEC, Japan ................................................................. 532

10:15-10:30 13A1-5 Optical Packet Interconnect System Using High-Speed SOA Switch for Peta-Scale Computing System
K. Sone, Y. Aoki, G. Nakagawa, Y. Kai, S. Yoshida, Y. Takita, S. Kinoshita, H. Onaka
Fujitsu, Japan ................................................................. 534

13A2: Optical Network Design (11:00-12:30)
Chair: M. Jinno

11:00-11:15 13A2-1 A Study on Fault Recovery of Optical Paths in Photonic Cross-Connect Systems
S. Yoshida, E. Horiiuchi, Y. Baba, Y. Akiyama, K. Onohara, T. Mizuochi, S. Seno
Mitsubishi Electric, Japan ................................................................. 536

11:15-11:30 13A2-2 A Study of Equipment Protection for High Availability of Control and Data Planes in Photonic Cross-Connect
K. Onohara, Y. Akiyama, T. Mizuochi, T. Ichikawa, H. Sato, K. Okubo, E. Horiiuchi, S. Yoshida, Y. Baba
Mitsubishi Electric, Japan ................................................................. 538
11:30-11:45  13A2-3  Fast Generation of Orthogonal Periodic Polynomials and Its Application to Smooth Approximation of the Internet Traffic  
H. Hasegawa, N. Matsusue, K. Sato  
Nagoya Univ., Japan ................................................................. 540

11:45-12:00  13A2-4  A Study of Optical Fiber Network Capacity Escalation due to Localized Extreme Traffic Increase  
H. Taga  
Natl. Sun Yat-Sen Univ., Taiwan .................................................. 542

12:00-12:15  13A2-5  Prefetching Protocol Proxy with Optimal Mirror Selection and Burst Transmission  
T. Tsuji, J. Honma, S. Shimizu, Y. Arakawa, N. Yamanaka  
Keio Univ., Japan ........................................................................ 544

12:15-12:30  13A2-6  An Algorithm for Resource Optimization of Consolidating Two Coexisting Networks  
The Chinese Univ. of Hong Kong, China ...................................... 546

13A3: Optical Packet Switching (14:00-15:15)  
Chair: T. Tanemura

14:00-14:15  13A3-1  IP over Optical Packet Switched Network  
H. Furukawa\(^{(1)}\), N. Wada\(^{(1)}\), H. Harai\(^{(1)}\), M. Naruse\(^{(1)}\), H. Otsuki\(^{(1)}\), T. Miyazaki\(^{(1)}\), K. Ikekazawa\(^{(2)}\), A. Toyama\(^{(2)}\), N. Itou\(^{(2)}\), H. Shimizu\(^{(2)}\), H. Fujinuma\(^{(3)}\), H. Iduka\(^{(3)}\), G. Cincotti\(^{(4)}\), K. Kitayama\(^{(5)}\)  
\(^{(1)}\)NICT, Japan, \(^{(2)}\)Yokogawa Electric, Japan, \(^{(3)}\)NTT Electr., Japan, \(^{(4)}\)Univ. Roma Tre, Italy, \(^{(5)}\)Osaka Univ., Japan ...................................................... 548

14:15-14:30  13A3-3  Demonstration of a Semiconductor-Based Multi-Wavelength Light Source and Its Application to Optical Label Processing  
K. Okamoto\(^{(1)}\), H. Uenohara\(^{(1,2)}\)  
\(^{(1)}\)Tokyo Inst. of Technol., Japan, \(^{(2)}\)JST-CREST, Japan .......................................................... 550

14:30-14:45  13A3-4  SubFrame-Based Slot Reservation Scheme for Minimizing Transmission Delay in Optical Slot Switching Network  
F. Uehara, T. Kasahara, M. Hayashitani, D. Ishii, Y. Arakawa, S. Okamoto, N. Yamanaka  
Keio Univ., Japan ........................................................................ 552

14:45-15:00  13A3-5  Analytic Model for Optical Packet Switch with Output Variable All-Optical Buffers under Asynchronous Variable-Length Packet Traffic  
H. Lee, C. Yun, W. Lim, K. Kim  
GIST, Korea .................................................................................. 554

15:00-15:15  13A3-6  Throughput Improvement of Deflection Routed Networks with an All-Optical Packet Scrambler  
C. Y. Li, P. K. A. Wai  
The Hong Kong Polytechnic Univ., China ..................................... 556

Room B  
13B1: Optical Signal Processing (II) (9:00-10:30)  
Chair: M. Hanawa

9:00-9:15  13B1-1  All-Optical NRZ-to-RZ Format Conversion Based on Optical Parametric Amplifier  
H. K. Y. Cheung, R. W. L. Fung, P. C. Chui, K. K. Y. Wong  
The Univ. of Hong Kong, China .................................................... 558

9:15-9:30  13B1-2  All-Optical RZ-OOK to RZ-BSPK Conversion Using Cross Phase Modulation  
N. Kulpawan\(^{(1)}\), P. Kaewplung\(^{(2)}\)  
Chulalongkorn Univ., Thailand ...................................................... 560

9:30-9:45  13B1-3  All-Optical NRZ-to-PRZ Converter Based on Cascaded Long-Period Fiber Gratings  
S.-W. Jeon\(^{(1)}\), T.-Y. Kim\(^{(1)}\), M. Hanawa\(^{(2)}\), Y. Chung\(^{(1)}\), C.-S. Park\(^{(1)}\)  
\(^{(1)}\)GIST, Korea, \(^{(2)}\)Univ. of Yamanashi, Japan .................................. 562

9:45-10:00  13B1-4  40 Gbit/s OTDM to 4x10 Gbit/s WDM Conversion Via Birefringence Switching  
L. Xu, P. S. Chan, Y. Liu, H. K. Tsang  
The Chinese Univ. of Hong Kong, China ...................................... 564

10:00-10:15  13B1-5  Performance of Wavelength Exchange in Anomalous-Dispersion Region  
R. W. L. Fung, H. K. Y. Cheung, K. K. Y. Wong  
The Univ. of Hong Kong, China .................................................... 566
10:15-10:30  13B1-6  All-Optical XNOR Gate Using Fiber Optical Parametric Amplifier
D. M.-F. Lai, B. P.-P. Kuo, K. K.-Y. Wong
The Univ. of Hong Kong, China ......................................................... 568

13B2: MLSE Technologies (11:00-12:15)
Chair: T. Mizuochi
11:00-11:30  13B2-1 (Invited)  Maximum Likelihood Sequence Estimation for Impairment Compensation in Advanced Modulation Formats
J. Zhao, L. K. Chen, C. C. K. Chan
The Chinese Univ. of Hong Kong, China ......................................... 570
11:30-11:45  13B2-2  Maximum Likelihood Sequence Estimation for Chromatic-Dispersion Compensation in 4-ASK Modulation Format
J. Zhao, L.-K. Chen, C.-K. Chan
The Chinese Univ. of Hong Kong, China ......................................... 572
11:45-12:00  13B2-3  Simultaneous Compensation of PMD and CD in a 10.7Gb/s Field Trial Based on the MLSE
(1)T-Systems Enterprise Services, Germany, (2)CoreOptics, Germany ........................................ 574
12:00-12:15  13B2-4  Mitigating Sampling Phase Sensitivity of the MLSE by Overlapping Branch Metrics
F. N. Hauske(1), B. Lankl(1), E.-D. Schmidt(2)
(1)Univ. of the Federal Armed Forces, Germany, (2)Siemens Networks, Germany ............ 576

13B3: Optical Signal Processing (III) (14:00-15:30)
Chair: M. Matsumoto
14:00-14:30  13B3-1 (Invited)  Optical Regeneration Technologies for Future Photonic Network
S. Watanabe
Fujitsu Labs., Japan ........................................................................ 578
14:30-14:45  13B3-2  DPSK Signal Restoration Using Four-Wave Mixing in a Dispersion-Flattened Highly Nonlinear Photonic Crystal Fiber
M. P. Fok, C. Shu
The Chinese Univ. of Hong Kong, China ......................................... 580
14:45-15:00  13B3-3  Experimental Demonstration of Filter-Free Wavelength Conversion Using Nonlinear Optical Loop Mirror at 10Gbit/s
M. Kannan, S. Oda, A. Maruta
Osaka Univ., Japan ......................................................................... 582
15:00-15:15  13B3-4  160 Gb/s Retiming Using Rectangular Pulses Generated Using a Superstructured Fibre Bragg Grating
(1)Technical Univ. of Denmark, Denmark, (2)Univ. of Southampton, UK ..................... 584
15:15-15:30  13B3-5  All Optical Clock Recovery at 10 GHz Using a Fabry-Perot Laser Diode
X. Fang, P. K. A. Wai, C. Lu, H. Y. Tam, K. K. Qureshi
The Hong Kong Polytechnic Univ., China ....................................... 586

13B4: Dispersion Compensation (16:00-17:15)
Chair: M. Hanawa
16:00-16:15  13B4-1  Block Turbo Code Encoded 24.8 Gbps RZ-DQPSK Experiment Using Parallel Prefix Network Based Differential Precoder
Y. Konishi, T. Mizuochi, K. Ouchi, K. Onohara, K. Kubo, S. Mitani, K. Ishida, K. Shimizu
Mitsubishi Electric, Japan ................................................................ 588
16:15-16:30  13B4-2  Optimization of Dispersion Compensation Value in Optical Fiber Transmission System
T. Shtaba, K. Kikushima
NTT, Japan .................................................................................. 590
16:30-16:45  13B4-3  Adaptive Steepest-Descent-Feedback Control of Tunable Dispersion Compensators Using a Three-Point Sampling Method in Time-Domain Waveforms
K. Tanizawa, A. Hirose
The Univ. of Tokyo, Japan ............................................................. 592
Proposal for Coordinate Transformed Electronic Pre-Compensator and Investigation of Its Robustness to Bias Error
T. Sugihara, T. Mizuochi, H. Kubo, K. Shimizu
Mitsubishi Electric, Japan..............................................................594

Record PMD Mitigation of 11 ps for 43 Gb/s RZ-DPSK by Distributed Polarisation Scrambling
A. Klekamp, H. Bülow
Alcatel Res. and Innovation, Germany........................................596

Room C
13C1: Fiber Nonlinearities (9:00-10:30)
Chair: K. S. Abedin

9:00-9:30 13C1-1 (Invited) Ultra-Compact Highly Nonlinear Fiber Module Technologies
M. Takahashi, Y. Mimura, J. Hiroishi, M. Tadakuma, R. Sugizaki, T. Yagi
Furukawa Electric, Japan..............................................................598

Comparison of Wavelength Converters Composed of HNL-DSF and Bi-HNLF by a New Figure-of-Merit
S. Yamashita, H. Kuno
The Univ. of Tokyo, Japan ............................................................600

SC Spectrum Broadening with Tapered Photonic Crystal Fiber
Y. Fuji(1), N. Karasawa(2), Y. Matsubara(2), S. Kobayashi(2)

Bragg Soliton Pulse Compression in Non-Uniform Fiber Bragg Gratings
K. Senthilnathan(1), Q. Li(1), P. K. A. Wai(1), K. Nakkeeran(2)
(1)The Hong Kong Polytechnic Univ., China, (2)Univ. of Aberdeen, UK......................604

Spatial Evolution of Supercontinuum Generation along a Varying Dispersion Tapered Fiber
Z. Wang(1), H. Sone(2), Y. Tsuji(2), M. Imai(1), S. Sato(1)
(1)Muroran Inst. of Technol., Japan, (2)Kitami Inst. of Technol., Japan.........................606

13C2: Fiber Sensors and Measurements (11:00-12:30)
Chair: S. Yamashita

11:00-11:30 13C2-1 (Invited) Recent Advances in the Fiber Optic Sensors Based on Stimulated Brillouin Scattering
X. Bao, C. Zhang, I. F. Ozkan, M. Mohareb, W. Li, F. Ravet, L. Chen
Univ. of Ottawa, Canada ................................................................608

Study on Low Cost Double-Fiber Model Sensor for Distributed Strain Monitoring
C. Wang, K. Shida
Saga Univ., Japan ........................................................................610

Measurement of the Brillouin Gain/Phase Characteristics in Optical Fibers Using a Double-Modulation Technique of a Single Light Source
K. Tsuji, J. Kim, T. Yamaguchi, N. Onodera, M. Saruwatari
Natl. Defense Acad., Japan ............................................................612

Measurement of Raman Gain Efficiency for an Optical Fiber Cable Installed in the Field by Using OTDR
K. Oro(1), H. Hatada(2), I. Yamashita(1), T. Yabu(2), M. Ohashi(2)
(1)Kansai Electric Power, Japan, (2)Osaka Prefecture Univ., Japan...............................614

13C2-5 Equivalent Optical Circuit Synthesis Using Polarimetric OTDR for Simulating Two-Pump Optical Fibre Parametric Amplifiers
T. Ozeki, T. Kanou, K. Hayashi, T. Kudou
Sophia Univ., Japan ..................................................................616

13C3: Fiber Gratings and Applications (14:00-15:30)
Chair: X. Bao

14:00-14:15 13C3-1 Recent Advances in the Design and Fabrication of High Channel-Count Fiber Bragg Gratings
M. Li(1), H. Li(1), Y. Sheng(2)
(1)Shizuoka Univ., Japan, (2)Laval Univ., Canada.........................................................618
14:15-14:30  13C3-2  Broadband Long Period Grating on Hollow Optical Fiber with Femtosecond Laser Pulses  
W. Ha(1), Y. Jung(2), J. Kim(2), W. Shin(2), I. Sohn(2), D. Ko(2), J. Lee(2), K. Oh(1)  
(1)Yonsei Univ., Korea, (2)GIST, Korea ................................................................. 620

14:30-14:45  13C3-3  Ultrasonic Hydrophone Based on Etched Distributed-Bragg-Reflector Fiber Laser  
L.-Y. Shao(1,2), S.-T. Lau(1), X. Dong(1), H.-Y. Tam(1), H. L. W. Chan(1), A. P. Zhang(2)  
(1)The Hong Kong Polytechnic Univ., China, (2)Zhejiang Univ., China ................. 622

14:45-15:00  13C3-4  All-Fiber Tunable Band Rejection Filter Based on Helicoidal Long-Period Fiber Grating Pair  
J. Lee(1)  
(1)GIST, Korea, (2)Yonsei Univ., Korea ................................................................. 624

15:00-15:15  13C3-5  A Cascadable Approach for Widely Tunable Optical Delay of Phase-Modulated Signals  
M. P. Fok, C. Shu  
The Chinese Univ. of Hong Kong, China .......................................................... 626

15:15-15:30  13C3-6  Polarization Dependency Reduction on Long Period Fiber Grating by Side Loading  
T. Kondo, M. Hanawa, K. Nakamura  
Univ. of Yamanashi, Japan .................................................................................. 628

13C4: Pulse Generation and Amplification (16:00-17:15)  
Chair: N. Kishi
16:00-16:15  13C4-1  Femtosecond Pulse Generation by Nonlinear Polarization Rotation in a Bismuth-Based Er-Doped Fiber Laser  
N. Tarumi, Y.-W. Song, S. Yamashita  
The Univ. of Tokyo, Japan .................................................................................... 630

16:15-16:30  13C4-2  10 GHz Regeneratively Mode-Locked SOA Fiber Ring Laser and Its Linewidth Characteristics  
M. Yoshida, A. Ono, M. Nakazawa  
Tohoku Univ., Japan .......................................................................................... 632

16:30-16:45  13C4-3  Stable Pulse Generation from a Rational Harmonic Mode-Locked Fiber Ring Laser Using Carrier-Suppressed Return-to-Zero Modulation Format  
S. Yamanaka, J. Maeda  
Tokyo Univ. of Sci., Japan .................................................................................... 634

16:45-17:00  13C4-4  Pulse Dropout and Subharmonic Locking in an Active Mode-Locked Birefringent Fiber Laser  
H. Q. Lam(1), P. Shum(1), L. N. Binh(2), Y. D. Gong(2), M. Tang(1), S. Fu(1)  

17:00-17:15  13C4-5  Development of a Small Size of Erbium Doped Optical Fiber Amplifier  
K. Haru, S. Shikii, S. Aoki, Y. Tamura, H. Takano  
Optohub., Japan ................................................................................................. 638

13C4-6  Withdrawn

Room E
13C5: Plastic Fibers and Dispersion Measurements (16:00-16:30)  
Chair: I. Sasaki
16:00-16:15  13C5-2  Dispersion Parameter and Fiber Length Measurements Technique over Multi-Wavelength Bands  
Y. C. Kim, H. D. Kim  
Kyungpook Natl. Univ., Korea ................................................................................. 642

16:15-16:30  13C5-3  Dispersion and Nonlinear Coefficient Measurements in Optical Fibres Using Soliton-Effect Compression  
T. N. Nguyen(1), T. Chartier(1), M. Thual(1), P. Rochard(1), L. Provino(2), A. Monteville(2),  
N. Traynor(2), V. Gaillard(3), C. Lupi(3), D. Leduc(3)  
(1)ENSSAT, France, (2)PERFOS, France, (3)Univ. de Nantes, France ............... 644

xxxi
Room D

**13D1: Semiconductor Lasers (I) (9:00-10:30)**

Chair: TBD

<table>
<thead>
<tr>
<th>Time</th>
<th>Session</th>
<th>Title</th>
<th>Authors</th>
</tr>
</thead>
<tbody>
<tr>
<td>9:00-9:30</td>
<td>13D1-1</td>
<td>Tunable Lasers for Optical Systems; Technologies and Commercialisation</td>
<td>J. Buus, Gayton Photonics, UK</td>
</tr>
<tr>
<td>9:30-9:45</td>
<td>13D1-2</td>
<td>Effect of Initial Phase in Wavelength Tunable DFB Laser with High Coupling Coefficient Gratings</td>
<td>N. Nunoya, Y. Shibata, H. Ishii, H. Okamoto, Y. Kawaguchi, Y. Kondo, H. Oohashi NTT, Japan</td>
</tr>
</tbody>
</table>

**10:00-10:15**

13D1-4 The Impact of InAlGaAs Barriers on Material and Differential Gain of Quantum Wells on Low Indium Content InGaAs Ternary Substrates

T. Fujisawa, M. Arai, T. Yamanaka, Y. Kondo, H. Yasaka NTT, Japan...

**10:15-10:30**

13D1-5 Analysis of Large Kink Mechanism in L-I Characteristics of Tunnel Injection Lasers

H. Nakajima, T. Miyamoto, T. Iwasaki, Y. Higa, F. Koyama Tokyo Inst. of Technol., Japan...

---

**13D2: Semiconductor Lasers (II) (11:00-12:30)**

Chair: J. Buus

<table>
<thead>
<tr>
<th>Time</th>
<th>Session</th>
<th>Title</th>
<th>Authors</th>
</tr>
</thead>
<tbody>
<tr>
<td>11:00-11:30</td>
<td>13D2-1</td>
<td>Photon Funneling from Photonic Crystal Nanolasers</td>
<td>Y.-H. Lee, I.-K. Hwang, S.-H. Kim KAIST, Korea</td>
</tr>
<tr>
<td>11:30-11:45</td>
<td>13D2-2</td>
<td>GaInNAs Distributed Feedback (DFB) Laser Diode with High Resistive Semiconductor Current-Blocking Layer</td>
<td>J. Hashimoto(1,2), K. Koyama(1,2), T. Ishizuka(1,2), Y. Tsuji(1), K. Fujii(1), T. Yamada(1,2), C. Fukuda(1), Y. Onishi(1), T. Katsuyama(1,2) Sumitomo Electric Inds., Japan, OITDA, Japan</td>
</tr>
<tr>
<td>11:45-12:00</td>
<td>13D2-3</td>
<td>1.55-μm-Wavelength λ/4-Shifted DFB Lasers with High-Density InAsSb Quantum-Dot Active Layers</td>
<td>M. Matsuda(1), K. Kawaguchi(1,2), A. Uetake(1,2), H. Kuwatsuka(1,2), M. Ekawa(1), T. Yamamoto(1), M. Sugawara(1,3), Y. Arakawa(1,2) Fujitsu Labs., Japan, Fujitsu, Japan, QDLaser, Japan, The Univ. of Tokyo, Japan</td>
</tr>
<tr>
<td>12:00-12:15</td>
<td>13D2-4</td>
<td>Fabrication and Optical Properties of Hybrid-Type Pillar Microcavity</td>
<td>T. Yamaguchi(1,2), T. Tawara(1), H. Gotoh(1), H. Kamada(1), H. Okamoto(1), H. Nakano(1), O. Mikami(1) NTT, Japan, Tokai Univ., Japan</td>
</tr>
<tr>
<td>12:15-12:30</td>
<td>13D2-5</td>
<td>Differential Gain in InGaAsN Quantum Well Structures</td>
<td>M. S. Wartak, P. Weetman Wilfrid Laurier Univ., Canada</td>
</tr>
</tbody>
</table>

---

**13D3: Optical Modulators (14:00-16:00)**

Chair: TBD

<table>
<thead>
<tr>
<th>Time</th>
<th>Session</th>
<th>Title</th>
<th>Authors</th>
</tr>
</thead>
<tbody>
<tr>
<td>14:00-14:30</td>
<td>13D3-1</td>
<td>Wavelength Tunable Laser Module Integrated with InP Mach-Zehnder Modulator</td>
<td>K. Tsuzuki, N. Kikuchi, H. Sanjo, Y. Shibata, H. Oohashi, H. Yasaka NTT, Japan</td>
</tr>
<tr>
<td>14:30-14:45</td>
<td>13D3-2</td>
<td>Multi-Carrier Light Generator Using Phase Modulators and Chirped Fiber Bragg Grating</td>
<td>T. Yamamoto, T. Komukai, K. Suzuki, A. Takada NTT, Japan</td>
</tr>
<tr>
<td>14:45-15:00</td>
<td>13D3-3</td>
<td>LiTaO3 Electro-Optic Polarization Modulator Utilizing Periodically Poled Structure</td>
<td>H. Murata, A. Takahashi, Y. Okamura Osaka Univ., Japan</td>
</tr>
</tbody>
</table>

---

xxxii
15:00-15:15  13D3-4  An 80-GHz Carrier-Suppressed Optical Pulse Generator Using Two Cascaded Phase Modulators Followed by a Delay-Interferometer  G.-W. Lu, T. Miyazaki  
NICT, Japan ................................................................. 672

(1)Keio Univ., Japan, (2)NTT, Japan, (3)Tokyo Univ. of Agriculture and Technol., Japan...... 674

NICT, Japan ................................................................. 676

15:45-16:00  13D3-7  50-nm Wavelength Tunable Ultra-Flat Comb Generation Using Single-Stage Mach-Zehnder Modulator  T. Sakamoto, T. Kawanishi, M. Izutsu  
NICT, Japan ................................................................. 678

Room E

13E1: Novel Waveguide Devices (II) (9:00-10:30)
Chair: M. Steel

(1)AIST, Japan, (2)Nippon Sheet Glass, Japan ................................................................. 680

(1)Yonsei Univ., Korea, (2)GIST, Korea ........................................................................ 682

9:45-10:00  13E1-3  Polarization Insensitive and Low-Loss Tunable 3D Hollow Waveguide for Tunable Photonic Devices  M. Kumar, F. Koyama  
Tokyo Inst. of Technol., Japan ...................................................................................... 684

10:00-10:15  13E1-4  Box-Like Filter Response of Quadruple Series Coupled Microring Resonator by Coupling Efficiency Control  M. Hisada(1), Y. Goebuchi(1), T. Kato(2), Y. Kokubun(1)  
(1)Yokohama Natl. Univ., Japan, (2)Tokyo Inst. of Technol., Japan ........................................ 686

NTT, Japan ..................................................................................................................... 688

13E2: Micro-Optics (11:00-12:15)
Chair: H. H. Hillmer

11:00-11:30  13E2-1 (Invited)  MEMS Devices and Technologies for Photonic Network  H. Toshiyoshi  
The Univ. of Tokyo, Japan .............................................................................................. 690

(1)The Univ. of Texas at Dallas, USA, (2)Southern Methodist Univ., USA .............................. 692

11:45-12:00  13E2-3  Beam Steering and Coupling in Tunable Hollow Waveguide with Narrow Air Core  A. Imamura, F. Koyama  
Tokyo Inst. of Technol., Japan .......................................................................................... 694

(1)Korea Advanced Inst. of Sci. and Technol., Korea, (2)Kyungpook Natl. Univ., Korea ...... 696

13E3: Waveguide Simulation (Tutorial + Oral Presentation) (14:00-15:30)
Chair: K. Kintaka

14:00-15:00  13E3-1 (Tutorial)  Finite-Difference-Time-Domain Calculation of Radiation Dynamics in Advanced Photonic Crystal Devices  M. J. Steel(1,2), D. P. Fussell(3), M. M. Dignam(3), C. M. de Sterke(2), R. C. McPhedran(2), F. Bordas(4), C. Seassal(4), A. Rahmani(4)  
(1)RSoft Design Group, Australia, (2)Univ. of Sydney, Australia, (3)Queen's Univ., Canada, (4)Inst. des Nanotechnologies de Lyon, France, (5)Macquarie Univ., Australia................. 698

xxxiii
Optimal Design and Analysis of Ultra Compact 1.3/1.55 μm Demultiplexer
Inha Univ., Korea ................................................................. 700

Design and Fabrication of Optical Multichannel Filter with Changeable Channel Spacing through Phase Sampling
L. Xia, P. Shum, C. T. Hiang
Nanyang Technological Univ., Singapore .................................. 702

Data Vortex Switch Network with Shared Cylindrical Traffic Control
Q. Yang
Harvey Mudd College, USA ..................................................... 704

On the Challenges of Transporting RF Signals over an FSO Channel
(1) Waseda Univ., Japan, (2) Osaka Univ., Japan ......................... 706

Proposal for Simple Bit Timing Synchronization Technique of Mapping Optical Burst Signals for Optical Burst Switching
A. Iwaki, M. Matsuura, N. Kishi, T. Miki
Univ. of Electro-Communs., Japan ........................................... 708

Millimeter-Wave True-Time Delay Measurement in WDM-Based Optically Controlled Array Antenna
W. Ohuchi(1), W. Chujo(2), Y. Fujino(2), Y. Koyama(1)
(1) Ibaraki Univ., Japan, (2) NICT, Japan ................................ 710

Two Novel Configurations and an Adaptive Round-Time MAC for Light-Trail IP-Centric Applications
N. C. Tran(1), E. Kim(1), K. Kim(2), J. Park(1)
(1) Korea Univ., Korea, (2) Electr. and Telecommuns. Res. Inst., Korea .................................................. 712

B. H. L. Lee(1,2), Z. Omar(1), S. M. Shukor(1), R. Mohamad(1), K. Dimyati(2)
(1) UPM-MTDC, Malaysia, (2) Univ. Malaya, Malaysia ........... 714

Employment of Optical Phase Conjugators in Transparent DWDM Single-Ring Network
N. Chuenprasertsuk, P. Kaewplung
Chulalongkorn Univ., Thailand .................................................. 716

Optimum Placement of Dispersion Compensating Unit for Transparent DWDM Ring Network
I. Mohara, P. Kaewplung
Chulalongkorn Univ., Thailand .............................................. 718

Reducing of Kerr Effect in Long-Haul Broadcast-and-Select Transparent DWDM Optical Network by Static Wavelength Assignment
P. Jarupoom, P. Kaewplung
Chulalongkorn Univ., Thailand ................................................ 720

Pseudo-Random Sequences for Modeling of Quaternary Modulation Formats
D. van den Borne(1), E. Gottwald(2), G. D. Khoe(1), H. D. Waardt(1)
(1) Eindhoven Univ. of Technol., The Netherlands, (2) Siemens Networks, Germany .............. 722

Chromatic Dispersion Compensation by Monitoring Bit-Pattern-Dependent Intersymbol Interference in NRZ Format
Y. Hayami, K. Iwashita
Kochi Univ. of Technol., Japan .............................................. 724

Proposal of Control Mechanism for Delay Sensitive Traffic in Simplified WFQ Scheduling
S. Taniguchi(1), R. Kawate(1), T. Yokotani(1), K. Motoshima(1), S. Yoshihara(2)
(1) Mitsubishi Electric, Japan, (2) NTT, Japan ....................... 726

Complexity Reduction of MSPE by an Iterative Algorithm for Optical D(Q)PSK
K. Piyawanno, F. N. Hauske, M. Kuschnerov, B. Lankl
Univ. of the Federal Armed Forces, Germany .......................... 728
13P-14  Heuristic Approximation of Transient Gain Dynamics due to Network Reconfiguration
S. Pachnicke(1), E. Gottwald(2), P. M. Krummrich(2), E. Voges(1)
(1) Univ. of Dortmund, Germany, (2) Siemens Networks, Germany
730

13P-15  Simple Fiber-Type Wavefront-Splitting Interferometer Using Hollow Optical Fiber
Y. Jung(1), J. K. Kim(1), B. H. Lee(1), K. Oh(2)
(1) GIST, Korea, (2) Yonsei Univ., Korea
732

13P-16  Common-Path Interferometer for Characterization of Fiber Bragg Gratings
L.-G. Sheu, H.-M. Fang
Vanung Univ., Taiwan
734

13P-17  Less Polarization Coupling and High Extinction Ratio of a Near-Elliptic Cladding Polarization-Maintaining Photonic Crystal Fibre
L. Wang, F. Yan, S. Jian
Beijing Jiaotong Univ., China
736

13P-18  Tunable Fiber Laser Using Brandband Fiber Mirror Integrated Tunable Fiber Bragg Gratings
(1) Natl. Taiwan Univ. of Sci. & Tech., Taiwan, (2) Natl. United Univ., Taiwan
738

13P-19  Gain-Clamped Fiber Amplifier with Reflective Configuration
H.-S. Huang(1), S. Hsu(2), T.-C. Liang(1)
(1) Natl. Kaohsiung First Univ. of Sci. and Technol., Taiwan, (2) Cheng Shiu Univ., Taiwan
740

13P-20  Dispersion Flattened Decagonal Photonic Crystal Fiber
S. M. A. Razzak, Y. Namihira, F. Begum, S. Kaijage, N. H. Hai, N. Zou
Univ. of the Ryukyus, Japan
742

13P-21  Nearly Chirp-Free and Pedestal-Free Pulse Compression
Q. Li(1), P. K. A. Wai(1), K. Nakkeeran(3), K. Senthilnathan(1)
(1) The Hong Kong Polytechnic Univ., China, (2) Univ. of Aberdeen, UK
744

13P-22  Optic Fiber Sensor Based on Etching Long Period Gratings of Photonic Crystal Fibers
(1) Feng-Chia Univ., Taiwan, (2) Air Force Acad., Taiwan, (3) Natl. Defense Univ., Taiwan
746

13P-23  Radiation Resonance of Transverse Magnetic Wave in Dielectric Micro-Spheres: An Asymptotic Solution
A. Rahman, S. Kumar
Polytechnic Univ., USA
748

13P-24  Fabrication and Characterization of GaP Two Dimensional Photonic Crystals for Terahertz-Wave Generation
(1) Tohoku Univ., Japan, (2) Semiconductor Res. Inst., Japan
750

13P-25  High Bandwidth Semiconductor Gain Material for Photonic Active Devices with a Stacked Quantum Dots Structure Grown by Strain-Compensation Technique
NICT, Japan
752

13P-26  Blue-Light Emission from Si Ion Implanted Fused-Silica Substrates Generated by High Temperature Annealing
K. Miura(1), T. Tanemura(1), O. Hanazumi(1), S. Yamamoto(2), K. Takano(2), M. Sugimoto(2), M. Yoshikawa(2)
(1) Gunma Univ., Japan, (2) Japan Atomic Energy Agency, Japan
754

13P-27  Eu3+ doped Polymer Waveguide Amplifiers
K. C. Tsang, E. Y. B. Pun
City Univ. of Hong Kong, China
756

13P-29  Optoelectronic Switch with S-Shaped Negative Differential Resistance
J.-H. Tsai(1), D.-F. Guo(2), M.-Y. Fu(2)
(1) Natl. Kaohsiung Normal Univ., Taiwan, (2) Air Force Acad., Taiwan
758

13P-30  Synthesis and Discussion of 3-hexylthiophene Related Rigid Copolymers for Optical Devices
P.-C. Huang(1), Y.-J. Lee(1), M.-Y. Chang(2), W.-Y. Huang(2), Y.-K. Han(1)
(1) Natl. Univ. of Applied Sci., Taiwan, (2) Natl. Sun Yat-Sen Univ., Taiwan
760
13P-31 Apodization Structure for Chirped Quasi-Phase-Matched Wavelength Converter
H. Okayama, H. Ono, Y. Okabe, Y. Ogawa
Oki Electric Ind., Japan .................................................................762

13P-32 Theoretical Investigation of Light-Guiding Structures of Surface Plasmon Resonance Waveguide Sensors
J. Shibayama, S. Takagi, T. Yamazaki, J. Yamauchi, H. Nakano
Hosei Univ., Japan .........................................................................764

13P-33 1-Input 3-Output Optical Interleave Filter with Group-Delay Dispersion Equalizer
S. Azam, T. Yasui, K. Jingui
Shimane Univ., Japan .....................................................................766

13P-34 Effect of Defect Pillars in T-Junction Photonic Crystal 1310/1550 nm Demultiplexer
A. J. M. Adnan(2), S. Shaari(1), I. A. Tengku(2), N. H. Z. Abedin(2)
(1)Univ. Kebangsaan Malaysia, Malaysia, (2)UPM-MTDC, Malaysia ......................768

13P-35 A Study of Three-Dimensional Fractal Cavity as a Photonic Resonator
H. Kakinishi, Y. Iida
Kansai Univ., Japan ........................................................................770

13P-36 On the Strain-Induced Wavelength Drifts of Thin DWDM Thin-Film Filters
A. K. Chu, M. C. Lee
Natl. Sun Yat-sen Univ., Taiwan ......................................................772

13P-37 Optimum Coupling Efficiency Condition of Quadruple-Series-Coupled Microring Resonator
T. Kato(1,2), Y. Kokubun(1)
(1)Yokohama Natl. Univ., Japan, (2)Tokyo Inst. of Technol., Japan ......................774
PLENARY TALKS

TUTORIAL

INVITED & CONTRIBUTED PAPERS
Fiber Technology in Early days and Future

Tatsuo Izawa
NTT Electronics Corp.
1-20-8, Nakacho, Musashino-shi, Tokyo, 180-0006, Japan
izawa@hqs.nel.co.jp

Abstract
In the 1970s, various issues of fiber technology including fabrication processes, materials, device structure and etc. were discussed to realize optical fiber communication systems. Now we can use the fiber technology both in trunk networks and access networks. However, some of these issues are still remained unsolved.

Recently, new technologies in materials and processing have been developed in other technical fields. For example, epitaxial growth of oxide crystal, transparent ceramics and ultra fine processing may be the key to develop new photonic devices.

In the talk, the fiber technology in early days and present will be reviewed and the possibility of the new technology will be mentioned.
Biography
Tatsuo Izawa

He joined Nippon Telegraph and Telephone Corporation (NTT) Research Laboratories and worked there for 28 years. He resigned from a board member and Vice President responsible for Research of NTT in 1998. During this time he made pioneering contribution in the area of low-loss optical fiber fabrication and planar lightwave circuits. He is a co-inventor of Vapor-phase Axial Deposition (VAD) process for making optical fiber performs. He also invented an ion electro-migration method for making planar optical waveguides. He performed early stage development of silica based planar lightwave circuits employing flame hydrolysis and CVD method.

He was appointed President and CEO of NTT Electronics Corporation in 1998. Since then he has focused his management activities on the development and manufacturing of photonic devices and electronic devices for Dense Wavelength Division Multiplexing networks including DFB lasers, Modulators, Detectors, AWGs, Optical Switches and GaAs ICs. He is now a Senior Advisor to the board of NTT Electronics Corp.

He served as a Vice President, in charge of members & regional activities and an elected member of Board of Governors, IEEE LEOS (1997-99). He also served as an elected member of Board of Governors, IEICE, the President of Electronics Society, IEICE (1998-99) and the President of IEICE(2006-07).

He is Life Fellow of IEEE, Fellow of Optical Society of America, Fellow of IEICE, and Foreign Associate of National Academy of Engineering. He received the John Tyndall Award, the David Sarnoff Award, Purple Ribbon Medal, Imperial Invention Award, and IEICE Achievement Award.
Fiber and lightwave technologies enjoyed an unprecedented growth over the 30-year period between the first and present IOOC conferences. The glass fiber has emerged to be the undisputed medium of transmission in long haul and metropolitan networks. Sophisticated undersea DWDM fiber systems criss-cross the bottom of the oceans, linking the continents together into a tight knit global information infrastructure. One of the remaining challenges in the completion of this infrastructure is the so-called “Last Mile”, “Last Kilometer”, or even “Last 100 meters” access to the core network, providing affordable, ubiquitous and unlimited broadband access to the end user. It was assumed early on that fiber is the medium of choice for these subscriber networks. One of the most ambitious goals in our industry was to run a fiber to every home. Fiber-to-the-Home (FTTH), however, suffered many false starts over the past two decades because of its high cost and the lack of a sound business model to justify the investment. The momentum of FTTH, or its variants known as FTTx, has picked up again over the past several years as a result of user demands and competitive pressure among broadband service providers. Japan is the leader in FTTH penetration, with an ambitious goal of providing FTTH services to 30 million homes by 2010. Leading service providers in North America and other regions have also begun their sizeable investments in FTTx, among them are the FiOS Project of Verizon Communications and Project Lightspeed of AT&T.

Despite the sizeable investment, the success of FTTH is by no means assured in many markets because of the wide range of competing technologies and economic constraints. In the United States, the dominant providers of broadband and video services are the cable companies which provide CATV services to 65 million subscribers and broadband Internet services to more than 30 million subscribers. This is followed by the DSL technologies of the phone companies, providing broadband services to over 25 million subscribers. The performance of DSL technologies continues to improve as a result of innovative signal processing and system design techniques. For example, VDSL2 is capable of supporting 80 Mb/s for single pair over short distances, and 160 Mb/s with bonded wires. Researchers and start-up companies have further expanded the techniques of dynamic spectrum management and vector processing, and proposed the revolution concept of “CuPON” as a copper alternative to the PON. CuPON promises to support a 100 Gb/s DSL network over the existing copper facility.

Equally impressive results are achieved in the broadband wireless access technologies. These include the worldwide effort in the definition of the Fourth Generation (4G) wireless networks which is a strong candidate to provide mobile access solutions to future subscribers. Mobile WiMAX based on IEEE 802.16e standards is often cited as a candidate for such 4G systems. Currently, Mobile WiMAX has an average throughput of 13 Mb/s downlink and 4 Mb/s uplink based on a 10 MHz carrier bandwidth. The performance of such systems continues to improve with advanced techniques in signal processing, multiple-input multiple-output (MIMO) and beam forming antenna technologies. The WiMAX Forum has grown to an industry group of 400-member organizations, constituting a large eco-system of service providers and components and equipment manufacturers.

This talk provides a survey of the growing family of access technologies. The technology and applications of FTTH and FTTx versus cable, xDSL, 802.11 and 802.16 families are compared. The possibility for these communications technologies moving towards one single converged network supporting triple- or quad-play will also be examined.
Biography

Nim K. Cheung

Dr. Nim K. Cheung is the 18th President of the IEEE Communications Society, a global professional organization with 45,000 members in 176 chapters around the world. He is a Telcordia Fellow and an Executive Consultant in Telcordia Technologies (formerly Bellcore). From 1999 to 2002, he was Vice President of Applied Research Government Program in Telcordia Technologies, managing a Government-funded research program that continues to grow at double-digit rate per year. He also serves as consulting professor of electrical engineering at Stanford University starting from 2004.

Dr. Cheung received his B.Sc. degree from University of Hong Kong in 1970 and Ph.D. degree from California Institute of Technology in 1976. After graduation, he joined Bell Laboratories to conduct pioneering research in early single-mode fiber optic systems. In 1984 he became District Manager of Advanced Lightwave Technology in Bellcore, where he created three world-class research programs in high-speed, coherent, and subcarrier-multiplexed lightwave systems.

In 1989, he was on leave as a visiting scientist at the School of Computer Science in Carnegie Mellon University. He helped create the Gigabit Nectar Testbed in Pittsburgh, one of the early gigabit testbeds sponsored by DARPA and NSF. He returned to Bellcore as an Executive Director in the Applied Research Area, where he managed a wide range of research programs in high-speed networking, terabit switching and network management systems. He was one of the originators of the Advanced Technology Demonstration Network Testbed (ATDNet), the first 2.5 Gb/s ATM testbed in Washington DC sponsored by six government agencies. ATDNet later evolved into MONET, the first large scale Dense Wavelength Division Multiplexing testbed. He chaired the SONET OC-192 (10 Gb/s) and ATM Self-healing Ring Consortium, a key Technology Reinvestment Projects (TRP) initiated by President Clinton. Dr. Cheung was the principal investigator of SuperNet Broadband Local Trunking, a Next Generation Internet (NGI) project sponsored by DARPA.

Nim Cheung was a member of the Executive Science and Technology Council of Science Applications International Corporation (SAIC), and has served on the EE advisory boards of Columbia University, Polytechnic University of New York, and New Jersey Institute of Technology. He is a Fellow of IEEE, and has served on different leadership positions of IEEE and IEEE Communications Society. He has also served on many editorial positions and organized numerous conferences and workshops.
Abstract
This paper first gives an overview of the opening and reform process of China’s telecommunications industry, especially regarding to public policies and regulations. It then describes the development of this sector in various aspects over the past decade, with China now having the largest telecommunications network and largest subscriber population in the world. The paper also reviews technology development and illustrates technologies currently adopted at transmission networks, access networks, mobile networks, and etc. In order to sustain the development and realize the potential of the networks, much research work has been carried out in this field, and this paper summarizes some on-going projects. China also stresses the importance of professional training and degree education in telecommunication so as to develop enough human resources for the coming of the information age. The paper ends with an introduction on key universities and major training/education activities related to the telecommunications sector.
Biography

LIN Jintong

Professor LIN Jintong, is currently President of Beijing University of Posts and Telecommunications (BUPT), China. He graduated from Peking University in physics, pursued his further studies in Electronic Engineering and earned MSc from BUPT and PhD from University of Southampton, U.K.

He worked at BUPT, University of Southampton, and King’s College of London University, in the field of high speed optical communications systems and broadband optical fiber access networks. He is a Fellow of The Institute of Engineering and Technology (IET, the former IEE), and has published over 300 academic papers and authored two books in China and USA. He has been invited to deliver speeches and presentations at various international conferences world wide. He was awarded honorary Doctor of Science by University of Southampton, UK in 2005.

Prof. LIN is vice Chairman of China Institute of Communications (CIC), and serves in many other professional organizations. In recent years, he served as Conference Chair or TPC chair in many international conferences, such as ICCT (International Conference on Communications Technology), APCC (Asia-Pacific Conference on Communications), OECC (Optic-Electronics & Communication Conference) and ITC (International Teletraffic Conference).
Flexible access technologies to upgrade PONs

Makoto Tsubokawa
NTT Access Network Service Systems Laboratories
1-6 NTT Bldg. Nakase Mihama-ku Chiba-shi, 261-0023 Japan
Tel: +81-43-211-2439, Fax: +81-43-211-8883, Email: tsubokawa.makoto@ansl.ntt.co.jp

Abstract: Access technologies designed to provide the next generation services and their migration scenarios are described. There is a strong need for both a flexible architecture that can overlay or coexist with a large number of FTTH facilities and simple high-end technologies.

Issues in system migration: Broadband services have been steadily spreading in Japan, and the target is now set at 30 million FTTH subscribers by 2010. Most FTTH facilities are constructed using GE-PON, and this will continue to exist as a network platform for a long time. PON is well known to be the most economical access architecture for the first stage of FTTH penetration; however, it will impose several constraints due to its unique architecture in the coming upgrade stage. For example, its lack of service flexibility, poor loss budget, very high-speed burst TDMA, and complex operation, administration, and maintenance (OAM) function should be carefully considered and overcome when developing new systems.

Upgrade scenarios and various technologies: The current GE-PON configuration is shown in Fig. 1. Thirty-two ONUs are slaved to one OLT through two-step optical power splitters. The system capacity of 1 Gbps is shared among all the ONUs. By adding a V-OLT and an OLT through an SCM-WDM at a wavelength of 1.55 μm, an optional RF-video distribution service can be provided to every user.

The upgrade scenario is shown in Fig. 2, where we simply assume three main branches. The first branch to appear indicates a partial upgrade when some users belonging to one PON require new services. The power splitter based WDM-PON is most suitable here as shown in Fig. 2(a). Each new ONU and OLT can exchange signals through an allotted wavelength filter whose wavelength can be chosen from a future band of the ITU-T G983.3 grid. With this additional WDM method, we must smoothly install new ONUs on customer premises to maintain good OAM. We have been developing a PnP mechanism whereby an OLT automatically configures the IDs and wavelengths of newly connected ONUs, and colorless ONU technology [1]. In addition to the above WDM systems, a simplex downstream system of ~10 Gbps, although it might be transitional, is also very attractive since broadband distribution services such as IP-video can be easily achieved with almost the same architecture as the RF-video distribution mentioned above. (Fig. 2(b)).

The second branch shows an extension of the area to be covered. This is typically required in rural areas to overcome the digital divide. Bidirectional optical amplifiers used as repeaters can extend the transmission distance and increase the number of branches supported by the optical power splitters. Last year BT, Mitsubishi Electric and NTT proposed long-reach PON technologies that represent enhanced versions of GE-PON or G-PON. BT achieved 60 km transmission distances with a 128-way splitter [2]. SOA-based repeaters were employed for both upstream 1.49 μm and downstream 1.51 μm wavelengths. NTT also achieved a 40 km transmission with a 128-way splitter by using T DFA for 1.31 μm and PDFA for 1.49 μm wavelengths. Furthermore, a 60 km transmission with 256 branches (64-way and 4-way splitters) can be realized by two-step amplification [3]. The wavelength arrangement in our experiment is equivalent to that in current GE-PON. PONs with more than 32 branches are also expected to be used for large housing complexes in place of VDSL. Optical devices, in particular optical amplifiers, are still too expensive for access network use, and the cost efficiency should be improved in future.

The last scenario means replacement or overlay by new access systems in the next generation. This is similar to the case of greenfield installations. The TDM-based PON will become mainstream with bit rates of around 10 Gbps and TDM, WDM and OCDM hybrid methods will become candidates for applications beyond a few tens of Gbps. Although the simple 10 Gbps has already been completed, 10G-PON cannot be fully achieved yet because of difficulties in obtaining burst signal processing devices with higher speeds and lower prices. This year, Mitsubishi Electric First announced that they had developed a 10 Gbps XFP transceiver module for PON [4]. NTT has also reported the burst-mode 3R receiver operating at 10.3125 Gbps using SiGe-BICMOS monolithic ICs, which achieved the highest sensitivity of -18 dBm, the widest dynamic-range of 16.5 dBm and instantaneous responsivity of 75 ns [5]. The competition is sure to be accelerated further in one or two years from now. From the viewpoint of upgradeability, a double/multi-rate 10 Gbps system accommodating both 1 and 10 Gbps services will be required in the future.

A lot of research (hybrid multiplexing, multilevel transmission, etc) has been reported with the goal of attaining much larger capacity for the long term. NTT proposed a high-density FDM (equivalent to WDM) system with coherent detection to expand capacity to
one digit or more. Customers can obtain the desired video signal from among a few hundred channels by optical frequency tuning. The application of coherent detection to conventional DWDM is expected to yield higher sensitivity and wavelength/efficiency. We have reported a 1 Gbps × 100 ch FDM transmission experiment with a simple ONU in which both the transmitter and local light source consist of a DFB-LD to lower the price [6].

CDMA technique is also now receiving a lot of attention because it can be flexibly applied to networks of any topology and offers high frequency/wavelength efficiency. Several basic experiments have been reported. OKI Electric has reported a system capacity of 2 Gbps (62.5 Mb/s × 32 users) with 100 km-distance by coding in electronic domain [7]. As for Optical CDMA, a system of 320 Gbps (10 Gbps × 32 users) with a sensitivity of 32 dBm has been realized by the University of California [8]. NICT and Osaka University have also reported a system capacity of 300 Gbps (3-WDM × 10-OCMDM users × 10.71 Gbps) with a sensitivity of 30 dBm [9]. These OCMDM experiments need highly accurate and large scale devices to support huge system capacities, therefore it may be important to ascertain OCMDM’s applicability to actual access networks in the future.

**Summary:** I have described system migration scenarios and several access technologies. Smooth migration from legacy PONs is a key factor for the next generation access systems, and this makes it important to consider various approaches including partial and/or complete replacement of systems to allow services to evolve.

**References:**

Fig. 1 Gigabit-Ethernet Passive Optical Network (GE-PON)

**Fig. 2 PON Upgrade Scenarios:**

(a) Partial Upgrade

(b) Replacement or Greenfield

(c) Area Extension

For rural customers or housing complexes.
Improvement of TDM-PON Upstream Bandwidth Utilization Adopting Two Wavelengths

Junhoi Chung, Jeauk Park, Youngil Park
School of Electrical Engineering, Kookmin University, Seoul, 136-702, Korea  email: ypark@kookmin.ac.kr
Byoung-whi Kim
Electronics and Telecommunications Research Institute, Daejeon, 305-350, Korea

Abstract
Two types of wavelengths are used for EPON ONU to increase the upstream bandwidth utilization. Bandwidth loss from Guard time and MPCP protocol can be removed by overlapping two consecutive packets from ONUs equipped with different wavelengths. Performance estimation and testbed results are presented.

1 Introduction
EPON is now deployed in several nations for commercial purpose, which will initiate broadband convergence network [1]. Current EPON system has 1 Gbps bandwidth for both up and down transmissions [2]. However, the current systems can’t utilize the full bandwidth, especially for the upstream, due to guard time and MPCP protocol. The bandwidth loss reaches up to 40% in the worst case. GPON, which is very promising especially in Europe and North America, requires very tight physical specification in its implementation. It specifies LD on/off time of 25 ns and while this parameter is 60~100 ns for EPON, which hinders cost reduction [3].

In this study, we propose to use two types of wavelengths at the ONU site to solve the above problems. About half of the ONUs are equipped with wavelength #1 while the other half with wavelength #2. If these two different wavelengths are sent in turn toward OLT, they don’t interfere with each other and can obtain enough guard time while maximizing the bandwidth utilization.

2 Principles of the Proposed Scheme
Fig. 1 illustrates the proposed two upstream wavelength PON (2W-PON) architecture. Each ONU is equipped with one of these wavelengths so that the two wavelengths are evenly used among ONUs. MPCP protocol orders different wavelengths are sent in turn from ONU site with part of the packets from consecutive ONUs being overlapped in time. Each wavelength is detected by two receivers at the OLT and then packets are realigned by adjustment logic. Details of the adjustment logic are shown in Fig. 2. Preambles of each packet are delimited. Then, the payload parts are buffered and multiplexed. Although this architecture needs additional receiver module at the OLT, it increases the upstream bandwidth efficiency by up to 40%. Furthermore, it lessens the requirements of OLT receiver and ONU transmitter, which can lower the total cost.

3 Performance analysis
Performance is calculated for the proposed 2W-PON and existing 1W-PON system. Considering all sources of bandwidth loss with 1W-PON, including LD on/off time, receiver settling time, preamble, MPCP operation time and so forth, 35 μsec interval is inserted between consecutive ONU’s windows. 10 Mbytes buffer, 16 ONUs with 8 ONUs have the same wavelength, self-similar traffic and 2 msec cycle time are assumed for the simulation. Fig. 3 shows that the throughput of the 1W-PON becomes 0.7, that is, 700 Mbps. For 2W-PON system, however, this time gap between time slots can be removed by overlapping windows, and the result shows that the throughput approaches 0.97. The reason that the
value is a bit less than 1.0 comes from the truncated window size by the DBA algorithm. Even this small loss can be compensated by overlapping larger part considering the vacancy occurring from the truncation. Fig. 4 compares delays for two PON systems. While 1W-PON shows sharp increase at the load of 0.35, the limited scheme of 2W-PON doesn’t have this slope until load 0.6.

4 Testbed Operation Results

Fig. 5 shows transmission test using a testbed. ONU blocks and an OLT adjustment part are emulated on FPGA chips. Preambles with one wavelength are sent before the transmission by the other wavelength is not finished yet. The OLT output shows a realigned packet stream with preamble parts stripped off.

**Conclusion:** 2W-PON is suggested to increase the EPON upstream bandwidth utilization. It is expected to lower the tight physical constraints of GPON. Implementation architecture and the test results are provided, which shows its validity.

5 References

Limited Bandwidth Allocation with Sliced Credit in Gigabit Passive Optical Network

Ying Shi, Hideya Yoshiuchi
IP Network System Laboratory, Hitachi (China) Research & Development
301, Tower C Raycom infotech Park
Kexueyuan Nanlu, Beijing 100080, P.R. China
Tel: 8610-8286-2918; Fax: 8610-8286-2919; Email: yshi@hitachi.cn

Abstract: Limited bandwidth allocation with sliced credit (LBA-SC) is proposed and used in Gigabit passive optical network (GPON). Simulation results show that packet delay of high priority service can be dramatically reduced by using predictive SC grants.

Introduction

Most current dynamic bandwidth allocation (DBA) algorithms concentrate on Ethernet passive optical network (EPON), such as IPACT [1], BGP [2], HSSR [3], LSTP [4] and D-CRED [5]. As EPON works in asynchronous way, optical line terminal (OLT) can send bandwidth allocation grant in downstream at any time, and require optical network unit (ONU) to report bandwidth request in upstream during the allocated time slot. However, in Gigabit PON (GPON), frames are aligned periodically in 125μs-length in both downstream and upstream directions. The bandwidth allocation grants and bandwidth requests are transmitted as frame headers instead of specific control frames in EPON. As a result, OLT and ONU can exchange bandwidth information with each other more frequently. Additionally, the upstream bandwidth must be allocated based on the 125μs period. Since the typical cycle time of polling DBA is set as 1 or 2 ms, due to the limitation of hardware capability, new DBA algorithm for GPON is necessary because allocations based on the relatively long cycle time may be cut by the 125μs frame length, leading to bandwidth waste and low efficient.

In this paper, limited bandwidth allocation with sliced credit (LBA-SC) is proposed for GPON system. Services from ONU side are sorted as high priority (HP) and low priority (LP) service, which are served by OLT in different way. Continuous time slots below a length limitation are allocated to both services and additional sliced credit (SC) slots are allocated to only HP, so that the efficiency of LP service is maintained and the packet delay of HP service is dramatically reduced.

Bandwidth Map and DBRu in LBA-SC

In GPON, bandwidth allocations are enclosed in upstream bandwidth map (US BW map) [6]. As shown in Fig. 1, each allocation includes allocation identifier (Alloc-ID), start and stop time, where Alloc-ID indicates a communication flow, e.g. a subscriber or an ONU. Since the minimum processing time of ONU is 16.384 μs [7] and within one frame, the allocation in downstream frame (DSFrame) i is acquisitely used in upstream frame (USFrame) i+1.

The grants for each USFrame include two parts: sliced credits (SC) and limited allocations (LA). SC part carries short pieces of payload coming from all users, with dynamic bandwidth report upstream (DBRu), and LA part carries long pieces of payload from partial users. So in each USFrame, OLT receives bandwidth reports from all users, meanwhile, in each DSFrame, all users receive SC grants and some users receive LA grants. After received grants, users transmit data accordingly so that mentioned operations are repeated frame by frame.

![Fig.1. Bandwidth allocation by SC and LA grants](image1)

Fig.1. Services with different priority in ONU

Limited Bandwidth Allocation with Sliced Credit

As [6] defines, ONU can choose to report total queue length in 1 byte or report separate queue length in 2 or 4 bytes message enclosed in DBRus. Therefore, from the simple 1-byte DBRu OLT can get only total bandwidth request, and from the complicated DBRu OLT can get separate bandwidth requests about HP and LP service. In LBA-SC, these two kinds of DBRus are processed differently when OLT assigns SC grants.

![Fig.2. Services with different priority in ONU](image2)

Fig.2. Services with different priority in ONU

Limited Bandwidth Allocation with Sliced Credit

LBA-SC assigns upstream bandwidth as two parts, SC and LA, as mentioned in the last section. SC grant is distributed to all users in every frame for the data coming after ONU transmitting the last DBRu. As shown in Fig.3, DBRu i-2 and DBRu i-1 are utilized to predict the data coming in Frame i-1.
The difference of HP bandwidth requests in these two DBRus is directly used as SC i in the case that DBRus includes separate HP/LP reports, or the difference of total bandwidth requests in these two DBRus is used as SC i after multiplying with a HR coefficient in the case that DBRus only includes the simplest report. In order to avoid sending too short SC, which will be omitted by ONU, or too long SC, which leads to possible bandwidth waste, the calculated SC grants are checked and limited in a range of 64–512 bytes, since 19440 bytes is the total length of a 125μs frame with 1.24Gbit/s bit-rate. After receiving SC i, ONU transmits HP data with DBRus in Frame i+1.

Fig. 3. Calculation and transmitting of sliced credit

Unlike sending SC grant, OLT sends LA grants to users in flexible polling cycle. All ONUs send DBRus in each frame, however, some of them are omitted in LA assignment. The length of LA grant is set as total bandwidth request in the last DBRus, and is limited by the minimum value of 64. When the surplus bandwidth of current frame is less than 64, OLT ignores it. Moreover, when the surplus bandwidth of current frame is more than 64 and less than the bandwidth request of current ONU, OLT assigns the surplus bandwidth to this user and keeps serving this user in the next frame. As the result, one user can obtain consecutive LA grants, however, up to 4 frames. In the example shown in Fig. 4, OLT begins from ONU1 in Frame i-1 by using DBRus i-2, and goes on with ONU2 when there is bandwidth left. After serving ONU2, there is no enough surplus bandwidth so OLT waits and assigns LA to ONU3 in Frame i by using DBRus i-1.

Fig. 4. Calculation and transmitting of limited allocation

By using mentioned SC and LA together, the packet delay of HP service can be dramatically reduced because of predictive SC grants sending in 125μs period, and the efficiency of LP service is maintained by LA grants. Although in practical system grants may be decided based on the DBRus received several frames ago because of the response time of OLT, SC grant still takes effect, however, when the variation period of the traffic is more than the OLT response time.

Simulation Result and Discussion

In the simulation, there are 16 ONUs connected to 1 OLT and the total upstream bandwidth is set as 1.24Gbit/s. The traffic for each ONU includes 20% HP packets and 80% LP packets, both in length of 512 bytes, chosen from the popular packet length of 64 bytes, 256 bytes, 512 bytes, and 1518 bytes. The OLT response time is set as one frame, as well as the ONU response time.

Fig. 5 shows the average and maximum packet delay, where the delay of LP service increases with total load beyond 400Mbit/s, and that of HP service keeps low level before the load reaches 800Mbit/s. The average delay of HP service is less than 0.2 ms with load from 0 to 800Mbit/s, and the variation of its maximum delay is below 1ms in the same range, indicating low jitter for real-time services. So LBA-SC can provide HP services with good performance.

Conclusions

In this paper, LBA-SC is proposed for dynamic bandwidth allocation in GPON system. The upstream services are sorted as HP and LP service, and are served in different way, where continuous time slots below the limitation are allocated to both and additional sliced credit time slots are allocated to HP. Simulation results show that the packet delay of HP service can be dramatically reduced by using predictive SC grants. The average delay of HP service is less than 0.2 ms when the total load increases from 0 to 800Mbit/s, and the variation of the maximum delay is below 1ms in the same range. At the same time, packet delay of LP service increases with total load more than 400Mbit/s.

References

Impact of Polarization Mode Dispersion on Analog Video Transmission over PON system

Youichi Akasaka, Paparao Palacharla, Martin Bouda, Takao Naito
Fujitsu Laboratories of America, Inc., 2801 Telecom Parkway, Richardson, TX 75082, USA
Tel. +1-972-479-2661, Fax +1-972-479-4482, youichi.akasaka@us.fujitsu.com

Abstract: Signal degradation of analog video optical transmission over fiber to the home system, caused by polarization mode dispersion, has been examined. It was revealed that dynamic changes of PMD degrade signal quality.

1. Introduction
Signal distortion in analog lightwave systems caused by polarization mode dispersion (PMD) has been investigated [1]. This degradation could be significantly reduced by use of low chirp lasers in optical transmitter, however, it is still considerable even with low PMD value of single digit [2,3]. Especially, recent widespread of Passive Optical Network (PON) system over the world rose issues related to RF video overlay seriously because its optical transmission distance (~20km) is relatively longer than those of common distance of analog optical transmission (a few kilometers) [4].

PMD is an effect to separate fundamental spatial mode into two orthogonal polarization modes, caused by birefringence, which is remained permanently in fiber through manufacturing process or stressed by external condition changes such as temperature changes. As environmental conditions vary constantly, PMD effects show stochastic behavior. It is well know that PMD values distribute along Maxwellian distribution and it has wavelength and time dependence. As a result, temporal changes of PMD might degrade signal quality of analog video as well as static PMD effects.

PON systems for Fiber to the home (FTTH) usually use OLT (Optical line terminal) and ONTs (Optical network terminal) to communicate a service provider with subscribers. To reduce service cost, information signal is equally divided by remote node between OLT and ONTs. The fiber feeding scheme to each subscribers depends on each individual situation but mostly the feeding fibers come under the influence of environmental changes such as temperature, rain, and winds. (Figure1)

Figure 1: Schematics of FTTH and environmental effects

Usually, video signal quality, degraded by impairments, is evaluated with CNR (Carrier to signal ratio), CSO (Composite second order beat), and/or CTB (Composite triple beat) [2]. And it states that the video signal less than certain amount of those numbers are bad quality. However, it is unclear how bad it is. For example, degradation over entire screen could not be recognized when the degradation amount is small. On the other hand, it is obviously felt uncomfortable if the screen has white flickering horizontal lines even though it is very minor. Thus, in this study, we evaluated video signal quality against temporal varying PMD by using subjective observation.

2. Experiment
To verify PMD effects against other fiber impairments, we used two experimental setups. Figure 2 (1) shows simple setup without PON configuration as well as 1490nm data signals. The second configuration (2) has a HGPON [5] remote node and data signals. Analog video signal operates at 1555nm and is amplified by an EDFA to 20dBm. The RF video input to the transmitter is a video modulated carrier of channel 3 @ 61.25MHz.
Signal input powers in case of configuration (1) and (2) to Rx were –6.5dBm, and –2.7dBm, respectively.

Only first order PMD (differential group delay: DGD) was introduced with General Photonics PMDE-301. Input state of polarization was fixed 45 degree (worst case) and changes instantaneous DGD value along Maxwellian distribution of given PMD. The time of each PMD states were varied from 10ms to 500ms.

3. Results and Discussion

Photo 1 shows the degradation (flickering white horizontal lines) caused by temporal changes of PMD. These lines came up when a few pico-second PMD was introduced. (The bad quality of the photo itself is not related PMD.) In case of very high speed changing each DGD states (10ms), white lines became annoying even with low PMD value of 2ps.

![Photo 1: Video screen with white horizontal lines caused by PMD](image)

Table 1 summarized observations. Frequency of white lines strongly depends on PMD value, and time of each DGD states. As aerial fibers, which are usually used for FTTH, tend to have shorter time of DGD states, those systems should consider total PMD carefully. Results after fiber transmission should be same, as those without transmission, however it looks slightly better because of other degradation made them not relievable.

<table>
<thead>
<tr>
<th>Test Setup</th>
<th>(1)</th>
<th>(1)</th>
<th>(1)</th>
<th>(2)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Time period</td>
<td>10ms</td>
<td>50ms</td>
<td>100ms</td>
<td>500ms</td>
</tr>
<tr>
<td>0ps PMD</td>
<td>v</td>
<td>v</td>
<td>v</td>
<td>v</td>
</tr>
<tr>
<td>2ps PMD</td>
<td>v</td>
<td>v</td>
<td>v</td>
<td>v</td>
</tr>
<tr>
<td>4ps PMD</td>
<td>v</td>
<td>v</td>
<td>v</td>
<td>v</td>
</tr>
<tr>
<td>8ps PMD</td>
<td>v</td>
<td>v</td>
<td>v</td>
<td>v</td>
</tr>
<tr>
<td>12ps PMD</td>
<td>v</td>
<td>v</td>
<td>v</td>
<td>v</td>
</tr>
<tr>
<td>16ps PMD</td>
<td>v</td>
<td>v</td>
<td>v</td>
<td>v</td>
</tr>
<tr>
<td>20ps PMD</td>
<td>v</td>
<td>v</td>
<td>v</td>
<td>v</td>
</tr>
</tbody>
</table>

Table 1: Frequency of flickering white horizontal lines on the video screen. ( v represents number of the lines in a screen)

Note: Although 2ps PMD is generated easily on a PON configuration (e.g. 20km fiber, EDFA, WDM filter, Splitter), rotating signal polarization by external changes can help to reduce degradation by PMD significantly against this experiment’s worst-case setup.

4. Conclusion

We investigated PMD effects on analog video optical transmission over PON system. Only a few PMD value could generate flickering white horizontal lines on video screens if the DGD value frequently changes with the worst input polarization state.

References
Properties of Optical Modulator Cascade for Multi-Channel Transmission Systems

Koji Kikushima and Toshihito Fujiwara

Access Network Service Systems Laboratories, NTT, 1-6 Nakase, Mihama-ku, Chiba-shi, Chiba, 261-0023 Japan
e-mail: kikushima.koji@ansl.ntt.co.jp

Abstract: This paper reports the properties of an optical modulation scheme that uses cascaded external modulators. Calculations show that modulation properties are improved by converting input signal frequency and format prior to optical modulation.

I. INTRODUCTION

We recently reported that cascaded modulation offers better modulation properties than the conventional single modulation approach for multi-channel transmission systems [1]. In the previous paper, we applied the FM and RF conversion to the multi-channel signals prior to modulation.

This previous paper, however, did not fully elucidate the effect of FM and RF conversion on the performance of the cascaded modulation scheme. We rectify this omission by calculations that compare the modulation properties with and without FM and RF conversion.

II. CALCULATIONS

The calculations considered the following setups. Fig.1 shows the schematic diagram of the optical modulator and modulation signal frequencies. Fig.1 (a) and (b) are for the test without and with FM and RF conversion, respectively.

FM and RF conversion altered the frequency plan and signal format of the input multi-channel signals as follows.

Input signals, 40 AM and 30 64-QAM CATV (from 91 to 770MHz) signal carriers, were converted into a single wideband FM signal (from 0 to 6GHz) by the FM converter [2]-[3]. This FM converted signal was input to the first EA modulator with optical modulation depth $m_1$ of 90%.

Pre-emphasis was applied to signal #1 in FM conversion in compliance with recommendation ITU-T J.185 [4] to compensate the triangle noise. Frequency deviation per carrier was set according to Eq. (1).

$$f_{\text{deviation}} = 70 \times \frac{12.9 (f - 47)}{16340} \times \frac{V_{\text{in}} - 85}{20} \quad (1)$$

Here, $f_{\text{deviation}}$ is the frequency deviation per carrier [MHz], $f$ is the carrier frequency [MHz], and $V_{\text{in}}$ is the signal voltage level input to the FM converter [dBmV]. In the experiment, the signal voltage level input to the FM converter was 85dBmV/carrier for AM signal and 75dBmV/carrier for 64-QAM signal. Frequency deviation was 126.7MHz$_{0-p}$/carrier for the highest frequency AM-channel carrier (373.25MHz) and 81.9MHz$_{0-p}$/carrier for the highest frequency 64-QAM-channel carrier (770MHz).

When FM conversion was not applied, optical modulation depth was set at 3.8%/carrier and 1.2%/carrier for AM and 64-QAM signals, respectively, and so the total optical modulation depth $m_1$ was set at 25%.

Input signals, 8 broadcast satellite (BS) and 12 communication satellite (CS) intermediate frequency (IF) signal carriers (from 1030 to 2070MHz), were frequency converted to BS and CS radio frequency (RF) signal carriers (from 11.7 to 12.7 GHz) by the up-converter. This up-converted signal was input to the second EA modulator.

For both cases, with and without RF conversion, the optical modulation depth was set at 7.3%/carrier and 4.1%/carrier for BS and CS signal, respectively, so the total optical modulation depth $m_1$ was set at 25%.

Fig.2 shows the calculated electrical power level of output signal carrier, and second-order intermodulation distortion, for the test without FM and RF conversion.

(a) CATV signal carriers and BS/CS-IF signal carriers are FM converted, respectively.

(b) Prior to optical modulation, input CATV signal carriers and BS/CS-IF signal carriers are FM converted and RF converted, respectively.

Fig.1 Schematic diagram of the optical modulator and modulation signal frequencies. LD is laser diode.
Fig. 2(a) is for the electrical power level of carrier and intermodulation distortion. Distortion is seen in the AM, 64-QAM, BS, and CS channels. Fig. 2(b) and 2(c) plot the ratio of carrier to total intermodulation distortion in a channel. In this measurement, the noise bandwidth was 4.2MHz for AM, 5.3MHz for 64-QAM, 28.9MHz for BS and CS signal. The calculated results indicate that unacceptable levels of distortion are created.

Fig. 3 shows the calculated results of the electrical power level of output signal carrier, and second-order intermodulation distortion level, for the test with FM and RF conversion. Fig. 3(a) shows the electrical power levels of FM converted signal, BS and CS signals, and intermodulation distortion. The FM converted signal spectrum has a wideband distribution. This makes distortion also wideband. The distortion, however, is quite low in the FM converted signal band and the BS/CS signal band.

Fig. 3(b) shows the ratio of carrier to total intermodulation distortion in a channel. By comparing the calculated values of carrier to distortion ratio shown in Fig. 2(c) and Fig. 3(b), we can see that the ratio in

Fig. 3(b) is larger than that in Fig. 2(c). The difference is 12dB at the worst values.

The above calculation indicates that the second-order intermodulation distortion created by the cascaded modulation can be reduced by FM and RF conversion.

III. CONCLUSIONS

FM and RF conversion is useful for reducing the second-order intermodulation distortion otherwise present in the cascaded modulation of multi-channel signals.

REFERENCES

A Review on WDM PON Technologies (Invited)

Y. C. Chung
Korea Advanced Institute of Science and Technology
Department of Electrical Engineering
373-1 Guseong-dong, Yuseong-gu, Daejeon 305-701, Korea
(phone) +82-42-869-3456, (fax) +82-42-869-3410, (email) ychung@ee.kaist.ac.kr

Abstract: We review the outstanding technical issues for the development of practical WDM PON and report the relevant technical progresses achieved at KAIST.

1. Introduction
Recently, FTTH has finally emerged from the R&D stage and became a commercial reality. For example, in 2006, the total number of FTTH subscribers exceeded 7 million in Japan and 1 million in US. It appears that this trend is rapidly spreading all over the world. At present, most of the FTTH deployments utilize TDM PON (such as EPON and GPON). However, WDM PON is beginning to attract significant attention, as Korea started a large-scale field trial in 2005. In fact, WDM PON has long been considered as an ultimate solution for the broadband access networks due to its large capacity, easy management, upgradeability, and network security. In this paper, we review the current technical issues involved in the development of practical WDM PON and report the relevant progresses achieved at KAIST. The topics covered in this paper include the low-cost WDM light sources, efficient network architectures, techniques for delivering broadcast services, wavelength-tracking techniques, fault monitoring techniques, and survivable network architectures.

2. Cost-effective WDM Light Sources
Although WDM PON has many advantages, it has been considered to be prohibitively expensive for the practical deployment. This is mainly due to the extra costs involved in the installation and maintenance of the wavelength-specific lasers required in WDM PON. To overcome this problem, there have been many efforts to develop the “colorless” light sources, so that we could implement WDM PON by utilizing the identical light sources at every subscriber’s site. In particular, the spectrum-sliced incoherent light sources (such as LED’s or ASE sources), ASE-injected FP lasers, and re-modulation scheme (using either external modulators or RSOA’s) draw significant attention [1]-[4]. We will review the characteristics of these light sources and discuss their potential problems and possible solutions.

3. Efficient Network Architectures
In general, WDM PON is implemented in double-star architecture, although its details can be slightly different from each other depending on the type of light sources, network size, required services, and necessity of protection, etc. The following is a list of exemplary architectures proposed for the implementation of WDM PON.

- WDM PON based on “broadcast and select” architecture [5]
- WDM PON based on DFB lasers and AWG’s [6]
- WDM PON based on spectrum-slicing [7]
- WDM PON based on re-modulation scheme [8]
- WDM PON based on ASE-injected FP lasers [9]
- Composite WDM PON [10]
- Hybrid WDM-TDM PON [12]
- Hybrid WDM-SCM PON [13]
- Hybrid WDM-CDM PON [14]
- Long-reach WDM PON [15]

In this paper, we do not intend to compare all the technical details of these network architectures. Instead, we will focus on the typical WDM PON architecture and discuss the possibility of reducing the cost associated with outside plant. For this objective, we will discuss only the issues involved in bidirectional transmission and the techniques for minimizing the cost of RN [16].

4. Delivery of Broadcast Services
WDM PON utilizes an AWG in the RN for the virtual point-to-point connectivity. As result, WDM PON is not well suited for the delivery of broadcast services. Several techniques have been proposed to solve this problem by using the WDM overlay on PON or broadband LED for the broadcast signals [17]-[18]. However, the first technique requires a large number of WDM filters and couplers, while the second one should use not only a pair of fibers but also the frequency up/down-conversion of video signals to overcome the limited modulation bandwidth of LED. At KAIST, we developed a new technique for the delivery of broadcast video signals by using the cyclic property of AWG [19]. For demonstration, we implemented a bidirectional WDM PON capable of transmitting both digital broadcast video signals and WDM data channels. This network utilized WDM lasers operating in the 1.55-µm region for the downstream data (up to 2.5 Gb/s), LED’s operating in the 1.3-µm region for the upstream data (155 Mb/s), and a laser operating at 1.53-µm for the broadcast of more than 70 digital video channels.

5. Wavelength Tracking Technique
In WDM PON, the spectral characteristics of the AWG located at the un-powered RN are vulnerable to the ambient temperature changes of the outside plant, which could be
larger than 100 °C. This temperature changes would cause the spectral misalignment between the WDM sources and the AWG at the RN (or between the AWG’s at the CO and RN) and result in the significant power loss and crosstalk. For example, it has been reported that the temperature-induced wavelength drift of AWG should be <6 GHz to maintain the corresponding power penalty within 1 dB (in a 155-Mb/s system using ASE-injected FP lasers) [20]. At present, it appears that a commercial athermal AWG could satisfy this requirement of 6 GHz in the temperature range of −5 °C ~ 70 °C [21]. However, athermal AWG is still more expensive than the conventional AWG. To solve this problem, several wavelength-tracking techniques have been proposed to align WDM sources to the wavelength comb of the WGR at the RN [22]-[24]. However, these techniques require an additional loop-back fiber with multiple couplers [22], an additional monitor channel with an analog pilot tone and a fiber grating [23], or an additional monitor channel with a loop-back fiber [24]. We solved this wavelength-tracking problem simply by monitoring the optical powers of upstream signals [25]. This technique compares the total power of the multiplexed upstream signals with the optical power of a demultiplexed upstream signal at the CO, and adjusts the operating temperature of the AWG located at the CO for the wavelength-tracking.

6. Fault Monitoring and Localization

For the practical deployments of WDM PON, it would be necessary to have the capability of detecting and localizing the fiber failures without delay. In general, an optical time-domain reflectometer (OTDR) is used to localize the fiber failures. However, since the conventional OTDR operates at single wavelength only, it cannot be used in WDM PON due to the AWG placed at the RN. For example, if a failure occurs in the drop fiber connecting the RN and a subscriber, the conventional OTDR cannot identify its location. To overcome this problem, it has been proposed to localize the fiber failures in WDM PON by using a wavelength-tunable OTDR or implementing additional paths at RN to bypass AWG [26]-[27]. However, these techniques require either an expensive OTDR or a complicate network configuration. To solve this problem, we developed a simple and cost-effective technique to detect and localize the fiber failures in WDM PON [28]. This technique detects the fiber failures by monitoring the status of the upstream signals. Once the failure is detected, the corresponding downstream light source is switched to transmit OTDR pulses instead of data. Thus, the failures in drop fibers as well as in feeder fiber can be localized without using the wavelength-tunable OTDR. In addition, this technique does not require any additional light sources for the localization of fiber failure, since it reuses the downstream light source of the failed channel to transmit the OTDR pulse.

7. Survivable Network Architectures

Most of WDM PON architectures do not provide the self-protection capability against fiber failures, despite its large capacity. To solve this problem, it has been proposed to duplicate the fiber links or re-route the disrupted traffic via the adjacent ONU [29]-[30]. However, these techniques could be too costly for the use in access networks, as they require the doubling of fiber installations or numerous additional optical components. For this problem, we developed new survivable network architectures for WDM PON by utilizing two different wavelength-bands for the neighboring WDM PON’s and the cyclic property of AWG [31]. Thus, in case of a fiber failure, the disrupted signals can be recovered by using the fiber links of the neighboring WDM PON. As result, the protection can be achieved with the minimum amount of protection fibers.

8. Summary

We review the current technical issues for the development of practical WDM PON and report the relevant progresses achieved at KAIST, including low-cost WDM light sources, techniques for delivering broadcast services, wavelength-tracking techniques, fault monitoring and localization techniques, and survivable network architectures.

8. References

2. D. K. Jung et al., IEEE PTL, 10 (1998), 1334-1336
4. K. H. Han et al., IEEE PTL, 16 (2004), 2380-2382
5. N. R. Dono et al., IEEE JSAC, 8 (1990), 983-994
7. D. K. Jung et al., IEEE PTL, 10 (1998), 1334-1336
8. P. Healey et al., EL, 37 (2001), 1181-1182
10. N. Kashima, JLT, 9 (1992), 113-119
11. G. Mayer et al., JLT, 18 (2000), 121-142
12. C. R. Giles et al., IEEE PTL, 9 (1997), 1283-1284
13. P. C. F. Manners et al., OFC (2003), paper ThB4-4
14. T. Pfeiffer et al., JLT, 18 (2000), 1928-1938
15. H. Rohde, APOC (2005), paper 6022-22
17. U. Hiltb et. al., EL, 32 (1996), 2162-2163
18. P. P. Iannone et. al., IEEE PTL, 10 (1998), 1328-1330
19. E. S. Son et al., JLT, 21 (2003), 1723 – 1727
20. D. J. Shin et al., OFC (2005), paper JWA54
22. D. Mayweather et al., IEEE PTL, 8 (1996), 1238-1240
23. G. Iannone et al., JLT, 9 (1997), 523-525
24. R. Monnard et al., IEEE PTL, 9 (1997), 1655-1657
27. U. Hiltb et al., OFC, (1997), paper TuK3
28. K. W. Lim et al., IEEE PTL, 17 (2005), 2691-2693
30. T. J. Chan et. al., IEEE PTL, 15 (2003), 1660-1662
31. E. S. Son, OFC, (2005), paper OF14
Performance of Wavelength Shared Hybrid PON

Paparao Palacharla, Martin Bouda, Youichi Akasaka, Alexander Umnov, Takao Naito
Fujitsu Laboratories of America, Inc., 2801 Telecom Parkway, Richardson, TX 75082
Tel: +1-972-479-2450, Fax: +1-972-479-4482, Email: paparao.palacharla@us.fujitsu.com

Abstract
Wavelength shared hybrid PON provides seamless evolution from existing TDMA PONs to deliver high bandwidth triple play services. The associated transmission convergence layer protocol and performance of analog video overlay are presented.

1 Introduction
The demand for bandwidth in the last mile access network is constantly increasing due to new applications such as HDTV, P2P applications, video-on-demand, interactive games/conferencing etc. As a result, network operators are interested in approaches for evolution from today’s TDMA-PON [1,2] systems to next generation PON systems with higher bandwidth. Next generation PON architectures can be categorized into three approaches: 1. Increasing the speed of TDMA (e.g. 10 Gb/s line rate), 2. Providing a wavelength to each user (WDM-PON) [3], and 3. Combining TDMA and WDM. In the short term, designing and manufacturing of 10 Gb/s burst mode transceivers at low cost is a challenge. The second approach of allocating a wavelength to each user entails higher cost due to the need for colorless ONU’s based on technologies such as spectrum slicing, reflective SOA etc. The third approach of combining TDMA and WDM relies on off-the-shelf components and technologies to share the fiber infrastructure using simple system configuration at lower cost. The recently proposed novel PON architecture based on the combination of TDMA and WDM, called Wavelength Shared-Hybrid PON (WS-HPON), satisfies future bandwidth demand as well as mitigates the impairment of analog video due to Raman crosstalk [4,5]. In this paper, we discuss the transmission convergence protocol and characterize the performance of analog video overlay in WS-HPONs.

2 Wavelength Shared Hybrid PON Architecture
The WS-HPON architecture shown in Figure 1 uses either CWDM or DWDM wavelengths for digital data transmission in the downstream direction, while in the upstream direction a single wavelength at 1310nm (same as standard PON systems) is used. Each downstream wavelength is shared by a group of ONU’s, thereby lowering the cost but increasing the downstream bandwidth N-fold, where N is the number of downstream wavelengths. Since the upstream wavelength is same as TDMA-PON, the existing TDMA-PON ONUs can be made WS-HPON ready for future upgrade. This WS-HPON architecture provides flexible bandwidth upgrade at lower cost and complexity compared to WDM-PON.

Figure 1. Wavelength Shared Hybrid PON architecture

In this architecture, the different downstream wavelength signals at the OLT are combined using a wavelength multiplexer and launched into the feeder fiber. The remote node for WS-HPON architecture shown in Figure 1 has two paths. The feeder fiber from the OLT is terminated with a 3-port WDM filter at the junction of the two paths. This filter separates out the downstream wavelengths of the digital signals and sends them to the lower path. The upper path is used for broadcasting the video overlay signal downstream at 1550nm to all ONUs and also for combining upstream 1310nm signal from all of the ONUs. The lower path consists of wavelength routing (filtering) component to direct each downstream wavelength to a different group of ONUs. The two paths are combined using couplers/splitters. Figure 1 shows an example configuration with 4 downstream wavelengths for digital data serving four groups of 8 ONUs each. One of the main advantages of this next generation PON architecture is that the optical loss for downstream data signals is lower than that of the overlay broadcast video signal and also to the typical downstream loss of a power splitter based PON (e.g. TDMA-PON, stacked PON), which allows the transmitter power at OLT to be 5.5 dB lower per wavelength. This lower transmitted power results in reduction of Raman crosstalk in
WS-HPON architecture compared to standard PON or stacked PON architectures.

3 Transmission Convergence Protocol for WS-HPON
The logical topology of WS-HPON is shown in Figure 2. The downstream traffic is sent to ONUs via multiple wavelength transmitters, while a single receiver at the OLT receives the upstream traffic. Each group of ONUs associated with a downstream wavelength can be viewed as a virtual PON. So the discovery phase (including serial number acquisition and ranging [1,2]), can be done sequentially for each downstream wavelength transmitter. This requires no changes to existing transmission convergence layer protocol [1, 2]. The reachability table associating the wavelength transmitter with the corresponding ONUs is established at the OLT during this discovery phase.

![Logical topology of WS-HPON](image)

Figure 2. Logical topology of WS-HPON

4 Experiment
The experimental setup used an analog video transmitter operating at 1554nm and amplified by an EDFA to 20 dBm. The RF video input to the transmitter is a combination of four un-modulated carriers (channel 2 @ 55.25MHz, channel 3 @ 61.25 MHz, channel 4 @ 67.25 MHz and channel 5 @ 77.25 MHz). The CWDM OLT transmitter used four SFPs operating at CWDM wavelengths of 1430 nm, 1450 nm, 1470 nm and 1490 nm. These SFPs were driven with a $2^7$ PRBS pattern. These signals were combined with analog video signal and launched into different feeder fiber lengths. At the end of the fiber, the video signal was separated and attenuated independently to 0 dBm at the input of the triplexer. Table 1 shows the experimental results of carrier-to-signal ratio measurements for various optical power levels launched into the feeder fiber of 10 km. Initially measurements were done for each CWDM downstream wavelength. The amount of Raman crosstalk (carrier-to-spur) is dependent on the wavelength. The 1450nm wavelength with a separation of ~100nm (at the peak of Raman gain spectrum) from the video signal has the highest crosstalk, while the 1430nm wavelength has the lowest since the separation is ~120nm (other side of the Raman gain spectrum peak). The acceptable carrier-to-spur is < -60 dB according to the CATV standard. In the experiment, the transmitter power levels for GPON/stacked PON and WS-HPON are set at 1.5 dBm and -4 dBm (due to 5.5 dB lower loss) respectively. The standard GPON downstream wavelength of 1490nm shows unacceptable carrier-to-spur even at the minimum launch power of 1.5 dBm. In the case of WS-HPON architecture, the launch power of -4 dBm results in acceptable carrier-to-spur of ~66.5 dB. For 4λ CWDM case, we estimate the carrier-to-spur for stacked PON architecture and WS-HPON architecture is around ~49 dB and ~60 dB respectively, a reduction of 11 dB in crosstalk. These measurements can be viewed as worst-case crosstalk since all wavelength channels are driven by the same traffic pattern. In actual network deployment each wavelength channel is driven by GPON formatted signals with different content, the crosstalk levels are expected to be lower.

<table>
<thead>
<tr>
<th>Wavelength (nm)</th>
<th>4λ CWDM</th>
</tr>
</thead>
<tbody>
<tr>
<td>GPON/Stacked PON</td>
<td>49</td>
</tr>
<tr>
<td>WS-HPON</td>
<td>60</td>
</tr>
</tbody>
</table>

Table 1. Experimental results showing carrier-to-spur (in dB) of video signal

5 Conclusion
Wavelength shared hybrid PON architecture delivers at least 4-fold increase of bandwidth to each user compared to standard PON architectures. We have shown that existing standard GPON protocol can be used without any changes in WS-HPON. This allows bandwidth upgrade without replacement of existing ONUs in the field. In addition, WS-HPON reduces the effect of Raman crosstalk on video overlay signal by more than 10 dB compared to stacked PON architecture.

6 References
[2] IEEE 802.3ah, “Media access control parameters, physical layers and management parameters for subscriber access networks”, 2004
Virtual Private Group Formation in a WDM Passive Optical Network

Qiguang Zhao and Chun-Kit Chan
Department of Information Engineering, The Chinese University of Hong Kong, Shatin, N. T., Hong Kong SAR, China
Tel: +852-2609-4325, Fax: +852-2603-5032, Email: qgzhao5@ie.cuhk.edu.hk

Abstract
We propose and demonstrate a multiple virtual private group (VPG) formation scheme for ONU-VPGs over a WDM-PON. By employing the novel remote node design, multiple independent ONU-VPGs can be realized with identical inter-ONU transceivers in the optical layer.

1 Introduction
With the increasing demand of the community networking and corporate networking, internetworking communication among optical network units (ONUs) has emerged as an access network design consideration in wavelength-division-multiplexed passive optical network (WDM-PON). Conventional strategies supporting inter-ONU communication would consume the bandwidth of both the downstream and the upstream signals [1]. Therefore, it is highly desirable to offer flexible and arbitrary formation of virtual private groups for ONU-VPGs in WDM-PONs. Electronic code-division multiple access (E-CDMA) technique [2] and subcarrier multiplexing (SCM) technique [3] were utilized to support multiple ONU-VPGs. However, these previous schemes required additional electronic techniques to distinguish among different ONU-VPG traffic and require the processing at higher layers. Moreover, these schemes multiplexed more than one data traffic on the same carrier. Thus the system may suffer from interference, which limits the transmission data rate for the inter-ONU traffic.

In this paper, a novel scheme is proposed and experimentally demonstrated to realize ONU-VPG formation in the optical layer. Compared with other schemes providing ONU-VPG configuration, our scheme can support high data rate up to 2.5-Gb/s for inter-ONU traffic in the optical layer with identical inter-ONU transceivers at all ONUs.

2 Proposed architecture and operation principle
Fig.1 illustrates our proposed architecture of multiple ONU-VPGs over a WDM-PON with N users, where N=8 for illustration. Similar to the conventional WDM-PON architectures, N downstream transceivers in the 1.55μm waveband are designated at the optical line terminal (OLT). The 1×2N (say 1×16) array-waveguide grating (AWG1) is employed at the OLT to multiplex and demultiplex the downstream and the upstream carriers, respectively. A formation example of two ONU-VPGs (ONU-VPG1 & ONU-VPG2) in the network is shown in Fig. 1(b), where five ONUs (ONU1-5) are involved in the ONU-VPG1 and the other three ONUs (ONU6-8) belong to the ONU-VPG2. At the remote node (RN), one 5×5 star-coupler and one 3×3 star-coupler are employed to duplicate and broadcast the inter-ONU traffic signals, via the 2N×2N AWG2 (say 16×16) within their respective ONU-VPGs. The connection patterns between the star-couplers and the AWG2 are consistent with the ONU-VPGs configuration. For ONU-VPG1, the input ports of the 5×5 star-coupler are connected to the 1, 11, 13, 15, 17, 19 ports of the AWG2, while their output ports are connected to 12, 14, 16, 18, 110 ports of the AWG2, respectively. Similarly, the connection pattern for the ONU-VPG2 is illustrated in Fig. 1(a). At the ONU, two transceivers are assigned for normal up-/downstream traffic transmission and inter-ONU traffic transmission, respectively. Fig. 1(c) illustrates the wavelength assignment plan. The downstream and the upstream carriers are in the blue-band free-spectral range (FSR) of the AWG2, while the inter-ONU traffic carriers are assigned as the first wavelength λ2N+1 (say λ17) in the red-band FSR of the AWG2. Therefore, Blue/Red (B/R) filters are utilized to combine and separate the inter-ONU carrier at the RN and the ONUs, respectively.

The formation of the ONU-VPGs is realized by flexible setting of the star-couplers configuration at the RN, λ17, as shown in Fig. 1(c), is the first wavelength channel in the red-band FSR of the AWG2, and it is assigned as the inter-ONU traffic carrier for all ONUs in the network. A suitable media access protocol, such as carrier-sense multiple access, is assumed to coordinate the data transmission by all ONUs among themselves. According to the wavelength routing principle of the AWG2, λ17 has a special correspondence feature between the input port and the output port of the AWG2. That is, when λ17 is fed into the kth input port of the AWG2, it
would be routed to its $k^{th}$ output port. Hence, when the star-couplers are employed to broadcast this inter-ONU carrier within their respective ONU-VPGs, this special correspondence feature of the AWG2 supports flexible and arbitrary formation of ONU-VPGs by simply connecting the output ports of the star-coupler to the input ports of the AWG2 whose corresponding output ports are connected to the destined inter-ONU receivers in the same ONU-VPG. Under this configuration, the inter-ONU traffic sent from ONU1 would be broadcasted to ONU1, ONU2, ONU3, ONU4, and ONU5 only in ONU-VPG1, while that sent from ONU6 would be broadcasted to ONU6, ONU7 and ONU8 only in ONU-VPG2.

3. Experimental demonstration
The experimental setup was similar to Fig. 1 so as to investigate the functionality and transmission performance of ONU-VPGs formation. Two DFB laser diodes directly modulated by 2$^{23}$-1 pseudorandom bit sequence (PRBS) non-return-to-zero (NRZ) data at 2.5-Gb/s were employed as the downstream and the upstream transmitters. The upstream signal was received by a p-i-n receiver at the OLT. Two 16×16 AWGs, with an FSR of 12.8 nm and 100-GHz channel spacing, were employed as the AWG1 at the OLT and the AWG2 at the RN. They were connected by a piece of 20-km standard single-mode fiber (SMF), as the feeder fiber. A 4×4 star-coupler was used at the RN, with its first input port connected to the Red/Blue filter (18-nm passband) while the others were directly connected to the designated input ports of the AWG2. The distribution fiber between the RN and the ONU was 4-km SMF. At the ONU, a tunable laser, lasing at 1529.2 nm and externally modulated by 2.5-Gb/s NRZ 2$^{23}$-1 PRBS, was employed as the inter-ONU traffic transmitter, due to component availability. The downstream signal and the inter-ONU traffic signal were received by a 2.5-Gb/s p-i-n receiver and a 2.5-Gb/s avalanche photodiode (APD) receiver, respectively.

The bit-error-rate (BER) measurement results for the upstream, the downstream as well as the inter-ONU traffic are depicted in Fig. 2. Compared to the back-to-back measurement, the negligible power penalty was less than 0.5 dB for both the downstream and the upstream traffic after 24-km SMF transmission, as depicted in Fig. 2(a). This could be attributed by the fiber chromatic dispersion. The receiver sensitivities at BER=$10^{-9}$ for the downstream and the upstream signals were measured as within -22.5 dBm and -23 dBm. Besides, the broadcasting transmission performance of inter-ONU traffic was also investigated in all four output channels of the star-coupler. As depicted in Fig. 2(b), error free operation was achieved in all cases, with the receiver sensitivities at about -28 dBm.

4. Discussion
The loss budget for the inter-ONU traffic signal was about 22 dB, which include the round-trip transmission between the OLT and the RN via the AWGs (10 dB). Thus it can provide more than 7-dB power margin at the receiver sensitivity of -28 dBm at 2.5-Gb/s, assuming the output optical power of inter-ONU traffic transmitter is 1 dBm. When the commercial AWGs with low insertion loss (3 dB) are employed in practical network deployment, the loss budget can be further reduced by 4 dB. Therefore, our proposed scheme can support up to sixteen ONUs within each ONU-VPG.

5. Summary
We have demonstrated a novel ONU-VPGs formation scheme in optical layer for WDM-PONs. The ONU-VPGs formation is realized by the special wavelength assignment and the star-couplers at the RN, where a cyclic AWG is employed. The connection patterns between the star-couplers and the AWG physically partition the network into multiple ONU-VPGs. It is compatible with the conventional WDM-PONs architectures and provides a flexible solution to upgrade it to the WDM-PONs with ONU-VPGs formation capability. This project is partially funded by a research grant from Hong Kong Research Grants Council (Project No. CUHK4142/06E).

6. References
A Novel Wavelength Router Architecture for Optical VPNs using Multistage AWGs in a WDM-PON

Kohei Okada1, Shohei Terada, and Kimio Oguchi2
Information Networking Lab., Graduate School of Engineering, SEIKEI University
3-3-1 Kichijoji-Kitamachi, Musashino, 180-8633 Japan
Tel: +81-422-37-3732, Fax +81-422-37-3871, Email: 1dm053203@cc.seikei.ac.jp, 2oguchi@st.seikei.ac.jp

Abstract
The WDM-PON architecture enables OLT-ONU connections while ONU-ONU full-mesh VPNs offer new functionality in PON-based access networks. A novel wavelength router architecture that yields multiple full-mesh VPNs is proposed. Measured optical power losses and crosstalk confirm the feasibility of the proposed architecture.

1. Introduction
Wavelength division multiplexing (WDM) was originally used to increase the capacity of optical networks, especially the core networks. It was recently extended to the routing function [1] e.g. GMPLS (General Multi-Protocol Label Switching). Access networks also use wavelengths in the WDM-passive optical network (PON) for service multiplexing and capacity enhancement [2, 3]. WDM also offers new functionalities in the WDM-PON, i.e. the creation of the Optical Network Unit (ONU)-ONU virtual private network (VPN) [4-7]. However, several issues with the WDM-PON remain. One is the need for architectures that can configure full-mesh VPNs for multiple groups of customers (ONUs). Scalability is also an issue for large scale wavelength routers that offer multiple VPNs.

This paper proposes a novel scalable wavelength router architecture for a WDM-PON that uses multistage arrayed waveguide gratings (AWGs); it can create multiple full-mesh VPNs. Verification of the wavelength routing function by the wavelength transfer matrix (WTM) method, together with some experimental results are described.

2. Proposed wavelength router architecture
Figure 1 shows the WDM-PON that uses the proposed wavelength router architecture. One optical line terminal (OLT) and sixteen ONUs A-P are connected to the wavelength router with optical fibers, as an example. This WDM-PON architecture offers four full-mesh VPNs based on wavelength as well as conventional OLT-ONU connections, simultaneously. The OLT and ONUs use only four wavelengths $\lambda_1$, $\lambda_2$, $\lambda_3$, and $\lambda_4$. The proposed router architecture with multi-stage structure is shown in Fig. 2. It consists of a $4 \times 1$ splitter/combiner and four $4 \times 4$ AWGs; its input-output wavelength characteristics are described in Fig. 3. AWG ports 1u, 2u, and 3u are connected to the other AWGs by looped-back optical fibers while port 4u is connected to the $4 \times 1$ splitter/combiner. AWG port 1d is connected to VPN #1, port 2d to VPN #2, port 3d to VPN #3, and port 4d to VPN #4. Logical network topologies of ONU-ONU connections are shown in Fig. 4 (b). All ONUs that belong to the same VPN can communicate with each other.

Figure 2 shows an example of sixteen customers for illustration. If $N$- $N \times N$ AWGs and one $N \times 1$ splitter/combiner are used, the number of ONUs, or customers can be scaled up/down to $N^2$ in a simple and easy way in the proposed router architecture.

3. Wavelength characteristics of the proposed wavelength router by the Wavelength Transfer Matrix
The wavelength transfer matrix (WTM) method is a powerful tool to design and verify complicated wavelength routing networks [6][8].

The WTM method is used to verify the input-output characteristics of the proposed wavelength router. Figure 5 shows WTM allocations for the proposed wavelength router. The WTM for optical fibers #1, $L_1$; the AWGs, $L_4$; optical fibers #2, $L_{22}$; and $4 \times 1$
An allocations of WTM for $B = A$ (of wavelength circuits and $(H)M$ on the wavelength transfer $G$).

Fig. 4. Logical network topologies of WDM-PON

Fig. 5. Allocations of WTM for proposed wavelength router

\[ O_{OLT} = L_S \cdot L_{F2} \cdot L_A \cdot L_{F1} \cdot L_{OUN} \]

4. Measurement of optical power losses and crosstalk

Optical power budget was designed with parameters such as user number = 256, 32 wavelengths, anchor wavelength of 1552.525nm, channel spacing of 100GHz, optical input power of 1.5dBm, bit rate of 2.5Gbps, optical sensitivity of -25dBm, insertion loss of 10dB for 1x8 optical splitter, insertion loss of 6dB for 32x32AWG, crosstalk of 32x32 AWG less than -25dB. According to measured optical power losses and crosstalk for the proposed router, the optical power losses were 14.2dB from OLT side to ONU side, and 12.0dB from ONU side to ONU side, and the cross talk from adjacent ports of the router was -35.4dB.

5. Conclusion

This paper presented a novel wavelength router architecture that offers excellent network scalability and can realize multiple full-mesh VPNs. It was also verified that it can model the wavelength characteristics of WDM-PONs that use the proposed wavelength router with multistage-AWGs. Measured optical loss and crosstalk for the proposed architecture verified its feasibility.

The authors would like thank to Dr. K. Noguchi, Dr. Y. Sakai, and Dr. O. Moriwake of NTT Photonics Labs. for their support in the experiments.

References

Demonstration of an 80-channel DWDM-PON based on Wavelength-locked Fabry-Perot Laser Diodes

Ki-Man Choi, Sil-Gu Mun, Jung-Hyung Moon, and Chang-Hee Lee

Dept. of Electrical Engineering and Computer Science, Korea Advanced Institute of Science and Technology
373-1 Guseong-dong Youseong-gu, Daejeon, 305-701, Korea
(Tel) +82-42-869-8063, (Fax) +82-42-869-3410, (e-mail) chl@ee.kaist.ac.kr

Abstract: An 80-channel DWDM-PON based on wavelength-locked F-P LDs with a 50-GHz channel spacing is demonstrated. We investigate the possibility of increasing the number of subscribers in DWDM-PON.

Recently, a change of paradigm occurs in access networks. Voice and text oriented services have evolved to data and image based services due to the growth of the Internet with its new variety of video services. This paradigm change requires new access networks that support high-speed, symmetric, and guaranteed bandwidth for future video services with high definition TV (HDTV) quality. The wavelength-division-multiplexing passive optical network (WDM-PON) has been considered as an ultimate broad-band access network to meet higher bandwidth requirement, high security, and bit rate/protocol transparencies. When we estimate the maximum TV sets per home was three and HDTV signals were encoded by MPEG2 (19.4 Mb/s per channel), the bandwidth requirement for downstream is about 73 Mb/s by including high speed internet (10 Mb/s) and some other services such as a video phone (3 Mb/s). For upstream, a game on demand and an education on demand need symmetric bandwidth with HD quality. We also assumed a symmetric bandwidth for high speed internet. Then, we come up with 52 Mb/s for upstream. Thus, the dedicated bandwidth per home can be determined as 100 Mb/s by adding future coming services [1].

Recently, a dense WDM-PON (DWDM-PON) guaranteeing a 125 Mb/s per channel was demonstrated with a 50-GHz channel spacing based on the wavelength-locked Fabry-Perot laser diodes (F-P LDs) [2]. The demonstrated DWDM-PON uses an erbium-doped fiber amplifier (EDFA)-based broadband light source (BLS) as a seed light and accommodates 64 subscribers with 8 Gigabit/s capacity in an upstream direction. The EDFA-based BLS can achieve about 32 nm bandwidth. Then, we can accommodate maximum 80 channels with a 50 GHz channel spacing.

In this paper, we demonstrate an 80-channel DWDM-PON based on the wavelength-locked F-P LDs with a 50-GHz channel spacing to investigate a possibility of increasing the number of subscribers to 80. From the 64-channel based on 125 Mb/s of the 100-Base Ethernet packet, 8-channel of 155 Mb/s NRZ data and 8-channel of CW light from high-power F-P LDs are added to demonstrate the 80-channel DWDM-PON. The DWDM-PON performance is measured with packet loss rate (PLR) for 125 Mb/s data and bit error rate (BER) for 155 Mb/s data. Then, we confirmed the packet-loss-free and error-free transmission.

The 80-channel DWDM-PON based on the wavelength-locked F-P LDs with a 50-GHz channel spacing is shown in Fig. 1. It is composed of an optical line termination (OLT) in a central office (CO), a 20-km SMF as a feeder fiber, a remote node (RN), 80 optical network terminations (ONTs). The OLT consists of transmitters, receivers, a cyclic arrayed waveguide gratings 1 (AWG 1), two BLSs for the C/L-band. The C/L-band BLS that provides the injection power for the upstream/downstream channels is fed into the system through coupling devices that consist of a circulator and two band-separating C/L wavelength-division multiplexers (C/L WDMs). The channel spacing of the AWG is 50 GHz and the BLS is an EDFA-based amplified spontaneous emission (ASE). Another cyclic AWG 2 is located at RN. At subscriber side, 64 ONTs for 125 Mb/s are located between channel 5 and channel 76.

Fig. 1. Experimental set-up for an 80-channel DWDM-PON based on the wavelength-locked F-P LD.

Fig. 2. The measured optical spectra of 8-channel CW light from the interleaved two high-power F-P LDs.
(ch. 5 ~ 40, ch. 49 ~ 76). To increase the number of ONTs to 80, 8 ONTs are inserted between ch. 1 ~ ch. 4 and ch. 76 ~ ch. 80. The 64 ONTs for 125 Mb/s data rate are directly modulated by 100 Base-x Ethernet packets with various packet lengths. The 8 ONTs are directly modulated with a 155 Mb/s NRZ data. The F-P LDs inside the transmitters are TO-can packaged and the mode spacing is 0.6 nm. The front-facet of the F-P LDs is anti-reflection coated and the temperature of the F-P LDs is controlled by a thermoelectric cooler (TEC) controller. In addition of these 72 channels, we use CW light from two high-power F-P LDs to make 80 channels in the upstream direction. Since mode spacing of the high-power F-P LD is 100 GHz, each mode of two high-power F-P LDs is interleaved with 50 GHz mode spacing and combined through 3 dB coupler. Then, each mode of the interleaved high-power F-P LD is filtered by AWG 3 with 50 GHz channel spacing. An isolator is used between 3 dB coupler and AWG 3. The measured optical spectra of the filtered CW light are shown in Fig. 2. Optical power of these 8-channel CW light is between -2 dBm ~ -11 dBm and each channel is connected to AWG 2 from ch. 41 to ch. 48.

To investigate the performance of 80-channel DWDM-PON based on the wavelength-locked F-P LDs, we measured PLR for the 64 ONTs (100-Base Ethernet packets of 125 Mb/s) and BER for the 8 ONTs (NRZ 155 Mb/s) in an upstream direction. Since mode spacing of the high-power F-P LD is 100 GHz, each mode of two high-power F-P LDs is interleaved with 50 GHz mode spacing and combined through 3 dB coupler. Then, each mode of the interleaved high-power F-P LD is filtered by AWG 3 with 50 GHz channel spacing. An isolator is used between 3 dB coupler and AWG 3. The measured optical spectra of the filtered CW light are shown in Fig. 2. The signal power of wavelength-locked F-P LDs varies according to detuning between lasing mode and the injection wavelength. To measure the PLR, a variable optical attenuator (VOA) is inserted between the feeder fiber and the RN, as shown in Fig. 1. The measured PLRs of the upstream 64-channel are shown in Fig. 4. Here, a PLR of $10^{-6}$ corresponds to a BER of $10^{-10}$, approximately. For the 8 ONTs with 155 Mb/s, we measured a BER as shown in Fig. 5. As shown in the PLT and the BER curves, there are no sign of error-floor. Thus, packet-loss-free transmission for the 64-channel and error-free transmission for the 8-channel are achieved. It may be noted that capacity is 9.2 Gb/s in the upstream direction (64-channel of 125 Mb/s and 8-channel of 155 Mb/s). Although we used a CW light for the 8-channel within 80-channel, we believe it is sufficient for demonstration of 80-channel transmission, since the induced crosstalk between WDM channels is small enough. It implies that the total capacity can be 12.4 Gb/s.

When a SLD with 52 nm bandwidth is used for the seed light instead of the EDFA-based BLS, the number of channels can be extended to 128 based on the DWDM-PON with 50 GHz channel spacing. In conclusion, we demonstrated an 80-channel DWDM-PON based on the wavelength-locked F-P LDs with a 50 GHz channel spacing. The split ratio (or number of channels) can be expanded to 128 with a SLD based the seed light. Therefore, DWDM-PON based on the wavelength-locked F-P LDs is useful for future access network.

**Acknowledgment**

This work was supported by the Korea Ministry of Science and Technology under the National Research Laboratory.

**References**


Ultra high capacity WDM transmission using CSRZ-DQPSK format and extended L-band optical amplifiers

Akihide Sano¹, Hiroji Masuda¹, Eiji Yoshida¹, and Yutaka Miyamoto¹,
(¹NTT Network Innovation Laboratories, 1-1 Hikari-no-oka, Yokosuka, Kanagawa, 239-0847 Japan,
Tel: +81-46-859-5541, Fax: +81-46-859-5214, Email: sano.akihide@lab.ntt.co.jp)

Abstract
WDM transmission system capacity has been dramatically increased to exceed 20 Tb/s. This paper describes high capacity transmission technologies focusing on high speed CSRZ-DQPSK modulation techniques and extended L-band amplification.

1 Introduction
Due to the rapid increase in IP data traffic, large capacity and cost-effective transmission systems are strongly required to construct future optical transport networks. In order to handle the increase in data traffic, future wavelength-division-multiplexing (WDM) systems are expected to achieve 10-Tb/s-class total capacity. High spectral efficiency (SE) modulation/demodulation schemes and wide band optical amplification technologies are indispensable to realize such high capacity transmission systems. Fig. 1 shows SE and signal bandwidth of recent large capacity WDM transmission experiments. High SE transmission has been demonstrated by using multi-level modulation techniques such as the differential quadrature phase-shift keying (DQPSK) format and polarization-multiplexing techniques. Broadband optical amplification is also indispensable to realize 10-Tb/s-class total capacity. Up to now, these techniques have been reported to realize total capacities of 25.6 Tb/s with dual gain-band amplifiers [1] and 20.4 Tb/s with single gain-band amplifiers [2].

Moreover, higher speed interfaces such as 100 Gb/s Ethernet are expected in the near future. Therefore, the future large capacity WDM system is expected to transport 100-Gb/s-class high speed channels.

In this paper, we describe high capacity and high speed WDM transmission technologies focusing on high SE modulation techniques and broadband optical amplification techniques.

2 High SE modulation
Multi-level modulation schemes are very effective to improve SE. In particular, the DQPSK format has been extensively investigated over the last few years. The advantages of DQPSK format are summarized as follows: 1) Spectral width is half that of binary modulation, it can

![Fig. 1. Spectral efficiency and signal bandwidth of recent large capacity WDM transmission experiments. Circles: single polarization. Squares: polarization interleaving. Triangles: polarization multiplexing. Open and filled symbols show binary and quaternary modulation, respectively.](image-url)
double the SE. 2) it can support line rates that are twice the operation speed of the electronics. 3) Tolerances against chromatic dispersion (CD) and polarization mode dispersion (PMD) can be improved because of their narrow spectral width and wide symbol interval. The latter two merits are quite important in high speed transmission such as 100 Gb/s. On the other hand, its drawbacks are mainly transmitter/receiver complexity and tolerance to fiber nonlinearity and laser phase noise.

Return-to-zero (RZ) modulation has been widely used in phase-modulated signal transmission. Carrier-suppressed (CS) RZ modulation is promising because the spectral broadening is small compared to other RZ modulation schemes, and so it is suitable for high SE WDM transmission. We have conducted 111-Gb/s carrier-suppressed return-to-zero (CSRZ-) DQPSK transmission by employing high-speed InP HBT digital ICs, DQPSK modulators based on the hybrid integration of a planar lightwave circuit (PLC) and LiNbO$_3$ (LN) lightwave circuit, and an InP balanced detector [4-6]. Fig. 2 shows measured and calculated CD and PMD tolerances of 111-Gb/s DQPSK format. The calculated performance of conventional non-return-to-zero on-off keying (NRZ-OOK) are also shown for comparison. The measured CD and PMD tolerances at 1-dB OSNR penalty are 30 ps/nm and 9 ps, respectively, and we can confirm that CSRZ-DQPSK format effectively improves both CD and PMD tolerances.

3 Broadband optical amplification techniques

One approach toward bandwidth-extension is the parallel configuration of optical fiber amplifiers with different gain bandwidth. 13.7-THz signal bandwidth has been obtained over the S-, C-, and L-bands by using three types of fiber amplifiers, and a 10.9 Tbit/s transmission experiment has been demonstrated [7]. In order to realize cost-effective WDM systems, on the other hand, extending the gain bandwidth of single optical amplifiers is an attractive approach. Extended L-band amplifiers such as phosphorous co-doped silicate EDFA (P-EDFA), bismuth-oxide-based EDFA and tellurite-based EDFA offer continuous wide gain bands. We have realized 7-THz bandwidth and demonstrated 14-Tb/s transmission by using a P-EDFA. Moreover, 10.2-THz bandwidth, 20.4-Tb/s transmission has demonstrated by using hybrid amplification technique of P-EDFA and Raman amplifiers. All Raman amplification is also attractive to obtain continuous wide gain bandwidth. 12.2-THz bandwidth has been obtained by using silica fiber Raman amplifiers [8], and hybrid silica/tellurite fiber Raman amplifiers have attained 15.7-THz bandwidth [9]. Further extension of total capacity will be expected by employing these Raman amplification technologies and high SE modulation techniques.

4 Conclusion

We have discussed high capacity and high speed WDM transmission technologies focusing on high SE modulation and broadband optical amplification. CSRZ-DQPSK modulation and extended L-band amplification techniques are promising to realize cost-effective WDM systems.

References

Performance improvement of DPSK signal transmission by a phase-preserving amplitude limiter

Tetsuji Kamio, Kenichi Sanuki, and Masayuki Matsumoto

Graduate School of Engineering, Osaka University, 2-1 Yamadaoka, Saita, Osaka 565-0871, Japan
kamio@procyon.comm.eng.osaka-u.ac.jp

Abstract

We show that a phase-preserving amplitude limiter using saturation of FWM in a fiber improves DPSK signal transmission through the reduction of nonlinear phase noise. Influence of imperfections of the limiter is also studied.

1. Introduction

Phase-shift keying (PSK) modulation has advantages including higher receiver sensitivity than on-off keying modulation and suitability for multi-level signaling such as QPSK or DQPSK [1]. The phase noise imposed on the signal generally determines the transmission performance of PSK signals. In long distance systems where higher signal power is required, nonlinear phase noise, which is caused by the translation of amplitude noise to phase noise through nonlinearity of transmission fiber, is especially harmful [2].

Some means for reduction of the nonlinear phase noise have been proposed [3,4]. In this paper, we study the effect of an optical limiter using saturation of four-wave-mixing (FWM) in a nonlinear fiber [5,6]. The limiter suppresses the amplitude noise which is responsible for the nonlinear phase noise. It is theoretically and experimentally shown that the phase noise at the receiver is reduced by inserting the limiter into a transmission line, leading to improvement of DPSK transmission performance.

2. Theoretical calculation of phase noise

We consider an amplified transmission system consisting of M spans. Amplified spontaneous emission (ASE) from the inline amplifiers is the dominant source of the phase noise. The quadrature component of the ASE relative to the signal makes direct phase fluctuation whose variance accumulates proportionally to the number of amplification stages. The in-phase component, on the other hand, makes amplitude fluctuation, which is translated to nonlinear phase noise through fiber nonlinearity. The nonlinear phase noise becomes large when transmission distance and/or the signal power are large.

An ideal limiter eliminates amplitude fluctuation perfectly without giving additional phase noise. Actual limiters, however, have several imperfections. Firstly, in the case of the limiter using fiber nonlinearity, signals having different input powers suffer different phase shift due to SPM in fiber although the signal power is equalized at the output of the limiter. The induced phase shift \( \delta \phi \) is proportional to the power fluctuation of the input signal \( \delta P \) to be suppressed by the limiter. The proportionality coefficient \( k \) is defined by the relation \( \delta \phi = k \delta P / P_{sat} \), where \( P_{sat} \) is the saturation input power. Secondly, actual limiters cannot eliminate amplitude fluctuations perfectly. This can be quantified by a residual power fluctuation ratio \( r \), where the input signal power with fluctuation \( P_{sig} + \delta P \) yields the output signal power \( P_{sat}(1 + r \delta P / P_{sat}) \). Considering these imperfections, the variance of phase noise when limiters are inserted every span is given by

\[
\langle \delta \phi^2 \rangle = \left\{ 1 + 4(k + P_{sig} \gamma L_{eff} \tau) \left[ (1 - r^M)/(1 - r) \right]^i \right\} \frac{N_{sat} B}{2P_{sig}} 
\]

\[
+ \left\{ M + 4(k + P_{sig} \gamma L_{eff} \tau) \sum_{i=1}^{M} \left[ (1 - r^{M-i})/(1 - r) \right]^i \right\} \frac{N_{sat} B}{2G P_{sat}} 
\]

\[
+ \left\{ M + 4(k + P_{sig} \gamma L_{eff} \tau) \sum_{i=1}^{M} \left[ (1 - r^{M-i})/(1 - r) \right]^i \right\} \frac{N_B}{2P_{sat}},
\]

where \( N_{sat}, B, P_{sig}, \gamma, L_{eff}, G, \) and \( N_B \) is the power spectrum density of the source noise, noise bandwidth, peak power launched into the transmission fiber, nonlinear coefficient and effective length of the transmission fiber, gain of an amplifier in the optical limiter, and spectrum density of ASE from each inline amplifier, respectively. \( N_s \) is related to the source OSNR (noise bandwidth of 0.1nm) as \( N_s = sP_{sig}'(12.5GHz \times OSNR) \), where \( s \) is the duty ratio of the signal (the averaged signal power is given by \( P_{ave} = sP_{sig} \)) while \( N_s \) is given by \( h v n_{sp}(G-1) \), where \( h v \), \( n_{sp} \), and \( G \) are the photon energy, spontaneous emission factor, and gain of inline amplifier, respectively.

Fig.1 is the phase noise versus average signal power \( P_{sig} \) when an imperfect limiter (\( k \neq 0 \) and \( r \neq 0 \)) is inserted every span. \( k \) and \( P_{sat} \) are assumed to be 0.8rad and 50mW, respectively. \( r = 0 \) corresponds to perfect amplitude limitation while no amplitude limitation exists for \( r = 1 \). Fig.1 shows that the phase noise is gradually reduced when \( r \) is small. However, when \( r \) is larger than ~0.4, the phase noise is larger than that without limiters.
Fig. 1. Standard deviation of phase noise at the receiver versus signal power. Amplitude limiters with imperfect noise suppression are inserted every amplifier span. Thin solid curve is the phase noise without using amplitude limiters. γ and α of transmission fiber, source OSNR, and noise figure of amplifiers are 3.5/W/km, 0.3dB/km, 24.5dB, 6dB, respectively. Span length and number of spans are 40km and 5.

3. Experiment

Fig. 2. Setup of 10Gbit/s short-pulse DPSK transmission. An amplitude limiter based on saturation of FWM is inserted in the recirculating loop.

Fig. 2 shows the setup of a DPSK transmission experiment. 10GHz pulses are generated by a mode-locked laser diode (MLLD) whose width is 6.8ps after optical bandpass filter (OBPF). The transmission fiber (40km) is a densely dispersion-managed fiber originally designed for 80Gbps short-pulse transmission. In this experiment, an amplitude limiter using saturation of four-wave mixing (FWM) in a highly nonlinear fiber (HNLF) is inserted in the recirculating loop. The limiter simply consists of an EDFA with gain G, a polarization controller, continuous-wave pump source, HNLF and an OBPF.

Fig. 3 shows bit-error rate measured after transmission of 200km (five spans) when the limiter is inserted inside the recirculating loop. OSNR of the input signal is 25.7dB. Fig. 3 shows that the BER is decreased in a wide range of signal power when the limiter is activated with the pump power for the FWM turned on. The behavior qualitatively corresponds to the calculated phase noise versus signal power as shown in Fig. 3.

Fig. 3. BER versus averaged signal power launched to the transmission fiber. Solid and dashed curves correspond to the cases where pump power in the limiter is on and off, respectively.

4. Conclusion

In this paper, we show both theoretically and experimentally that an amplitude limiter using saturation of FWM in a fiber is effective in improving DPSK signal transmission performance.

References

Suppression of BER fluctuations by electrical equalizer in the ultra-long 43Gb/s transmission experiments

Toshiharu Ito
System Platform Research Laboratories, NEC Corporation.
1753, Shimonumabe, Nakahara-ku, Kawasaki, 211-8666, Japan, Tel: +81-443963096, E-mail: t-ito@dl.jp.nec.com

Abstract
The use of Electrical Equalizer for PMD tolerance increase in the 43Gb/s RZ-DPSK signal transmission successfully suppresses the BER fluctuations due to time-varying PMD, which is essential for maintaining the appropriate condition of the receiver.

Introduction
In the long distance optical transmission systems using high speed channel speed of more than 40Gb/s, the signal waveform distortion due to the polarization mode dispersion (PMD) in the transmission line can be a dominant limiting factor of transmission distance.

It can be roughly said that there are two issues caused by PMD-induced signal degradation. One is that the system operation will be stopped in case when the time-varying PMD value exceeds the PMD tolerance of the receiver. The other is that the adjustment or control of receiver condition become complicated or accuracy of these processes can be spoiled. The receiver used in such high-speed transmission systems needs the precise adjustment or control of itself to cope with various degradation factors, however, only the PMD-induced degradation can be changeable with the speed faster than the ordinary control speed.

We have been experimentally investigating an EE (Electrical Equalizer) [1] applied on the 43Gb/s RZ-DPSK signal for the increase of PMD tolerance as well as the stabilization of BERs in the range of this increased PMD tolerance[2]. In these studies, we did some long-term transmission experiments using ultra long length straight transmission line in order to observe the BER fluctuations due to time-varying PMD. In this paper, we will introduce some examples of the obtained data to show how the BER fluctuations were suppressed by the EE.

Experiments
The straight transmission line used for this test had the length of 4,300km, and composed of hundred 43-km DMF spans [2]. All the experimental setup including transponder was placed in the laboratory room with stable environment. Fig.1 shows the main part of the 43Gb/s RZ-DPSK transponder used in following evaluations. An EE, consisting of four gain stages in a four-tap transversal filter topology and a gain limiting block, was inserted at the output of the dual O/E converter.

Wavelength of the signal light from this transponder was 1550.9nm, and the input signal power into the transmission fiber was set to be -4dBm/ch.

<table>
<thead>
<tr>
<th>Setting of EE</th>
<th>Polarization scrambling</th>
<th>Fluctuation of Q-factors [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a) High sensitivity</td>
<td>OFF</td>
<td>2.31</td>
</tr>
<tr>
<td>(b) High sensitivity</td>
<td>ON</td>
<td>1.36</td>
</tr>
<tr>
<td>(c) Large PMD tolerance</td>
<td>OFF</td>
<td>0.88</td>
</tr>
<tr>
<td>(d) Large PMD tolerance</td>
<td>ON</td>
<td>0.58</td>
</tr>
</tbody>
</table>

We have conducted long-term BER measurements for four cases, described in Table.1. The EE setting of “High sensitivity” was the setting to achieve best receiver sensitivity for non-distorted signal. With this setting, the receiving performances were nearly equal to the case without EE. A polarization scrambler was inserted at the output of transponder,
and the dependence of input polarization state into the transmission line was averaged when the scrambler was ON. The speed of polarization scrambling was fast enough to remove the input polarization dependency in the BER averaging interval of 10 seconds. “Fluctuation of Q-factors” in Table.1 stands for the [Best Q-factor] - [Worst Q-factor] in each case. Every “Q-factor” used here was the value obtained by the calculation from BER data.

Fig.2 shows the measured Pre-FEC BER data for four cases. The measurement time was 48 hours (2 days) for each case. Fig.2 (a) shows the measured results of case (a) where the BERs fluctuated with large vibrations (2.31dB) and with relatively short time intervals. In case (b), the fluctuations with short intervals were reduced by the adoption of polarization scrambling while the large vibrations (1.36dB) still remain. As shown in fig.(a) and (b), adoption of “Large PMD tolerance” setting effectively suppressed the BER fluctuations. In case (d), “Fluctuation of Q-factors” was only 0.58dB during two days.

Discussions
The present design of the transmission systems mainly uses the average of BERs, their deviation, and the margin considering the FEC capability. The BER fluctuations which are intentionally suppressed by the EE should not be used directly in such design method since this suppressing function is effective in the finite range determined by the capability of EE. The use of some complementary techniques like in-service PMD estimation method [2] will be needed for the design with the consideration of the PMD margin.

Conclusion
We have introduced some experimental results showing how an EE affects the BER fluctuations due to the time-varying PMD in the transmission line. According to our evaluations using 4,300km straight transmission line and 43Gb/s RZ-DPSK signal, “Fluctuation of Q-factors” in continuous 48-hour measurements were improved from 2.31dB to 0.58dB. These stabilized performances will contribute to execute an accurate adjustments of the receiver conditions, and thus the further stable transmission performances will be realized.

Reference
Abstract  Polarization mode dispersion (PMD) is a potentially limiting impairment in high-speed long-distance fiber optic communication systems. Using extensive data on buried fibers used in long-haul high speed links, we discuss the proposition that most of the temporal PMD changes that are observed in installed routes arise primarily from a relatively small number of "hot spots" along the route that are exposed to the ambient environment, whereas the buried, shielded, sections remain largely stable for month-long time periods. It follows that the temporal variations of the differential group delay (DGD) for any given channel constitutes a distinct statistical distribution with its own channel-specific mean value. The impact of these observations on outage statistics is analyzed and the implications for future optoelectronic fiber-based transmission are discussed.
Abstract

We describe signal transmission and processing technologies for 160 Gbit/s OTDM systems, focusing especially on the time-domain optical Fourier transformation technique. Improvements in 160 Gbit/s OOK and DPSK transmission are described in detail.

1. Introduction

Motivated by the increasing demand for high capacity optical networks, many groups have performed 160 Gbit/s OTDM transmission experiments. Intensive efforts have recently been made to extend the transmission distance and increase the system margin by employing polarization multiplexing or advanced modulation formats such as DPSK. In this talk, we first review progress on ultrahigh-speed OTDM transmission. Then, we describe a new technique for distortion-free transmission employing the time-domain optical Fourier transformation (OFT) technique that we recently proposed [1].

2. Key technologies for ultrahigh-speed optical pulse transmission

In parallel with the increase in bit rate, various advanced modulation formats such as CS-RZ, DPSK, and DQPSK have recently been proposed to extend the transmission distance and increase the system margin. Figure 1 shows the way in which ultrahigh-speed optical transmission experiments have shifted from OOK to DPSK and DQPSK in which data are modulated in optical phase changes between bits and demodulated with a delay-line interferometer at the receiver. Since DPSK improves the SN ratio by 3 dB compared with OOK, more stable and long-distance transmission can be realized. In addition, the SN improvement allows transmission at a lower optical power, and thus it is possible to reduce nonlinear optical effects.

At bit rates faster than 40 Gbit/s, the signal pulse width is only a few ps. Therefore, the waveform distortion caused by chromatic dispersion and polarization-mode dispersion imposes a severe limitation on the transmission performance. In addition, even a small variation in dispersion, which is easily caused by environmental changes in the fiber, has a crucial effect on transmission performance. Adaptive dispersion equalization or the optical 3R (Retiming, Reshaping, Regeneration) technique will become the key approach to eliminating such time-varying dispersion.

3. Distortion-free transmission using time-domain OFT

3.1 Principle of time-domain OFT

We recently proposed a novel distortion-free transmission scheme that uses transform-limited pulses and time-domain optical Fourier transformation (OFT). With this technique, we take advantage of the fact that the spectral shape is maintained throughout the transmission, even when the fiber has linear perturbations, i.e., $|U(z,\omega)|^2 = |U(0,\omega)|^2$. The unchanged spectral profile is then converted into a time-domain waveform at the receiver by using the OFT technique. Time-domain OFT can be achieved ideally by employing a linear chirp (with a chirp rate $K$) and a GVD medium (with $D = 1/KL$). When $D = 1/K$, the output waveform after OFT, $v(t)$, becomes proportional to the spectrum before OFT, $U(z,\omega)$, and hence $U(0,\omega) :$

$$v(t) = \sqrt{\frac{i}{2\pi D}} \exp\left(\frac{-iKt^2}{2}\right) U(z, t/D)$$

$$\sqrt{\frac{i}{2\pi D}} \exp\left(\frac{-iK}{2}t^2 + i\beta(t/D)z\right) U(0,t/D)$$

where $\omega$ is converted to $t/D$. This enables us to reconstruct the original undistorted pulse shape.

3.2 160 Gbit/s-600 km OOK transmission

By applying this technique to 160 Gbit/s OTDM transmission, we successfully demonstrated a...
single-polarization, OOK transmission over 600 km [2]. The experimental setup is shown in Fig. 2(a). The optical pulse source was a 40 GHz mode-locked fiber laser (MLFL) emitting a 1.7 ps transform-limited Gaussian pulse train at 1550 nm. The 160 Gbit/s OTDM signal was transmitted over 8 spans of 75 km-long dispersion-managed fiber link. After the transmission, the 160 Gbit/s signal was demultiplexed to 40 Gbit/s by using an electro-absorption modulator (EAM) and launched into an optical Fourier transform circuit (OFTC). In the OFTC, the signal was first sinusoidally phase-modulated at an LN phase modulator (PM) with $K = -0.3$ ps$^{-2}$. The chirped signal then passed through a 36 m SMF. Finally the OFTC output was detected with a photo detector and the bit-error rate (BER) was measured. Figure 2(b) shows the BER curves versus received power after 524 and 600 km transmissions. Without OFTC, the BER curves had an error floor. The error floor at 524 km was suppressed by employing OFT, and the penalty at BER = $10^{-9}$ with respect to the back-to-back curve was reduced from 4.5 to 2.5 dB. At 600 km, there was a residual error floor even with OFT, but the power penalty was reduced by more than 2 dB.

3.3 160 Gbit/s-750 km DPSK transmission

In the next step, we applied the OFT technique to DPSK transmission and successfully extended the distance to 750 km. The experimental setup shown in Fig. 3(a) is similar to that in Fig. 2(a) except that a dual-drive push-pull Mach-Zehnder modulator driven by $2V_z$ was employed as a DPSK modulator, and a delayed interferometer (DI) and a balanced PD were included in the receiver. The BER performance after a 750 km transmission is shown in Fig. 3(b). The power penalty after a 600 km transmission was 3 dB, which is 3.5 dB lower than that with OOK. At 750 km, error-free transmission was successfully achieved with a penalty of 5.5 dB. By employing OFT, the power penalty was reduced to 3.5 dB. The improvement in the power penalty is attributed to the reduction of the jitter and pulse width, as shown in the inset of Fig. 3(b).

4. Conclusion

Recent progress on ultrahigh-speed OTDM signal transmission and processing technologies was described. A combination of the time-domain OFT technique and the DPSK format is particularly useful for extending the transmission distance and increasing the system margin in a 160 Gbit/s OTDM transmission.

References
Low power-penalty 160 Gbit/s DPSK transmission over 600 km

Kou Osawa, Masatada Okazaki, Toshihiko Hirooka, and Masataka Nakazawa
Research Institute of Electrical Communication, Tohoku University
2-1-1 Katahira, Aoba-ku, Sendai, 980-8577 Japan
Tel: +81-22-217-5525, Fax: +81-22-217-5524, E-mail: osawakou@riec.tohoku.ac.jp

Abstract Improved 160 Gbit/s-600 km OTDM transmission performance was realized by employing DPSK. The power penalty was only 3 dB. Error-free transmission over 750 km was also achieved with a penalty of 5.5 dB.

1. Introduction
Differential phase shift keying (DPSK) transmission has substantial advantages over on-off keying (OOK) in extending the transmission distance and increasing the system margin. DPSK enables us to obtain a 3 dB improvement in receiver sensitivity due to balanced detection. DPSK also leads to a reduction in optical nonlinear effects, since all the bit slots are occupied by an optical pulse and therefore the peak power can be reduced to one half that for OOK for a fixed average optical power. Another advantage of DPSK is that the decision threshold level can be fixed at a constant value independently of the signal power. This can simplify the system operation and offers long-term stability.

In particular, ultrahigh-speed OTDM transmission exceeding 40 Gbit/s benefits from DPSK because of these attractive features. Indeed, several long-haul OTDM transmission experiments employing DPSK have recently been reported at 160 Gbit/s and above, and substantial performance improvements have been realized. For example, 160 Gbit/s DPSK transmission over 650 km has been successfully achieved by adopting hybrid EDFA and Raman amplification in the transmission line [1]. The transmission distance has been extended to 4320 km in a recirculating loop configuration with the aid of polarization multiplexing and forward error correction (FEC) [2].

In this paper, we report 160 Gbit/s OTDM-DPSK transmission based on a simple configuration that does not employ Raman amplification or polarization multiplexing. A bit error rate (BER) of less than $10^{-10}$ was obtained at 600 km with a power penalty of only 3 dB (at BER = $10^{-9}$) without adopting FEC, which is 3.5 dB lower than that obtained with OOK. Error-free transmission over 750 km was also successfully demonstrated with a penalty of 5.5 dB.

2. Setup for 160 Gbit/s OTDM-DPSK transmission experiment
Figure 1 shows the experimental setup for 160 Gbit/s OTDM-DPSK transmission. The optical pulse source was a 40 GHz mode-locked fiber laser (MLFL) emitting a 1.7 ps transform-limited pulse train at 1550 nm. The pulse train was DPSK modulated at 40 Gbit/s with a $2^{15}-1$ PRBS. We used a dual-drive LiNbO₃ Mach-Zehnder modulator operated in a push-pull mode driven by $2xV_{s}$. The DPSK signal was optically multiplexed to 160 Gbit/s in a fiber delay-line multiplexer.

The OTDM signal was transmitted over 8~10 spans of 75-km dispersion-managed fiber link that consisted of standard single-mode fiber (SMF) and inverse dispersion fiber (IDF), in which the dispersion and dispersion slope were compensated simultaneously. The residual dispersion and dispersion slope were $-0.1 ~ -0.5$ ps/nm and $0.9 ~ 1.0$ ps/nm², respectively. The fiber loss was compensated at each span by using an EDFA. The average input power to each span was +6 dBm.

The DPSK receiver comprises a demultiplexer, a demodulator, and a detector. The transmitted 160 Gbit/s signal was first demultiplexed to 40 Gbit/s by using an electro-absorption modulator (EAM) with a switching window of 4.6 ps. The clock signal used for driving the EAM was obtained with an EAM-based 10 GHz clock recovery (CR) unit [3], whose frequency was upconverted to 40 GHz. The demultiplexed DPSK signal was then launched into a one-bit delay interferometer (DI) where the DPSK signal was converted to two complementary OOK signals that were output at two separate ports. Finally the two signals were detected with a balanced photodetector (PD) and the BER was measured.
3. Experimental result

The DPSK signal waveform and the optical spectrum before transmission and after 600 and 750 km transmissions are shown in Fig. 2 (a)-(c), respectively. The pulse width was broadened from 1.7 to 2.5 ps due to the residual dispersion, and the pulse suffered from jitter. Extending the transmission distance from 600 to 750 km does not result in significant degradation, indicating the high margin of the DPSK transmission.

Figure 3 shows the BER performance against the received optical power. The relation between the transmission distance and power penalty is shown in Fig. 4. The result with OOK is also shown for comparison. With OOK transmission, the penalty was as large as 6.5 dB already at 600 km, and the transmission distance could not be extended. With DPSK, on the other hand, the penalty was reduced to 3 dB at 600 km. When the length was extended to 750 km, the BER curve had an error floor, and the penalty was increased to 5.5 dB, but a BER of less than $10^{-10}$ was obtained for all the OTDM tributaries. A clear eye opening was obtained at the output of the balanced PD with a received optical power of –20 dBm.

![Fig. 2 Pulse waveform and optical spectrum of DPSK signal (a) before transmission and (b) after 600 km and (c) 750 km transmissions.](image)

![Fig. 3 BER measurement. The inset shows the detected eye diagram after a 750 km transmission, measured at the output of the balanced PD with a received optical power of –20 dBm.](image)

![Fig. 4 Change in power penalty with transmission distance.](image)

4. Conclusion

We have successfully demonstrated an error-free 160 Gbit/s OTDM-DPSK transmission over 600–750 km with a power penalty of 3–5.5 dB. Further improvement is expected by employing the time-domain optical Fourier transformation technique [4], which can eliminate transmission impairments caused by linear perturbations such as dispersion, dispersion slope and jitter.

References

10B2-3

100 Gb/s Ethernet over multimode fiber based on MIMO with spatial pre-coding

Maxim Greenberg, Student Member IEEE, Moshe Nazarathy, Senior Member IEEE, Member OSA, and Meir Orenstein

Abstract: We propose a novel MIMO scheme over multimode fiber, acting as a distributed random code generator fed by spatial codes, using silicon photonics in the transmitter and maximum-likelihood electronic detection in the receiver. Index Terms — Optical MIMO, Maximum likelihood decoding, Integrated Optics, Phase shift keying (PSK), Optical fiber communication.

I. INTRODUCTION

In light of the large-scale proliferation of 10 GbE (10G-BASE-SR) optical transmission in datacom applications, interest has recently peaked in extending the 802.3 suite of standards to 100 GbE over Multi Mode Fiber (MMF), currently envisioned to be realized by means of CWDM technology. Here we explore an alternative optical Multiple-Input-Multiple-Output (MIMO) transmission architecture for short-range 100 GbE over MMF, based on the paradigm that the MMF channel matrix random nature provides a unique opportunity to generate and manipulate random codes, reliably conveying vast amounts of information, in parallel over a single MMF interconnect. In this paper we extend and further improve the performance of our recently proposed novel optical technique [1] for massive parallel transmission over MMF. In effect we convert the fiber into an extended random code generator providing a closer-to-ideal novel opto-electronic realization of Shannon’s mathematical construct of random coding [2].

Our approach capitalizes on the statistical multi-path characteristics of the MMF propagation modes [3-4], as further clarified in our recent research [5] applied at the input of the optical MIMO system exerts a beneficial effect upon the quality of the random codes generated at the output side. Further to the novel optical realization of the communication-theoretic concepts of spatially coded signaling, the opto-electronic feasibility of our system is made possible at this juncture by advances in opto-electronic components, in particular silicon photonics [6], using potentially simpler optics, eliminating the multiple CWDM sources in favor of a single CW laser coupled to a silicon-based array of high-speed PSK waveguide modulators.

II. CONCEPT AND SYSTEM ARCHITECTURE

Consider a multimode fiber configured as a MIMO system, randomly coupling \( n_r \) modulated sources at its input and terminated in \( n_t \) detectors at its output (Fig. 1). The transmitter opto-electronic structure comprises \( n_t \) optical input ports synchronously modulated by binary PSK, each at bit-rate \( r = 12.5 \text{ Gb/s} \) (10 GbE + 2.5 Gb/s overhead), injecting the optical field complex amplitudes into the MMF, randomly coupled to the propagation modes of the MMF. The multiple PSK modulated optical sources are realized as silicon-based plasma-effect modulators – a recently emerging technology amenable to compact integration.

In each symbol interval \( T \), a \( k \)-bit message \( m = m_1m_2...m_k \in Z_2^n \) (with \( Z_2 = \{0,1\} \)), selected out of a space of \( M = 2^k = 2^n \) messages, is initially encoded as an \( n_r \)-dim. binary bitstring \( g = g_1g_2...g_\eta_p \in Z_2^{\eta_p} \) with \( n_r \geq k + 1 \), selected out of a Gilbert-Varshamov code \( G \) of specified minimum distance as discussed below. The input binary codewords are then mapped into bipolar complex field vectors \( E' \), with phase-modulated elements \( \pm 1 \) respectively launched into the fiber input ports.

The complex field vector incident at the \( n_t \) detectors is \( D = H E' \), with \( H \) the MMF random channel matrix of size \( n_t \times n_r \) and \( D \) the number of speckle/modal optical Degrees of Freedom (DOF) at each detector. Upon optical detection, the components of the received field are absolute-squared, summed over the DOFs and one-bit quantized with thresholds \( B \) selected offline such that \( \Pr(0) = 0.5 \)

Fig. 1. 100 Gbps MIMO over MMF with random coding naturally generated by MMF channel matrix statistics, assisted by G-V precoding.

Technical Digest, July 2007, Pacifico Yokohama
The detected \( n_g \)-bit codewords \( \mathbf{c} = c_1 c_2 \ldots c_{n_g} \) form realizations of a random code \( \mathbb{C} \subset \mathbb{Z}_{2}^{n_c} \) of size \( M \). As the channel matrix \( \mathbf{H} \) is random, so will be the code \( \mathbb{C} \langle \mathbf{H} \rangle \). End-to-end, we have synthesized a coded vector binary channel, \( \mathbf{r} = \mathbf{c} \oplus \mathbf{e} \), \( \mathbf{c} \subset \mathbb{C} \langle \mathbf{H} \rangle \), with \( \mathbf{e} \) the error vector induced by the thermal noise.

The receiver side electronic ML processing and the channel matrix estimation procedure illustrated in Fig. 1, were discussed in [1]. The main novelty here consists in the introduction of a transmit-side spatial code \( \mathbb{C} \subset \mathbb{Z}_{2}^{n_c} \) of size \( M \). Let \( n_r = k + 1 = 11 \) (including a reference input to remove PSK phase ambiguity). It was shown that the fiber acts as a distributed random code generator, with the modal fluctuations naturally generating random codewords \( \mathbf{c} \subset \mathbb{C} \) at the fiber output, however the code joint statistics was outside the designer’s control.

Here we significantly improve system performance by deliberately introducing spatial coding redundancy at the input side, replacing the uncoded input messages of [1] with a spatial code, \( \mathbb{C} \subset \mathbb{Z}_{2}^{n_c} \), requiring to increase the number of input ports to \( n_r > k + 1 \) in excess of the minimum necessary to carry the \( M \) messages. Such novel application of a controlled spatial redundancy at the input side embeds our \( k \)-dim. transmission message subspace into a higher \( n_r \)-dim. coupled with the natural random coding effect induced by the fluctuations of channel matrix, \( \mathbf{H} \). This architectural modification is shown here to substantially reduce the correlation between the randomly generated codewords at the fiber output, generating a code \( \mathbb{C} \subset \mathbb{Z}_{2}^{n_c} \) with nearly independent codewords, hence improving BER and/or link budget.

To be effective, the input spatial code introduced here requires that its codewords be sufficiently spread apart, i.e. their minimum normalized Hamming distance, \( \delta(n_r) = \min d_{H}(\mathbf{g}, \mathbf{g}) / n_r \), be equal to a substantial fraction, of the maximum possible distance \( n_r \). Let \( p_{BF}(\delta) = \Pr\{ d_{H} = 1 \} \) be the probability of a bit flip (BF) in any bit position between two output codewords, s.t. \( d_{H}(\mathbf{g}, \mathbf{g}) / n_r = \delta(n_r) \). The simulation presented in Fig. 2(a) demonstrates that \( p_{BF}(\delta(n_r)) \rightarrow 0.5 \) i.e. a perfect random code is asymptotically attained upon increasing \( \delta(n_r) \). Fig. 2(b) shows the Gilbert-Varshamov Sphere-Packing bound on \( \delta(n_r) \) [5]. Finally, Fig. 2(c) evaluates BER vs. the output detectors count, indicating that the error rate performance is successively improved upon increasing \( \delta(n_r) \), over the three analyzed cases \( \delta(n_r) = 1/11, 3/14, 7/21 \).

We conclude that the random codes naturally induced by the MMF channel matrix statistics, further assisted by spatial input coding with specified minimum distance, as proposed and analyzed here for the first time, enable ultra-high bitrate parallel transmission of 100 GbE, with manageable complexity, of \( n_r \) independently modulated signals all superimposed into the MMF at the same wavelength.

**Fig. 2.** (a) Optoelectronic multi-port phase modulator (\( n_c = 11, 14, 21 \) ports) based on Si photonics. (b) Bit-flip prob. vs. normalized min. distance \( \delta(n_r) \) defined as Hamming distance over \( n_r \). (c) BER performance for three systems with \( \delta(n_r) = 1/11, 3/14, 7/21 \) assuming raw per bit error prob. \( P_e = 0.01\).

**REFERENCES**

4. M.Greenberg, M. Nazarathy, M. Orenstein LEOS 2006 WXS.
Increased Transmission Bandwidth of Multimode Fiber by using Mode-Field Matched Center Launching Technique

D. H. Sim, Y. Takushima, and Y. C. Chung
Korea Advanced Institute of Science and Technology
Department of Electrical Engineering
373-1 Guseong-dong, Yuseong-gu, Daejeon 305-701, Korea
(Phone) 82-42-869-3456, (Fax) 82-42-869-3410, (E-mail) ychung@ee.kaist.ac.kr

Abstract  The mode-field matched center launching technique can drastically increase the transmission bandwidth of multimode fiber (MMF). In this paper, we evaluate the frequency response of the MMF link using this launching technique, and demonstrate the transmission of 10-Gb/s signal over 12.2 km of MMF.

I. Introduction
Recently, there have been growing interests in the high-speed multimode fiber (MMF) transmission for the use in the next-generation local area network (LAN). It has been reported that the maximum bit-rate and transmission distance of a MMF link depends highly on the launching conditions. Thus, extensive studies have been carried out to develop the proper launching techniques [1]-[3]. Recently, we have proposed the mode-field matched center launching technique for this purpose [4]. This technique requires matching the beam profile of the incident light to the fundamental mode of MMF. However, the mode field of single-mode fiber (SMF) matches well with the fundamental mode of MMF. Thus, this condition can be achieved simply by splicing the MMF to the SMF-pigtailed transmitter.

In this paper, we measure the frequency response of MMF after applying the input beam by using the mode-field matched center launching technique. The result shows that the frequency response of 12.2-km long MMF link did not roll off even at 10 GHz. Thus, by using the proposed launching technique, we could successfully demonstrate the 10-Gb/s transmission over 12.2 km of MMF.

II. Transmission bandwidth of the MMF link utilizing the mode-field matched center launching technique
It is well-known that a few lower-order modes of MMF can be selectively excited by launching the input beam at the center of MMF. However, to achieve the quasi single-mode transmission, it is necessary to match the input beam profile to the mode profile of LP01 mode (so that the incident power is dominantly coupled into LP01 mode). We designated this launching method as the “mode-field matched center launching technique” in our previous report [4]. To describe the fundamental mechanism of this launching method, we calculated the coupling efficiencies of several lower-order LP modes of a MMF as a function of the mode field diameter (MFD) of the input beam. Fig. 1 shows the result for a graded-index MMF with a core diameter of 50 μm. In this calculation, we used the index profile of MMF measured at 1304 nm, and assumed that the incident beam had a Gaussian profile. The result showed that, when the MFD was about 14 μm, most of optical power was coupled into LP01 mode. This was because the mode field of the launched beam matched well with LP01 mode of MMF. On the other hand, we noted that the MFD of the conventional SMF is in the range of 9 - 11 μm. Thus, it would be possible to achieve this mode-field matched launching condition simply by splicing the SMF to the MMF. For example, when we launched the light into the MMF through a conventional SMF, more than 80 % of the incident power could be coupled into LP01 mode.

The most important limiting factor on the transmission bandwidth of a MMF is the differential modal delay (DMD). However, by using the mode-field matched center launching technique, we can mitigate this limitation and improve the transmission bandwidth significantly. Although this technique can also excite some higher-order modes propagating in different group velocities, their total power is much smaller than that of LP01 mode. Because of their small power, these higher-order mode components broaden only the pedestal of the pulse, which can be resulted in the small incoherent crosstalk to the LP01 mode component.

To evaluate the effectiveness of the mode-field matched center launching technique, we measured the frequency response of MMF using an intensity-modulated small signal. In this experiment, we sinusoidally modulated the laser output by using a SMF-pigtailed LiNbO3 modulator, and launched the modulated light into OM-3 MMF. The MFD of SMF was 9.3 μm. The lateral offset between the MMF and SMF was adjusted by using a precision motion controller. We measured the amplitude of the modulation component as a function of frequency and lateral offset. Fig. 2(a) shows the frequency response measured after transmission of 1-km long MMF for different offset conditions. In the cases of using the offset launching techniques (10 and 16 μm in Fig. 2(a)), the measured frequency responses had low-pass characteristics, and the bandwidth decreased with the offset due to the DMD among the excited higher-order modes. On the other hand, when the mode-field matched center launching technique was used, the bandwidth limitation was moderate. For further understanding, we repeated the same measurement by using 12-km long MMF. Fig. 2(b) shows the result. Unlike in the case of using the offset launching technique (offset = 10 μm),

Fig. 1. Calculated coupling efficiency by varying the MFD of launched beam.
the frequency response did not roll off even at 10 GHz when we used the mode-field matched center launching technique. This result indicates that a long-distance MMF transmission at the speed higher than 10 Gb/s would be possible by using the mode-field matched center launching technique. However, although the measured frequency response did not roll off, it had non-negligible fluctuations. These fluctuations were caused by the modal interference between LP_{01} mode and higher modes, and could result in the penalties in the receiver sensitivity. In this experiment, LP_{01} was dominant among the excited higher modes. Thus, the period and depth of the fluctuation were determined mainly by the DMD and power ratio between LP_{01} and LP_{m2} modes.

III. 10-Gb/s MMF transmission experiments

Using the mode-field matched center launching technique, we demonstrated the transmission of 10-Gb/s signal over up to 12.2 km of commercial grade MMF. Fig. 3 shows the experimental setup. We modulated the output of the DFB laser operating at 1304 nm with 10-Gb/s non-return-to-zero (NRZ) signal by using a LiNbO_3 modulator. For the mode-field matched center launching, we simply spliced the output fiber of the modulator to the MMF. The lateral offset between these two fibers was less than ±1 μm after splicing. The core diameter and the overfilled launch (OFL) bandwidth of the MMF were 50 μm and 700 MHz-km, respectively. We measured the BER curves of the 10-Gb/s signal after the transmission over 5.6 km and 12.2 km of this MMF. It should be noted that, when the offset launching technique is used, the maximum transmission distance of this MMF is limited to only about 300 m at 10 Gb/s [4]. At the receiver, we detected the whole power from the MMF by using a free-space coupled photodetector. We did not use any mode-stripping devices to eliminate the effects of higher-order mode. The back-to-back receiver sensitivity was -18 dBm (@ BER = 10^-9). Fig. 4 shows the measured BER curves. The power penalty was only 1.6 dB when we transmitted the 10-Gb/s signal over 5.6 km of MMF. The power penalty remained to be almost same (1.7 dB), even when we increased the transmission distance to 12.2 km. To our knowledge, this result represents the longest transmission distance ever achieved in the MMF link using 10-Gb/s signals. We attributed the small power penalties in Fig. 4 to the ripples in the frequency response of MMF. As described in the previous section, the frequency response of the MMF did not roll off even when the fiber length was as long as 12.2 km. Thus, the power penalty should not increase with the transmission distance until the mode coupling to higher-order modes becomes significant. We also found that the performance of this MMF link was not sensitive to external perturbations such as bending.

IV. Summary

We have demonstrated that the transmission bandwidth of MMF could be drastically increased by using the mode-field matched center launching technique. This was because we could selectively excite the fundamental mode of MMF by using this technique and avoid the limitation imposed by DMD. As a result, when we used this launching technique, the frequency response of the 12.2-km long MMF did not roll off even at 10 GHz. As a result, we could transmit 10-Gb/s NRZ signal over 12.2 km of MMF without using any dispersion compensation. To our knowledge, this result represents the longest transmission distance achieved in the MMF link using 10-Gb/s signals.

References

10 Gb/s-2 km photonic crystal fiber transmission at 850 nm with a directly modulated single-mode VCSEL

Hideaki HASEGAWA, Yosuke OIKAWA, and Masataka NAKAZAWA
Research Institute of Electrical Communication, Tohoku University, 2-1-1, Katahira, Aoba-ku, Sendai-shi 980-8577, Japan E-mail: hasehide@riec.tohoku.ac.jp

Abstract
We report a 10 Gb/s transmission at 850 nm using a 2 km-long single-mode PCF and a directly modulated single-mode VCSEL. Power penalty-free transmission has been successfully achieved without an optical amplifier or a modulator.

Introduction
Recently, photonic crystal fiber (PCF), which has a core region surrounded by multiple air holes, has received a lot of attention because of its novel characteristics such as endlessly single-mode operation and arbitrary dispersion control [1], [2]. A zero dispersion shift toward the shorter wavelength region is particularly interesting in terms of realizing high-speed optical communication in a new wavelength region. For example, the 800 nm region is attractive, since inexpensive and high-speed optical devices such as an AlGaAs vertical cavity surface emitting laser (VCSEL) and Si-avalanche photo diode (APD) are available. A VCSEL has many advantages over edge emitting lasers including a low threshold current, a single longitudinal-mode, and high-density integration, due to its two-dimensional structure and short cavity length [3], [4]. By combining VCSEL, Si-APD and PCF, we can construct simple and high-speed communication systems such as 10 Gb/s Ethernet. Recently, a 10 Gb/s single-channel transmission over 5 km at 850 nm was demonstrated using PCF [5], [6]. However, these experiments employed an LN (LiNbO3) modulator and an erbium-doped fluoride fiber amplifier (EDFFA).

In this paper, we report a 10 Gb/s transmission over 2 km with an 850 nm directly modulated single-mode VCSEL with a view to realizing a simple and high-speed transmission system. We also describe the influence of chirp in a directly modulated VCSEL on the transmission characteristics.

High-speed characteristics of a directly modulated single-mode VCSEL
We used an oxide-confined VCSEL with an oxide aperture diameter of 3.5 μm in our transmission experiment. The VCSEL has single transverse and longitudinal-modes. The side-mode suppression ratio was more than 40 dB.

Figure 1 (a) shows the light output and voltage versus current (L-I-V) characteristics of the single-mode VCSEL. The threshold current and voltage were 0.4 mA and 1.6 V, respectively. The maximum output power was 3.4 mW. The slope efficiency and device resistance were 0.87 W/A and 200 Ω at an injection current of 3.0 mA, respectively. The VCSEL was TO-packaged with a 50 Ω impedance SMA connector. Figure 1 (b) shows the measured small signal modulation response of the VCSEL. The maximum 3 dB bandwidth was 6.3 GHz at a current of 3.0 mA, which was sufficient to modulate a 10 Gb/s NRZ signal. The relaxation oscillation frequency (ROF) of the laser estimated from the measured relative intensity noise (RIN) spectrum was 11.7 GHz at a driving current of 3 mA. The ROF was much higher than the maximum modulation bandwidth. On the other hand, the total capacitance C and resistance R of the TO-packaged VCSEL measured with an impedance analyzer were 0.4 pF and 50 Ω, respectively. The cutoff frequency f_c given by

\[ f_c = \frac{1}{2\pi CR} \]  

was 6.4 GHz. The calculated value agrees well with the measured modulation bandwidth. This means that the cutoff frequency originating from the parasitic components of the VCSEL limits the maximum bandwidth.

In our experiment, the bias and modulation currents chosen for the best bit error rate (BER) performance under a back-to-back condition were 3.0 and 2.5 mA, respectively. Under this modulation condition, the chirp parameter \( \alpha \) estimated from the chirp to modulated power ratio (CPR) [7] was 4, which is similar to that of a distributed-feedback (DFB) laser at 1.55 μm [8].

---

Fig. 1 (a) L-I-V characteristics and (b) the measured small signal response of the single-mode VCSEL.
10 Gb/s transmission experiment at 850 nm

We used a PCF fabricated by the capillary method. The air-hole pitch \( \Lambda \), air-hole diameter \( d \), \( d/\Lambda \) and core diameter \( 2\Lambda - d \) of the PCF were 3.4 \( \mu m \), 1.2 \( \mu m \), 0.35 and 5.6 \( \mu m \), respectively. Since the \( d/\Lambda \) value is less than 0.43, this PCF has an endlessly single-mode property [9]. The zero dispersion wavelength was 1098 nm. The measured dispersion value \( D \) of the PCF and a conventional step-index fiber (SIF) at 850 nm were -62.8 and -97.5 ps/nm/km, respectively. We calculated the dispersion distance \( L_D \) defined as

\[
L_D = \frac{\alpha + \sqrt{1 + 2 \alpha^2}}{1 + \alpha^2} \frac{\alpha T_0^2}{2 \ln 2 L D}
\]  

(2)

to estimate the maximum transmission distance. Here, \( c \) is the velocity of light in a vacuum and \( \lambda \) is the transmission wavelength. \( L_D \) corresponds to the distance at which the pulse width broadens to the twice the input as a result of chromatic dispersion. When the pulse width \( T_0 \) is 50 ps, roughly corresponding to a bit rate of 10 Gb/s signal, \( L_D \) is 21.4 km. For a 2 km transmission, which is much shorter than \( L_D \), the influence of dispersion is negligible.

Figure 2 shows the experimental setup for a 10 Gb/s transmission at 850 nm. The single-mode VCSEL was directly modulated by a 10 Gb/s NRZ signal with a \( 2^{15}-1 \) PRBS length. An input signal with a power of 2 dBm from the VCSEL was coupled into the PCF with a coupling efficiency of 70%. The signal was transmitted through a 2 km-long PCF with a loss of 5.2 dB/km. The output signal was detected with a Si-APD, and the BER performance was measured. Figure 3 (a) shows the BER performance after a 2 km transmission without an EDFFA. The BER performance was almost power penalty-free. This means that PCF dispersion had negligible influence on the transmission characteristics. We also undertook 3 and 4 km transmission experiments using an EDFFA as a pre-amplifier. Figure 3 (b) shows the BER performance after 3 and 4 km transmissions. Figure 3 (c) shows the eye patterns of the input and output signals after a 4 km transmission. In Fig. 3 (b), there was no power penalty for a 3 km transmission. However, for a 4 km transmission, there was a power penalty of 0.8 dB at a BER of \( 10^{-9} \). Figure 3 (c) shows the eye patterns measured at an optical signal-to-noise ratio (OSNR) of 25 dB. The distortion in the eye pattern after the 4 km transmission was caused by the chirp-induced pulse broadening, resulting in the power penalty.

Conclusion

We have demonstrated a 10 Gb/s transmission at 850 nm by using a single-mode PCF and a directly modulated single-mode VCSEL. The influence of the PCF dispersion started to occur at a transmission distance of 4 km. This result indicates that simple, inexpensive and high-speed transmission over short distances appears to be feasible by combining these devices.

References

Recent Progress on Silica-based Highly Nonlinear Fiber
Masaaki Hirano, Tetsuya Nakanishi, Toshiaki Okuno and Masashi Onishi
Sumitomo Electric Industries, Ltd., 1, Taya-cho, Sakae-ku, Yokohama, 244-8588 Japan
(E-mail: masahirano@sei.co.jp)

Abstract: Advantages of highly nonlinear fibers with silica-based glass against those with other glass materials are explained from the practical viewpoint, and recent progresses on silica-based highly nonlinear fibers in our research group are described.

1. Introduction
Optical fibers are essential for current information societies as signal transmission media, and are also quite attractive as functional devices. Especially, the 3rd-order optical nonlinear effects generated in fibers, such as self phase modulation (SPM), cross phase modulation (XPM), four-wave mixing (FWM) and stimulated Raman scattering (SRS) have been tried to be applied to all-optical devices. In this area, highly nonlinear fibers were developed and have been actively studied for realization of such applications as optical amplifications, optical signal processing, optical monitoring and supercontinuum light sources.

In this paper, advantages of silica-based highly nonlinear fibers (HNLFs) are explained compared to highly nonlinear fibers with other materials from the viewpoint of practical use. Then recent progresses on HNLFs and their applications in our research group are described.

2. Silica-based highly nonlinear fibers
HNLF has a nonlinear coefficient (γ) more than ten times larger than that for a standard single mode fiber (SMF). Here, γ is a parameter indicating how efficiently nonlinear optical effects are generated in the fiber. To our knowledge, the highest γ for HNLF is 30W/km3. Its refractive index profile is shown in Fig. 1, consisting of a GeO2-doped core, a fluorine-doped depressed cladding and an outer cladding. The relative refractive index difference of the core to the outer cladding (Δn) is 3.5%.

![Fig. 1 Refractive index profile of HNLF with γ of 30W/km^3](image)

Recently, holey fibers (HF) and nonlinear fibers using non-silica glass materials have been proposed because of their 1 to 2 orders of magnitude higher γ as compared to that of HNLFs. Their characteristics are listed in Table 1.

<table>
<thead>
<tr>
<th>Fiber Type</th>
<th>γ [W/km]</th>
<th>Transmission Loss [dB/km]</th>
<th>Dispersion [ps/nm/km]</th>
<th>Coupling Loss [dB/endpoint]</th>
</tr>
</thead>
<tbody>
<tr>
<td>HNLF</td>
<td>30</td>
<td>1.3</td>
<td>0</td>
<td>0.1</td>
</tr>
<tr>
<td>Silica HF</td>
<td>70</td>
<td>190</td>
<td>0</td>
<td>---</td>
</tr>
<tr>
<td>Bi2O3 fiber</td>
<td>1100</td>
<td>800</td>
<td>-260</td>
<td>2.9</td>
</tr>
<tr>
<td>Bi2O3 HF</td>
<td>730</td>
<td>1800</td>
<td>-15</td>
<td>---</td>
</tr>
<tr>
<td>TeO2 HF</td>
<td>675</td>
<td>400</td>
<td>5</td>
<td>---</td>
</tr>
<tr>
<td>Pb-Silicate HF</td>
<td>1860</td>
<td>3000</td>
<td>210</td>
<td>---</td>
</tr>
<tr>
<td>As2Se3 HF</td>
<td>1200</td>
<td>1000</td>
<td>-504</td>
<td>2.9</td>
</tr>
</tbody>
</table>

Even with such a smaller γ, HNLFs have great advantages for practical use as listed below, thank to application of the matured technologies used in a silica-based SMF manufacturing.

1/ Low transmission loss (long effective length)
2/ High controllability for chromatic dispersions
3/ Low splicing loss to a SMF, about 0.1dB/splice
4/ High environmental reliability and durability

To clarify the advantages addressed in 1/ low transmission loss and 2/ high dispersion controllability, FWM-based wavelength conversion efficiency (ηFWM) for each highly nonlinear fiber was calculated. For high ηFWM, the phase-matching condition should be realized and the phase-mismatching parameter of Δβ is 0.

\[ \Delta \beta = \beta_2 \cdot \frac{4 \pi^2 c^2}{\lambda_p} \left( \frac{1}{\lambda_p} - \frac{1}{\lambda_S} \right)^2 - \beta_4 \cdot \frac{4 \pi^4 c^4}{3} \left( \frac{1}{\lambda_p} - \frac{1}{\lambda_S} \right)^4 \]

Here, \( \lambda_p \) and \( \lambda_S \) are the pump and the signal wavelength, respectively, and \( \beta_2 \) and \( \beta_4 \) are the 2nd and the 4th-order dispersions at the \( \lambda_p \), respectively. In addition, the ηFWM is known as a function of (γ × pump power × effective length)2. Figure 2 shows the calculated ηFWM when the (\( \lambda_p + \lambda_S \)), the \( \beta_4 \) and the pump input power is assumed to be 5nm, 0.7W/m and 0dBm, respectively. It is clearly shows in Fig. 2 that HNLF has the larger ηFWM in the hundreds meters long range due to its low transmission loss and small Δp. On the other hand, ηFWM for other nonlinear fibers have much smaller maxima at less than 100m long, as compared with hundreds meters ηFWM of long HNLF.

3. Zero-dispersion wavelength uniformity
Zero-dispersion wavelength (\( \lambda_0 \)) uniformity in the longitudinal direction is very important, because efficiency and bandwidths for generation of nonlinear effects are very sensitive to the dispersion at the \( \lambda_0 \) and relation between \( \lambda_0 \) and \( \lambda_p \). Today, we can control the \( \lambda_0 \) within +/- 1nm over several-km long owing to progresses in fabrication processes, as shown in Fig. 3.

![Fig. 3 Longitudinal zero-dispersion wavelength uniformity](image)
4. β₄ design for FWM-based wavelength conversion

For an efficient FWM generation, Δβ in eq.(1) should be 0 as described above. The 4th-order dispersion of β₄ had been little considered in the actual HNLF fabrication, but is quite important to functional wavelength conversions. For this purpose, we have designed and actually fabricated HNLFs having appropriate β₄ values, as listed in Table 2. In this table, HNLF-A is a typical HNLF with γ of 20/W/km.

| Table 2 Characteristics of β₄ improved HNLF at 1550nm |
|-----------------|-----------------|-----------------|-----------------|
| HNLF-A 11 | 20 | 0.5 | -1×10⁻⁶ | 1559 | +0.035 |
| HNLF-B 9.4 | 25 | 1.0 | +2×10⁻⁶ | 1562 | +0.019 |
| HNLF-C 12 | 18 | 1.3 | -2×10⁻⁵ | 1528 | +0.049 |

4-1. Ultra broadband wavelength conversion

In order to realize a simultaneous wavelength conversion of multiple signals over a broad wavelength range, Δβ should be 0 in broad λₑ range. Therefore, from eq.(1), |β₄| should be small. We fabricated a HNLF with decreased |β₄| of 2×10⁻⁶ s/m (HNLF-B in Table 3) that is one-fifth of a conventional HNLF-A. Evaluated normalized ηFWM spectra using 100m long HNLFs are shown in Fig. 4. Due to small |β₄|, HNLF-C has broad bandwidth of 222nm in 100m long. Conversion bandwidth were also evaluated in the fiber-length of 100 to 1000m, and we confirmed that the bandwidths are extending those with previously reported works12,13 to a factor of about 2 in any length.

Fig.4 Normalized conversion efficiency spectra

4-2. Selective wavelength conversion

On the other hand of broadband wavelength conversion, there are growing demands for wavelength selective devices from multiple signals in the optical processing. It was derived from eq.(1) that there would be a specific range of signal wavelength with high efficiency of FWM generation. This will occur in the wavelength region apart from the pump wavelength λₚ where the value balance between β₂ and β₄ is optimized. Therefore, we have fabricated HNLF with increased |β₄| of 2×10⁻⁵ s/m (Fiber-C in Table-2)14. Using Fiber-C, we evaluated FWM and successfully realized wavelength selective and tunable conversion by shifting the pump wavelength λₑ (Fig. 5).

Fig.5 Spectra of normalized conversion efficiency

6. SBS suppression in HNLF

Efficiency of nonlinear effects is enhanced as the higher launched pump power to HNLFs. Stimulated Brillouin scattering (SBS), however, limits the launched power. SBS is also one of nonlinear effects, and then the SBS threshold power for a HNLF is about 10 times lower than that for a SMF. A HNLF with improved SBS threshold has been strongly expected, but was very challenging while keeping γ value. Recently, we have fabricated an Al₂O₃ - SiO₂ cored highly nonlinear fiber (AI-HNLF)15, and the characteristics are summarized in Table 3. It is successfully realized high γ, λₒ shifted in the C-band and low PMD. Its transmission loss is still high, which is expected to be improved to a comparable level as that of a conventional HNLF.

| Table 3 Characteristics of Al₂O₃ - SiO₂ cored HNLF at 1550nm |
|-----------------|-----------------|-----------------|-----------------|
| Al-HNLF 11.5 | 19 | 37 | 0.03 | 1544 | +0.027 |

We experimentally confirmed in Fig. 6 that a 100m long Al-HNLF has 8.9dB higher SBS threshold than that for a conventional HNLF. This includes influences of the higher transmission loss and structural fluctuation. We also evaluated Brillouin gain spectra with 1m long fibers, and we confirmed that the SBS in Al-HNLF is intrinsically improved by +6.1dB.

Fig.6 SBS threshold power for HNLFs

7. Summary

Recent progresses on HNLFs are reviewed. Although HNLFs are classical silica-based fibers with a simple profile, new applications have been continuously proposed. Very recently, actual commercializing of optical sampling oscilloscopes using HNLFs is ongoing, and greater uses of HNLFs for future key devices are strongly expected.

Acknowledgement We would like to thank Dr. Inaba and Dr. Onae in National Institute of Advanced Industrial Science and Technology (AIST) for FWM measurement.

References

6. A. Mori et al., ECOC2004, Th.3.3.6, (2004).
Nonlinearity and Dispersion Characteristics of Bismuth-based Highly Nonlinear Fiber

Naoki Sugimoto, Tatsuo Nagashima, Tomoharu Hasegawa and Seiki Ohara
Research Center, Asahi Glass Company, 1150 Hazawa-cho, Kanagawa-ku, Yokohama 221-8755, Japan
naoki-sugimoto@agc.co.jp

Abstract  Bismuth-based PCF was fabricated to obtain high optical nonlinearity and low group-velocity-dispersion. The fiber structure was designed to have moderately low GVD. It was experimentally revealed that this PCF shows $\gamma \sim 780 \text{ W}^{-1}\text{km}^{-1}$ and GVD $\sim -25 \text{ ps/nm/km}$.

Introduction
High nonlinear optical fibers have attracted much attention because of their various applications in ultrafast optical processing and nonlinear optical devices. The conventional silica fibers are widely used as a standard nonlinear medium. However, the third order optical nonlinearity of silica is intrinsically low, and long fiber length is required to obtain a sufficient nonlinear phase shift. Since the third order optical nonlinearity of glass material strongly depends on the linear refractive index (Miller's rule), several types of glass materials, such as, lead-silicate, chalcogenide glass, tellurite glass systems have been developed to obtain a high refractive index [1-3]. High fiber nonlinearity can be obtained by applying these high refractive index glasses as nonlinear medium [4, 5]. It is known that Bi$_2$O$_3$-based glass shows remarkably high refractive index $n > 2$ at 1550 nm. We have succeeded to fabricate bismuth-based nonlinear optical fiber (Bi-NLF), and reported that Bi-NLF shows significantly high optical nonlinearity $\gamma \sim 1000 \text{ W}^{-1}\text{km}^{-1}$. On the other hand, group-velocity-dispersion (GVD) of Bi-NLF is as large as -300 ps/nm/km [6]. So the applications using Bi-NLF is limited. High refractive index glasses generally show large dispersion due to electronic transition in UV ~ VIS region. GVD is the summation of material dispersion $D_m$ and waveguide dispersion $D_w$, therefore, Photonic crystal fiber structure is favorable to GVD management. In this paper, we report on the fabrication and optical measurements in bismuth-based photonic crystal fiber (Bi-PCF). The comparison with Bi-NLF shows the drastic effect of GVD reduction induced by PCF structure.

Table 1. B036 and B037 properties

<table>
<thead>
<tr>
<th>Bi$_2$O$_3$ [mol%]</th>
<th>$n$</th>
<th>$\chi_{(3)}$ [C]</th>
<th>Thermal expansion coefficient</th>
</tr>
</thead>
<tbody>
<tr>
<td>B036 (Cladding)</td>
<td>53.5</td>
<td>2.111</td>
<td>370</td>
</tr>
<tr>
<td>B037 (Core)</td>
<td>57</td>
<td>2.139</td>
<td>368</td>
</tr>
</tbody>
</table>

Table 2. Obtained structural parameters

<table>
<thead>
<tr>
<th>Core diameter (CD)</th>
<th>Inscribed circle diameter (ID)</th>
<th>Cladding diameter</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.5~1.6 $\mu$m</td>
<td>2.7~2.8 $\mu$m</td>
<td>125 $\mu$m</td>
</tr>
</tbody>
</table>

Fabrication Process
We have developed bismuth-based glasses B036 and B037. The refractive indices of B036 and B037 are shown in Tab. 1. The refractive index is determined by Bi$_2$O$_3$ concentration. The fabricated Bi-PCF has core of B037 and cladding of B036 with air holes (Fig. 1). The B037 core guides the propagating light to prevent mode leak at the fusion splicing point. At the fusion splicing point, air holes collapse because of surface tension. In the fabrication process, at first, the fiber preform was fabricated by redrawing the B037 rod and B036 cladding tube. Six holes were drilled in B036 cladding tube before redrawing. In redrawing process, the glass preform experienced three times of heating cycle, therefore, the thermal properties of these glasses were well adjusted shown in Tab.1. The air pressure is applied to inflate the air holes in redraw processes [7]. The SEM image of fabricated Bi-PCF is shown as Fig.2 and structural parameters are listed in Tab.2.

We performed a fusion splicing on Bi-PCF fabricated in this work. The best splicing loss 4.6 dB/point was obtained and averaged splicing loss was 6.8 dB/point. The ideal loss is predicted to be less than 2 dB/point if we assume an adiabatic mode conversion. We consider that the large extra splicing Fig.1 Schmatic picture of Bi-PCF preform with six holes.

Fig.2 (a) SEM image of Bi-PCF.  (b) Structural parameters.
loss could attribute to the mode-leak. In fact, we observed hole collapsing at the fusion splicing point, and the length of hole collapsing area ~ 20 μm is too short to make mode-field conversion adiabatic [8].

**GVD and Nonlinearity of Bi-PCF**

The core diameter (CD) of Bi-PCF was designed to have moderate normal dispersion ~ 25 ps/nm/km. The full-vector BPM was employed to obtain GVD in PCF structure. GVD of PCF was experimentally obtained by a homodyne interferometric method (Agilent 81910A). In Fig. 3, the experimental result is compared with Bi-NLF of step-index structure (GVD~300 ps/nm/km). $D_m$ of B037 is estimated to be ~ -170 ps/nm/km, from the wavelength dependence of refractive index. GVD is obviously well reduced by PCF structure. On the other hand, Bi-NLF shows significantly large GVD mainly due to large $D_m$ ~ -210 ps/nm/km. We can estimate the fiber nonlinear coefficient $\gamma$ of Bi-PCF by four-wave-mixing (FWM) measurement. Fig. 4 shows the results of FWM experiment and its wavelength detuning dependence in 97 cm-long Bi-PCF. The signal wave is converted into FWM wave via third order nonlinear process induced by pump wave. It should be noted that the significant FWM is observed in short Bi-PCF. From power conversion efficiency, we can calculate the nonlinearity of Bi-PCF as $\gamma \sim 780 \text{ W}^{-1}\text{km}$. This nonlinearity shows good agreement with the prediction of Miller’s rule which describes the relationship between the linear and nonlinear refractive index of glasses. This wide wavelength tunable range is the result of well reduced GVD.

**Conclusion**

We fabricated Bi-PCF which shows low GVD and high nonlinearity simultaneously. The guiding core of high refractive index glass B037 enables us fusion splice easily. The extra loss due to mode-leak must be improved to construct more efficient nonlinear devices.

**References**

7. T. Nagashima, et al., ECOC2006, We1.3.2.

![Fig. 3 GVD of Bi-PCF and Bi-NLF](image)

![Fig. 4 FWM observed in 97 cm-long Bi-PCF.](image)
Highly nonlinear non-silica fibers for efficient slow light devices

Kazi Sarwar Abedin, Guo-Wei Lu, and Tetsuya Miyazaki
Photonic Network Group, National Institute of Information and Communications Technology
4-2-1, Nukui-Kitamachi, Koganei, Tokyo 184-8795, Japan
Tel: +81-42-327-6792, Fax: +81-42-327-7035, Email: abedin@nict.go.jp

Abstract
We report the superior Brillouin scattering properties of single mode tellurite and chalcogenide fibers. We found large Brillouin gain coefficients, one to two orders of magnitude larger than silica fibers, and demonstrated the suitability of these fibers for efficient generation of slow light in 1.55 μm-wavelength region.

1 Introduction
Optical buffers with variable delay are highly desirable for avoiding packet contention within the routing nodes of the communication system. Among the various ways reported so far for controlling the speed of light, such as electromagnetically induced transparency, coherent population inversion, slowing of light via stimulated Brillouin scattering (SBS) in optical fiber [1, 2] is the most attractive for optical communication systems, because of the ease in incorporation with fiber-optic systems and operability at an arbitrary wavelength and in room temperature. As the optical group delay is linearly proportional to the exponential Brillouin gain, fibers with large Brillouin gain coefficient is highly desirable to reduce the pump power and also the device length.

In this paper, we report SBS properties of single mode chalcogenide and tellurite glass fibers. Large Brillouin gain coefficients, about 1–2 orders of magnitude larger than that of silica fibers, were measured in tellurite and chalcogenide fibers that allowed efficient generation of slow light in fibers of only a few meter in length.

2 Experiment
We first investigated the SBS properties of chalcogenide and tellurite glass fibers. The As$_2$Se$_3$ chalcogenide fiber [3] used in the experiment had a core diameter of 6 μm and a NA of 0.18, which allowed the single mode propagation in 1.55 μm wavelengths region with a transmission loss of about 0.84 dB/m. The tellurite fiber used here was a single-mode fiber with a core diameter of 2.6 μm, Δn of 1.6%, and cutoff wavelength of about 1.3 μm. The fiber was developed particularly for use in L-band amplifiers for which the core was doped with erbium with a concentration of about 1000 ppm (wt%). The transmission loss in the Er-doped tellurite fiber at saturation was measured to be 0.51 dB/m at 1542 nm.

To study the Brillouin scattering, cw light with a wavelength of ~1.55 μm was launched into a 2–5-m-long fibers though an optical circulator, and the backscattered radiation from the fiber was observed with an optical spectrum analyzer. The intensity of the Stokes component at different pump power levels could be directly read from the optical spectrum analyzer. Brillouin shift $\nu_B$ was determined with a high resolution using heterodyne detection of the beat signal produced by the pump and the Stokes wave.

The small-signal SBS gain was determined by measuring the intensity change of a probe signal that was launched into the tellurite/AsSe glass fiber in a direction opposite to the strong pump wave as shown in Fig. 1 [3]. The probe signal was created from the output of the same laser used for pumping, by generating optical sidebands using an electro-optic phase modulator. The change in the intensity of the lower frequency sideband (Stokes wave) was measured using the optical spectrum analyzer while the modulation frequency ($f_m$) was varied over a frequency range of ~200 MHz centered on the Stokes frequency shift $\nu_B$.

Fig. 1 Experimental setup used for measuring Brillouin gain.

In order to demonstrate effectiveness of these fibers for slowing light, continuous wave laser radiation at ~1.55 μm from an external cavity semiconductor laser was
divided into two parts. One part was amplified for use as a cw pump, while the rest was phase-modulated using an RF oscillator at a frequency equal to the Brillouin shift $v_B$ in the fiber. The output from the phase modulator was then intensity-modulated using a Mach-Zehnder modulator to produce Gaussian probe pulses with about 50–60 ns width at a 1-MHz-repetition-rate, and launched into the fibers in a direction opposite to that of the pump. The waveform at the output of the fiber was detected at different pump powers to measure the delaying of pulse.

3. Results

![Graph showing small signal gain as a function of frequency shift](image)

Fig. 2. Small signal gain as a function of the frequency shift ($f_m$) measured in a 5-m-long chalcogenide fiber. Pumped power is 68 mW.

![Graph showing intensity of probe signal](image)

Fig. 3. Intensity of the probe signal at the fiber output as a function of the frequency shift ($f_m$) measured in a 2-m-long Tellurite fiber. Pump power is 360 mW.

Figure 2 shows the small signal gain as a function of the frequency shift from the pump wave, as observed in chalcogenide and tellurite fibers. A maximum Brillouin gain of 42 dB was obtained when the modulation frequency was 7.958 GHz, which corresponded a gain of 0.62 dB/mW of pump power. In tellurite fiber, as shown in Fig. 3, a peak Brillouin gain of 30 dB was obtained at a modulation frequency of 7.882 GHz using a 2-m-long fiber and a pump power of 360 mW, which corresponded to a gain of 0.083 dB/mW of the pump power.

Fig. 4 shows the waveforms, normalized to the peak power, of the output probe pulses from a 2-m-long tellurite fiber observed at different launched power. A shift in the peak by as much as 67 ns (1.1 times the pulselength) was measured for the 60-ns pulses, as the pump was raised from 0 to 630 mW (a gain change from 0 to 52.5 dB).

![Waveforms of probe pulses](image)

Fig. 4. Waveforms of probe pulses showing pulse delaying for different amount of pump powers in a 2-m-long tellurite fiber.

From similar experiments performed with 5-m-long AsSe chalcogenide fiber, a pulse delaying by 37 ns with a pump power of 60 mW was observed [4]. While AsSe fiber is found to be the most efficient for low power operation, tellurite fiber is attractive for its relatively low loss. Tellurite fiber with a background loss as low as 0.054 dB/m (at 1.2 μm) was reported for similar fibers [5], which suggests that successful drawing of low-loss undoped tellurite fibers will make it possible the realization of compact slow light fiber devices with improved performances.

3. Conclusions

In conclusion, we have reported large Brillouin gain coefficient in single mode chalcogenide and tellurite fiber and demonstrated its application for efficient generation of slow light. A delay of 67 ns (pulse width 60 ns) was achieved in a 2 m-long tellurite fiber (pump power: 630 mW) exhibiting a delay generation efficiency about half of what was achieved in a 5m-long chalcogenide fiber. Fibers with large Brillouin gain coefficients as we have demonstrated here are expected to be highly useful for realization of compact slow light devices.

4. References

Resonant Coupling in Photonic Crystal Fibers

Kunimasa Saitoh, Shailendra K. Varshney, and Masanori Koshiba
(Graduate School of Information Science and Technology, Hokkaido University, Sapporo 060-0814, Japan,
Tel: +81-011-706-6542, Fax: +81-011-706-7892, Email: ksaitoh@ist.hokudai.ac.jp)

Abstract
We discuss the resonant coupling phenomenon in hollow-core photonic band-gap fibers for suppression of higher-order-modes, large-mode-area index-guiding photonic crystal fibers with low bending losses for high-power laser applications, and hole-assisted fibers for broadband dispersion compensation.

1 Introduction
After the first proposal and demonstration of photonic crystal fibers (PCFs) in 1996 [1], they have received a considerable attention in optical community due to their superior characteristics and large degrees of freedoms in designing the core and cladding geometries according to the application. PCFs, where tiny air-holes running down its length build the cladding, can be categorized in two main classes; (i) solid-core PCFs where a single or multiple missing air-hole forms the guiding core and the light is guided via total internal reflection, and (ii) hollow-core PCFs (HC-PCFs) where several missing air-holes construct the air-guiding core and the light is guided through photonic band-gap (PBG) effect [2].

In this talk, we focus on hollow-core as well as solid-core PCFs and discuss the main aspects of them for optical communication and industrial applications. Firstly, we will consider the case of an HC-PCF where we point out the suppression of higher-order-modes (HOMs) by creating outer-cores in the cladding through resonant coupling mechanism. And in the second half of this presentation, we will focus on the dispersion compensation, Raman amplification, and lasing performances of solid-core PCFs which use resonant coupling phenomenon effectively. We have adopted a full-vectorial finite element method (VFEM) to model the proposed PCF configurations.

2 Suppression of higher-order modes in HC-PCFs
The potential use of PCFs becomes clearer when the unique properties can be directly related to qualitatively novel functionalities. For example, the ability to guide light in a hollow-core via the PBG effect suggests a new technological regime of transmission with low attenuation, high power delivery with low nonlinearities, or nonlinear optics applications. The main technological issue that exists when guiding the light through HC-PCFs is the inevitable presence of HOMs. Although suppression of HOMs can be observed in small-core PCFs, there exists a fundamental limit of scattering losses due to the surface roughness scattering, which prohibits the further reduction of the fiber’s attenuation to the level of the conventional fiber. One possible solution to overpass this limit is to consider large HC-PCFs, leading the multi-mode operation. In order to enhance the suppression of HOMs in HC-PCFs, a resonant coupling mechanism between the HOMs in the central core and the much leaky outer-core modes around a certain wavelength range can be adapted for effectively single-mode operation. In Fig. 1(a) we model an idealized but realistic standard 7-unit-cell HC-PCF [3]. The 7-unit-cell hollow-core size is not so large, and therefore it is difficult to suppress the scattering losses due to the surface roughness. In order to form larger HC-PCFs, three rows of air-holes can be removed and the resulting structure is depicted in Fig. 1(b) (19-unit-cell HC-PCF). We found that the effective index of the fundamental mode in the small-core PCF is quite close to that of the HOMs in the large-core PCF. This fact can enable the index-matching mechanism to take place by bringing two cores (19-unit-cell central core and 7-unit-cell outer core) close to each other, thus enabling resonant coupling over a certain wavelength regime.

In order to enable the index matching mechanism between the HOM in 19-unit-cell core and the fundamental mode in 7-unit-cell core, we proceed by incorporating six 7-unit-cell cores (shown in Fig. 1(c)) into the cladding surrounding the central 19-unit-cell core to expect polarization independent operation. Introducing the several outer cores close to the central core would lead in the enhancement of the confinement loss (CL) of the HOMs in comparison to that of the fundamental mode. Through numerical simulations, it was observed that the CL of HOMs in 19-unit-cell HC-PCF with outer cores stays above 100 dB/km which is 3 times higher in order than that of HOMs in 19-unit-cell HC-PCF without defected outer cores, whereas the CL of fundamental mode remains below 0.1 dB/km. This ensures the effectively single mode operation of the device.
3 Design of a single-mode and large-mode-area PCF and its lasing characteristics

One of the major trends in optical fiber science is to obtain fibers with large-mode-area (LMA), optimized for various applications such as high power delivery, fiber amplifiers, and fiber lasers. In order to ensure the high beam quality and the ultimate controllability of the damage threshold in the fiber’s material, it is required to have a LMA property with maintained single-mode operation. We propose an interesting design method for realizing LMA single-mode PCFs, small allowable bending radius, and high beam quality factor. Figure 2 (a) shows the proposed structure [4], where the central core is formed by seven missing air-holes surrounded by 12 air-holes with diameters of $d_1$ and $d_2$ arranged alternatively with hole-pitch $\Lambda$.

The enhancement of the CL of the higher-order central-core mode can be achieved through the resonant coupling mechanism between the central-core and the leaky ring-core modes. The air-hole diameter-$d$ in the outermost air-hole ring is determined such that it can match the effective index of the $LP_{11}$-like HOM in the central core with that in the ring core, at the desired wavelength $\lambda = 1064$ nm. The optimized fiber shows large effective mode area of 1400 $\mu m^2$ at 1064-nm wavelength, good beam quality factor of 1.15, and CL exceeding the 1 dB/m for the HOM at 1064-nm wavelength. We then investigated the lasing performances of the proposed fiber by doping the fiber core with ytterbium ions in 25 $\mu m$ doping region radius and noted the influence of the bending on the overlap factor. It was found that the confinement of the pump or signal decreases as the fiber is bent into 5 cm bending radius, whereas the laser output power does not deteriorate so much on bending. This demonstrates the bend-insensitive operation of designed Yb$^{3+}$-doped, LMA-PCF laser.

4 Dispersion compensating hole-assisted fibers

Another variant of solid-core PCFs is the hole-assisted fibers (HAFs) which have a germanium-doped core surrounded by either one or two rings of air-holes. HAFs are the simplest design of PCFs and easy to fabricate. This fact motivated us to design a HAF which can provide a very large negative dispersion with relatively large mode area. Figure 2(b) shows the intriguing design of dispersion-compensating HAF (DC-HAF), where only one air-hole ring surrounds the doped core. The optimized DC-HAF exhibits the largest negative dispersion around $-550 \, \text{ps/nm/km}$ at 1550 nm with matched relative dispersion slope of transmission fiber as shown in Fig. 3(a). This large negative dispersion is also based on the index-matching mechanism between the effective index of the fundamental mode in the highly-doped core and the cladding index at longer wavelength. The described DC-HAF can act as a broadband dispersion compensator as it satisfies the condition for broadband dispersion compensation. The length of DC-HAF was computed as 2.52 km to compensate for the positive dispersion accumulated over 80-km long conventional single mode fiber link. It is evident from Fig. 3(a) that the DC-HAF leaves a residual dispersion of $\pm 20 \, \text{ps/nm}$, which is below than $\pm 60 \, \text{ps/nm}$; a limit for 40 Gbps transmission systems.

5 Concluding remarks

We have modelled and simulated different designs of PCFs using VFM. The suppression of HOMs in large HC-PCFs was achieved by resonant coupling and introduction of six outer-cores. The resonant coupling phenomenon was also employed to design a single-mode, LMA solid-core PCF and its lasing characteristics were demonstrated numerically. Finally, a simple and intriguing design of dispersion compensating HAF was proposed and it was shown that the DC-HAF module can compensate for the dispersion in C-band and can also acts as inline Raman amplifier.

6 References

Periodic refocusing in microstructured fiber for efficient quasi-phase matched optical nonlinearity

Makiko Hisatomi¹, Michael C. Parker¹, and Stuart D. Walker²

¹Fujitsu Laboratories of Europe, Columba House, Adastral Park, Ipswich, IP5 3RE, U.K.
Tel: +44(0)1473 614143, Fax:+44(0)1206 762916, email: Makiko.Hisatomi@uk.fujitsu.com,
²University of Essex, Department of Electronic Systems Engineering, Wivenhoe Park, Colchester, CO4 3SQ, U.K

Abstract
We demonstrate periodic light refocusing within a solid-core photonic bandgap microstructured fiber design. This technique offers enhanced efficiency in quasi-phase matched optical non-linear applications. Spatial periods varying from 17µm – 43µm are possible.

1 Summary
Microstructured fiber (MSF), such as photonic crystal fiber (PCF) [1], Bragg fiber [2] and zoned MSF [3], offers numerous design possibilities compared to conventional single-mode fiber (SMF). Recently, new MSF designs have also emerged with enhanced optical nonlinear properties. This has been achieved by a combination of novel materials (e.g. bismuth glasses), and very small mode sizes [4]. In addition, the use of quasi-phase matching (QPM) techniques has also been developed to enhance nonlinear interaction efficiencies, e.g. via periodic-poling [5]. Recent work by Pfeifer et al. [6], has also demonstrated the feasibility of QPM in hollow-core fiber, by means of two-mode beating effects. In this paper, we present a novel MSF design which exploits the inherent fiber geometry to enhance the efficiency of optical nonlinear effects by QPM. This is achieved by careful phase-matching of all modes in a multimoded waveguide structure, so that the modes become spatially and periodically “mode-locked”. In previous work, we exploited a similar MSF geometry to demonstrate photonic bandgap light confinement in hollow-core radially-chirped Bragg fiber (RCBF), featuring an aperiodic geometry [7]. We continue to exploit the photonic bandgap effect to confine the light within our MSF design, so highlighting the flexibility of our approach in possible choices of optical nonlinear media. In the RCBF case, the zoned MSF can also be regarded as the longitudinal embodiment of a binary Fresnel zone plate [8], such that light is guided by periodic refocusing back within the fiber structure. Alternatively, the periodic refocusing of the light within the fiber core is arguably analogous to the Talbot self-imaging effect, exploited within multi-mode interference (MMI) couplers. However, rather than employing an essentially periodic grating as is required for the 1D case, the cylindrical geometry of a fiber requires aperiodic gratings [9] to achieve the same self-imaging effect.

Fig.1: Schematic diagrams of solid-core RCBF and refractive index profiles with: (a) & (b) \( r_1=1.67\mu m \); (c) & (d) \( r_2=3.92\mu m \)
We have adopted a simple Fresnel zoning geometry to achieve the refocusing effect, although the zeroes of a Bessel function could also be exploited; this being the subject of future research. Figure 1 shows schematic cross-sections of the two solid-core RCBF waveguide geometries that we have simulated. For a Fresnel zone configuration, the general expression for the radius $r_m$ of the $m^{th}$ zone is given by the equation $r_m = \sqrt{m r_1}$, where $r_1$ is the central zone radius. This simple square-root relationship presupposes closely similar refractive indices for the different zones, such that overall optical path lengths are in the same proportion as geometric path lengths. In the geometry of fig. 1(a)&(b), the center (geometric) radius is given by $r_1 = 1.67 \mu m$, with center refractive index given by $n_2 = 2.25$, whilst the subsequent index is $n_r = 2.35$, and there are 9 zones. Fig. 1(c)&(d) shows a MSF geometry with a larger central zone radius, $r_1 = 3.92 \mu m$. However, rather than following the conventional zone radii of the Fresnel formula, we have maintained the same zone radius increments of the previous geometry in a manner similar to that used in [7], such that the subsequent zone radii are given by $r_{m} = r_{1} + (r_{m-1} - r_1)$, where $r_{1} = 1.67 \mu m$, and $r_{m}$ follows from the square-root Fresnel formula. We have used a fully vectorial 3D finite-difference time-domain (FDTD) package, *Poynting for Optics*, to simulate electric field evolution in our MSF designs.

Fig.2(a) shows the evolution of a linearly-polarised $D_s=9\mu m$ width Gaussian input mode source ($\lambda = 1.55\mu m$) along a $55\mu m$-length of the fiber shown in Fig.1(a). The periodic refocusing of the light is a marked effect, with a spatial period of $\Lambda = 17 \mu m$ within the low refractive index core of the fiber. This is the key result of our paper. The vertical (green) lines indicate the zone boundaries. In the FDTD simulation, time steps were $dt = 133 \text{as}$, whilst a minimum spatial resolution of $dx = dy = 100 \text{nm}$ for transverse dimensions, and $dz = 50 \text{nm}$ (along the fiber axis) were employed. Fig. 2(b) and 2(c) show how the spatial period $\Lambda$ can be tuned by varying both fiber geometry and also the Gaussian diameter $D_s$ of the illuminating light. In this case, the period $\Lambda$ can be increased by either increasing the radius of the central zone, e.g. $\Lambda$ can be increased to $\Lambda = 30 \mu m$ for the central zone radius of $\rho_s = 3.92 \mu m$. Furthermore, reducing the diameter of the source light to $D_s = 6 \mu m$ also increases the spatial period further to $\Lambda = 43 \mu m$. In these two latter cases, the FDTD simulation parameters were increased slightly to $dt = 144 \text{as}$ and $dx = dy = 131 \text{nm}$, with $dz = 50 \text{nm}$, in order to accommodate the larger fiber structure within the computing memory resources. In summary, we have shown how fiber geometry can be exploited to further enhance within-fiber optical non-linearities with QPM techniques via periodic Talbot-like imaging. Optimisation of the QPM spatial periodicity $\Lambda$ is made possible by appropriate MSF geometric design, whilst dynamic tuning of $\Lambda$ is also available by varying the diameter $D_s$ of the illuminating light source.

2 **Acknowledgement**

The authors would like to thank Dr Namiki (Fujitsu Ltd.) for the *Poynting for Optics* software used in this work.

3 **References**

Square Photonic Crystal Fiber (SPCF) with Flattened Chromatic Dispersion and High Nonlinearity

Feroza Begum\textsuperscript{a)}, Yoshinori Namihira\textsuperscript{b)}, S.M. Abdur Razzak, Shubi Kaijage, Nguyen H. Hai, Tatsuya Kinjo and Nianyu Zou

Graduate School of Engineering and Science, University of the Ryukyus, 1 Senbaru, Nishihara, Okinawa 903-0213, Japan
Email: \textsuperscript{a)} k048612@eve.u-ryukyu.ac.jp, \textsuperscript{b)} namihira@eee.u-ryukyu.ac.jp

Abstract
An index-guiding SPCF for ultra-flattened dispersion of $0 \pm 0.7$ ps/(nm.km) and low confinement loss $10^{-4}$ dB/m in a wide wavelength range is presented. The nonlinear coefficient is $27$ [Wkm]$^{-1}$ at 1.55 µm.

Key words: fiber design, photonic crystal fiber (PCF), chromatic dispersion, nonlinearity, confinement loss.

1. Introduction
Index-guiding photonic crystal fibers (PCFs) are a new class of optical fibers in which the solid pure silica core region is surrounded by a cladding region that contains air-holes running down their length [1]. It is possible to design various novel PCFs required in many applications such as optical communication systems [2], nonlinear optics [3], and ultra-broad supercontinuum generation [4].

It has been known that in index-guiding PCFs, since the periodicity in the cladding region is not essential to confine the guiding light into the core region, it is possible to control both the dispersion and dispersion slope in wide wavelength range by varying the cladding structure, air-hole diameter of each air-hole ring and the hole-to-hole spacing. Using a triangular and square PCF with a small core and large air-hole diameter, various flattened-dispersion PCFs with tight mode confinement and nonlinearity have been reported [3, 5, 6]. To the limit of our knowledge, without significantly increasing the number of the control parameters, this type of highly nonlinear dispersion flattened square PCFs (HNDF-SPCF) at 1.55 µm wavelength range has not been reported yet.

In this paper we propose and numerically characterize a simple structure of index-guiding HNDF-SPCF with flattened dispersion characteristics, low confinement loss and small effective area. It is shown through numerical results that the simple HNDF-SPCFs are obtained with $0 \pm 0.7$ ps/(nm.km) flattened-dispersion and less than $10^{-4}$ dB/m confinement loss in the wavelength range of 1.2 µm to 1.7 µm. It also exhibits large nonlinear coefficient of about $27$ [Wkm]$^{-1}$ at 1.55 µm, which is found to be better than that of the reported triangular PCFs [4].

2. Proposed Design and Simulation Method

Fig. 1 shows the schematic cross-section of the proposed SPCF with two rings, which is composed of circular air-holes in the cladding arranged in a square array. Where $\Lambda$ is the center-to-center spacing between the air-holes and $d$ is the air-hole diameter, $d/\Lambda$ is the normalized diameters of the air-holes in the cladding and $2a = 2\Lambda - d$ is the core diameter. Fig. 2 shows the proposed HNDF-SPCF, with the air-hole diameters in the first ring are varied and the rest of the air-hole diameters remained same. In the first air-hole ring, two different types of air-hole diameter are used. In this study, the finite difference method (FDM) [7] with anisotropic perfectly matched layers (PMLs) is used to analyze the various properties of HNDF-SPCFs. To demonstrate the proposed design, the chromatic dispersion $D(\Lambda)$, confinement loss $L_c$, effective area $A_{eff}$, and nonlinear coefficient $\gamma$ are analyzed based on [7-9].
3. Simulation Results and Discussion

Fig. 3 shows the chromatic dispersion, the confinement loss, and the effective area as a function of wavelength for the HNDF-SPCF in Fig. 2 at $\Lambda = 1.0 \mu m$. The flattened chromatic dispersion is $0 \pm 0.7 \text{ ps/(nm.km)}$ and confinement loss is less than $10^{-4} \text{ dB/m}$ in a wavelength range of $1.2 \mu m$ to $1.7 \mu m$. The effective area is $3.42 \mu m^2$ and the nonlinear coefficient is about $27 \text{ [W/km]}^{-1}$ at $1.55 \mu m$ wavelength.

4. Conclusions

A simple structure of index-guiding HNDF-SPCF with ultra-flattened dispersion, low confinement loss, small effective area and large nonlinear coefficient is investigated in this research. The numerical results revealed that it is possible to get flattened chromatic dispersion of $0 \pm 0.7 \text{ ps/(nm.km)}$ and the confinement loss is less than $10^{-4} \text{ dB/m}$ in a wavelength range of $1.2 \mu m$ to $1.7 \mu m$. In this case, flattened dispersion wavelength bandwidth is $500 \text{ nm}$. It means that our proposed fiber can be used in S, C, and L bands for controlling chromatic dispersion. Besides the nonlinear coefficient is about $27 \text{ [W/km]}^{-1}$ at $1.55 \mu m$ wavelength. For these properties the proposed index-guiding HNDF-SPCFs may be suitable for applications such as, broadband optical communication systems, broadband supercontinuum generation, nonlinear optical loop mirror, etc.,.

References

Square-Lattice Photonic Crystal Fiber with High Negative Dispersion and Low Confinement Loss for Dispersion Compensating Applications

Shubi Kaijage\textsuperscript{a}, Yoshinori Namihira\textsuperscript{b}, Feroza Begum, Nguyen H. Hai, S. M. Abdur Razzak, Tatsuya Kinjo, Jitsuryo Nakahodo, Miki Naruse and Nianyu Zou
Graduate School of Engineering and Science, University of the Ryukyus, 1 Senbaru, Nishihara, Okinawa, 903-0213 Japan
E-mail: a) k068455@eve.ryukyu.ac.jp, b) namihira@eee.u-ryukyu.ac.jp

Abstract
A designed structure of six rings square-lattice PCF (SPCF) having both, high negative dispersion characteristics of -295 ~ -342 ps/(nm.km) and low confinement losses within the wavelength range of 1.53 \( \mu \)m-1.565 \( \mu \)m has been proposed.

Key words: photonic crystal fiber(PCF), Chromatic dispersion, residual dispersion, confinement loss.

1. Introduction
Photonic Crystal Fibers, PCFs are transparent strands (usually of silica) with an array of micron-scale holes that run along the length of the fiber [1]. It is the presence of air holes, rather than index variation in the glass, that confines light to the core. Such PCFs have lot of unique properties which are not realized in conventional optical fibers, like large waveguide dispersion, high birefringence and so many [2, 3].

The chromatic dispersion in single mode optical fibers causes light pulses to spread, limiting the data transmission rate and length of optical fiber links. To overcome these limits, various techniques have been used to eliminate the dispersion effects. Optical fibers presently installed in telecommunication links show dispersion of about 17 ps/(nm.km) at the currently preferred wavelength of 1.55 \( \mu \)m. This large dispersion can be compensated by a short length of a special type of optical fibers, such as dispersion compensating fibers (DCF) with a dispersion of opposite sign so that the net dispersion of two fibers in series becomes zero [4]. DCF is one of the most utilized techniques for upgrading to the high bit rates and to the wideband transmission systems for its ability to compensate for dispersion and dispersion slope, simultaneously [4]. However, the conventional DCFs do exhibit small negative dispersion coefficient of about -100 ps/ (nm.km) at 1.55 \( \mu \)m wavelength and also with high losses [5, 6].

Circular air holes arranged on the square array in the cladding with eight air holes in the first ring surrounding the core is named square photonic crystal fiber (SPCF). Besides square PCFs, there are also hexagonal [1], octagonal PCFs [7] and many other structures proposed to design PCFs. The presence of air holes in the cladding makes PCFs versatile for many dispersion managed-applications. In particular, dispersion compensating designs of PCFs compel them to be an indispensable candidate to be employed in fiber optic communication link due to their large negative dispersion coefficient.

In this paper, we numerically investigate and propose the SPCF structure for dispersion compensating applications. It has shown that for the particular pair of values of air hole diameters, high negative dispersion coefficient, negative dispersion slope with low confinement loss over C-band could be realized. The Finite difference method (FDM) with anisotropic perfectly matched layer (PML) has used to calculate dispersion properties [7].

Fig.1: Cross section of the designed SPCF showing pitch, \( \Lambda \) and the air hole diameters \( d \) and \( d_1 \) in the 4-rings case.

2. Fiber design and simulation results
In conventional PCFs (the ones with same air-hole diameters in the cladding region) it is difficulty to control dispersion and dispersion slope at the same time in a wide wavelength range. In this design we considered the structure of the Square PCF shown in Fig. 1, which shows the transverse cross section of the designed SPCF, where by \( d \) and \( d_1 \) are the air-hole diameters for large and small holes respectively and \( \Lambda \) is the pitch, centre to centre separation between two consecutive air holes.

Chromatic dispersion, \( D(\lambda) \) of the SPCF is calculated by using [8,9];

\[
D(\lambda) = -\frac{\lambda}{c} \frac{d^2 \text{Re}[n_{\text{eff}}]}{d\lambda^2}
\]

With unit of ps/(nm.km), where; \( \lambda \) is the operating wavelength, Re \( (n_{\text{eff}}) \) is the real part of the refractive index, \( n_{\text{eff}} \) and \( c \) is the velocity of light in a vacuum.
Confinement loss, \( L_c \) is then obtained from the imaginary part of \( n_{\text{eff}} \) as follows [8]:

\[
L_c = 8.686 k_0 \Im\{n_{\text{eff}}\} 
\]

(2)

With unit of dB/m, where, \( k_0 \) is the free space wave number.

Fig.2 depicts the dispersion coefficient for the designed SPCF and conventional single mode optical fibre (SMF), where the parameters of the designed SPCF are \( d=1.5 \mu m \), \( d_v=0.64 \mu m \) and \( \Lambda=2 \mu m \). By using 6 air-hole rings, high negative dispersion are realized (-295 ~ -342 ps/(nm.km)) in 1.53 \( \mu m \) ~ 1.565 \( \mu m \) wavelength range. Dispersion coefficient of -322 ps/(nm.km) at wavelength 1.55 \( \mu m \) has been obtained. The designed SPCF exhibits a negative dispersion slope and has a relative dispersion slope, RDS= \( \frac{dD}{d\lambda}D \) of 0.0038 nm\(^{-1}\) which is close to matching the value of SMF’s RDS=0.0036. For the multiple wavelength compensation within C-band, RDS values for the dispersion compensator should match as closely as possible to that of the fiber being compensated (transmission fiber) [4, 5]. Fig.3 shows the spectral variations of confinement loss which is less than \( 10^{-4} \) dB/km for the operating wavelength range and varies from \( 3.5 \times 10^{-5} \) dB/km ~ \( 5.8 \times 10^{-5} \) dB/km within C-band. Also Fig.3, shows calculated residual dispersion, \( D_R \) for 4.15km long of designed SPCF in the link with 80km long conventional SMF. The residual dispersion varies from -7.7 ps/nm to 28.3 ps/nm for wavelengths of interest (C-band), which is suitable in 20Gbps WDM networks applications [6].

3. Conclusion
It was found that to be possible achieving both high negative dispersion coefficient and low confinement losses from a simple structure of index-guiding square-lattice PCF. A 6-ring square-lattice PCF with high negative dispersion coefficient of about -295 ~ -342 ps/(nm.km) and low confinement losses within C-band were obtained. The designed SPCF clearly depicts its potentiality as a candidate for multiple wavelength dispersion compensating in 20 Gbit/s WDM networks systems.

References

Fig.2: Graph for dispersion coefficients of the designed SPCF and conventional single mode fiber (SMF) with wavelength.

Fig.3 Variations for the confinement loss and residual dispersion for 4.15 km-long of designed SPCF compensating 80 km-long SMF.
This Paper has been withdrawn.
In this paper an overview of semiconductor lasers, covering the 2-to-3 µm wavelength range, will be given. There is an increasing demand for compact near-to-mid infrared semiconductor to serve a variety of applications including spectroscopic sensing, medical diagnostics, laser surgery and materials processing such as welding of transparent plastics. The (AlGaIn)(AsSb) materials family is well suited for the fabrication of quantum well (QW) diode lasers covering the 2-to-3 µm wavelength range. GaInAsSb is used as the active region QW material, which can be grown either lattice-matched or deliberately strained onto GaSb as substrate material. The separate confinement layers as well as the cladding layers consist of AlGaAsSb lattice matched to GaSb with typical Al-contents of 20-40 % and 50-90 %, respectively.

For broad area (BA) lasers emitting at 2 µm, high power efficiencies (~25%) and output powers of 2 W in cw mode (>9 W in pulsed mode) have been achieved at room-temperature. Laser bars with 19 BA emitters show at the same wavelength and temperature a cw output power of 21 W. To serve applications which require a better slow-axis beam quality than that provided by BA lasers, the tapered laser concept has been adopted. Such lasers yield at 1.9 µm a nearly diffraction limited output with a beam quality factor of $M^2 < 1.7$ up to an output power of 1.5 W.

Optically pumped Vertical-External-Cavity Surface-Emitting Lasers (VECSELs) are attracting considerable current interest as an alternative to edge-emitting diode lasers. VECSELs combine a high quality circular output beam, which is a feature of classical solid state lasers, with the wavelength versatility of a gain medium composed of semiconductor quantum structures. VECSELs consisting of a GaInAsSb/AlGaAsSb QW active region and a GaSb/AlAsSb distributed Bragg reflector have been demonstrated to emit at 2.3 µm a maximum cw output power of 1.5 W ($M^2 < 3$) with just thermoelectric cooling to -20°C. The VECSEL chip was bonded to an intra-cavity diamond heat-spreader for efficient heat removal.
Joachim Wagner received his Ph.D. degree in physics in 1982 from the University of Stuttgart (Germany). Then he worked at the “Max-Planck-Institut für Festkörperforschung” in Stuttgart in the group of Prof. M. Cardona before joining the Fraunhofer-Institute for Applied Solid State Physics in Freiburg (Germany) in 1985. There he is currently head of the department “Optoelectronic Moduls”. He is also professor at the Physics Department of the University of Freiburg and an associated member of the Materials Research Center Freiburg (FMF). His current research interest is in III/V-semiconductor heterostructures and their application in optoelectronic devices both for the infrared and the visible/uv spectral range. Specifically he is heading various R&D projects on group III-antimonide based infrared lasers as well as on group III-nitride based LEDs and diode lasers. He is author or co-author of more than 300 scientific publications including several review papers and book chapters.
Abstract: A hybrid silicon evanescent racetrack laser with integrated photodetectors has been demonstrated running continuous-wave (c.w.) at 1590 nm with a threshold, maximum output power, and maximum operating temperature of 175 mA, 29 mW and 60 °C, respectively.

1. Introduction

Silicon photonics research has had many advancements with hopes to enable the introduction of integrated photonics in new applications due to silicon’s low cost manufacturing infrastructure [1]. Recently, we demonstrated a hybrid AlGaInAs-silicon evanescent laser that fulfills the need for an electrically pumped laser source that can be integrated on a wafer scale with a silicon photonic platform [2]. The first hybrid laser demonstration relied on the dicing and polishing of straight hybrid waveguides to define a Fabry-Perot laser cavity. Here we describe a monolithic hybrid AlGaInAs-silicon evanescent laser based on a racetrack-resonator-topography [3]. The laser runs continuous-wave (c.w.) with a threshold of 175 mA, a maximum total output power of 29 mW and maximum operating temperature of 60 °C. Moreover, the integration of this laser with a hybrid AlGaInAs-silicon evanescent photodetector is used to measure the laser output [4].

Fig. 1. The hybrid silicon-evanescent device cross section structure.

2. Device Structure and Fabrication

The hybrid AlGaInAs-silicon evanescent device cross section is shown in figure 1. The devices are fabricated using an AlGaInAs quantum well epitaxial structure that is bonded to a low-loss silicon rib waveguide. The silicon waveguide was fabricated with a final height, width, and rib-etch depth of 0.69 μm, 1.65 μm, and 0.5 μm, respectively. This results in a calculated overlap of the optical mode with the silicon waveguides and the quantum wells of 64 % and 4.2 %, respectively. The fabrication procedure and III-V epitaxial structure details can be found in Ref. [2].

Fig. 2. a) The layout of the racetrack resonator and the photodetectors. b) A top view SEM micrograph of two racetrack resonator lasers. The racetrack resonator lasers on the top and bottom have radii of 200 and 100 microns, respectively.

The laser layout is shown in Figure 2. It consists of a racetrack ring resonator with a straight waveguide length of 700 microns. A directional coupler is formed on the bottom arm by placing a bus waveguide 0.5 micron away from the racetrack. Four device designs were fabricated with varying ring radius, and coupler interaction lengths \( L_{interaction} \). Table 1 shows the device layout breakdown with the corresponding cavity lengths \( L_{cavity} \) and the computed coupling percentage to the bus waveguide. The laser power is collected into the two 440 micron long photodetectors. These photo-detectors have the same waveguide architecture as the hybrid laser, the only difference being that they are reverse biased to collect photo-generated carriers.
3. Experiment and Results

The laser is driven by applying a positive bias voltage to the top p-probe contact while the optical power is measured by the two photodetectors on each side of the coupler. The photocurrent is measured while reverse biasing the photodetectors at -5V. We use a responsivity of 1.25 A/W such that the laser power values are on the conservative side. Since the testing of the lasers are done all on chip without polishing and dicing, the lasing spectrum is measured by collecting scattered light near the bends of the ring through a fiber probe.

![Fig. 3. The LI curve for a laser with radius R = 200 microns, and L\textsubscript{interaction} = 400 microns for various temperatures](image)

Figure 3 shows the measured total c.w. laser output power which is the sum of the optical power measured at both detectors as a function of injected current for various operating temperatures ranging from 15 to 60 °C for the laser with a ring radius and coupling interaction length of 200 microns and 400 microns respectively. As can be seen from Fig. 3, the laser threshold is 175 mA with a maximum output power of 29 mW at 15 °C. The maximum power is limited by the available drive current to the device. The laser has a 60 °C maximum lasing temperature with a characteristic temperature of 55 K. The laser has a threshold voltage of 1.75V and a series resistance of 3.5 ohms.

The spectrum was measured with an HP 70952A optical spectrum analyzer with a resolution bandwidth of 0.1 nm. The lasing wavelength is 1592.5 nm with a 0.21 nm mode spacing corresponding to a group index of 3.67.

![Table 1, Ring dimensions and coupling parameters](image)

<table>
<thead>
<tr>
<th>Radius (µm)</th>
<th>L\textsubscript{ring} (µm)</th>
<th>L\textsubscript{interaction} (µm)</th>
<th>Computed Feedback Coupling %</th>
</tr>
</thead>
<tbody>
<tr>
<td>200</td>
<td>2656</td>
<td>600</td>
<td>3%</td>
</tr>
<tr>
<td></td>
<td>2400</td>
<td>300</td>
<td>12.6%</td>
</tr>
<tr>
<td>100</td>
<td>2028</td>
<td>100</td>
<td>85%</td>
</tr>
</tbody>
</table>

Table 2 shows the maximum output powers, differential efficiencies, threshold currents, and maximum output temperatures of the four device designs. The injection efficiency, modal loss, and g\textsubscript{0} were found to be 70%, 15 cm\textsuperscript{-1}, and 1500 cm\textsuperscript{-1}, respectively. The additional bend loss for 200 micron and 100 micron bends were found to be 0 cm\textsuperscript{-1} and 50 cm\textsuperscript{-1}, respectively.

4. Conclusion

The integration of a racetrack laser with a photodetector on the hybrid silicon evanescent device platform demonstrates the potential to realize practical photonic integrated circuits on a silicon substrate. These two types of photonic devices are fabricated on a single active region design showing the flexibility of the hybrid silicon evanescent device platform. On-chip testing and characterization of the laser simplifies the testing by eliminating facet polishing and characterization uncertainties caused by coupling losses. We have demonstrated a monolithic laser with output powers up to 29 mW operating up to 60 C in the range of 1590nm. The integrated photodetector shows a responsivity of ~1.11 A/W.

Acknowledgments

We thank Intel & Jag Shah & Wayne Chang through DARPA for supporting this research through contracts W911NF-05-1-0175 and W911NF-04-9-0001.

References

High speed all-optical packet switching

H. J. S. Dorren1, J. Herrera1,2, O. Raz1, E. Tangdiongga1, Y. Liu1, J. Marti3, F. Ramos3, G. Maxwell3, A. Poustie2, H.C. Mulvad1,4, M. T. Hill1, H. de Waardt1, A. M. J. Koonen1, G. D. Khoe1

1: CODRA Research Institute, Eindhoven University of Technology, PO. Box 512, 5600MB – Eindhoven, The Netherlands
2: Nanophotonics Technology Center, Universidad Politécnica de Valencia, Camino de Vera, s/n, 46022 – Valencia, Spain
3: Centre for Integrated Photonics (CIP) Ltd., Adastral Park, IPS 3R – Ipswich, United Kingdom
4: COM•DTU, Technical University of Denmark, Building 345V, DK-2800 Kgs. Lyngby, Denmark

Abstract: We discuss a packet switched node that accomplishes packet routing solely by photonic control circuits. All the function blocks are discussed and packet routing over a 110 km field installed fiber link is shown.

Introduction

All-optical packet switching has been proposed as a future technology to solve the mismatch between the fibre bandwidth and the router forwarding capacity, especially when the data bit-rates exceed the hundreds of Gb/s. At these very high speeds, electronics is not capable for coping with the high amount of data and require parallel processing which is complicated to realize, and costly in terms of size, volume, power consumption. All-optical technologies arise as a viable alternative, especially when the main sub-systems can be integrated in a monolithically or hybrid technology which results in reduced footprint and cost-effective solutions. In this paper, we discuss all-optical packet switching employing in-band wavelength labelling. The packet switch employs an optical filter as label extractor, a hybrid-integrated optical flip-flop, and a wavelength converter that utilises nonlinear chirp dynamics in a single SOA. Error-free packet switching at acceptable penalties is demonstrated over 110 km field installed fiber links.

System

Fig. 1 shows the packet switch configuration. At the transmitter side, packet payload is generated by time-multiplexing 10-Gb/s pre-defined 768 bits return-to-zero data consisting of 1.6 ps optical pulses at \( \lambda_{\text{F}1}=1546 \text{ nm} \), up to 160 Gb/s using a passive fibre-based pulse interleaver. The resulting payload consists of a 62 ns data burst at a repetition rate of 78.6 ns, 2.6 ns squared optical pulses at two alternating wavelengths are generated as labels and placed prior to the payload. The label and the payload are separated in time with about 7 ns guard times. The labels wavelength are \( \lambda_{\text{L}1}=1547.4 \text{ nm} \) and \( \lambda_{\text{L}2}=1544.6 \text{ nm} \), within the 20-dB bandwidth of the data payload.

The advantage of an optical label that is in-band with the packet payload is that the label can be extracted easily by a narrow-band optical filter, that a large number of optical labels can be assigned (commercially available AWGs allow for separation of up to 25 labels within the 5-nm packet payload spectrum), and the label can be recovered with low distortion due to the spectral spreading of the payload energy.

In the switch the incoming power is split in two, one half drives the wavelength converter while the other half is processed in the control plane. The header processor, a Gaussian filter with (3 dB bandwidth of 0.3 nm) bandwidth, is tuned to filter out the packet labels. When a packet with a label at the appropriate wavelength arrives, the filter extracts the label with a high extinction ratio. This pulse is then split in two. Half of the power is delayed for about 66 ns with respect to the other half to set and reset the optical flip-flop.

The flip-flop is a fully- packed, hybrid-integrated all-optical device and consists out of two coupled Mach-Zehnder interferometers. The bias currents of the SOAs in the optical flip-flop are 172 mA and 185 mA respectively. The flip-flop outputs a signal at a wavelength either at \( \lambda_{\text{F}1}=1554 \text{ nm} \) or at \( \lambda_{\text{F}2}=1560 \text{ nm} \), depending on the flip-flop state. The state is set and reset using externally injected pulses with interval of 66.4ns. This signal shows an optical power of 0.2 mW, dynamic contrast-ratio of 6 dB and a rise/fall time of 2 ns. Finally, after amplification up to 1.9 dBm with high extinction ratio. This pulse is then split in two. Half of the power is delayed for about 66 ns with respect to the other half to set and reset the optical flip-flop.

Fig. 1. Experimental set-up.
In order to recover the clock for demultiplexing, the initial predefined sequence at 10 Gb/s is also transmitted in an out-of-band wavelength, $\lambda_C=1549.5$ nm, through the whole transmission link. The clock signal is detected and feeds a phase-locked loop which drives the optical demultiplexer. Both packets and clock are separated using optical filters at the packet switch in the intermediate point and at the receiver side.

A packet switch experiment is carried out by launching the output of the transmitter into a 54.3 km of field installed fiber link with appropriate dispersion compensation and amplification at the fibers output to restore signal fidelity. This signal is fed into the packet switch and output into a second field installed fiber link. To compensate for chromatic dispersion carefully selected lengths of dispersion compensation-fiber (DCF), and SMF are used. To control polarization mode dispersion, the polarization of the payload and labels is controlled in the transmitter, and a polarization controller plus a polarization beam-splitter are placed after the first link, as for the second link the light will be already polarized.

Results and discussion

Fig. 2a-f and Fig 2g-h show optical traces and eye diagrams in different points of the system set-up. As observed from the corresponding eye diagram in Fig 2k, further degradation of the eye occurs but still considerable eye-opening, suggests error-free operation.

To estimate the error performance of the system, the data was routed through the whole link using static wavelength conversion. That is, without labels in the packets and dynamic operation of the flip-flop. Fig 3a-d show the BER performance of the 16 multiplexed channels (solid lines) measured in different points of the experimental set-up compared to the basic 10 Gb/s channel (dashed line). Fig. 3e-h shows the eye diagram of a demultiplexed channel in different points across the link. It is visible from Fig 3a that the penalty at BER=$10^{-9}$ is visible from Fig 3a to 3b, due to the accumulated noise and dispersion issues. After the wavelength conversion the penalty range is from 3 to 4.5 dB (Fig 3b), further degradation of the eye occurs but still considerable eye-opening, suggests error-free operation.

References


Fig. 2. Traces of (a) the transmitted packet; (b) the packet after the first link; (c) the recovered labels; (d) the flip-flop output; (e) the packet-switched payload; (f) the packet payload after the second link. Eye diagrams of: (g) the transmitted packet payload; (h) the packet payload after the first link; (i) the packet-switched payload; (j) the packet payload after the second link.

Fig. 3. BER curves of: (a) the packet payload in back-to-back; (b) the packet payload after the first link; (c) the packet payload after the wavelength conversion; (d) the packet payload after the second link. Eye diagrams of a demultiplexed basic channel from: (e) the packet payload in back-to-back; (f) the packet payload after the first link; (g) the packet payload after the wavelength conversion; (h) the packet payload after the second link.
Double Wavelength Conversion with Multi-Resonant Saturable Absorber-Based Vertical-Cavity Semiconductor Gate

C. Porzi\textsuperscript{1}, L. Poti\textsuperscript{2}, A. Bogoni\textsuperscript{2}, L. Orsila\textsuperscript{3}, and M. Guina\textsuperscript{3}
\textsuperscript{1}Scuola Superiore Sant’Anna, Via Moruzzi 1, 56124 Pisa, Italy
\textsuperscript{2}CNIT, Via Moruzzi 1, 56124 Pisa, Italy
\textsuperscript{3}Optoelectronics Research Centre, Tampere University of Technology, Tampere, Finland
email: claudio.porzi@sssup.it

Abstract—We report simultaneous wavelength conversion at two different wavelengths using a multi-resonant, all-optical vertical-cavity semiconductor gate. The device is compact, passive, polarization independent and ensures wavelength conversion with high extinction ratio and low OSNR penalty.

Cost-effective and easy to implement solutions enabling wavelength conversions are thought to be key elements for increasing flexibility, throughput, and transparency. For such purposes, semiconductor devices seem to be the most promising candidates; their main advantages are compactness, low power operation, and relatively fast response times. In recent years, beside implementation based on the popular semiconductor optical amplifiers (SOAs), wavelength conversion with passive nonlinear gates exploiting saturation of absorption in semiconductor multiple-quantum-wells (MQWs) has attracted increased attention \cite{1}–\cite{3}. In particular, in \cite{3} we have demonstrated that a vertical-cavity semiconductor gate (VCSG) comprising MQWs embedded in a resonant asymmetric Fabry–Pérot (FP) cavity can be highly effective in converting external input data to a wavelength close to one cavity resonance. Good quality for the wavelength converted data over a relative broad range of input signal wavelengths was demonstrated. In this work, we extend the same principle to a multi-resonant (MR) FP VCSG to simultaneously convert the input data pattern into two output wavelengths. The operation could in principle be extended to a higher number of wavelengths. Our demonstration shows the potential of this technology for effective all-optical broadcasting with a passive device.

In our experiment we have used a VCSG grown by solid-source molecular beam epityoxy on n-type InP (100) substrate. The sample consisted of Burstein–Moss shifted distributed Bragg reflector (DBR) with 19.5 pairs of n+-Ga\textsubscript{0.47}In\textsubscript{0.53}As/InP \cite{4}, a 480-nm InP spacer layer, an active region, and a \SI{11.1}{\mu m} InP cap layer. The active region was comprised of four groups of seven Ga\textsubscript{0.47}In\textsubscript{0.53}As QWs with a thickness of 9 nm and 10 nm thick InP barriers. The QWs have been centred at the antinodes of the FP cavity defined between the DBR and top surface of the gate. The DBR reflectivity was over \SI{96}{\%} in the wavelength range from 1525 to 1610 nm. The DBR reflectivity and the low-intensity reflectivity spectrum of the gate are depicted in Fig. 1, where the presence of several resonances with a free spectral range of \SI{25}{nm} can be seen.

Fig. 1: Low-intensity reflectivity spectrum for the MR-VCSG (dotted line) and DBR reflectivity spectrum (solid line).

To increase the nonlinear effects, a \SI{63}{\%} reflective dielectric mirror was deposited on the top of the device \cite{5}. The experimental set-up we adopted for demonstrating double wavelength conversion using the MR-VCSG is shown in Fig. 2. Two external-cavity semiconductor tunable laser sources (TLSs) delivered the probe CW signals at two different wavelengths using a multi-resonant, all-optical vertical-cavity semiconductor gate. The operation could in principle be extended to a higher number of wavelengths. Our demonstration shows the potential of this technology for effective all-optical broadcasting with a passive device.

In our experiment we have used a VCSG grown by solid-source molecular beam epityoxy on n-type InP (100) substrate. The sample consisted of Burstein–Moss shifted distributed Bragg reflector (DBR) with 19.5 pairs of n+-Ga\textsubscript{0.47}In\textsubscript{0.53}As/InP \cite{4}, a 480-nm InP spacer layer, an active region, and a \SI{11.1}{\mu m} InP cap layer. The active region was comprised of four groups of seven Ga\textsubscript{0.47}In\textsubscript{0.53}As QWs with a thickness of 9 nm and 10 nm thick InP barriers. The QWs have been centred at the antinodes of the FP cavity defined between the DBR and top surface of the gate. The DBR reflectivity was over \SI{96}{\%} in the wavelength range from 1525 to 1610 nm. The DBR reflectivity and the low-intensity reflectivity spectrum of the gate are depicted in Fig. 1, where the
minimized by using a notch filter centred at the pump wavelength.

Fig. 2: The set-up for the double wavelength conversion experiment.

Fig. 3 shows the optical spectra at MR-VCSG output with and without the pump data signal applied. It can be seen that in the absence of pump signal the probe fields were strongly attenuated. When pump was applied, the absorption in the MQWs was further saturated and the reflectivity experienced by the probe fields was increased. The measurements revealed an average reflectivity change of $\sim 18$ dB for both probe beams.

Fig. 3: Optical spectra at the output of the MR-VCSG with pump Off (solid line) and On (dashed line).

After being amplified and filtered, the two probe signals at the MR-VCSG output were sent to a 125 MHz photoreceiver and monitored on a sampling oscilloscope. The oscilloscope traces, showing the input and the converted eye diagrams, are shown in Figure 4. An extinction ratio larger than 10 dB was measured at both converted wavelengths. Finally, a BER vs. OSNR measurement was performed for the input and converted data. The OSNR could be varied by means of a local ASE source and was measured on an OSA within a 0.1 nm bandwidth. The results, shown in Fig. 4, indicated an OSNR penalty (at BER=10$^{-9}$) of 1.9 dB and 2.7 dB for the 1541 nm and 1566 nm signals, respectively. The difference in the OSNR penalty is ascribed to the different EDFA related noise figures for the two output wavelengths.

Fig. 4: BER vs. OSNR for the input and output converted data.

These results show the potential of MR-VCSG to perform effective multiple wavelength conversion with passive, polarization independent, cost-effective vertical-cavity technology for all-optical networking and broadcasting applications. Advanced designs will be implemented to reduce the spacing between adjacent channels and to increase the operation speed.

REFERENCES


Frequency-dependent electric dc power consumption in ultrafast all-optical semiconductor gates

Jun Sakaguchi¹, Ferran Salleras¹ and Yoshiyasu Ueno¹

¹University of Electro-Communications, Dept. of Electronics Engineering, 1-5-1, Chofugaoka, Choho, Tokyo 182-8585, Japan, Tel: +81-424-43-5207, Fax: +81-424-43-5207, Email: sakaguchi@ultrafast.ee.uec.ac.jp

Abstract: We developed a model of the power consumption in the all-optical gates in frequencies from 10 to 160 GHz. In it, we assumed a realistic electron-photon conversion efficiency that we experimentally evaluated from up-to-date series of SOA samples.

1 Introduction

All-optical signal-processing gates based on nonlinearities of semiconductor optical amplifiers (SOAs) are promising devices to realize future OTDM networks, for their capability of ultrafast signal processing [1-5] and low power-consumption [6]. As its capability of ultrafast gating has become clear, the lower limit of electric power consumption and its origin have been important issues from a physical and design viewpoint. There has been no method to design the electric power consumption requirement, however, with practical reliability and simplicity. Also no systematic measurement has been reported, to the best of our knowledge.

For the power consumption, we have insisted on that conversion efficiency from injected-carrier to photon in the SOA should have significant role [7]. In this report we propose a new method to estimate the amount of power consumption of an SOA on the ultrafast gating, from its fundamental parameters including conversion efficiencies. Then we will show measured results of conversion efficiencies for our SOA samples with different structures. Using these results we calculated SOA carrier dynamics, and power consumption requirement. Part of the calculated results will be compared with measured results.

2 Estimation method of the electrical power consumption

In our SOA model [7], the injected carriers suffer losses before stimulated recombination, as shown in Fig. 1. Since our target frequency is 160 GHz or higher, the third process should be taken into account. Then total efficiency ηT depends on the width of the control pulse. To describe the carrier dynamics of the SOA including these effects, we expanded the rate equation that we used in the former analysis [8,9] to the form below:

\[
\frac{dn_{\text{pulse}}}{dt} = \frac{1}{qV} \eta \eta; \eta \eta_{\text{pulse}} \frac{1}{\tau_c} - n_{\text{pulse}} \left[ \exp \left( \frac{1}{L} \frac{dg_{\text{pulse}}}{dn_{\text{pulse}}} \right) - 1 \right] \frac{P_{\text{pulse}}}{h\nu V}
\]

(1)

\( n_{\text{pulse}} \) carrier density available for ultrafast gating, \( I_{\text{OP}} \) injection current, \( \eta \sim \eta; \eta \sim \eta_{\text{pulse}} \) conversion efficiencies, \( \tau_c \) carrier lifetime, \( P_{\text{pulse}} \) and \( P_{\text{cw}} \) input light intensities, \( V \) and \( L \) active region volume and length, \( q \) elementary charge, \( f \) confinement factor, \( dg_{\text{pulse}}/dn \) differential gain). Note that \( \eta \) is multiplied to the cw-light term, in view of the difference between \( n_{\text{pulse}} \) and \( n_{\text{cw}} \) (carrier density available for cw light). Both cw gain and pulse gain are supposed to be determined by \( n_{\text{pulse}} \), since we did not observe much difference between them unless gain saturations occur.

Injection of \( P_{\text{pulse}} \) causes carrier recombination \( \Delta n_{\text{pulse}} \), and nonlinear phase shift

\[
\Delta \phi = k_0 \int n_{\text{pulse}} \Delta \nu \Delta n_{\text{pulse}}.
\]

(2)

Injection of holding beam \( P_{\text{cw}} \) accelerates carrier recovery after depletion by control pulses, at the cost of \( n_{\text{pulse}} \) and \( \Delta \phi \). Since carrier recovery rate \( 1/\tau_{\text{eff}} \) should be larger than the operation frequency \( B \) to avoid pattern-dependent intensity noise and \( \Delta \phi \) should be as large as \( -0.3\pi \) for an XPM-based gate [9], maximum frequency can be determined for given \( I_{\text{OP}} \) and SOA parameter sets. Numerical simulation shows that the maximum frequency is almost proportion to the total conversion frequency \( \eta_{\text{T}} \) and \( dg_{\text{pulse}}/dn \), but less sensitive to the intrinsic carrier-recovery rate \( 1/\tau_c \).

3 Conversion efficiencies of the SOA samples

To observe the influence of SOA structure on the power

Figure 1. Loss model of the injected carriers and definition of the conversion efficiency
consumption, we studied several custom-designed SOA samples made by two manufacturers A and B. The SOAs in each series have same active-region width w, thickness d, g but different L. The measured conversion efficiencies are summarized in Fig. 2. We see large dependence of total efficiency ηt on the SOA structure. It resulted mostly from ηt. A-series and B-series show opposite ηt dependence on L. Nevertheless, longer SOAs tend to show larger ηt as IOP increases, due to the contribution of ηt.

4 Calculation of holding-beam effect with our model
Before the power estimation based on the model in section 2, we checked to how extent our SOA model can explain the change of SOA property caused by the holding beam. Fig. 3(a)–3(C) show the example of measured and calculated carrier recovery rate, gain saturation against ultrashort pulses, and nonlinear phase shifts. We see that measured and calculated results approximately agree with each other, though for some quantities accuracies are not quite good (~5dB) at the moment.

5 Electric power consumption in SOA
Fig. 4 shows the power consumptions P_Op calculated for each SOA samples, against carrier recovery rate 1/τ_eff. We see that B-series show better performance than A-series. When P_Op is small, B#3 shows best performance around 20 GHz. As the target frequency increases over 100 GHz, B#1 sample becomes preferable. For A-series, longer sample has higher frequency and shortest sample could not be used for the gating. These characteristics result from those of ηt discussed in section 3. Except for the difference of scale factors, most of the P_Op/1/τ_eff profiles in Fig. 4 show some resemblance to corresponding P_Op/1/τ_C profiles. It is due to complicated correlations of SOA parameter changes. Measured P_Op for B#3 sample were also plotted in the same figure, and they agree quite well with the calculated results.

6 Conclusion
We proposed a new model of the electric power consumption in the SOA used for ultrafast all-optical gating, from SOA fundamental parameters. The power consumptions for actual SOA samples were estimated in good agreement with measured results. Due to their larger efficiencies, longer SOAs are expected to consume less electric power when operation frequency is around 100 GHz. This method will provide the fundamental of all-optical gate design.

7 Acknowledgement
We gratefully thank Dr. Koosuke Nishimura and Mr. Tonomori Yazaki in KDDI Research Laboratories for valuable technical discussions.

8 References
Wavelength Conversion Characteristics of SOA-MZI Based All-Optical NRZ-OOK/RZ-BPSK Modulation Format Converter

Suresh M. Nissanka 1, Ken Mishina 1, Akihiro Maruta 1, Shunsuke Mitani 2, Kazuyuki Ishida 2, Katsuhiko Shimizu 2, Tatsuo Hatta 1,2,3, Ken-ichi Kitayama 1

[1]Graduate School of Engineering, Osaka University, 2-1 Yamada-oka, Suita, Osaka 565-0871 Japan, TEL: +81-6-6879-7728, FAX: +81-6-6879-7688, suresh@pn.comm.eng.osaka-u.ac.jp

Abstract Wavelength conversion characteristics of SOA-MZI based NRZ-OOK/RZ-BPSK modulation format converter was investigated experimentally. The error free operation in the whole range of C-band has been successfully demonstrated.

Introduction Optical communication systems mainly employ conventional on-off keying (OOK) signals in either non-return to zero (NRZ) or return to zero (RZ) signal format. Recently, differential phase-shift keying (DPSK) modulation formats have been extensively studied for their high robustness for fiber nonlinearities, which enhance the performance of the long haul transmission [1],[2]. Therefore, it is likely that DPSK modulation formats would be employed in the long haul backbone network, whereas OOK modulation formats would be employed in the metro area networks (MAN) in future optical networks. To transparently connect the long haul backbone network and the MANs, all optical modulation format conversion becomes a key technique. An all-optical NRZ-OOK/RZ-BPSK (Binary PSK) modulation format conversion using semiconductor optical amplifier based Mach-Zehnder interferometric(SOA-MZI) wavelength converter was proposed and experimentally demonstrated with 10.7Gb/s operation [3]. However, to flexibly connect different networks operating with different wavelength ranges, all-optical wavelength conversion becomes an important technique. In this paper we experimentally investigate the wavelength conversion characteristics of the proposed NRZ-OOK/RZ-BPSK modulation format converter.

Experiment Gain spectrum of the SOAs used in the experiment are shown in Fig. 1. 400mA current was injected and CW light was launched into the SOA to measure the gain spectra. The 3dB gain range of SOA1 is from 1523.0nm to 1570.2nm and 1529.0nm-1565.0nm for SOA2. SOAs’ gain spectrum are major limiting factors for the NRZ-OOK/RZ-BPSK modulation format and wavelength conversions.

The experimental set-up is shown in Fig. 2. The NRZ-OOK data signal that acts as the control pulse was generated by modulating a CW light at 1548.1nm in a lithium niobate (LN) modulator with a 10.7Gb/s, 2\(^{11}\)-1 long pseudo random binary sequence(PRBS). RZ clock pulse with a duty cycle of 50% was generated by modulating the CW light in an LN modulator by using the clock recovered from the control pulse. Each input polarization was optimized by a polarization controller (PC). The RZ clock pulse was coupled with a CW assist light at 1551.5nm. Control pulse and probe pulse were synchronously launched into SOA1. Using the cross phase modulation (XPM) in SOA1, the intensity modulated data was transferred to phase modulated data. Due to the simultaneous cross gain modulation (XGM) in SOA1, the output RZ-BPSK signal consists of different peak intensity levels depending on the “0” or “1” in OOK data signal. SOA2 was used to adjust the peak intensity level of the converted RZ-BPSK signal thanks to in-phase or anti-phase interference. We changed the wavelength of the RZ clock pulse from 1530nm to 1569nm with approximately 5nm intervals and conducted the modulation format and wavelength conversions. SOA1 was driven at 400mA injection current. We adjusted the phase of the probe pulse on the lower arm of the MZI by the phase shifter after fixing the injection current of the SOA2. After amplified by EDFA4, the converted RZ-BPSK signal was demodulated using a 1 bit delay interferometer and a balanced receiver.

The signal waveforms observed by the balanced receiver are shown in Fig. 3 for wavelengths of 1530.33nm, 1546.12nm, 1555.75nm, and 1569.59nm. Clear eye opening was observed for all wavelengths.
Fig. 3. Signal waveforms observed by the balanced receiver. Figure 4 shows the measured bit error rate (BER) for the above-mentioned wavelengths. The linearity of the BER curves indicates the normal operation of the modulation format and wavelength conversions and error free operation can be confirmed for all wavelengths. At 1530.33nm wavelength, 3.45dB power penalty was observed for BER of $10^{-9}$. This power penalty is induced by the noise characteristics of EDFA4.

<table>
<thead>
<tr>
<th>Wavelength [nm]</th>
<th>OSNR [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1525</td>
<td>14</td>
</tr>
<tr>
<td>1535</td>
<td>13</td>
</tr>
<tr>
<td>1545</td>
<td>12</td>
</tr>
<tr>
<td>1555</td>
<td>11</td>
</tr>
<tr>
<td>1565</td>
<td>10</td>
</tr>
<tr>
<td>1575</td>
<td>9</td>
</tr>
</tbody>
</table>

Fig. 5. Received OSNR for BER of $10^{-9}$.

**Conclusions**

In this paper, we have experimentally investigated the wavelength conversion characteristics of the SOA-MZI based all-optical NRZ-OOK/RZ-BPSK modulation format converter. The maximum OSNR penalty in the whole range of C-band is only 0.96dB, and it is fairly small.

**References**

Modeling of Optical Nonlinear-Effect Compensator With Vertical-Cavity Saturable Absorber
Satoshi Suda, and Fumio Koyama
Microsystem Research Center, P&I Lab., Tokyo Institute of Technology
4259 Nagatsuta, Midori-ku, Yokohama, 226-8503 JAPAN
Phone. +81-45-924-5026, Fax. +81-45-924-5961,
E-mail: suda.s.ab@m.titech.ac.jp

Abstract: We present the modeling of the optical compensation for fiber nonlinearity in 40 Gb/s long-haul transmission using a vertical-cavity nonlinear-effect compensator. The modeling result shows that the self-phase modulation and associated waveform distortion can be reduced.

1. Introduction
Linear and nonlinear propagation effects in fibers primarily limit the performance of long-haul fiber transmission systems for 40 Gb/s or beyond. We proposed and demonstrated a vertical-cavity nonlinear-effect compensator for optical compensation of fiber nonlinearities [1]. The proposed device gives a large intensity-dependent phase shift, which enables us to compensate optically self-phase modulation in fibers.

In this paper, we present the modeling on the nonlinear phase shift of a vertical cavity saturable absorber. The effect on the waveform distortion due to the self-phase modulation is analyzed. The proposed device would be useful for mitigating fiber nonlinearities in optical domain.

2. Principle and Modeling
Figure 1 shows the schematic structure of nonlinear etalon for optical nonlinear compensation [1]. This nonlinear etalon consists of 3 pair SiO₂ / Ta₂O₅ (top-DBR) and 40 pair InGaAlAs/InP (bottom-DBR). An intensity-dependent refractive index change takes place due to photo-carriers in the saturable absorber, which is enhanced by a resonant cavity. InP/InGaAlAs is sandwiched and serves as the saturable absorber region. If the input light power through this device increases, the phase difference can occurs resulting from refractive index changes in the saturable absorber. The absorption coefficient of saturable absorber, α, can be approximated by α=α₀/(1+Iin/I₀), I₀=ḥωτ/τA, where Iin, α₀, A denote, absorbed power density in saturable absorber, low-intensity-material absorption coefficient, and material-dependency absorption cross section, respectively. The saturation coefficient I₀ is in inverse proportion to carrier lifetime τ and related to the model of the intensity dependent absorption shown in [2]. The reverse bias can be applied in order to reduce the carrier lifetime for high bit rate signals. Also, the phase induced by self-phase modulation (SPM) in fibers is proportion to light intensity. For example, the intensity dependence of the phase induced by self phase modulation (SPM) is a +0.14 radian at 5 mW input light for the wavelength of 1550 nm [3]. The proposed compensator can be included in each repeater as shown in Fig. 2. The SPM in each fiber span can be optically compensated by our compensator.

3. Result
We calculated the phase induced by SPM for 100 km fiber transmission with a nonlinear Schrödinger Equation. Figure 3 shows the phase of the output light
after 100 km fiber transmission with dispersion compensator and optical nonlinear-effect compensator for dispersion-shifted fiber (DSF) and large effective area fiber (LEAF). We assumed an input pulse of 5 mW peak power and the parameters of these fibers are shown in Table 1. The phase induced by SPM can be compensated completely in case of DSF. On the other hand, the phase induced by SPM cannot be perfectly compensated in case of LEAF because of the interaction between chromatic dispersion and SPM.

Figure 3 shows the eye diagram of NRZ modulation formats for 500 km transmission (5 spans) with LEAF with and without optical nonlinear-effect compensator (N=5 in Fig.2). The result shows that SPM induced waveform distortion can be avoided and an extinction ratio can be improved by 3.5 dB with the nonlinear-effect compensator.

### Table 1 Parameters for certain fibers

<table>
<thead>
<tr>
<th></th>
<th>LEAF</th>
<th>DSF</th>
</tr>
</thead>
<tbody>
<tr>
<td>$D$ [ps/nm/km]</td>
<td>4.0</td>
<td>0.0</td>
</tr>
<tr>
<td>$S$ [ps/nm²/km]</td>
<td>0.090</td>
<td>0.077</td>
</tr>
<tr>
<td>$A_{\text{eff}}$ [μm²]</td>
<td>72</td>
<td>52</td>
</tr>
<tr>
<td>Loss [dB/km]</td>
<td>0.21</td>
<td>0.25</td>
</tr>
</tbody>
</table>

(D: Dispersion / S: Dispersion slope / $A_{\text{eff}}$: effective area)

4. Conclusion

We presented the effect of the optical compensation of SPM for 40Gbps fiber transmission with a vertical cavity nonlinear-effect compensator. We found the SPM in LEAF can be reduced by a factor of 7 with the optical nonlinear compensation. We exhibited improvement of +3.5 dB in an extinction ratio from calculated eye diagrams with a nonlinear Schrödinger equation.

Reference


Abstract

Recent results on photonic device research are introduced which are conducted as a research theme in MEXT funded project of Priority Areas named INGOC, i.e., Innovation for New Generation Optical Communications.

1. Introduction

A number of subscribers who can access to the networks with a bandwidth of 10 Mb/s or higher exceeded 25 millions at September 2006 in Japan[1]. Among the broadband access such as ADSL, FTTH and CATV, fiber optic version (FTTH) has shown a stronger growth rate than the other schemes. A positive feedback between the growth of the high volume contents, such as movies, pictures, and music, and the growth of the broadband access capability has led to the strong increase in the information traffic. The trend seems to continue and to be further enhanced, because transition of the contents toward those with the higher quality, such as high-density TV(HDTV) and in a future Super Hi-Vision[2] will likely to appear. This situation will induce strong motivation to realize networks with an extraordinary higher performance. Although photonics has been contributing to construct global information and communication networks through the technological breakthrough of optical fiber communications since 1970’s, extensive efforts are still required to cope with the issues where two to three orders of magnitude increase in the network performance is anticipated. As one of the efforts, a new project named INGOC, Innovation for New Generation Optical Communications has started since 2005. The conceptual image is depicted in Fig. 1 with the project logo mark. It is a project funded by MEXT (Ministry of Education, Culture, Sports, Science and Technology) as a Grant-in-Aid for Scientific Research on Priority Areas. In this talk, recent results are introduced on the photonic device research in this project.

Fig.1 The conceptual image of the project

2. About the Project

The project consists of three research groups and an executive committee. The purpose is to create a new area of science and technology and to achieve innovation for the transmission and the signal processing for the future optical communication networks through breakthrough of photonic devices. In this project, we would like to explore new functions of light which have not been yet fully utilized. Among them are phase, polarization, quantum state of light, light velocity and so on. Twelve principal research topics are divided into the three research groups, i.e., A01: Function Innovation Group, A02: Structure Innovation Group and A03: Integration Innovation Group. These names do not
look popular. In a more usual way, they can be called fundamental, device and system-oriented research groups for A01, A02 and A03, respectively.

3. Photonic Device Breakthrough

As examples of photonic device breakthrough in our project, the principal research topics are shown schematically in Figs. 2 and 3. In A01, several approaches have been tried to explore the potential of light toward the new generation optical communication. Twin photons with cross-polarization state were generated using a Ti: LiNbO3 waveguide with domain-inverted grating pumped by a semiconductor laser[3]. Single photon and photon-pair generation were achieved with an InAs quantum dot covered by GaNAs[4]. Slow light effects in a photonic crystal waveguide with a chirped structure was investigated to apply to an optical buffer memory[5]. Ultra-fast optical non-linearity of optical fibers was utilized to produce ultra-short pulse and very wide continuum spectrum[6].

In A02, emphasis is laid on photonic device innovation. Waveguide ring resonators were utilized to devise a multi-wavelength multi-port wavelength selective switching circuit for RODAM applications[7]. A widely tunable dispersion compensator was realized by combining a unique hollow waveguide with MEMS technology[8]. Ultra-fast all-optical switching was demonstrated in InGaAs/AlAs/AlAsSb multi-quantum well structure by using inter-sub-band transition[9]. A variety of optical devices such as tunable filters and reflectors were fabricated by exploiting MEMS technology[10].

4. Concluding Remarks

Due to the limited space, the research themes of the candidates for the photonic device breakthrough in our project are only briefly described in this paper. In the talk, the latest results of these themes will be addressed in addition to those of system oriented topics such as digital coherent optical receiver[11], multi-port optical spectrum analyzer[12], and all-optical signal processing and buffer memory[13][14].

References

Fig. 2 Principal research topics of A01 group

Fig. 3 Principal research topics of A02 group
Monolithically Integrated Tandem Waveguide-Type Acoustooptic Frequency Shifter Driven by Surface Acoustic Waves

Shoji Kakio*, Motoki Kitamura*, Yasuhiko Nakagawa*, Takefumi Hara†, Hiromasa Ito‡, Tetsuya Kobayashi† and Masayuki Watanabe†

*Interdisciplinary Graduate School of Medicine and Engineering, University of Yamanashi, Takeda-4, Kofu 400–8511, Japan
†Research Institute of Electrical Communication, Tohoku University, Katahira-2, Aobaku, Sendai 980-8577, Japan
‡Optoquest Corporation, Haraichi–1335, Ageo-shi 362–0021, Japan

Abstract—A monolithically integrated tandem waveguide-type acoustooptic frequency shifter driven by surface acoustic waves was fabricated on 128° Y-cut LiNbO₃ for an optical wavelength of 1.55 μm. A peak doubly diffraction efficiency of 63% was obtained.

I. INTRODUCTION

A new type of laser, called the frequency-shifted feedback (FSF) laser, with unique spectral properties has been developed and has received attention owing to its potential applications[1]-[7]. The FSF operation is achieved by feedback of the first-order diffracted light of an intracavity acoustooptic frequency shifter (AOFS), and the FSF laser output consists of periodically generated chirped frequency components whose chirp rate is higher than 100 PHz/s[2].

To realize a compact and stable FSF fiber laser, we have proposed a waveguide-type AOFS driven by surface acoustic waves (SAWs) using guided-optical waves in a tapered crossed-channel proton-exchanged (PE) optical waveguide on a 128° rotated Y-cut LiNbO₃ substrate for an optical wavelength of 1.55 μm[8]. An 84% diffraction efficiency was obtained for an AO interaction region length of 2 mm and a driving frequency of 200 MHz[9]. The wide frequency-shift range makes it attractive for various applications in optical measurements.

In this study, the monolithically integrated tandem AOFS driven by SAWs was fabricated and the diffraction properties were measured. Moreover, an optical frequency domain ranging using the FSF fiber laser with the tandem AOFS was demonstrated.

II. DESIGN AND FABRICATION OF TANDEM AOFS

The overall configuration of the monolithically integrated tandem waveguide-type AOFS is shown in Fig. 1. Several waveguide-type AOFSs driven by SAWs in the tapered crossed-channel PE optical waveguide are connected in tandem on a 128°-rotated Y-cut LiNbO₃ substrate. The structure considered is a 2×4 optical switch consisting of one 2×2 and two 1×2 switches. The input ports of the two second 1×2 switches are connected to the two output ports of the first 2×2 switch on the same substrate. Using this structure of the tandem AOFS, the optical frequency shifts for the sum of two driving frequencies and the difference between two driving frequencies can be obtained by combining the upward and downward frequency shifts in the first and second SAWs.

The waveguide shape was designed using a beam propagation method (BPM). To increase the length of the AO interaction region, the 10 μm width of the single-input four-output waveguide is increased to 100 μm using a tapered waveguide 3.3 mm long. The first and second switches are connected together with the 100-μm-wide straight waveguide.

The fabrication conditions of the proton exchanged (PE) waveguide were the same as those given in the previous paper[10]. An interdigital transducer (IDT) with a period length Λ of 32 μm and an overlap length of 2 mm was formed in both steps. The substrate size after polishing the end face was 32 mm × 16 mm.

III. MEASURED DIFFRACTION PROPERTIES

Diffraction properties were measured using a 1.55 μm laser diode at a driving frequency of approximately 120 MHz. The maximum values of the diffraction efficiency of 92% and 83% were obtained by the first and second SAWs, respectively. Figure 2 shows the dependences of the diffraction properties on the input voltage supplied to the second SAW at the maximum diffraction efficiency by the first SAW. The peak diffraction efficiency of 63% was obtained by driving both steps at the same frequency.
IV. OPTICAL FREQUENCY DOMAIN RANGING USING FSF FIBER LASER

The experimental setup of optical frequency domain ranging using the FSF fiber laser with the tandem AOFS is shown in Fig. 3. The tandem AOFS was incorporated into a module and optically connected to the input polarization-maintained (PM) and output single-mode fiber arrays through collimating and condensing lenses. The erbium-doped fiber (EDF) was used as a gain medium. The tandem AOFS was driven by the first and/or second SAW. The light in the cavity is frequency shifted every round trip by feeding the diffracted light of the tandem AOFS back into the EDF. The FSF fiber laser output was split into the two arms of the Mach-Zender interferometer. The adjustable path difference was obtained by including the propagation length of spatial light to one arm. The beat signal was detected with a photodetector and the beat spectrum was analyzed using an RF spectrum analyzer. Figure 4 shows the beat frequency change as a function of the path length change. The linear relationships between the beat frequency change and the path length with three slopes were obtained. One is due to the optical frequency shift of 120 MHz generated by driving only the first or second SAW. The other two slopes correspond to the optical frequency shifts for the sum of two driving frequencies (244 MHz) and the difference between two driving frequencies (3 MHz), respectively.

V. CONCLUSION

A monolithically integrated tandem waveguide-type AOFS driven by SAWs was designed and fabricated for an optical wavelength of 1.55 μm. A peak diffraction efficiency of 63% was obtained by driving both steps. Moreover, the optical frequency domain ranging using the FSF fiber laser with the tandem AOFS was demonstrated.

ACKNOWLEDGMENT

This work was supported by the Cooperative Research Project of the Research Institute of Electrical Communication, Tohoku University and the Strategic Information and Communications R&D Promotion Program of the Ministry of Public Management, Japan.

REFERENCES


Email: kakio@yamanashi.ac.jp
Temperature Dependence of Long Term Modulation Applied to Electro-optic Polymer Waveguide

Shigetoshi Nakamura, Takashi Kikuchi, Roshan Thapliya
Corporate Research Group, Fuji Xerox Co., Ltd. 2274, Hongo, Ebina-shi, Kanagawa 243-0494, Japan
E-mail: Shigetoshi.Nakamura@fujixerox.co.jp

Abstract: We evaluated the long-term stability of EO-polymer based Mach-Zehnder modulator at 25°C, 50°C and 70°C. We confirmed the existence of phase-drift and derived the activation-energy to be equal to 0.52 eV.

Introduction
Electro-optic (EO) polymers have attracted renewed interest as practical material platform for realistic devices. One of the major issues of these materials is long-term behavior and stability. There have been reports on the $\xi$ relaxation and DC-drift for organic\(^1,2\) as well as inorganic materials\(^3-5\). However, important factors, which are necessary for long-term extrapolation, such as Arrhenius curves, activation energies, have not been sufficiently studied.

In this talk, we report for first time, to the author’s knowledge, the temperature dependence of the drift of EO polymer based Mach-Zehnder modulator (MZM) and discuss on the feasibility of this material system with respect to commercially available lithium niobate (LN) based MZM.

Design

As shown in Fig. 1 (a), the modulator consists of a silicon substrate, a bottom electrode, a lower clad layer, a core layer with a rib-structure etched on top of it, an upper clad layer and, finally, the top electrodes. We used chromophores called IPC-E\(^6\). The IPC-E was dispersed in the host material polysulfone (PSU) such that 20wt% loading density of IPC-E was mixed in PSU. We fabricated the devices using a standard semiconductor process scheme, which has been reported in our past works\(^7,8\). The planer structure of the device is shown in Fig. 2 (b). The coupling loss and the refractive index difference, $\Delta n$, for the 0.6μm rib height and 5.0μm channel width is 3.3dB/side and 0.26% (ns=1.6409, nc=1.6451) for the TM-mode, respectively, which is calculated from the refractive indices shown in Fig. 1 (a).

Experiment and Discussion
A commercial laser at 1552nm wavelength provided a continuous wave generation which was launched into a polarization maintaining fiber and connected to a calibrated half-wave plate (HWP). The HWP allowed us to select the TM-mode which was launched into the MZM. The multifunction data acquisition (DAQ) card installed to the personal computer generated a triangular wave form with 10Hz frequency, and the voltage output from the DAQ card was amplified to ±20V without DC bias by high speed bipolar amplifier for driving the MZM. The optical output from MZM was monitored using a power-meter. Both the driving signal and optical response were stored into the personal computer simultaneously using the DAQ card, and subsequently the optical power versus applied voltage was plotted to
produce the MZM output. As shown in Fig.2, the response of our MZM for t=0 and t=80 hrs at T = 25°C, is illustrated in order to define the phase-drift, Δφ(t, T).

![Figure 2: Typical response curves of our MZM at t=0 and t=80 hours at T=25°C](image)

As shown in Fig.3, we present the behavior of the phase-drift dependence on the operation temperatures, 25°C, 50°C and 70°C. It is interesting to see that at room temperature, T=25°C, the drift shifts toward negative side and then reverses for the first hundred hours which is similar to LN devices\(^4\). This effect is also visible in higher temperature T=70°C. It is, therefore, speculated that the cause of this behavior in polymers might be similar to that in LN’s. The details of this phenomenon will be discussed in our talk.

![Figure 3: Δφ(t, T) characteristics for T is 25°C, 50°C and 70°C data respectively](image)

Fig. 4 shows the Arrhenius curves obtained using the results in Fig. 3. The characteristics show a linear property which allow us to deduce the activation energy, \(E_a=0.52\) eV. This value is similar to LN which is approximately 1 eV\(^4\). Therefore, we believe that with optimization realistic device similar to LN-based modulator can be achieve using DC-compensation circuit.

![Figure 4: Arrhenius plot of phase-drift rate.](image)

**Summary**

We have evaluated the phase drift of EO polymer device and found that it is comparable to LN-based MZMs. Although, further repeatability test is necessary, by directly measuring the phase-drift at 25°C, 50°C and 70°C, we deduced the activation-energy to be equal to 0.52 eV. This degree of performance is expected to be sufficient for realistic applications using EO polymer devices combined with commercially available DC-compensation circuits.

**Reference**

2) H. Park, et al, SPIE Vol.2852, 286
All-optical tunable wavelength filter using photorefractive material

Satoshi Honma¹, Mitsuhiro Komatsu¹, Hirokazu Ito¹, Shinozo Muto¹, Terumasa Ito², Atushi Okamoto²
¹ Interdisciplinary Graduate School of Medicine and Engineering, Yamanashi University, Takeda, 4-3-11, Kofu, Yamanashi, 400-8511, Japan,
² Graduate School of Engineering, Hokkaido University, Kita 13-Nishi8, Kita-ku, Sapporo, 060-8628, Japan
Tel: +81-55-220-8412, Fax: +81-55-220-8412, Email: shonma@yamanashi.ac.jp

Abstract
An optical tunable wavelength filter using photo-induced gratings is proposed. It is shown that the central wavelength and the bandwidth of the output signal can be controlled by pure optical operation.

1 Introduction
Optical switching devices and filters have been expected to be developed for all-optical network systems. Photorefractive (PR) effect is known well as one of the highest nonlinear optical effect. It is promising technique to realize all-optical switches and filters controlled by relative low beam power.

In recent years, many researchers have studied about fabrication of photorefractive waveguides. Irradiating two coherent beams on the waveguide, periodical refraction distribution i.e. index grating can be induced in the waveguide. We propose a tunable optical filter by using the photo-induced grating in the PR waveguide. We analyze beam propagation and refractive distribution, and then investigate the filtering property. It is shown the central wavelength and the bandwidth of the output signal is controllable by adjusting the period and the width of grating with pure optical operation.

2 Tunable optical filter with photorefractive waveguide
Fig. 1 shows the schematic diagram of a tunable optical filter. A wavelength division multiplexed optical signal propagates in a PR planar waveguide. Two control beams irradiate on the waveguide. A periodical modulated refractive index is induced in the waveguide corresponding to interference pattern of the control beams. The modulation works as an index grating. The wavelength components in the input signal which satisfy the Bragg condition with the grating is extracted to output port. It is able to adjust a period of a grating by changing incident angel, wavelength or phase distribution of the control beams. The central wavelength of an output signal can therefore be controlled by pure optical operation. It is also possible to control the bandwidth of the output signal by adjusting modulation depth and broadening of the index grating.

3 Diffraction efficiency and bandwidth of optical filter using transmission grating
We investigated diffraction efficiency and bandwidth of the tunable filter using FD-BPM with consideration of permittivity variation via photorefractive effect.

Fig. 2 shows the analysis model. Refractive index of photorefractive materials is modulated dynamically by irradiation of light. We assumed that the signal beam propagates along z-axis, the beam intensity distribution and the electron density distribution change predominately along y-axis. The spatial electric field $E_s$ in the PR material is given by the following equation

$$E_s = \frac{1}{k_D} \frac{\partial^2 E_I}{\partial y^2} + \frac{1}{k_D} \frac{\partial E_I}{\partial y} = -\frac{E_D}{k_D} \frac{\partial}{\partial y} \ln(f)$$  \hspace{1cm} (1)

, where $E_D = k_D T k_0 / e$ is diffusion field[1]. $k_D$ is Debye wave number approximated by $k_D^2 = e^2 N_A (\epsilon_0 k_0 T)$. $N_A$, $\epsilon$ and $\epsilon_0$ are the acceptor density, the electric charge and the permittivity to direct electric field, respectively. $k_0$ is the Boltzmann constant. $T$ is the temperature. $I$ is the intensity distribution of the signal beam and the control beam. The relative permittivity $\epsilon$ is modulated by $E_s$ via Pockels effect which is described by

$$\Delta \epsilon = -\frac{\epsilon \Delta \left( \frac{1}{n^2} \right)}{\epsilon_0} = -\frac{\epsilon_0 r_{eff} E_s}{\epsilon_0}$$  \hspace{1cm} (2)

, where $n$ and $r_{eff}$ are the refractive index and the effective Pockels coefficient, respectively. On the other hand, applying the Fresnel approximation to Helmholtz equation, we obtain the wave equation as

$$\frac{\partial \phi}{\partial z} = \frac{1}{2 j \beta} \left( \frac{\partial^2}{\partial x^2} + k_0^2 \epsilon_s + \Delta \epsilon - n_{eff}^2 \right) \phi$$  \hspace{1cm} (3)

, where $\beta$ is the propagation constant toward z-axis, $k_0$ and $n_{eff}$ are the free space wavenumber and the effective refractive index, respectively. Applying FD-BPM method for wide-angle propagation with Douglas scheme to Eq. (3) and the finite difference method to Eq. (1), we calculated the beam propagation in the PR material.

We assumed a KNbO$_3$ crystal for a photorefractive material. The input signal beam was regarded as Gaussian beam whose maximum value of intensity $I_s$ and...
radius are 1.0[W/cm²] and 0.5[mm], respectively. The incident angle is \( \theta_i = 5.5^\circ \). On the other hand, the intensities of control beams are equal. We gave the intensity ratio of the control beam to the signal beam as \( I_c / I_s = 5 \). An index grating which satisfies the phase matching condition with a input signal beam whose wavelength is \( \lambda = 532 \text{ nm} \) is induced by the control beams at spatial range of \( z = 0 \text{[mm]} \) and \( z = 3.8 \text{[mm]} \).

\[
\begin{align*}
\theta_{s,z} &= 45^\circ \\
\theta_i &= 5.5^\circ
\end{align*}
\]

Fig. 2 Analytical model for calculation

Fig. 3 shows the calculation results where (a) and (b) indicate the intensity distribution of the signal beam and the relative permittivity distribution, respectively. Not only grating induced by the control beams but also that induced by the input signal and its diffracted beam exists in the material. In case that the signal beam is diffracted toward c-axis, the two index gratings superpose in phase. As a result, the diffraction efficiency is improved because of amplification of index grating. The diffraction efficiency has the maximum value 0.92 under the assumption that the irradiation length of the control beams is equal to 3.8[mm]. If the irradiation length is larger than the optimum value, the diffraction efficiency decreases as shown in Fig. 4.

Fig. 4 shows the relationship of the diffraction efficiency and the full width at half maximum (FWHM) of the output signal against the irradiation length of the control beams. In this calculation, we assumed that the total intensity of the control beams with respect to the intensity of the signal beam is steady value. We gave the intensity ratio of the two control beams as \( I_c / I_c = 1, 5 \) and 10. Index modulation in photorefractive crystals is proposal to intensity contrast ratio of interference fringes. Therefore it makes possible to control the modulation depth of the index grating by changing the intensity ratio between the two control beams. The optimum interaction length for high diffraction efficiency increase with increasing the intensity difference between the two control beams because the modulation depth of the grating become small. We found that the irradiation length for the minimum FWHM almost coincide with the value at which the diffraction efficiency have the maximum value. Therefore it is possible to control the bandwidth of the output signal without significantly decreasing the diffraction efficiency by adjusting the exposure region and the intensity ratio of the control beams. The minimum values of FWHM are 2, 1.8 and 1.5[mm] at \( L = 5, 6 \) and 6[mm] in case of \( I_c / I_c = 1, 5 \) and 10, respectively. The maximum diffraction efficiencies are 0.91, 0.88 and 0.85, respectively at the optimum interaction length.

4 Conclusion

We proposed the tunable optical filter using photorefractive planar waveguide. The central wavelength of the output signal is controllable by adjusting the period of interfering fringe of the control beams. The bandwidth can be controlled by adjusting the intensity ratio and the irradiation length of the control beams.

References


Acknowledgments

This study was supported by development of optical communication module using photorefractive KNbO₃, optical thin-film waveguide in 2006 from New Energy and Industrial Technology Development Organization (NEDO) of Japan.
Integrated OCDM En/Decoder Module for Practical Deployment
Shuko Kobayashi, Kensuke Sasaki and Takashi Ushikubo
Corporate Research & Development Center, Oki Electric Industry Co. Ltd.,
550-5 Higashiasakawa-cho, Hachioji-shi, Tokyo 193-8550 Japan (e-Mail: kobayashi535@oki.com)

Abstract
We have studied a novel module structure integrating multiple optical code division multiplexing (OCDM) encoders, and experimentally demonstrated that the new module meets the practical deployment by evaluating the wavelength tuning characteristics.

Introduction
Optical code label offers applications for optical code division multiplexing (OCDM) and photonic packet switching, due to its all-optical signal processing, adaptable capacity, and robust information security [1,2]. The coherent phase coding employing binary phase shift keying (BPSK) exhibits remarkable improved correlation properties as compared with the incoherent coding [3]. Superstructured fibre Bragg grating (SSFBG) attracts much attention as a BPSK signal generation and recognition component, because of compactness, compatibility with fibre optics, and potentially low cost. In principle, SSFBG en/decoder particularly can handle having long code length, which realizes larger number of multiplexing and higher signal security than other technique, such as planar light-wave circuit devices.

In order to recognize an autocorrelation peak in the OCDM system, an encoder and a decoder should have about the same reflective wavelength in a Pico meter order. Since the reflective wavelength of an encoder changes with a manufacture error or environmental temperature, we have developed the module in which wavelength adjustment in a Pico meter resolution is possible by temperature control. For installing in a device of OCDM-PON system, we have tackled a development of a module that has parallel mounting of two or more encoders for the purpose of space-saving of encoders. In this paper, we introduce a novel module structure integrating up to 85 mm long four SSFBGs. And the wavelength tuning characteristics of novel module are demonstrated.

Design of integrated encoder module
In order to adjust the reflective wavelength of an encoder, uniform temperature control of while SSFBG is required. Since it is necessary to control temperature of each encoder individually when two or more encoders are mounted in parallel, we have developed the module is available for having different setting temperature between adjacent encoder. If temperature distribution in encoder arose by the difference in preset temperature, the characteristic of an encoder would change. In order to design the parallel structure of encoders, the mutual influence on the temperature distribution of each encoder by the encoder interval was studied in thermal analysis of the module structure. Consequently, when the encoder interval was 2 mm or more, it turned out that a temperature distribution is not influenced even if the encoder with which 40 degrees C of setting temperature differ adjoins each other. Moreover, also when setting temperature of encoder differed from 65 degrees C of environmental temperature, it has also studied that the temperature distribution of an encoder was less than 1 degree C in thermal analysis.

Based on these results, the integrated encoder module was manufactured experimentally. Four encoders were integrated in the module. The appearance of a four-encoder module is shown in Fig. 1. The installation area of an encoder module is a half for four conventional single encoders, and has attained space saving.

Fig. 1. Appearance of an encoder module.

Wavelength tuning characteristic
The relation between the setup temperature and reflective wavelength in each encoder of the integrated encoder module is shown in Fig. 2. The wavelength-tuning ratio by the setup temperature of each encoder is about 13 pm/degree, and wavelength control in Pico meter resolution made temperature control of 0.1 degree C possible for them. For example, the wavelength of all encoders can be made in agreement with the arbitrary wavelength between 1549.7 nm and 1549.9 nm by adjusting setup temperature from 15 degrees C before 45 degrees C in the case of this module.
y = 0.0130x + 1549.3771
y = 0.0128x + 1549.4043
y = 0.0127x + 1549.3669
y = 0.0126x + 1549.5210

Fig. 2. Wavelength tuning characteristics of encoder module.

The reflection spectrum of each encoder was measured at changing setup temperature. The correlation coefficient of the measured reflection spectrum was calculated on the basis of the measured reflection spectrum in the setup temperature of 25 degrees C. Even if it changed setting temperature, as shown in Fig. 3, the correlation coefficient of a reflection spectrum was 0.98 or more. When the correlation coefficient of reflection spectrum was 0.9 or more in the case of the encoder mounted in this module, it has checked that the ratio of the peak and sub-peak on autocorrelation waveform hardly decreased.

It turned out that only wavelength of each encoder in this module is changed in Pico meter resolution, without influencing the reflective characteristics. Therefore, this encoder module meets the practical deployment.

Fig. 3. Correlation coefficient of reflection spectrum at the different setting temperature.

Temperature uniformity of each encoder

The temperature distribution state at the time of changing 10 degrees C of setup temperature of adjacent encoder was measured by thermography, as shown in Fig. 4(a), and the temperature plot of a thermal image on each encoder is shown in Fig. 4(b). Even when the setup temperature of an adjacent encoder differed, it has checked that temperature distribution of each encoder was less than 1 degree C.

Fig. 4(a). Thermal image of an encoder module.

Fig. 4(b). Temperature plot of a thermal image.

Conclusions

Integrated encoder module that carries out parallel mounting of four optical code division multiplex (OCDM) encoders for the purpose of space-saving has been developed. The installation area of an encoder module is the half of the four conventional single encoders. Integrated encoder module can perform adjustment of wavelength in Pico meter resolution individually by temperature control for each encoder, without affecting the reflective characteristic by wavelength tuning, and meets practical deployment.

Acknowledgement

This work was supported in part by National Institute of Information and Communications Technology (NICT) of Japan.

References

Optical interconnections integrated in Printed Circuit Boards

P. Van Daele¹, G. Van Steenberge¹, N. Hendrickx¹, E. Bosman¹, P. Karioja², C. Debaes³
¹IMEC-Ghent University, INTEC-TFCG Microsystems, Belgium,
²VTT, Finland, ³Dept. Applied Physics, Vrije Universiteit Brussels, Belgium

ABSTRACT

An overview will be given of different solutions that are being pursued to introduce optical interconnections on a short distance, i.e. 1 m or below and integrated in printed circuit boards.

1. INTRODUCTION

In present and future broadband networks, multigigabit transmission over longer distances is only feasible via optical interconnections that form the very heart of the network. Worldwide ongoing research aims at the extension of the optical interconnect to the board level and to the switching level. In spite of repeated predictions that the all optical interconnect is soon going to replace electrical interconnect on the board level, this turned out not to be yet the case. Several reasons can be given for this, but the two most important ones are that it turned out to be more difficult than expected to integrate optical interconnect in an easy and cost effective way into or onto a board. The second reason is that, as can typically be expected from a technology with a large investment base, the possibilities in terms of maximum bitrates of electrical board interconnect are continuously being upgraded, postponing the need for replacement by optical interconnect. Many systems are rack-based backpanel configurations with interconnection lengths ranging from a few centimeters to a few meters. It is crucial that, in order to be accepted by system engineers and designers and in order to be low cost, the introduction of optical interconnections on this level should be completely compatible with existing board-technology. Therefore optical interconnections should be integrated in FR4-based PCB’s both in view of the optical layer itself as well as the coupling of light to and from this optical layer to optoelectronic components and/or fibers. The same can be said of the interconnections at even shorter distances (on MCM-level) where compatibility should be sought with existing MCM-technology, chip-packaging and Chip-on-Board technologies.

The work and the results of the work described in this paper is carried out at the INTEC Dept. of Ghent University (Belgium) and in the framework of the European funded “Network of Excellence on Micro-Optics” (NEMO, www.micro-optics.org) and especially part of the Workpackage 9 which objective is to investigate the potential of micro-optics for achieving the goal of integrating optical interconnections on printed circuit boards and on MCM-level. As stated above, it is clear that for achieving a low cost solution, compatibility with existing technology should be sought. This implies, in most cases, alignment tolerances of traditional electrical packaging and mounting technologies which are not in line with required optical performances. Therefore micro-optics will play a crucial role in achieving good optical performances, together with existing tolerances within the board manufacturing technology.

When discussing optical interconnections at short distance, one should keep in mind the application one is looking for:

1. the situation where optical signals are arriving into the cabinet from optical fibers coming from long haul interconnections, and which signals have to be switched, repeated or decoded. In this case the arriving signal will be on a wavelength of 1.3 or 1.55 um and the fibers will be SM, requiring similar wavelengths and modal structures on the boards, and

2. the situation where processors on the same or different boards within the same system have to communicate at high bitrates, but over short distances, which will imply a MM structure and most probably also a less stringent requirement on wavelength dependent losses.

2. FREE SPACE OPTICAL INTERCONNECTIONS

Optical interconnections can be realized through free space, and even free space interconnections between boards have been shown. However due to alignment tolerances, this solution is hard to be made cost-effective. Free space optical interconnections are therefore only considered, within NEMO, for inter- and intra-chip interconnections through optical bridges, structures through which the light is traveling non-guided, but which allows good and reliable alignment between the different components. Such solutions find application in massive parallel and high-density interconnections in e.g. Multi-Chip-Modules. These will however not be discussed here.
3. FIBER-BASED INTERCONNECTIONS

In order to realize optical interconnections on a board-level, the first option people have been considered, already many years ago, was to integrate optical fibers into the FR4-stack. This poses however considerable problems regarding integration of deflecting and coupling structures and connectorisation of the optical interconnections. Therefore this option has been used mainly for passive optical backpanels in telecommunication, where no deflection is required and connectors can be fabricated at the end of long lips of fiber ribbons stretching out of the board.

To overcome the problem of deflecting and coupling to embedded fibers, one of the routes followed within NEMO is to use clips inserted in holes in the board where the fiber is made accessible.

4. WAVEGUIDE-BASED INTERCONNECTIONS

The other route which is being followed within NEMO is to seek for the integration of waveguide-based interconnections where the waveguiding layer is integrated in the PCB-stack or deposited on top of it. The main requirements regarding these materials are the temperature stability in view of the FR4-processing steps and mounting / soldering steps of the PCB, and, naturally, low optical losses.

Several materials are being investigated from the commercially available SU-8, ORMOCER® en Truemode® materials to in-house fabricated materials at some partners. Once these materials are deposited on the PCB-stack, the waveguides can be defined. Again several technologies are available. These can vary from traditional UV-lithography over laser-written, laser-ablated to molded or embossed structures. Deflecting structures can then either be defined in the optical layer or can be an insert which is separately fabricated and placed with required accuracy in the board.

To achieve optimum coupling efficiency, micro-optical elements such as micro-lenses can be used in between the waveguides and the components or fibers. Again depending on the application and the materials used, these micro-lenses can be fabricated in the board, or separately on an extra element or can be integrated with the insert.

Figure 1: Coupling and 90 degree deflecting element placed in a laser-ablated trench in front of a waveguide array on a PCB.

5. FUTURE TRENDS

As the interest PCB industry is clearly moving towards flexible circuits, the question can be raised whether it is also necessary to integrated optical interconnections in flexible substrates. Applications lie e.g. in sensing applications and wearable substrates, integrated in clothing. Otherwise multiple, stacked optical layers can ease the routing of optical signals and can fully benefit from the 2-D character of most of the available optoelectronic circuitry. In Figure 2 examples are shown of both trends.

Figure 2: Cross section of stacked optical layers with laser-ablated coupling structures (left) and a flexible print with arrays of optical waveguides and embedded VCSEL- and detector arrays (right)

CONCLUSIONS

Optical interconnections have shown and proven their use in long-haul communications but are also gradually being introduced on shorter distances. Currently the step towards board-level integration is to be taken. Fiber-based interconnections have proven their effectiveness but show disadvantages regarding coupling, connectorisation and deflection of the light towards opto-electronic arrays or fiber arrays. Special techniques can be used to overcome this problem, but also waveguide-based interconnections are being investigated and within NEMO several options are being considered and investigated.

ACKNOWLEDGEMENTS

Material presented in this overview has been realized within the framework of the NEMO-project and exhibits work carried out by NEMO WP9-partners: Ghent University (Belgium), VTT Electronics (Finland), FZK (Germany), Vrije Universiteit Brussel (Belgium), Herriott Watt University (UK)

The work carried out at Ghent University (Belgium) is also part of SBO project “FAOS: Flexible Artificial Optical Skin” funded by the Institute for the Promotion of Innovation by Science and Technology (IWT), Flanders, Belgium.
An Efficient Optical Coupling Method for Multilayer Optical Printed Circuit Boards

Y. Matsuoka, T. Ban, T. Shibata*, A. Takahashi*, and M. Shishikura
Central Research Laboratory, Hitachi, Ltd., 1-280 Higashi-koigakubo, Kokubunji-shi, Tokyo 185-8601, Japan
*Advanced Materials R&D Center, Hitachi Chemical Co., Ltd., 48 Wada, Tsukuba-shi, Ibaragi 300-4247, Japan
TEL:+81-42-323-1111, FAX:+81-42-327-7673, E-mail: yasunobu.matsuoka.av@hitachi.com

Abstract
An efficient optical coupling method has been developed for multilayer optical printed circuit boards. Excess coupling loss of less than 0.2 dB in a two-layer waveguide was achieved using the proposed cube-core optical confinement structure.

Introduction
The rapid increase in data processing in routers and servers requires higher speed and larger capacity interconnections. Optical interconnections are suitable for such a high bandwidth signal transmission because they have both a low propagation loss and a small inter-channel crosstalk even at a data rate of 10 Gbit/s or higher [1-3]. In particular, a large number of wirings more than a thousand will be needed to achieve Tbit/s-class interconnection; thus, a multilayer optical printed circuit board (OPCB) using a polymer waveguide is advantageous to such high-density interconnection. With the multilayer OPCB, maintaining the high coupling efficiency between the optical device and the waveguide, especially the lower layer, is very difficult because the insertion loss increases due to the spread of the light beam. To solve this problem, an effective optical coupling structure using optical micro-lenses has been reported [4]. In this paper, we propose an efficient optical coupling method using a newly optical confinement structure called a cube core. The excess optical coupling loss of less than 0.2 dB in a two-layer waveguide has been demonstrated in the fabricated OPCB.

Structure and analysis
Figure 1 shows the structure of the multilayer OPCB. It consists of multilayer multimode polymer waveguides on an FR-4 substrate, 45-degree mirrors, and optical I/O packages that integrate a vertical cavity surface-emitting laser (VCSEL) or photo diode (PD) and an IC. A light beam from the VCSEL propagates in a vertical direction to the board, and is coupled to waveguides after being reflected by the 45-degree mirror.

Fig. 1 Structure of the multilayer OPCB

In this structure, the insertion loss between the VCSEL (or PD) and the lower waveguide layer increases as the number of layers increases. To solve this issue, we proposed a cube-core structure made of the same material as the waveguide core. The proposed optical coupling system using the cube-core structure is illustrated in Fig. 2. As shown in Fig. 2 (a), with a conventional optical coupling system without the cube core, the spreading light beam from the VCSEL extends past the mirror that is mounted on layer 3; thus, excess loss increases. With the proposed cube core shown in Fig. 2 (b), the beam spread is confined in the high-refractive-index cube-core structure, which is formed just above the mirror. Therefore, the optical coupling efficiency in a multilayer OPCB can be improved with the proposed optical coupling method using the cube-core structure. Further, this cube core can be simultaneously formed in the same process as the upper-layer waveguide core. Thus, it also has the advantage of being less-costly to manufacture.

Fig. 2 Proposed optical coupling system
The effect of our proposed cube-core structure was confirmed in an optical simulation of the ray-tracing method. Figure 3 plots the simulated results of optical coupling efficiency between the VCSEL and the waveguide with and without the cube-core structure. These results show that the excess optical coupling loss that occurs when increasing the number of waveguide layers from one to two is very low, less than 0.2 dB, with the cube core, compared to 0.8 dB without the cube core. Furthermore, the excess loss at three layers is estimated to be less than 0.5 dB compared to 2.0 dB without the cube core. These results indicate that our proposed cube-core structure provides effective optical coupling for the multilayer OPCB.

![Optical coupling efficiency graph](image)

**Fig. 3 Simulated results of optical coupling loss versus layer number of waveguides**

**Fabrication**

To verify the effectiveness of this structure, we fabricated single or double-layer waveguides with the cube core on FR-4 substrates. A photo image of the fabricated OPCB is in Fig. 4. The waveguide core using photosensitive polymer material was fabricated on a FR-4 substrate by film lamination and UV photolithography, which are suitable for conventional PCB fabrication. The cube core was formed just above the mirror in the layer-1 waveguide.

![Fabricated OPCB image](image)

**Fig. 4 Photo image of the fabricated OPCB**

**Experimental Results**

Figure 5 plots the measurement results of the optical coupling efficiency between the VCSEL and layer 2 optical waveguide in a two-layer OPCB. The black line indicates the coupling efficiency of the single-layer waveguide, and the red and blue lines are those of the two-layer waveguides with and without the cube-core structure. The optical coupling efficiency with the cube core was about -1.3 dB compared to that of -1.9 dB without the cube core. The excess loss by using the cube-core structure was suppressed to as little as 0.18 dB from one to two layers compared to 0.82 dB without the cube core. The 1-dB optical coupling tolerance was also improved from +/- 13 μm without to +/- 17 μm with the cube core. These measurement results of the coupling efficiency in a two-layer waveguide are nearly agree with the simulation data, as shown in Fig. 3. This indicates that the excess loss can be further improved by the cube core in cases where there are more than three layers.

![Optical coupling efficiency graph](image)

**Fig. 5 Optical coupling efficiency between the VCSEL and layer-2 waveguide (2-layer OPCB)**

**Conclusion**

We proposed an efficient optical coupling method using a cube-core optical confinement structure integrated into a multilayer OPCB. With this structure, we succeeded in suppressing excess optical coupling loss to less than 0.2dB in a two-layer waveguide. This OPCB is suitable for large-capacity optical interconnections.

**References**

Replication of Multimode Polymer Optical Waveguides from Flexible Film Stamps

Shogo Yagi¹, Yutaka Hatakeyama², Naomi Kawakami², and Junya Kobayashi¹
¹ NTT Photonics Laboratories, NTT, Wakamiya 3-1 Morinosato, Atsugi, Kanagawa 243-0198 JAPAN
² NTT Advanced Technology Corporation, Nakacho 1-16-10, Musashino, Tokyo 180-0006 JAPAN

Abstract
A low cost replication method of multimode polymer optical waveguides using flexible film stamps is presented, which achieved a low propagation loss of 0.06 dB/cm. Mirrors and ridge waveguides are simultaneously replicated with high fidelity.

1. Introduction
As the bandwidth of data transmission soars in high-end routers, servers and mobile phones, optical interconnections is getting attractive aiming at the replacement of metal wiring. A flexible polymer optical waveguide film can be applied to it as an optical flexible circuit sheet as long as the fabrication cost is enough low. Therefore, some simple replication methods have been studied elsewhere¹,².

This paper reports on a new replication method using a flexible film stamp, namely, the flexible stamping method, which has several features that previous methods do not have, and also reports on the fabricated waveguides.

2. Flexible stamping method
2.1 Replication
Figure 1 illustrates the flexible stamping method. First, a grand master is prepared by engraving designated waveguide patterns on a PMMA plate with a dicing saw or a machining center, so that it is much economical than conventional grand masters. Then, a flexible film stamp is replicated from the master. Next, a liquid resin coats over the grand master, and then a flexible film covers it. After curing the resin, it sticks to the film, which is then peeled off from the master.

The replicating processes from the film stamp to a waveguide are illustrated in Fig.2. Although the conventional methods need temporary substrates with hard and flat surfaces such as Si wafer, the cores are directly formed over the cladding films without hard substrate in this method. (a) UV curable epoxy, which will form the core, is coated on the cladding film. (b) Then the film stamp is carefully laminated on it so that they do not include air bubbles. (c) Excess epoxy fluid is squeezed out by pushing a roller. (d) The epoxy is cured and the film stamp is peeled off. Finally the over cladding is formed if necessary.

As long as the ridge waveguides are concerned, it is a remarkable feature that mirrors can be simultaneously formed in this method because the cores are directly replicated on the under-cladding surface.

2.2 Optical properties
Figure 3 is the waveguide length dependence of the insertion loss, where the core size is 50 x 50 μm square and 850 nm laser light is coupled with the
The fidelity of the replication and the removal of the excess core resin are confirmed by Fig. 4. Since the guided wave dissipates by micro-bumps at the core-cladding interface and the residual core resin between over and under claddings, Fig. 4 agrees with the obtained low propagation loss.

![Fig.4 Cross-section of the fabricated waveguide](image)

**2.3 Simultaneous mirror fabrication**

Figure 5 shows a SEM micrograph of multi-channeled ridge waveguides with 45 degree mirror end faces. The faces of mirrors and the core ceilings are flat. Every corner is very sharp. They are also special features of this method.

**2.4 Possibility of wide area waveguide films**

It is easy to enlarge a waveguide film because the film stamp is flexible which gives good transfer uniformities in large area. We successfully fabricated an A4 size waveguide film as shown in Fig. 6.

![Fig.5 SEM micrograph of mirror-end waveguides](image)

Fig.5 SEM micrograph of mirror-end waveguides

![Fig.6 Replicated A4 size waveguide film](image)

Fig.6 Replicated A4 size waveguide film

**3. Conclusion**

A low cost replication method of multimode polymer optical waveguide films using flexible film stamps is proposed. The propagation loss of fabricated waveguide film is fairly low of 0.06 dB/cm. This result shows that application to the inter-shelf connection with this waveguide film is in the target scope.

References

1) Sugihara et. al., C-13-7, Gen.Conf. of IEICE2004.
2) Terakawa et. al., C-3-29, Gen.Conf. of IEICE2003.
Self-Written Optical Funnel for Optical Interconnect by

Microlens-transfer Method

H. Kubo, M. Kanda, O. Mikami
School of Information Science and Technology, Tokai University
1117 Kitakaname, Hiratsuka, Kanagawa, 259-1292 JAPAN,
6adgm008@keyaki.cc.u-tokai.ac.jp

Abstract – We have proposed a new self-written optical funnel fabricated by using a microlens transfer for board level optical interconnect. The optical characteristics of these optical funnels were investigated experimentally and theoretically.

I. INTRODUCTION

Recently, the information and communication technology equipments have faced a serious problem of "Metallic Interconnection Crisis" due to a drastic increase of information flow. Therefore, "Optical Interconnect" which uses optical signal for the board level wiring has attracted much attention [1]. However, there are still some problems in introduction of optical interconnect, such as small tolerance in positional alignment. To solve this problem, we have proposed optical coupling with a taper shape optical funnel, as shown in Fig. 1. In order to achieve this new shape of funnel on the optical device, we have applied a microlens transfer method. The shape and characteristics of fabricated optical funnels are described.

II. OPTICAL INTERCONNECTION

Fig. 2 shows an application model of optical surface mount technology (O-SMT) [1]. Optical devices with optical connecting rods are inserted into via-holes on an O/E printed wiring board and optical signal is transferred between optical devices and optical wiring waveguide. A space gap between the optical waveguide and the optical device mounted on OE-board can be minimized by applying Self-Written Optical rod (SW-Optical rod) on optical devices such as VCSEL and PD. Therefore, optical coupling with higher efficiency is expected.

A conventional SW optical rod has been fabricated by a photomask transfer method [2]. A straight shape optical rod is successfully made by this method. Because the emitting window of VCSEL is small, a sufficient tolerance in positional alignment is possible even with a straight shape optical rod of around 50 μm diameter when the laser beam is coupled into the connection rod.

However, as for the receiver side, the condition is reverse; a signal laser beam comes from the connection rod into the PD. The receiving-area of PD usually becomes smaller as the required response speed is faster. Therefore, the positional alignment becomes difficult when higher optical coupling between the PD and optical wiring is tried to be obtained. We have proposed a taper SW-optical funnel to realize higher efficiency optical coupling between optical wiring and PD. By applying SW-optical funnel, the positional tolerance becomes larger and the higher optical coupling efficiency is expected to be obtained.

In order to obtain an SW-optical funnel, the microlens transfer method is investigated. A multi-channel funnels can be also achieved by using a microlens plate as same as the photomask transfer method.
FABRICATION OF OPTICAL FUNNEL

Fabrication process is schematically shown in Fig. 4. In the experiment, two kinds of microlens plate A, B are used. The diameter and the pitch of lens A are 84\(\mu\)m and 250\(\mu\)m, respectively. Those of lens B are 84\(\mu\)m and 250\(\mu\)m, respectively. The focus lengths are 100\(\mu\)m (lens A) and 120\(\mu\)m (lens B). In the test, we used 500\(\mu\)m thick spacer in lens A and 120\(\mu\)m thick spacer in lens B.

Fig. 5 (a) shows a fabricated sample when a 500\(\mu\)m thick spacer was used. Unique shape of SW rod was obtained. It is probably due to a mismatch between the spacer thickness and the focus length of microlens. Fig. 5 (b) is a fabricated sample when a spacer of 120\(\mu\)m thickness was used. A typical funnel shape was achieved.

Next, the positional tolerance was analyzed using a ray tracing method. Two types of SW-optical rods, funnel and straight shapes, were assumed, as shown in Fig. 6 (a) and (b). The analysis model is shown in Fig. 6 (c). The optical coupling efficiency was calculated when the fiber with VCSEL was shifted against the PD. The diameter of PD receiving area was set to be 50\(\mu\)m. The calculation results are summarized in Fig. 7. The tolerance of the taper SW-optical funnel is quite larger than that of the straight one. Therefore, the proposed optical funnel is expected to be quite effective in optical coupling.

CONCLUSION

We have proposed SW-optical funnels fabricated by using a microlens transfer. A funnel type was successfully realized and larger tolerance is expected by a ray tracing method.

REFERENCES


Polymeric Waveguide Optical Switch
Using Rotary Drive Mechanism

Toshitsugu Uesugi, Shiho Zaizen, Atsushi Sugitatsu and Tatsuo Hatta
Information Technology R & D center, Mitsubishi Electric Corporation
4-1, Mizuhara, Itami, Hyogo 664-8641, Japan
Tel: +81-72-780-2653 / Fax: +81-72-780-3774
Email: Uesugi.Toshitsugu@ak.MitsubishiElectric.co.jp

Abstract
We propose a polymeric waveguide optical switch using a novel drive mechanism. A trench formed at a cross-point in the flexible polymeric waveguides is gated by rotating the near-trench arm and switching operation is demonstrated.

1 Introduction
Optical switches are key devices for optical fiber communication systems because they can change light paths without optical/electrical conversion. The important features of optical switches are not only the performance but also the cost saving.

In our past works [1-3], we proposed and demonstrated the bascule structure optical switch using polymeric waveguides, suitable for low-cost mass production. The switch also has good optical properties such as low-loss transmission and relatively fast switching speed. Fig. 1 shows the schematic of the optical switch.

The switch uses a flexible polymeric waveguide film where trenches are formed at cross-points of the waveguides. When the trench opens, light path changes by total internal reflection between the air gap and the polymeric waveguide. Previously, piezo actuators just below the cross-points were used to push up and open the trenches. However, such drive mechanism requires precise horizontal alignment between the individual actuators and the trenches, and micrometer-scale flatness of both the polymeric waveguide film and the actuators over the large film area in order to overcome the limitation of the actuators’ strokes. In this paper, we propose a novel drive mechanism to solve the abovementioned problem and demonstrate the switching operation.

2 Switching mechanism
Instead of the vertical drive described above, we adopt a rotary drive mechanism shown in Fig. 2.

Fig. 1: Schematic of the optical switch.
Fig. 2: Principle of rotary drive in a cross state (cross section). Inset: In a bar state.
A vertically inserted rotating arm into the film is set near the trench. By rotating the arm in the direction normal to the trench (the direction of an arrow with $\Delta x$ in Fig. 2), the trench changes from bar (transmission) to cross (reflection) state. This rotary drive mechanism does not require precise alignment or flatness between the film (trenches) and the drive units. Therefore, it will be more cost-effective than the previous vertical drive mechanism especially for $N \times N$ matrix switches.

3 Design

Due to the flexibility of the polymeric film, the distance between the rotating arm and the trench in the film affects the force and the amount of rotation required for switching operation. If the arm is set far away from the trench, the rotational force necessary for switching operation will become large because it does not directly act on the trench. Using structural analysis, we found that when the arm is set 500 $\mu$m away from the trench in the normal direction, the required rotational force at the drive unit becomes 5 times larger than that when the arm is set 150 $\mu$m away. According to the structural analysis with the appropriate arm size, we chose the distance between the rotating arm and the trench to be 250 $\mu$m.

4 Experimental Results

The rotating arm of 125 $\mu$m diameter was set 250 $\mu$m away from the trench in the normal direction in the film. We observed the intensity change of the transmission light (Output T in Fig. 1) and the reflection light (Output R in Fig. 1) with the change of a horizontal displacement $\Delta x$ at the level of 3 mm high from the film surface (see Fig. 2). When $\Delta x = 1$ mm, the attenuation of the transmission light reached 21 dB and the reflection light intensity increased by 27 dB as shown in Fig. 3. Switching operation with the rotary drive mechanism was successfully demonstrated. As the mechanism uses rotary force, the required displacement $\Delta x$ can be reduced by lowering the height level of the force.

5 Conclusion

We proposed and demonstrated the polymeric waveguide optical switch based on the bascule structure with the novel rotary drive mechanism. We observed 21 dB attenuation of the transmission light and 27 dB increase of the reflection light intensity when the trench at the cross-point of the waveguides was switched from bar to cross state by the rotary drive. This drive mechanism will offer the polymeric waveguide optical switch with both high productivity and low cost.

Acknowledgements

This work was partly supported by a project of the National Institute of Information and Communications Technology (NICT), as part of a program of the Ministry of Public Management, Home Affairs, Posts and Telecommunications of Japan.

References

Burst-mode Bidirectional Optical Amplifier
with single-EDF configuration for L-band use in 10Gbit/s-PON systems
Yasuhiro NAKANISHI, Takashi NAKANISHI, Youichi FUKADA, Ken-Ichi SUZUKI and Katsumi IWATSUKI
NTT Access Network Service Systems Laboratories, NTT Corporation
1-6 Nakase, Mihama-ku, Chiba-shi, Chiba, 261-0023 Japan
Phone: +81 43 211 3295, Fax: +81 43 211 8250, E-mail: y-naka@ansl.ntt.co.jp

1. Abstract
We propose a novel burst-mode bidirectional L-band optical amplifier with single-EDF configuration, show its gain characteristics, and clarify the impact on downstream transmission properties of the gain deviation caused by upstream burst signal amplification.

2. Introduction
There has been increasing interest in broadband optical access services using PON (Passive Optical network) systems due to their cost-effectiveness; the transmission fiber and the central office equipment are shared by many subscribers. GE-PON(Gigabit Ethernet PON)[1] and G-PON(Gigabit-capable PON)[2] systems, which can provide gigabit-capable FTTH services, have been deployed in Japan and North America, respectively. Moreover, the next generation PON systems with 10Gbit/s capacity are being discussed in IEEE and FSAN [3, 4]. In these associations, coexistence with current PON systems has been recognized as indispensable to achieve smooth migration. A wavelength allocation plan for next-generation PON systems in the L-band wavelength region (1560 to 1625 nm) has been proposed for realizing coexistence because the L-band wavelength region is not used by existing PON systems that comply with current standards.

Optical amplifiers bring some benefits to PON systems such as long-reach, high splitting ratios, and high-sensitivity. For existing PON systems that comply with current standards, we have investigated a bidirectional optical amplified PON repeater that consists of an O-band (1260 to 1360 nm) optical burst-mode amplifier and an S-band (1460 to 1530 nm) optical amplifier [5]. This amplifier utilizes gain-clamping to suppress optical surges; such surges can well cause failure of the optical receiver as well as interfering with the reception of normal signals at the OLT due to gain dynamics.

In this paper, we propose a novel burst-mode bidirectional optical amplifier with single erbium-doped fiber (EDF) configuration for L-band amplification. The proposed amplifier employs downstream optical signals as the gain-clamp light to suppress optical surges. This eliminates the additional gain-clamp light source and simplifies the optical amplifier because we can use the same gain medium to amplify both upstream and downstream optical signals. We also show the gain characteristics of the proposed optical amplifier and clarify the impact on downstream transmission properties of the gain deviation due to upstream burst signal amplification.

3. Proposed burst-mode bidirectional optical amplifier with single-EDF configuration
Figure 1 shows the configuration of the optical access system with the proposed amplifier and the structure of the amplifier. To suppress optical surges, the downstream optical signal is stronger than the burst-mode signal light, so proposed amplifier is allocated near the OLT. The proposed amplifier consists of just an erbium-doped bismuth-based fiber [6] to amplify upstream and downstream signals in the L-band wavelength region. Downstream optical signals are utilized to suppress optical surges, which are usually triggered in the upstream optical burst signals by optical amplification. Therefore, the proposal yields a simple bidirectional optical burst-mode amplifier that does not need any additional LD for gain-clamping and thus offers significant cost benefits. On the other hand, the gain variation caused by the upstream burst-mode optical signal may degrade the downstream optical signals.

4. Experiment
We conducted an experiment to clarify the influence of upstream and downstream signals as well as basic gain properties. Figure 2 shows gain properties as a function of input power. Both forward and backward pump powers were set at 500 mW. The downstream signal was set at the relatively high power of 0 dBm to realize gain clamping of the upstream signals. The wavelengths of upstream and downstream signals were 1571 nm and 1591 nm, respectively, these wavelengths lie on the ITU-CWDM grid [7].
As shown in Figure 2, although the downstream signal suppress the upstream gain from 23dB to 16dB, good gain linearity is obtained because the 1 dB gain suppression point is improved to -10 dBm and gain deviation of the upstream signal is suppressed compared to upstream signal amplification (unidirectional amplification). The downstream signal gain is around 10 dB at input powers less than -10 dBm while the downstream gain is decreased with upstream signal input powers greater than -10 dBm. These results show that the proposed bidirectional optical amplifier works well at upstream averaged signal powers under -10 dBm.

Next we confirmed that the downstream signal could suppress optical surges in the upstream signal. Figure 3 shows the optical surge ratio (OSR) as a function of burst repetition frequency. We define OSR as the ratio of optical surge peak level to restored normal signal level. Bidirectional amplification well suppresses OSR compared to unidirectional amplification, see Figure 3.

Figure 4 shows the bit-error rate (BER) as a function of averaged receive power of upstream signal at the optical receiver; the parameter is burst repetition frequency. We evaluated the power penalties at the BER of 10^-12 for each burst repetition frequency. In the experiment, downstream 10.3125 Gbit/s (PRBS was 2^15-1) signals were generated by a LiNbO intensity modulator; extinction ratio was 9.8 dB and averaged power was -3 dBm. A PIN-PD was used in the optical receiver. The maximum power penalty of only 1 dB is observed at the burst repetition frequency of 100 Hz and averaged optical input power to the amplifier of -13 dBm.

5. Conclusion

We proposed a novel burst-mode bidirectional L-band optical amplifier with single-EDF configuration, and demonstrated its gain characteristics. First we confirmed the effect of optical surge suppression using gain-clamp light. We also clarified the impact on downstream transmission properties of the gain deviation caused by upstream burst signal amplification. Measurements showed that the upstream signal had no significant influence at averaged optical input powers to the optical amplifier of under -13 dBm which confirms the validity of our proposed bidirectional amplifier.

Reference

[1] IEEE 802.3ah
Abstract: Fiber loop buffer with EDFA had relative low noise accumulation. For long packet-delay, suppression of relaxation oscillation was a key issue. This paper realizes the power stability by using SOA following EDFA in fiber loop.

1. Introduction

Optical packet switching (OPS) technology is the ultimate future-proof solution for a next-generation optical transport network that allows direct forwarding of data packets in the optical domain [1]. Optical buffers are key components in OPS systems that avoid packet collision along the same virtual path and wait for header processing time. Fiber delay lines are a practical candidate for all-optical packet buffers. Use of recirculating fiber loops [2, 3] can reduce the components needed to implement buffers, thereby reducing their size. In our previous paper, we described the use of an optical frequency shifter in the fiber loop to prevent unexpected oscillation caused by ring structure with including optical amplifiers and demonstrated, experimentally, an optical delay line that is tunable by the optical fiber ring [4].

Optical packets undergo many round-trip circulations in the loop. Optical amplifiers, then, are necessary to compensate the optical loss of the ring. The conventional configuration as shown in Fig. 1(a) uses an EDFA because of its high gain and low noise. The BPF rejects amplified spontaneous emission (ASE) noise and residual pump light from the EDFA. An AOS is inserted into the ring to prevent oscillation caused by ring structure [4].

2. Configuration of fiber loop buffers

Figure 1 shows the configuration of an optical tunable delay line using a fiber loop with an acousto-optic frequency shifter (AOS) and an erbium-doped fiber amplifier (EDFA). The storage and delay functions can be provided by the fiber loop. The optical switch located at the loop output plays the role of a delay-time selection function. Figure 2 shows the timing chart of the tunable delay line. The fiber length of the ring delays the packet when the optical packet signal pulse is incident into the fiber loop. The loop output is an optical pulse sequence constructed from replicas of the incident packet. If the time duration of packet $T_p$ is shorter than the round trip time of the ring $T_{ring}$, then the optical switch following the ring selects the packet when the packet route is available.
Figure 3(a) shows the optical power waveform from the fiber loop with conventional configuration of Fig. 1(a). In this experiment, the rectangular pulse was used as an incident optical packet. The wavelength of the packet was 1552.5 nm. The center wavelength of BPF in the loop was identical to that of the packet. Packet length $T_p$ was 5.0 $\mu$s. Optical loss of ATT was 0 dB in this case. Optical power waveform was measured by an oscilloscope after O/E conversion. The relaxation oscillation can be found with 4.3 kHz. Optical power variation in the oscillation was 2.8 dB. The top of each pulse was tilted by the oscillation after the eighth circulation. The basic origin of the oscillation is interplay between the oscillation field in the ring circuit and the atomic population inversion in the amplification medium. An increase in the field intensity induces a reduction in the population inversion because of the increased rate of stimulated transitions, causing a reduction in the amplification gain that tends to reduce the field intensity.

![Figure 3. Optical power waveform from the fiber loop.](image)

3. Fiber loop with EDFA followed by SOA

In order to suppress the relaxation oscillation, a low polarization dependent SOA module followed the EDFA in the fiber loop as shown in Fig. 1(b). In this experiment, the injection current of the SOA was a constant value of 100 mA. Then, the small signal gain and noise figure of the SOA module were 8.7 dB and 10.1 dB, respectively. The SOA saturation output power was 5.6 dBm. A variable optical attenuator (ATT) was inserted between each optical amplifier to change the operating point of optical amplifiers. Optical loss of $\alpha_1$ is defined as the power loss from EDFA output to SOA input. $\alpha_2$ represents the optical power loss from SOA output to EDFA input. In our setup, $\alpha_1$ was 5.0 dB. Figure 3(b) shows the optical power waveform from the fiber loop with both EDFA and SOA. No relaxation oscillation is apparent in the case of $(\alpha_1, \alpha_2) = (23.7$ dB, 5.0 dB). Power variation at the top of the pulse was 0.9 dB in our measurement time range. This power difference is caused by the combination of signal frequency shift from AOS, EDFA gain tilt, ASE accumulation in EDFA saturation. Suppression of the relaxation oscillation can be explained by the gain saturation and fast response of the SOA [6].

![Figure 4. Optical power variation and output power with changing optical loss of $\alpha_1$.](image)

When the optical loss of $\alpha_1$ is increased, then the SOA is more deeply saturated. Figure 4 shows the power variation of the fiber loop with changing the $\alpha_1$. Almost identical power variation of 1 dB can be obtained with changing the SOA saturation. However, the output power tends to be small with low saturation. The pattern effect of SOA can be mitigated by using Manchester coding for signal.

4. Conclusion

Output power fluctuation of fiber loop buffer including a low noise EDFA can be suppressed by a following SOA. Because an EDFA is used as the first stage of the tandem configuration, the noise performance can be improved. In our system, optical amplifiers of the EDFA and SOA are driven by a constant current. No sophisticated control is needed for mitigation of power fluctuation.

References

Self-seeded Reflective Semiconductor Optical Amplifier for Upstream Access and Local Area Networking in Passive Optical Networks

Nishaanthan Nadarajah, Ka Lun Lee, and Ampalavanapillai Nirmalathas
National ICT Australia, Victoria Research Laboratory, Dept. of Electrical & Electronic Engineering, The University of Melbourne, VIC 3010, Australia.
Tel: +61 3 8344 6095; Fax: +61 3 8344 6678; email: nnad@ee.unimelb.edu.au

Abstract: A scheme for upstream access and local area networking in passive optical networks using a single self-seeded reflective semiconductor optical amplifier placed in customer premises is proposed and experimentally demonstrated.

Introduction: Passive optical network (PON) technology is considered as an efficient solution to facilitate high bandwidth, low cost, and fault-tolerant next generation broadband access networks [1]. Apart from the upstream and downstream transmissions between the central office (CO) and the optical network units (ONUs), customers of a PON may require communication links between themselves for various services. To facilitate these services, a number of optical layer local area networking (LAN) schemes have been demonstrated [2]. Recently, there has been a lot of interest in the elimination of a laser source at the ONU, thus avoiding its stabilization and provisioning all ONUs with wavelength independence for the future access network. PONs based on a reflective semiconductor optical amplifier (RSOA) placed at the ONU enable a simpler, more cost effective, and easily upgradeable infrastructure for access networks [3]. We recently demonstrated schemes for LAN capabilities using a single RSOA placed at each ONU [4, 5]. One scheme uses the broadband spectrum of RSOA for the LAN traffic transport and therefore suffers from dispersion while the second scheme requires unmodulated optical carriers delivered to ONUs from the CO for seeding the RSOA. In this paper, we propose and experimentally demonstrate a scheme for upstream access and LAN using a self-seeded RSOA, whereby self-seeding of RSOA is performed with the use of fiber Bragg gratings (FBGs) placed at each ONU. Compared to previous schemes, this scheme does not require high speed RF electronics, stable laser sources, and modulators at each ONU. Moreover, LAN traffic transport can be performed at any time rather than in designated time slots. We experimentally demonstrate the proposed scheme with 2.5 Gb/s downstream traffic, 1 Gb/s upstream traffic, and 1 Gb/s LAN traffic.

System Architecture: The proposed scheme for implementing upstream access and local networking using a self-seeded RSOA is shown in Fig. 1. A \((N+1) \times (N+1)\) star coupler (SC) is used to split the optical signals to each ONU, whereby the number of ONUs attached to the SC is \(N\). As shown in Fig. 1, each ONU is connected to the SC via two distribution fibers. At the ONUs, two FBGs and an optical switch (OSW) are used to select the wavelength channels for the seeding of the RSOA. The reflective FBG slices broadband spectrum of RSOA and continuously feeds the sliced channel back to the RSOA for wavelength seeding. The self-seeded wavelength channel can therefore be used for data transport. In the upstream transmission mode, the OSW is set to ‘cross’ state such that upstream wavelength channel \(\lambda_{U}\) is used to wavelength seed the RSOA. In the LAN mode, the OSW is set to ‘bar’ state such that LAN wavelength channel \(\lambda_{LAN}\) is used to wavelength seed the RSOA. In both operating modes, the self-seeded RSOA is directly modulated with the data and transmitted in the upstream direction. LAN data transmission to each ONU is performed using a secondary distribution fiber and is not expected to increase costs more than 0.3% [6]. A WDM coupler is used at each port of the SC facing the CO to prevent \(\lambda_{U}\) reaching the ONUs and therefore LAN is completely isolated from the PON and hence provides improved security. Upstream access follow the time division multiple access (TDMA) protocol, while LAN may follow any media access control (MAC) protocol.

Experimental demonstration: The experimental setup to demonstrate the feasibility of the proposed scheme is similar to that shown in Fig. 1. A downstream signal of 2\(^{31}-1\) pseudo random binary sequence non-return to zero (PRBS NRZ) data at 2.5 Gb/s was modulated onto downstream wavelength channel \(\lambda_{D} = 1540.172\) nm using a Mach-Zehnder modulator and transmitted to the ONUs through a 10 km feeder fiber, a 4 x 4 SC and a 3 km distribution fiber. At the ONUs, \(\lambda_{D}\) was separated from self-seeded channels \(\lambda_{U} (=1549.32\) nm) and \(\lambda_{LAN} (=1554.07\) nm) using a coarse wavelength division multiplexing (CWDM) coupler. A 2 x 2 OSW was used to select the seeding wavelength channels depending on the mode of operation. The RSOA used in the experiment has a small signal gain of 25 dB and a noise figure of 9 dB at a bias current of 50 mA. An optical isolator is used in front of a port of the 3 dB coupler to prevent back reflections.
into the RSOA. In both transmission modes, 2^{31}-1 PRBS NRZ data at 1 Gb/s was directly modulated onto the RSOA when the RSOA was biased at 50 mA. A variable optical attenuator (VOA) is used next to the OSW to vary the seeding optical power into the RSOA. The signals on all three wavelength channels were detected using a 2.5 Gb/s p-i-n receiver. An optical filter with a bandwidth of 1 nm was used before the LAN and upstream data receiver to prevent the out-of-band ASE noise from the RSOA entering the receiver. However, the use of this optical filter can be avoided by employing a narrowband CWDM filter at the SC. A series of experiments were conducted and bit error rates (BERs) for all signals were measured.

Results and Discussions: Fig. 2 shows the observed optical spectra at the upstream and LAN data receivers. The optical power difference of 6 dB at both receivers was due to the lower transmitted power and feeder fiber loss for the upstream wavelength channel. Note that an optical signal to noise ratio (OSNR) of more than 36 dB was observed for both channels. Fig. 3 shows the measured BER curves for the signals. The transmission penalty for 2.5 Gb/s downstream data compared to the back-to-back (B-B) measurements was less than 0.1 dB. The penalty for the 1 Gb/s upstream data and LAN data transmissions compared to B-B measurements was less than 0.4 dB. To measure the effects of the seeding optical power into the RSOA when the RSOA was biased at 50 mA. A variable NRZ data at 1 Gb/s was directly modulated onto the RSOA. In both transmission modes, 2^{31}-1 PRBS NRZ data at 1 Gb/s was directly modulated onto the RSOA when the RSOA was biased at 50 mA. A variable optical attenuator (VOA) is used next to the OSW to vary the seeding optical power into the RSOA. The signals on all three wavelength channels were detected using a 2.5 Gb/s p-i-n receiver. An optical filter with a bandwidth of 1 nm was used before the LAN and upstream data receiver to prevent the out-of-band ASE noise from the RSOA entering the receiver. However, the use of this optical filter can be avoided by employing a narrowband CWDM filter at the SC. A series of experiments were conducted and bit error rates (BERs) for all signals were measured.

Fig. 2: Observed optical spectra at upstream and LAN data receivers.

Fig. 3: Measured BER curves for all signals.

Fig. 4: Measured BER curves for LAN data with varying seeding power for a bias current of 50 mA.

Conclusions: We have proposed and experimentally demonstrated a scheme for upstream access and LAN capability in PONs using a self-seeded RSOA at the ONU. The wavelength seeding of the RSOA can be performed using FBGs. The experimental results show that all signals can be recovered with minimal penalty.

References:
Reflection Tolerance of RSOA-based WDM PON  
Korea Advanced Institute of Science and Technology  
373-1 Guseong-dong, Yuseong-gu, Daejeon 305-701, Korea  
(Phone) 82-42-869-3456, (Fax) 82-42-869-3410, (E-mail) ychung@ee.kaist.ac.kr  
*KDDI R&D Laboratories Inc., 2-1-15 Ohara, Fujimino-shi, Saitama 356-8502, Japan

Abstract
We investigate the effects of back-reflection in RSOA-based WDM passive optical networks. The results show that the upstream signal can tolerate the back-reflection in the range of up to -27 ~ -25 dB, depending on the data rate of the downstream signal.

I. Introduction
For the realization of the practical WDM passive optical network (PON), the wavelength-independent operation of the optical network unit (ONU) is indispensable. Various types of colorless light sources have been proposed to achieve this objective, including the spectrum-sliced light sources, ASE-injected Fabry-Perot lasers, and reflective semiconductor optical amplifiers (RSOAs)'s [1]-[4]. In particular, the use of RSOA’s is attractive since the received downstream signal can be reused for the upstream transmission [3]-[4]. However, in such a network, the back-reflections from splices and connectors can severely degrade the performances of the upstream signals. This is because the upstream signal reflected back to the ONU is re-amplified by the RSOA, and induces a large intensity noise. Recently, this effect of back-reflection on the upstream performance has been studied in WDM single-fiber loopback networks [5]. However, in this study, it is assumed that additional WDM light sources are used at the central office (CO) to inject the cw seed light into each ONU (implemented by using an optical amplifier and an optical modulator). Thus, each ONU in this network amplifies and modulates the injected cw light from the CO for the upstream transmission.

In this paper, we investigate the effect of the back-reflection on the upstream performance in the RSOA-based WDM PON. Unlike the previous report [5], we assume that the modulated downstream signal is injected to the RSOA instead of the cw seed light. The results show that the upstream signal can tolerate the back-reflection up to -27 ~ -25 dB, depending on the data rate of the downstream signal.

II. Experiment setup
Fig. 1 shows the experimental setup. We assumed a bidirectional WDM PON system, and evaluated the impact of a discrete reflection occurred near the ONU. At the CO, we used a DFB laser operating at 1551 nm for the downstream transmission. We directly modulated this laser with 1.25-Gb/s or 2.5-Gb/s non-return-to-zero (NRZ) signals. The extinction ratio of the downstream signal was set to be 2 dB to ensure the saturated operation of the RSOA even at ‘0’-level [4]. The downstream signal was sent to the ONU and reused for the upstream transmission. At the ONU, an RSOA was used as an upstream light source. The small signal gain was 17 dB (bias current = 80 mA), and its polarization dependence was 0.4 dB. We directly modulated the gain of the RSOA with a 155-Mb/s NRZ signal. In order to improve the upstream performance by compressing the downstream signal, we set the optical power of the downstream signal incident on the RSOA to be -11 dBm (at which the RSOA gain was compressed by 3 dB) [4]. The upstream signal was then sent to the upstream receiver at the CO. To evaluate the effect of the back-reflected light into the RSOA, we used the reflection module composed of a mirror and a variable optical attenuator. The fiber length between the mirror and RSOA was about 15 m. The effect of back-reflection was highly dependent on the state-of-polarization of the reflected light. Thus, we used a polarization controller, PC2, for the worst case analysis (i.e., worst BER).

III. Results and Discussions
We first evaluated the impact of the back-reflection by observing the eye diagram of the upstream signal using a 2.5-GHz photodetector. Fig. 2 shows the eye diagrams of the upstream signal measured with and without applying the back-reflection. In this figure, the downstream signal was modulated at 1.25 Gb/s except in the case of Fig. 2(a), which was measured without modulating the downstream signal for a reference. Fig. 2(b) shows the eye diagram measured without back-reflection. In this diagram, the
‘1’-level of the upstream signal was split into two levels due to the residual downstream signal. This thick ‘1’-level is one of the major impairment factors for the upstream signal in the loopback network utilizing remodulation [3]-[4]. When we applied the back-reflection (reflectivity = -27 dB), the eye diagram was further degraded (i.e., the thickness of ‘1’-level increased) as shown in Fig. 2(c).

To evaluate the impact of these noises (induced by the back-reflection) quantitatively, we measured the BER of the upstream signal by using a 155-Mb/s optical receiver. Fig. 3 shows the measured receiver sensitivity (@ BER = 10^-3) as a function of the reflectivity. No significant degradation was observed in the receiver sensitivity when the reflectivity was smaller than -35 dB. However, as we increased the reflectivity, the sensitivity was drastically degraded. For example, when the downstream data rate was 1.25 Gb/s, it was not possible to achieve the error-free transmission of the upstream signal if the reflectivity exceeded -27 dB. On the other hand, the same result was observed for the 2.5-Gb/s signal at the reflectivity of -25 dB. These results indicate that the reflection tolerance of the upstream signal increases with the downstream data rate due to its broadened spectral width.

Previously, it has been reported that the back-reflected light in a single-fiber loopback system could generate extra intensity noises due to the optical beat interference (OBI) between the upstream and reflected signals [5]. In this report, the bandwidth of the generated intensity noise was assumed to be identical to the signal’s bandwidth (since cw light from the CO was used as a seed for the RSOA at the ONU), and estimated the degradation of the signal-to-noise ratio (SNR) caused by the back-reflection. However, in our case (i.e., the downstream signal is not cw but directly modulated), the downstream signal has much broader bandwidth than that of the upstream receiver. Thus, the intensity noise (induced by the OBI) is spread over a wide spectral range, and, consequently the SNR degradation caused by the back-reflection is alleviated.

To estimate the SNR degradation caused by the back-reflection, we evaluated the intensity noise parameter \( r_i \) by using \( \delta_i = -10\log(1 - r_i^2 Q^2) \), where \( Q \) is the Q-factor (=6) and \( \delta_i \) is the power penalty in dB [6]. The error-free transmission (i.e., BER<10^-3) cannot be achieved if \( r_i \) exceeds 0.167. Fig. 4 shows the intensity noise parameter obtained by using this equation and the measured receiver sensitivities in Fig. 3. This parameter, \( r_i \), was nearly constant at the low reflectivity. However, \( r_i \) increased with the reflectivity, when it was higher than -35 dB. These results suggested that the intensity noise parameter could be expressed in the following form:

\[
r_i^2 = r_{remod}^2 + r_{ref}^2 = r_{remod}^2 + aR^2
\]

where \( R \) is the reflectivity, \( r_{remod} \) and \( r_{ref} (=\sqrt{a} \cdot R) \) are the intensity noise parameters resulting from the residual downstream components and the OBI caused by the back-reflection, respectively. The constant \( a \) is a fitting parameter originating from the spectral distribution of the upstream signal. The dashed curves in Fig. 4 are the calculated values by using equation (1). The results show that these curves agree well with the measured values. In comparison, the solid line in Fig. 4 represents the intensity noise parameter calculated by using the theory in [5]. It is clearly shown that this simple theory overestimates the SNR degradation (caused by back-reflection) when the modulated downstream signal is injected to the RSOA.

IV. Summary

We investigated the effects of the back-reflection on the upstream signal in a RSOA-based WDM PON. The results showed that this network could tolerate the back-reflection up to \(-27 \sim -25 \) dB (when the RSOA gain is 14 dB), depending on the data rate of the downstream signal injected to the RSOA.

References

Wavelength Monitoring of Remote Node by Using Self-Locked Reflective Semiconductor Optical Amplifier for Bidirectional WDM-PON

Seung Heon Han, Tae-Young Kim and Chang-Soo Park
Department of Information and Communications, Gwangju Institute of Science and Technology
1, Oryong-dong, Buk-gu, Gwangju, 500-712, Republic of Korea
Tel: +82-62-970-3150, Fax: +82-62-970-3151, E-mail: csp@gist.ac.kr

Abstract

A wavelength monitoring technique of remote node (RN) is proposed using RSOA self-locked by Bragg grating at the RN. The temperature variation of RN can be monitored by measuring the lasing wavelength of the RSOA.

I. INTRODUCTION

Wavelength division multiplexing - passive optical network (WDM-PON) is highly concerned with future broadband access networks. For commercial deployment, many studies have been focused on the WDM-PON with simple and low cost solutions. Bidirectional WDM-PON is one of the solutions in terms of fiber capacity and maintenance. In this WDM-PON, however, how to discern and offset the wavelength drift of an arrayed waveguide grating (AWG) at the remote node (RN) without complicated wavelength monitoring technique is important [1].

Previously, a stable optical source such as multi-frequency [2,3] or distributed feedback - laser diode [4] should be used to monitor the temperature-induced wavelength drift. Its optical power is monitored and fed back to the optical line terminal (OLT), and can be adjusted to accommodate the misalignment of the wavelength. However, these optical sources are expensive. Moreover, those techniques require an additional output port of the AWG at the RN to monitor the wavelength drift [3,4].

In this paper, we propose a simple monitoring technique on the temperature-induced wavelength drift at the RN using a reflective semiconductor optical amplifier (RSOA) and a Bragg grating (BG) (Fig.1). The RSOA is self-locked by the BG, and its lasing wavelength follows the center wavelength of the BG, reflecting the temperature-induced wavelength drift of the AWG. Thus, the temperature variation of the RN can be monitored by measuring the lasing wavelength of the self-locked RSOA. It can be used to control the wavelengths of the transmitters and the AWG. The proposed technique is simple and cost-effective, and the additional port is not needed for the wavelength monitoring technique.

II. OPERATION PRINCIPLE

Silica-based Bragg grating has the same temperature-induced wavelength drift ratio as that of silica-based AWG, 0.01nm/˚C [2]. When the RSOA is self-locked by the reflected light from the BG, its lasing wavelength is shifted with the same amount of the wavelength drift in the AWG. This drift is detected by measuring the optical power of the self-locked RSOA passed through a narrow optical band-pass filter (BPF). The detected wavelength drift of the AWG can be used to control WDM channel.
wavelengths at the OLT to minimize the misalignment along the optical path [4]. In bidirectional transmission, a cyclic AWG with skip-band is mostly used and, as a result, the lasing light from the self-locked RSOA could be transmitted to the optical network unit (ONU) and the OLT. Therefore, we select a wavelength for monitoring in our experiment within the skip-band, not a pass-band of the AWG.

III. EXPERIMENT AND RESULTS

The experimental setup is shown in Fig. 2. The RSOA is added through a 3-dB coupler to the downstream. One output from the coupler was coupled to the feeder fiber. To investigate length-dependent wavelength change, various lengths of single mode fibers were tried. The peak wavelength of the self-locked RSOA was measured by the optical spectrum analyzer (OSA) instead of the optical BPF and the optical power meter. The bias current of the RSOA was set at 70mA. The fiber Bragg grating (FBG) with 3-dB bandwidth of ~0.35nm was used as a BG, and its center wavelength was 1550.82nm at 30°C. The temperature of the RN was changed by the oven over a temperature range from 20°C to 60°C.

Fig. 3 shows the spectra of the reflection of the FBG and the self-locked RSOA under the back-to-back operation. Side-mode suppression ratio (SMSR) and the 3-dB bandwidth of the self-locked RSOA appeared to be ~36dB and ~0.15nm, respectively. By comparing of two spectra, the peak wavelength of the self-locked RSOA was placed at the falling edge of the FBG. The inset of Fig. 3 shows the spectra of the self-locked RSOA according to different fiber lengths at 30°C. As a result, the peak wavelength of the RSOA does not change according to the fiber length. Therefore, the proposed technique can be used independently of the length of the feeder fiber.

Fig. 4 shows that the temperature-induced wavelength drifts of the self-locked RSOA, the reflection curve of the FBG, and the AWG were measured from 20°C to 60°C. Their drift ratios were about 0.0136nm/°C, 0.0141nm/°C and 0.0117nm/°C, respectively. This slight difference was within a measurement error range due to the resolution of the OSA.

IV. CONCLUSION

We have proposed and experimentally demonstrated a simple wavelength monitoring technique of the RN based on the RSOA, which is self-locked by the FBG at the RN. By measuring the wavelength drift in the lasing wavelength of the self-locked RSOA, we can effectively reduce the misalignment of wavelength by temperature variation of the RN in WDM-PON.

ACKNOWLEDGEMENT

This work is partially supported by GIST Technology Initiative (GTI) and by the BK-21, Republic of Korea.

REFERENCES

MZI-SOA based all-optical router implementation with OCDMA header recognition

P. Teixeira¹, L. Oliveira¹, T. Silveira¹², P. André¹, R. Nogueira¹, M. Lima¹, A. Teixeira¹
¹Instituto de Telecomunicações, Campus Universitário de Santiago 3810-193 Aveiro, Portugal
²Siemens Networks SA, Rua Irmãos Siemens 1, 2720-093 Amadora, Portugal

Phone: +351-234377900, Fax: +351-234377901, e-mail: pteixeira@av.it.pt, lmoliveira@av.it.pt, tiago.silveira@siemens.com, pandre@av.it.pt, rmogueira@fis.ua.pt, nlima@det.ua.pt, teixeira@ua.pt.

Abstract — A router implementation based on OCDMA header recognition and MZI-SOA switching, allowing ps range packet switching capabilities is presented in this paper. Architecture and simulation results are presented.

I. INTRODUCTION

In this paper an all-optical router architecture is proposed, where the label detection is based on time domain Optical Code Division Multiple Access (OCDMA) header correlation. To perform routing, the data labels from different users/channels have to be detected and recognized in order to allow all-optical forwarding to the required port. An OCDMA packet header/trail/label, unique for each route, is coded prior/following to each data packet. This label, by all-optical correlation identifies the port where to forward the packet, where a new or the same label can be added. Fig. 1 illustrates the proposed router block diagram.

Fig. 1: All-optical router block diagram.

II. CODING SYSTEM

The encoders can be achieved by, eg., coding orthogonal sequences (eg. Kasami, Gold, Hadamard) with good auto- and cross-correlation properties in Super-Structured Fiber Bragg Gratings (SSBFG), while each decoder is achieved by writing the reversed code sequence of the matching encoder. This is a coherent bipolar system that uses the phase dimension to encode the header/trail [1]. The SSBFGs are divided into segments, one for each bit of the code sequence. Each different bit represents a different phase shift: 0 or π. Thus, between adjacent segments with different phase shifts, a phase discontinuity occurs on the grating periodic variation of the refractive-index modulation (Fig. 2).

Coding system performance can be evaluated by measuring the auto-correlation peak at the decoder output (Pa), the Pa to maximum wing level ratio (P/W) and the Pa to cross-correlation level (P/C). For low values of the grating refractive index, the relation between the grating spatial refractive index modulation profile and the impulse response shape is given by a Fourier-transform [2]. Thus, the refractive index must be kept low, which introduces a trade-off between high reflectivity (high Pa) and high P/W, P/C ratios. Though better correlation values can be obtained by using longer optical codes, the requirements of this system permit to partially disregard this constrain, maintaining system simplicity level. The simulations were run using two 16 chip Hadamard codes. Shorter chip length will achieve better pulse shape fidelity to the code, for the energy decaying effect along the grating [3], and thus higher P/W and P/C. Though, inter-symbolic interference effects due to dispersion are more important to shorter chip time. Hence, our choice is a chip length of 2.3mm which corresponds to a chip time of 22.2ps. Moreover, a minimum 125ps pulselwidth is required to trigger the flip-flops [5], so pulse broadening is needed in a following stage. A study of the coding system performance for different input pulselwidth values is presented in subsections a) and b). The study includes the effect of the apodization of the SSBFG.

a) Apodization

The try-and-error method suggested in [1] was used to apodize the gratings. Here, the reflectivity of different chip along the grating is adjusted in order to maintain the pulse response peak variation within a small range. The apodization was designed for the maximum Pa, occurring at 16ps pulselwidth. The obtained apodization profile is shown in the grating structure depicted in Fig. 2.

In Fig. 3, Pa and P/W, P/C ratios are plotted versus the pulselwidth. Apodization performance is also shown. While the Pa maximum corresponds to 16ps, P/W and P/C present an overall tendency to decrease with pulselwidth as predicted in [4]. For 64ps, P/W shows an abnormal increase due to an observable code distortion by pulse overlapping within code, which introduces a higher
right-side wing whereas for all other pulsewidth, the higher wing is the one in the left.

The effect of apodization, also shown in Fig. 3, is to improve \( P_a \) and \( P/W \), \( P/C \) ratios. For 16ps, the increase is about 5dB for \( P_a \), 2dB for \( P/W \) and 1dB for \( P/C \) and the total power loss, measured as the ratio of auto-correlation peak to input pulse peak, is of 16.6dB for the apodized case and of 20.3dB for the non apodized.

\[ \text{Fig. 3: OCDMA system: pulsewidth vs. performance.} \]

\[ \text{b) Pulse expansion} \]

Once the previous system cannot achieve 125ps wide pulses for flip-flop triggering, we use, after the decoder, a 60mm long apodized BFG acting as a pulse expander to convert the pulsewidth from 16ps to 130ps which introduces a power penalty of about 8dB.

II. THE SWITCH SUBSYSTEM

After correlation, the header becomes a pulse. Both initial and final header pulses are input into the RS flip-flop [5]. When correlated, the initial header sets on the flip-flop and the final header sets it off. During this time, the packet is all-optically switched in the switch subsystem.

The switch subsystem is divided in three MZI-SOA stages for each input. The first stage receives a packet and its respective header signal. This stage decides if the packet should be bar or cross ported, according to the existence or not of the header signal. In case of “not”, the packet will cross the switch by passing through a second MZI-SOA structure. The second stage receives the crossing packet and decides if the packet should be wavelength converted or not. The decision is based on the other packet header’s information. However this MZI-SOA does the XOR operation between the two header signals, since the XOR operation improves the packet pass-to-packet drop ratio (\( P_1/P_0 \)) in about 10dB, according to the simulations. The third MZI-SOA stage is used to implement the wavelength converter.

One important aspect in the switch subsystem is the reminiscent power of the blocked packets. It can be an issue in the system performance. Since it is verified in every MZI-SOA stage, the effect can be cumulative and, consequently, increase the blocked packet power in the system output, making it comparable to the unblocked packet or increasing the crosstalk effect, degrading the unblocked packet.

Considering only output1, Fig.4 shows the variation of the \( P_1/P_0 \). The \( \lambda_o \) (original wavelength) curve represents the variation of \( P_1/P_0 \) for both packets in the original wavelength. The \( \lambda_c \) (converted wavelength) curve represents the same variation for packets that had to be wavelength converted. In (A), the difference, in the best case, between the original and converted wavelength are about 10dB. The crossing packet2 has to pass by the third MZI-SOA stage, decreasing the performance of the \( \lambda_c \) curve. Anyway, a \( P_1/P_0 \) of about 25dB is possible for the \( \lambda_o \) curve, if the header powers are optimized. Even for the \( \lambda_c \) curve, \( P_1/P_0 \) of about 15dB is also a good result.

\[ \text{Fig. 4: Results for output1 (O1). (A) } P_1/P_0 \text{ as a function of the header1 (H1) power when header2 (H2) power is fixed. (B) } P_1/P_0 \text{ as a function of both headers Extinction Ratio (ER).} \]

The results referred in Fig. 4 (A) considered the ER of both header signals as ideal. By varying the ER of both header signals, Fig. 4 (B) illustrates that, for output1, \( P_1/P_0 \) for the original wavelength packets decreases along with the ER of the header signals. The decreasing is mostly caused by the packets that are being bar ported, which are directly dependent of the corresponding header power. The packets that are crossing the switch, although in the same wavelength, pass through the XOR gate, which helps them to maintain the same amplitude.

III. CONCLUSION

A model to implement an all-optical router was demonstrated by means of simulation. It showed good performance in terms of Packet pass-to-Packet drop ratio. A study about the viability of using OCDMA encoding based on SSFBG to implement the headers correlators was made. It also showed promising results.

REFERENCES


Crosstalk in Downlink Carrier Reused WDM-PONs
Based on Subcarrier Modulation

Z. Xu\textsuperscript{1,2}, Y. J. Wen\textsuperscript{1}, W.-D. Zhong\textsuperscript{2}, X. F. Cheng\textsuperscript{1}, Y. Wang\textsuperscript{1}, M. Attygalle\textsuperscript{3}, J. Shankar\textsuperscript{4}, and C. Lu\textsuperscript{4}

1. Institute for Infocomm Research, A*STAR, 21 Heng Mui Keng Terrace, Singapore 119613
2. School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore 639798
3. National ICT Australia Ltd, Department of EEE, the University of Melbourne, Australia
4. Department of Electronic and Information Engineering, Hong Kong Polytechnic University, Hong Kong

Email: yjwen@i2r.a-star.edu.sg

Abstract
We demonstrate that, in carrier reused WDM-PONs based on subcarrier modulation, residual downlink data exists in the reused optical carrier. Both experiments and simulations show that uplink suffers from impairments due to the crosstalk of the residual data.

1 Introduction
Today, wavelength division multiplexed passive optical networks (WDM-PONs) are very attractive due to its potential for future broadband access networks. WDM-PONs do not suffer from power-splitting losses and allow enhanced reliability and privacy because of their virtual point-to-point connections [1]. Centralized light sources of the uplink make WDM-PONs more convenient for wavelength arrangement and upgrade. Some techniques have been proposed to achieve centralized light sources, such as remotely seeding Fabry-Perot lasers [2] and reflective SOAs [3] by spectrum-sliced amplified spontaneous emission (ASE) noise. However, these schemes need two wavelengths arrangement for uplink and downlink, respectively. Recently, subcarrier transmission technique was applied to PONs, where the downlink data was carried on the subcarriers [4, 5]. At each ONU, the subcarriers were filtered out for downlink data detection, while the optical carrier was considered as CW light and remodulated for upstream transmission.

In this paper, we analytically and experimentally investigate the optical subcarrier modulation and show that the reused optical carrier still carries residual baseband data, which cannot be ignored especially for large signal modulation. This residual data degrades the uplink performance when the downstream optical carrier is reused for uplink data encoding. Both simulations and experiment are carried out to examine this degradation

2 Downlink Carrier Reused WDM-PON
In subcarrier modulation (SCM), the baseband data is mixed with a local oscillator (LO) and upconverted to a subcarrier. The mixed signal is then modulated onto a CW light by a Mach-Zehnder modulator (MZM) in amplitude shift keying (ASK) format and transmitted to ONU via a feeder fiber, as shown in Fig. 1. Here only a single channel is considered for principle demonstration. In this system, ONU can have different configurations as shown in Fig 1. In the first two configurations, the carrier and subcarriers are separated by a fiber Bragg grating (FBG) and a circulator. In configuration 3, a delay interferometer (DI) is used to separate the carrier and subcarriers. At each ONU, the subcarriers are detected by downstream receiver and the separated optical carrier is remodulated as uplink light source. In configuration 1, an external modulator is used for data encoding, while the last two schemes use RSOA for data encoding.

Assuming the bias voltage of the MZM is $V_b$, the downstream output optical field of the modulator is [6]

$$E(t) = E_m \cos (\alpha) J_\nu (\beta) \exp (j \omega t) + V_b$$

where $E_m$ and $\omega$ are the amplitude and carrier frequency of the input light, respectively, $\beta$ is the frequency of the local oscillator, $V_b$ is the half-wave voltage of the modulator, and $V(t) = \sum b_k A_k f(kT_s) \leq t \leq (k+1)T_s$ is the baseband signal voltage, where $T_s$ is the bit duration of the downstream data, $b_k$=0 or 1 is the bit value, and $f(t)$ is the data waveform.

Expanding Eq.(1) in terms of Bessel functions leads to

$$E(t) = E_m \cos (\alpha) \sum_{n=0}^{\infty} (-1)^n J_{\nu+n} (\beta) \exp [j 2n(\omega t + \omega_b) t]$$

$$A \cos (\alpha) \sum_{n=0}^{\infty} (-1)^n J_{\nu+n} (\beta) \exp [j \omega (2n+1) \omega_b t]$$

Here, $\alpha = \pi V_b / V_m$ and $\beta = \pi V_m / V_b$. Eq. (2) shows that the modulated optical field contains many frequency components, according to $\omega_{2n+1} = \omega_b$ or $\omega_{2n} = \pi \omega_b$, $n=0,1,2,\ldots$. The first

![Fig. 1 PON based on downlink optical carrier reusing with different ONU configurations.](image-url)
carrier due to the residual data also increases with the RF signal amplitude, which is not desirable. Experimental results show that when $V_o = V_c/4$, the separated downlink optical carrier indeed carries residual baseband data with an extinction of 1.7 dB as shown in the inset of Fig. 3. It introduces crosstalk when the carrier is reused for upstream transmission. In this work, we demonstrated the transmission of the first two ONU configurations, as shown in Fig. 1, to verify the impact of the residual data.

Fig. 4 shows the BER performance of the upstream transmission based on downlink carrier remodulation. Here, both downstream and upstream data rates are 1.25 Gb/s. Local oscillator is 12.5 GHz. The optical carrier to subcarrier ratio (OCSR) is 11 dB. The transmission link is 21 km single mode fiber. In the first configuration, we use an ideal FBG to filter out the optical carrier perfectly and simulate the bidirectional transmission. Considering the receiver sensitivity at 10$^{-10}$, simulation results showed there was about 3 dB power penalty due to the crosstalk of the residual downlink data. The second configuration was experimentally demonstrated using a simple apodized FBG with a 3 dB bandwidth of 0.15 nm. The OCSR of the seeding light was increased to 18 dB after FBG. The seeding power launched into the RSOA was −20 dBm. Experimental results also show a 1 dB power penalty, which is smaller than the first case. This reduced power penalty is due to the gain saturation of RSOA and the residual subcarriers due to the imperfect filtering.

3 Summary

We have investigated optical subcarrier modulation and demonstrated that the residual downstream data exists in the reused optical carrier. This residual data induces crosstalk and degrades the uplink performance when the downlink optical carrier is reused for upstream data encoding. Simulations and experiments demonstrated there were 3 dB and 1 dB power penalties when the carrier was remodulated by an external modulator and RSOA, respectively.

References

Power Penalty Dependency on Sideband Suppression Ratio in Optical SSB Signal Transmission

Toshihito Fujiwara and Koji Kikushima
Access Network Service Systems Laboratories, NTT, 1-6 Nakase, Mihama-ku, Chiba-shi, Chiba, 261-0023 Japan
E-mail: fujiwara.toshihito@ansl.ntt.co.jp

Abstract—This paper analyzes the relationship between the sideband suppression ratio (SSR) and the power penalty theoretically and experimentally. The transmission distance limitation is discussed from the viewpoint of SSR.

I. INTRODUCTION

Optical single sideband (SSB) and vestigial sideband (VSB) have been studied for more than a decade to enhance bandwidth efficiency and transmission distance [1]-[5]. The power penalty caused by chromatic dispersion limits transmission distance of optical double sideband (DSB) signals. Optical SSB and VSB have been mainly compared against optical DSB for use in baseband transmission, and typical evaluation metrics are eye pattern and bit error rate (BER).

Subcarrier multiplexing (SCM) systems are now being considered as one of the applications of optical SSB; this is necessitated by the increase in transmission frequencies. For optical SSB SCM systems, multiple radio frequency (RF) signals or frequency division multiplexed (FDM) signals are transferred sideward by optical carrier. Previous studies has discussed distortion and suppression incompletely. Therefore, the sideband suppression ratio (SSR) directly determines transmission quality through the effect of the power penalty. To the best of our knowledge, the impact of the power penalty has not been studied with regard to SSR.

In this paper, we theoretically and experimentally investigate the power penalty as determined by the SSR.

II. PRINCIPLE

The power penalty is caused by the group velocity dispersion (GVD) between upper sideband (USB) and lower sideband (LSB). We assume that \( \omega_m \) is optical radian frequency of subcarrier, \( m \), which is neighborhood of the optical carrier, \( \omega \). The phase of the subcarrier in GVD component is expressed as

\[
\theta(\omega_m) = -\frac{z\lambda D\omega_m^2}{2\omega}
\]  

(1)

where \( z \) is the distance traveled, \( D \) is the dispersion parameter, and \( \lambda = 2\pi c/\omega \). \( c \) is the velocity of light in free space [10].

Considering the optical carrier and sidebands, the electrical field of the light, \( e \), at time \( t \) is described as follows.

\[
e = \cos(\omega t) + a_U \cos(\omega t + \omega_m t + \theta(\omega_m)) + a_L \cos(\omega t - \omega_m t + \theta(-\omega_m))
\]

(2)

Here, \( a_U \) and \( a_L \) is the amplitude of USB and LSB, respectively. There are harmonics components in practical modulation systems, but we neglect them here because they are very small when the modulation depth is low.

Direct detected electrical current, \( i \), is written as follows.

\[
i \propto |e|^2
\]

(3)

The current of frequency \( m \) component, \( i_m \), is as follows.

\[
i_m \propto \frac{1}{2}(a_U + a_L) \cos(\theta_m) \cos(\omega_m) - \frac{1}{2}(a_U - a_L) \sin(\theta_m) \sin(\omega_m)
\]

(4)

Here, \( \theta_m = \theta(\omega_m) = \theta(-\omega_m) \). Hence, \( i_m \) is at the maximum value when \( \theta_m = 0 \), and at the minimum value when \( \theta = \pi/2 \). We call the difference between the maximum and minimum value as the maximum power penalty, \( P \), here.

\[
P = 20 \log 10 \left( \frac{a_U + a_L}{a_U - a_L} \right)
\]

(5)

SSR is defined as following equation.

\[
SSR = 20 \log 10 \left( \frac{a_U}{a_L} \right)
\]

Equation (5) is rewritten as follows.

\[
P = 20 \log 10 \left( \frac{10^{\frac{SSR}{20}} + 1}{10^{\frac{SSR}{20}} - 1} \right)
\]

(7)

III. EXPERIMENTS AND RESULTS

We conducted experiments to confirm the theoretical performance.

Fig. 1 shows the experimental setup and optical spectrum for the SSR of 20dB. Sideband suppression was varied by a tunable optical filter. Light from the laser diode (LD) was modulated at 12 GHz continuous wave. The wavelength of the LD was set to 1559.07nm. A tunable dispersion compensator of a virtually imaged phased array (VIPA) [11] was used as the dispersion generator.
LD: Laser Diode
EAM: Electro-Absorption Modulator
TDC: Tunable Dispersion Compensator
PD: Photo Diode
SA: Spectrum Analyzer

**Fig. 1. Experimental setup**

- **Fig. 2.** Relationship between dispersion and electrical power: Plots are measured values. Lines are theoretical relationships.

- **Fig. 3.** Maximum electrical power penalty: Points are measured values. The line is the theoretical relationship.

**Fig. 4.** Transmission distance limit versus SSR

Fig. 4 plots the transmission distance limitation versus the SSR. Here, the dispersion is assumed to be 17ps/nm/km.

Assuming that a system has a penalty limit of 1dB, the distance must be limited to 7.5km and 14.6km when the SSR is 0dB and 20dB, respectively; assuming that chromatic dispersion is the only factor to be considered, the transmission distance is infinite at the maximum power penalty of 24.9dB.

**V. CONCLUSION**

We elucidated the relationship between the power penalty and SSR. Experimental results well reproduced the theoretical relationship. We also discussed the maximum transmission distance from the viewpoint of SSR. The information provided here enables us to design systems.

**REFERENCES**


The dependence of the power penalty on dispersion was described above. For the system design, the maximum dependency must be determined.

- **Fig. 1.** Experimental setup

- **Fig. 2.** Relationship between dispersion and electrical power: Plots are measured values. Lines are theoretical relationships.
Simple Multi-wavelength Stabilization Technique Using a Periodic Optical Filter for WDM Access Networks

Tetsuya Suzuki¹, Hiro Suzuki¹, Masamichi Fujiwara¹, Naoto Yoshimoto¹, Katsumi Iwatsuki², and Takamasa Imai¹

¹NTT Access Network Service Systems Laboratories, ²NTT Service Integration Laboratories, NTT Corporation
1-6 Nakase, Mihama-ku, Chiba-shi, Chiba, 261-0023 Japan
Phone: +81 43 211 3038, Fax: +81 43 211 8250, E-mail: suzuki.tetsuya@ansl.ntt.co.jp

Abstract
This paper proposes a multi-wavelength stabilization technique based on a periodic optical filter and maximum power searching. The operating principle is described and the performance is experimentally estimated for four 100-GHz spaced transmitters.

Introduction
The rapid growth of the Internet has drastically increased the total number of data packets being transferred and has necessitated the emergence of broadband access networks. To meet this demand, several wavelength division multiplexed (WDM) access networks employing dense-WDM (DWDM) technology were reported [1-2]. In these systems, wavelength stabilization is required to provide stable services to users. The conventional approach is to incorporate a Fabry-Perot (F-P) Etalon filter into each transmitter to stabilize the output wavelength [3]. However, in this approach the cost increases in proportion to the number of wavelengths.

In this paper, we propose a simple multi-wavelength stabilization technique that is suited to WDM access networks. Each wavelength is stabilized in the desired ITU-grid one-by-one using a single optical filter with a periodic peak profile. We conduct an experiment to confirm the feasibility of this technique.

WDM Transmitter Configuration
Figure 1 shows the WDM transmitter configuration employing the proposed multi-wavelength stabilization technique. This configuration includes N transmitters, an optical multiplexer (MUX), an optical coupler, a periodic optical filter, an optical power monitor, and a wavelength stabilization circuit. Each transmitter consists of a laser diode (LD) with an auto power control (APC) circuit and a temperature controller. The wavelength stabilization circuit determines the LD temperature from the optical power obtained through the filter. The transmission spectrum of the optical filter is a periodic peak profile that has minimum losses at the desired ITU-grid. Optical filters such as a Mach-Zehnder interferometer (MZI) filter and a F-P Etalon filter can also be applied. Each wavelength is stabilized one-by-one by searching for the LD temperature at which the optical power obtained through the filter is the maximum within the spectral range involving the target spectral peak.

Wavelength stabilization is achieved by varying the LD temperature and searching for the local maximum. Figure 2 shows the operation scheme when searching for the local maximum. The optical power through the filter is measured each time while changing the LD temperature. The LD temperature is varied in steps based on arbitrary units as indicated in Fig. 2. Initially, the optical power is measured at the initial LD temperature and then compared to the measurements at temperatures above and below that point. Thus, the direction that the LD temperature changes when the optical power increases is determined. Next, the LD temperature is varied until the optical power decreases. Thus, three points surrounding the local maximum are selected. Finally, the LD temperature corresponding to the local maximum is determined by interpolating a parabola against the three points.

Fig. 1: WDM transmitter configuration employing proposed multi-wavelength stabilization technique

Fig. 2: Operation scheme for local maximum search
Experiments

To confirm the feasibility of this technique, we conducted an experiment. Four Distributed-Feedback LDs (DFB-LDs) are connected to a 100-GHz spaced MUX with a flat profile. The multiplexed WDM signal is tapped using a 3-dB optical coupler, and input to a 25-GHz spaced MZI filter. The optical power input to the filter is –12 dBm/ch. The optical power through the filter is measured using an optical power meter. The one-by-one local maximum search operation program is fabricated on a PC.

Figure 4 shows the experimental results for the case where the initial frequency deviations are set to +10 GHz for ch. 1, -10 GHz for ch. 2, +10 GHz for ch. 3, and –10 GHz for ch. 4. The frequency deviation of each channel is expressed by the relative frequency against the desired ITU-grid. Based on the one-by-one local maximum searching operation, the frequency deviation of each channel is suppressed to +0.8 GHz for ch. 1, +1.2 GHz for ch. 2, +1.2 GHz for ch. 3, and +0.6 GHz for ch. 4. Regardless of the sign of the initial frequency deviation, we find that the frequency deviation after the stabilization is suppressed to 1.2 GHz for the four transmitters.

Discussion on Effect of Modulation

In the proposed technique, the wavelength is stabilized in the wavelength corresponding to the local maximum. In the case of a symmetric optical spectrum such as a continuous wave (CW) and externally modulated signal, the wavelength corresponding to the local maximum corresponds to the desired ITU-grid. On the other hand, in the case of an asymmetric optical spectrum such as a directly modulated DFB-LD signal, which is caused by an adiabatic chirp, there is the possibility that the wavelength corresponding to the local maximum is somewhat far from the desired ITU-grid.

Therefore, we investigated the effect of a directly modulated DFB-LD signal on the proposed technique. Figure 4(a) shows the output spectrum of the 1.25-Gbit/s directly modulated DFB-LD used in this investigation. As shown in the figure, the directly modulated signal exhibited adiabatic chirp dominant behavior [4], i.e., strong and weak peak powers in the optical spectra. For simplicity, we define the frequency with the strongest peak power as the relative frequency of zero. Figure 4(b) shows the relationship between the optical power through a 25-GHz spaced MZI filter and the relative frequency against the peak frequency of the filter. The peak frequency in the case of the CW (dashed line) corresponds to the peak frequency of the filter. In the case of the directly modulated signal (solid line), the peak frequency deviates by approximately 0.4 GHz from that of the CW. This is because the weighted center of the output optical spectrum lies on the frequency between the two spectral peak powers. Considering the modulation effect, we conclude that the frequency deviation after the stabilization in the case of a directly modulated DFB-LD signal is expected to be less than approximately 2 GHz.

Conclusion

We proposed a simple multi-wavelength stabilization technique for WDM access networks. This technique reduces the number of optical filters used for wavelength stabilization compared to the conventional method. Each wavelength is stabilized in the ITU-grid one-by-one by searching for the LD temperature at which the optical power obtained through a periodic optical filter is the maximum within the spectral range involving the target spectral peak. We experimentally confirmed that regardless of the sign of the initial frequency deviation the frequency deviation against the desired ITU-grid is suppressed to 1.2 GHz using four 100-GHz spaced CW lights. We quantitatively estimated the effect of the asymmetric profile due to direct modulation and concluded that the proposed technique stabilizes the frequencies of directly modulated signals within approximately 2 GHz.

References

Local Area Network Emulation in Passive Optical Networks by Wavelength Switching the Distributed Feedback Laser

Nishaanthan Nadarajah and Chang-Joon Chae
National ICT Australia, Victoria Research Laboratory, Dept. of Electrical & Electronic Engineering, The University of Melbourne, VIC 3010, Australia.
Tel: +61 3 8344 6095; Fax: +61 3 8344 6678; email: mnad@ee.unimelb.edu.au

Abstract: A novel scheme for upstream transmissions and local area network emulation amongst customers in a passive optical network by fast current-tuning a distributed feedback laser placed at the customer premises is proposed and experimentally demonstrated.

Introduction: Passive optical network (PON) technology is considered as the most promising candidate for next generation broadband access due to its cost effectiveness, simplicity, and easier upgradeability [1]. Apart from the upstream and downstream transmissions between the central office (CO) and the optical network units (ONUs), customers of a PON may require private communication links between themselves for various services. To facilitate these services, a number of optical layer local area networking (LAN) schemes have been demonstrated [2, 3]. One of the earlier proposals for the optical layer LAN emulation scheme uses a separate optical transceiver at each ONU [2]. The use of a separate optical transceiver at each ONU for the LAN emulation is not effective in a cost-sensitive access networks. RF subcarrier multiplexing has also been used to carry LAN traffic whereby a single laser source is used at each ONU [4, 5]. However, this scheme requires a narrow linewidth laser and high frequency RF electronics at each ONU making the cost and complexity at the ONUs high. In this paper, we propose and experimentally demonstrate a simple and cost efficient solution for optical layer LAN emulation using a single uncooled distributed feedback (DFB) laser placed at each ONU. The laser is used for upstream lasering channel of the DFB laser is switched by changing the bias current of the DFB laser and therefore two separate traffic transport can be carried out. Compared to the previous schemes, this scheme does not require high speed RF electronics, stable laser sources, and modulators at each ONU. Moreover, this scheme enables that the time slots allocated to each ONU are efficiently used [6]. We experimentally demonstrate the proposed LAN emulation scheme with 2.5 Gb/s downstream traffic, 1.25 Gb/s upstream traffic, and 1.25 Gb/s LAN traffic.

System Architecture: The proposed scheme for implementing upstream transmission and LAN emulation by fast current-tuning a DFB laser is shown in Fig. 1. A (1xN) star coupler (SC) is used to split/combine the optical signals to/from each ONU, whereby the number of ONUs attached to the SC is N. An uncooled DFB laser is used at each ONU and the operating wavelength channel of this DFB laser is changed by varying the bias current of the laser. For the bias current of I1, the DFB laser is tuned to upstream wavelength channel (\(\lambda_u\)) and therefore upstream transmissions to the CO can be carried out. When the current is changed to I2, the DFB laser is tuned to LAN wavelength channel (\(\lambda_{LAN}\)) and therefore LAN transmissions to other ONUs in the PON can be performed. A FBG is placed in feeder fiber close to the star coupler (SC) such that it reflects LAN wavelength channel (\(\lambda_{LAN}\)) back to the ONUs. Moreover, as data for CO and LAN are separated using the FBG, no further filtering is required at CO and ONUs. Unlike previous LAN emulation schemes, this scheme enables higher bandwidth LAN traffic transport. Upstream access follow the time division multiple access (TDMA) protocol, while LAN may follow any media access control protocol.

Experimental demonstration: The experimental setup to demonstrate the feasibility of the proposed scheme is shown in Fig. 2. A downstream signal of 2^{31}-1 pseudo random binary sequence non-return to zero (PRBS NRZ) data at 2.5 Gb/s was modulated onto downstream wavelength channel \(\lambda_D = 1540.08\) nm using a Mach-Zehnder modulator and transmitted to the ONUs through a 10 km feeder fiber, a 4 x 4 SC and a 3 km distribution fiber. At the ONUs, \(\lambda_D\) was separated from wavelength channels \(\lambda_U = 1550.56\) nm and \(\lambda_{LAN} = 1550.92\) nm using a coarse wavelength division multiplexing (CWDM) coupler. For the upstream transmission mode, the DFB laser was biased at 30 mA, while the bias current was

\[ \lambda_U = 1550.08 \text{ nm} \]

\[ \lambda_D = 1540.08 \text{ nm} \]

\[ \lambda_{LAN} = 1550.92 \text{ nm} \]

\[ \lambda_U = 1550.56 \text{ nm} \]

---

11A2-4
increased to 45 mA for the LAN transmission mode. Therefore, a wavelength change of 0.36 nm was obtained from the DFB laser for 15 mA difference in the bias current. A FBG with a Bragg wavelength of 1550.92 nm and a 3 dB bandwidth of less than 30 GHz was placed in the feeder fiber next to the SC for the reflection of $\lambda_{\text{LAN}}$ back to the ONUs. In both transmission modes, $2^1-1$ PRBS NRZ data at 1.25 Gb/s was directly modulated onto the DFB laser and transmitted in the upstream direction. The signals on all three wavelength channels were detected using a 2.5 Gb/s p-i-n receiver. The unused ports of the SC were anti-reflection treated. A series of experiments were conducted and bit error rates (BERs) for all signals were measured.

**Results and Discussions:** Fig. 3 shows the observed optical spectra at the DFB for upstream and LAN wavelength channels. A suppression of more than 18 dB was observed for $\lambda_{\text{LAN}}$ between the reflected and transmitted portion of the signals. For $\lambda_{\text{U}}$, more than 28 dB suppression was observed between the transmitted and reflected spectra. Fig. 4 shows the measured BER curves for the signals. For the 2.5 Gb/s downstream data, no penalty was observed when the signals were transmitted through the link in the presence of upstream signals compared to the back-to-back (B-B) measurements. Similarly, no penalty was observed for the 1.25 Gb/s upstream data compared to B-B measurements. A penalty of less than 0.2 dB was observed for the 1.25 Gb/s LAN data compared to B-B measurements.

The switching time of the DFB laser is required to be fast to achieve higher transmission efficiency for the signals. An experiment was carried out to measure the switching time of the DFB laser used in the experiment. The DFB laser was directly modulated by a square wave signal at two different current levels (representing two different wavelength channels) and the signals were detected at the LAN data and upstream data receivers. For this purpose, the fibers used in experiment were removed. Fig. 5 shows the timing diagrams for the received upstream and LAN data signals when the wavelength difference between $\lambda_{\text{U}}$ and $\lambda_{\text{LAN}}$ was approximately 0.4 nm. From the experiment, it was observed that 280 $\mu$s pulse width signal can be recovered without much distortion. The switching of the desired wavelength channel is performed in smaller steps and an additional 75 $\mu$s rise/fall time is required. Therefore, the DFB laser is capable of wavelength switching at a rate of 355 $\mu$s for a wavelength difference of 0.4 nm. The residual signals observed in the timeslots in Fig. 5 are due to the imperfect filtering of signals. Using a narrowband FBG for filtering the signals, clean traces can be obtained. It should be noted that LAN data traverses through the SC twice and therefore the bandwidth of the LAN data is limited by the power budget. However, using a $(N+1) \times (N+1)$ SC and a secondary distribution fiber between the SC and each ONU, the bandwidth limitations can be overcome.

**Conclusions:** We have proposed and experimentally demonstrated a scheme for upstream transmissions and LAN emulation capability in PONs using a single DFB laser placed at the ONU. The experimental results show that all signals can be recovered with minimal penalty.

**References:**
Remote Nodes for Wavelength Shared Hybrid PON Supporting Video Overlay

Martin Bouda, Paparao Palacharla, Youichi Akasaka, Alexander Umnov, Takao Naito
Fujitsu Laboratories of America, Inc., 2801 Telecom Parkway, Richardson, TX75093, USA
Tel. +1-972-479-2691, Fax +1-972-479-4482, martin.bouda@us.fujitsu.com

Abstract: A novel architecture has been proposed to enhance performance of power-splitting PON systems cost-effectively, with reuse of existing equipment. Here we demonstrate the feasibility of a modular and cost-effective WS-HPON remote node supporting analog video overlay.

1. Introduction
Passive Optical Networks (PONs) are being deployed widely. The wide adoption of these passive, power-splitting PON architectures, typically serving 32 users by means of Time Division Multiple Access, and operating at serial rates of up to 2.5Gbps rates, indicates that this set of technologies is the most cost-effective and mature for providing broadband access today. Applications such as video conferencing, and immersive on-line shopping will continue increasing the demand for bandwidth. Operators are considering next generation access architectures to provide higher capacities at lower cost per user than existing systems [1].

2. Wavelength-Shared Hybrid PON
WDM with a relatively small number of wavelengths can provide a very flexible, low-risk, cost-effective upgrade path for power-splitting PON systems. The proposed Wavelength-Shared Hybrid GPON (WS-HGPON) is an architecture that can be cost-effectively reconfigured to behave like a standard GPON system, a stack of GPON systems, or hybrids of these, depending on particular needs in the future [2].

Figure 1 shows an example of a symmetrical WS-HGPON with four downstream channels and four upstream channels. The downstream WDM signals from the Optical Line Termination (OLT) are distributed by the Remote Node (RN) to groups of Optical Network Units (ONUs). All ONUs in the PON have identical optical receivers to be able to receive any one of the downstream wavelengths.

![Fig. 1 Symmetrical (4λ+4λ) WS-HGPON.](image_url)

In the upstream, four different wavelengths are used. However, ONUs with fixed wavelength transmitters can be deployed anywhere in the PON, because optical signals outside the downstream wavelength band containing the downstream sub-bands are passed through power splitters just like in the case of a standard GPON. The upstream wavelengths are demultiplexed at the central office with low-loss thin-film filters. The problem associated with WDM-PON systems of having to use more complex technologies to realize colorless ONUs is avoided this way.

This architecture is very cost-effective because of the very small number of additional off-the-shelf components used to quadruple the capacity in both directions over a conventional power-splitting PON. Enhancement band services such as video overlay or other future access systems operating outside the GPON downstream wavelength band can easily be supported in this approach.

3. WS-HGPON Remote Nodes
A WS-HGPON Remote Node can be realized cost-effectively with only five thin-film filters. We have built a modular version of a WS-HGPON RN consisting of six cards as shown in Fig. 2. There are four conventional 2x8 power splitter modules, one conventional 2x4
power splitter module, and one HGPON upgrade band-drop/MUX module. The first five modules can be used to build a standard 1x32 RN (a cascade of 1x4 and 1x8).

![Fig. 2 A modular WS-HGPON RN.](image)

The WS-HGPON upgrade module can be inserted between OLT and an input of the 2x4 module to drop and demultiplex the four downstream wavelengths. Each individual wavelength output is then connected to one unused 2x8 splitter module input to distribute single downstream wavelengths to groups of 8 ONUs. All other wavelengths outside the drop band containing the four downstream wavelengths are passed by the WS-HGPON upgrade module to the input of the 2x4 module. Video overlay signals are broadcast to all ONUs in this way, and all upstream signals are collected as if the RN were a standard power splitting RN.

### Table 1 Measured Characteristics.

<table>
<thead>
<tr>
<th>Module output loss (insertion)</th>
<th>1550nm (dB)</th>
<th>1491nm (dB)</th>
<th>1496nm (dB)</th>
<th>1501nm (dB)</th>
<th>1505nm (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>MUX through</td>
<td>0.4</td>
<td>-35</td>
<td>-31</td>
<td>-25</td>
<td>-20</td>
</tr>
<tr>
<td>MUX downstream ports #1–#4</td>
<td>N.A.</td>
<td>0.9</td>
<td>0.7</td>
<td>1.1</td>
<td>1.0</td>
</tr>
<tr>
<td>2x8 #1, port #1</td>
<td>16.5</td>
<td>9.5</td>
<td>-40</td>
<td>-41</td>
<td>-30</td>
</tr>
<tr>
<td>2x8 #2, port #1</td>
<td>15.8</td>
<td>-49</td>
<td>8.8</td>
<td>-40</td>
<td>-38</td>
</tr>
<tr>
<td>2x8 #3, port #1</td>
<td>16.2</td>
<td>-49</td>
<td>-40</td>
<td>10.8</td>
<td>-39</td>
</tr>
<tr>
<td>2x8 #4, port #1</td>
<td>16.1</td>
<td>-49</td>
<td>-40</td>
<td>-41</td>
<td>11.1</td>
</tr>
</tbody>
</table>

All values including loss from SC connectors and patch cords between modules.

The requirements on the optical components are easy to meet. Standard Thin Film Filters have temperature coefficients below 3pm/°C, equivalent to a shift of the filter edges by less than 0.38nm over a −40...+85°C temperature range. The isolation between the grid of downstream wavelengths, and other wavelengths such as video overlay and upstream is realized by the filters in the demultiplexer and the band drop filter together. Even with a drop filter with relatively low isolation, a very high isolation can be achieved between multiplexer ports and through port of −75dB, which is important for a high quality analog video signal distribution. The band drop filter has to operate from 1260nm to over 1600nm and have low insertion loss outside the drop band, which was 100nm wide in this work. The losses due to the band drop filter were low, about 0.4dB at 1.55μm, and about 0.2dB at 1.3μm, which is within the measurement accuracy due to connectors.

The RN insertion losses vary between 9.5dB and 11.1dB. Compared to the insertion losses of 15.8dB...16.5dB at 1.55μm, and 16.2dB at 1.3μm we find that the distribution loss for WDM signals in the downstream is at least 4.7dB lower. One possible use of this is to reduce downstream power levels accordingly, to eliminate Raman cross-talk to analog video signals in the 1.55μm band [3].

### 4. Conclusion

We have demonstrated the feasibility of a remote node supporting our proposed WS-HGPON architecture, with analog video overlay. Only five additional thin-film filters are needed compared to a standard power splitting remote node and the loss of an WS-HGPON remote node will be only slightly higher than that of a GPON power splitting node.

### References


Optical-Label and Code Empowered Systems for Next Generation Photonic Networks

S. J. Ben Yoo

Department of Electrical and Computer Engineering, University of California, Davis, California 95616, U. S. A.
yoo@ece.ucdavis.edu

Abstract: This paper discusses optical-label and optical-code based networking technologies for next generation photonic WAN, MAN, and LANs. Optical-Label switching is expected to play a key role in photonic WANs and MANs, and Optical-CDMA offers flexible network reconfiguration in photonic LANs. All-optical code translation capability helps cascaded operation of secure optical networks.

Introduction

Optical-label switching (OLS) [1] and optical-code-division-multiple-access (O-CDMA) [8,9] technologies both emerged as viable future networking technologies. Optical-Label switching (OLS) facilitates introduction of optical packet switching by providing a shim layer between IP and WDM, and allows seamless upgrades from optical circuit switching and burst switching networks towards agile OLS networks supporting optical packet/burst/circuit traffic. OLS routers deployed in the future WANs and MANs can facilitate this transition. The O-CDMA technology also provides flexible networking in LANs. It offers flexible assignment and access of bandwidth using optical-codes, and the data plane can support IP and other format traffic. This paper discusses successful systems integration of OLS routers, O-CDMA nodes, and network demonstrations incorporating OLS and O-CDMA technologies.

Optical Label and Optical Code Empowered Networks

The key underlying networking concept behind Optical-Label Switching [1] is an efficient and transparent packet forwarding method using an optical-label switching mechanism which can co-exist with legacy WDM technologies on the same fiber. Figure 1 shows OLS networking in the WAN and MANs working together with O-CDMA LANs. New signaling information is added in the form of an optical signaling label which is carried in-band within each wavelength in the multi-wavelength transport environment. The optical-label containing routing and control information such as the source, the destination, the priority, and the length of the packet, will propagate through the network along with the data payload. Each optical-label switching router will sense this optical-label, look-up the forwarding table, and take necessary steps to forward the packet. If the packet is to be routed to a wavelength/path where there is already another packet being routed, the optical label switching router (OLSR) will seek routing by an alternate wavelength, by buffering, or by an alternate path. This wavelength, time, and space domain contention resolution is a key to implementing optical-router without heavily relying on time-buffers as conventional electronic routers do [2].

Optical-label switching accommodates data packets of any length, flows of an arbitrary number of packets, a burst of a long datagram, and even a circuit-connection. Fig. 2 illustrates the schematic of the integrated OLSR [5]. The OLSR consists of the optical router controller, the optical label extractor, the optical label rewriter, the optical label detector, a switch fabric, and client interfaces. The optical router controller, implemented by a field programmable gate array (FPGA) incorporates the wavelength-time-space domain contention resolution algorithm. The switching fabric consists of rapidly tunable wavelength converters and arrayed wavelength grating routers (AWGR) [4] and fixed wavelength converters. With the GMPLS extension, the OLS system is designed to interoperate with MPLS, MPLambdaS and IP [2]. Successfully integrated OLSRs achieved 1,001-hop cascaded operation thanks to all-optical 3R burst mode regeneration incorporated in the OLSR. Fig 3 shows the experimental data. Recent experiments also
showed a successful field trial across 477 km San Francisco bay dark fiber NTON-Sprint networks, and a demonstration of IP-client-to-IP-client multimedia demo with multicast packet switching across an all optical label switching network using the OLS edge routers.

For O-CDMA, we employed Spectral Phase Encoded Time Spreading (SPECTS) method for coherent O-CDMA. The eventual goal is to achieve flexible optical code based networking using integrated O-CDMA chips like one shown in Fig. 4.

Using an InP based integrated SPECTS O-CDMA encoders and decoders, we achieved all-optical code translations. Figure 5 shows the cross-correlation traces of O-CDMA code decoding and translation achieved all-optically. [11]

Figure 5 Cross-correlation traces of (a) the cascaded encoder-decoder output with only phase error compensation applied to the phase modulators, (b) the encoder output under W5 encoding, (c) the decoder output for correctly decoded signal, and (d) the decoder output for incorrectly decoded signal. (Solid lines are experimental results and dotted lines are simulated results. W5 code is [11100000]). [11]

Using a bulk optics testbed, we have demonstrated 320 Gb/s O-CDMA LAN in support of 32 users at 10 Gb/s data rates [12]. Fig. 6 shows the BER performance of the testbed (a) with Forward-Error-Correction (FEC) and (b) without FEC.

Fig. 6. The BER performance of the 32 user 10 Gb/s testbed (a) without FEC and (b) with FEC. [12]

Conclusion
OLS and O-CDMA technologies can offer powerful means to realize flexible and unified next generation networks spanning WAN, MAN, and LANs with agility to support packet/burst/circuit traffic on a unified platform with full interoperability with legacy networks.

Acknowledgments
This work was supported in part by DARPA and AFRL under agreement number F30602-00-2-0543, by DARPA and SPAWAR under agreement number N66001-01-1, by NSF under grant number ANI-998665, and by the support of OIDA JOP program supported by DARPA and NSF.

References
Enabling Techniques for Multi-user Asynchronous OCDMA System

X. Wang1, N. Wada1, G. Cincotti2, and K. Kitayama3

(1. National Institute of Information and Communication Technology (NICT), 4-2-1 Nukui-Kitamachi, Koganei, Tokyo 184-8795 Japan, Tel: +81-42-327-7052 Fax: +81-42-327-7035, Email: xwang@nict.go.jp)
(2. Department of Applied Electronics, University of Roma Tre, Rome, Italy)
(3. Department of Electrical, Electronics and Information Systems, Osaka University, 2-1 Yamadaoka, Suita, Osaka 565-0871, Japan, Tel:+81-6-6879-7692, Fax:+81-6-6879-7688)

Abstract
Key techniques enabling asynchronous coherent OCDMA include: lowering the interference level by ultra-long superstructured FBG, optical threshtolder and multi-port encoder/decoder; enhancing noise tolerance of system by forward-error-correction and differential-phase-shift-keying with balanced detection; and security enhancement by code-shift-keying.

Introduction
Optical code division multiple access (OCDMA) is one promising technique for next-generation broadband access network [1]. Figure 1 shows the architecture and working principle of an OCDMA network. It has advantages of full asynchronous transmission, low latency access as well as soft capacity on demand. Recently, coherent OCDMA using ultra-short optical pulse is receiving increasing attention with the progress of reliable and compact encoder/decoder (E/D) devices, such as spatial light phase modulator (SLPM) [2-4], micro ring resonator [5], planar lightwave circuit (PLC) [6], superstructured FBG (SSFBG) [7-8] and arrayed waveguide grating (AWG) multi-port device [9-11].

A multi-user coherent OCDMA system may suffer from severe signal-interference (SI) beat noise if the signal and interferences overlap each other. If the receiver is fast enough to perform the chip-rate detection, the SI beat noise will be the dominant noise source and eventually limits the maximum number of active users that can be supported [1]. However, in a practical multi-user coherent OCDMA system, data-rate detection accompanies with the recovered signal pulse.

Enabling techniques

a. Ultra-long OC generation/recognition with SSFBG

Employing ultra-long optical code (OC) with uniform cross-correlations to lower the interference level \( \xi \) is an effective approach to suppress the noise level. Theoretical analysis has predicted that to support up to \( K = 10 \) error free (BER<10^-6) transmission with chip-rate detection, \( \xi \) should be lower than -28 dB [1]. Phase-shifted SSFBG E/D is one desired candidate that has the capability to process OC as long as 511-chip with chip-rate as high as 640 Gchip/s [8]. SSFBG also has advantages of polarization independent performance, low and code-length independent insertion loss, which is also essential for ultra-long OC generation compared to PLC type E/D, inherent compatibility with fiber-optic system, high compactness as well as potential low cost for mass producing.

b. Multi-port E/D with high PCR

AWG-based multi-port OCDMA E/D has the unique capability of simultaneously processing multiple time-spreading OCs with one device [9], which makes it a
potential cost-effective device to be used in the central office of OCDMA network to reduce the number of E/Ds [10-11]. Another attractive feature of the AWG E/D is that it has very high power contrast ratio (PCR) between auto- and cross-correlation signals compared to other coding devices. The AWG E/D can reach 15~20 dB PCR, while the PCR of the SSFBG is around 1 dB and SLPM is 0 dB. That means the \( \xi \) value could be significantly reduced (up to 20 dB) using the AWG decoder with the same length of code [10]. Moreover, flexibility of the OCDMA network can be improved by hybrid using different type of the E/D in a network [11].

c. Optical thresholding In the coherent OCDMA with ultra-short optical pulse, the properly decoded signal is rather narrow (in a chip time) compared with the bit duration. In a practical system that employs “data-rate” instead of “chip-rate” detection, applying optical thresholding (OT) technique to remove the MAI noise is essential to enable data-rate detection for achieving a practical asynchronous OCDMA system [12]. Optical thresholding techniques have been demonstrated by using nonlinear effect in periodically-poled lithium niobate (PPLN) and optical fibers to enable data-rate detection.

Another kind of approaches to improve the multi-user capability is to enhance the system noise tolerance by:

d. Forward-error-correction (FEC) which is a powerful technique to enhance the system performance.

e. Differential-phase-shift-keying (DPSK) with balanced detection

Coherent OCDMA with DPSK modulation format and balanced detection (DPSK-OCDMA) offers the advantages of improved receiver sensitivity, better tolerance to beat noise and MAI noise without OT, and no need for dynamic threshold level setting [11,14]. Theory predicts that for a given \( K \), DPSK-OCDMA can tolerate about 4 dB higher interference level comparing to OOK-DPSK with optimal threshold, therefore, DPSK-OCDMA is superior to OOK-OCDMA for multi-user coherent OCDMA system.

f. Security enhancement in by CSK-OCDMA

If an eavesdropper can tap the signal in an OCDMA network before multiplexing, it could easily break the security by simple data-rate power detection without any information about the OC. Therefore the OOK-(also DPSK)-OCDMA has vulnerable security.

Coherent OCDMA with code-shift-keying data modulation and balanced detection (CSK-OCDMA) scheme has better multi-user capability and simpler threshold setting in the receiver than OOK-OCDMA. In particular, CSK-OCDMA has superior security to both OOK- and DPSK-OCDMA that the security can not be broken without code information as it is shown in Fig.3 [15]. Therefore, CSK-OCDMA is an attractive candidate for secure transmission systems.

**Multi-user asynchronous OCDMA experiments**

The general experimental setup of multi-user asynchronous OCDMA system is shown in Fig. 4 [8, 10-11]. Fixed fiber delay lines with different lengths were used to randomly set the transit delays and de-correlate the signals from different OCDMA users. Tunable optical delay lines (TODL) were used to investigate the impact of different phases of signal-interference overlapping. Variable optical attenuators were employed to balance the optical power of signals from different users and optical switches were placed to adjust the number of active users. Polarization controllers (PC) were placed as well for investigating the system performance in different scenario. To guarantee that the system can operate in a truly-asynchronous environment, we tested the performances in the worst scenario by adjusting the PCs and TODLs. In the experiment, up to 12 active OCDMA users have been supported at 10 Gbps with BER<10^\(-9\)[11].

**Summary**

Several approaches to enable multi-user asynchronous OCDMA in the presence of SI beat noise as well as MAI with data-rate detection are discussed. Further interesting research topics could include high performance OT, OCDMA with advanced modulation formats, and the application of OCDM technique in secure communication.

**References**

5. A. Agarwal, et al., OFC’05 postdeadline, PDP 6, 2005.
11. X. Wang et al., OFC 2006 post-deadline, PDP 44. 2006.
15. X. Wang et al, OECC’06, 6E2-2, 2006.
**Optical Code Label Processing Using Multi-Port Optical Spectrum Synthesizer and Frequency Comb Generator**

*Mitsuko Mieno (1), Fumi Moritsuka (1), Yuki Komai (1), Shimako Anzai (1), Kashiko Kodate (1)
1: Japan Women’s university, 2-8-1 Mejiro-dai bunkyou-ku, Tokyo 112-8681, Japan, awg@fourier.jwu.ac.jp

Naoya Wada (2), Takahide Sakamoto (2), Tetsuya Kawanishi (2), Masayuki Iatsu (2)
2: National Institute of Information and Communications Technology (NICT), 4-2-1, Nukui-Kita, Koganei, Tokyo 184-8795, Japan, wada@nict.go.jp

**Abstract:** 80 different optical-code(OC)-labels are generated simultaneously using an optical spectrum synthesizer (OSS) and multi-wavelength electro-optic frequency comb generator. 40 parallel matched-filtering is performed by an OSS.

1. **Introduction**

Recently, in spite of immaturity of optical technology, many label generation and processing technologies and related technologies have been developed because of their potential abilities of ultra high-speed processing and many applications in future photonic network. In optical packet switching (OPS) and optical code division multiple access (CDMA) system, label generation and processing are key technologies [1,2]. Although OPS system using multiple OC-label processor has been proposed and experimentally demonstrated [1], tunability of the processing device is not sufficient. Optical CDMA system with tunable spectral domain processing device has been proposed and demonstrated [2]. However, multiple processing capability is not shown.

In this paper, we propose experimentally demonstrate a large scale, multiple, OC-label generation and processing technology based on multi-port, fully-tunable OSS and multi-wavelength electro-optic frequency comb generator. The feasibility is confirmed by simulation.

2. **Multi-port, tunable optical spectrum synthesizer**

Figure 1 shows multi-port, tunable OSS [3]. The multi-port OSS consists of 20x20 ports cyclic AWG, variable optical attenuators (VOA), variable optical phase shifters (VOPS), mirror, and circulators. All components are fully integrated by planar lightwave circuit (PLC) technology except circulators. OSS can control amplitude and phase of each spectrum components. The channel spacing of AWG is 10GHz. The spectrum focused to each output waveguides shifts by changing the input port when cyclic AWG with 20x20 in/output port is employed in the OSS. The proposed OSS enables multiple OC-label generation and processing. This OSS performs spectral en/decoding in all wavelength band matching with free spectral range of cyclic AWG.

3. **Simulation of optical encoding and decoding**

We use prime code of optical orthogonal code for simulation. We set prime number $p$, cycle $p^2$, processing code $P=\{C_0, C_1, \ldots, C_{(p^2-2)}\}$, When $i=S_j \cdot j \cdot p^j (j=0,1, \ldots, p-1, S_j=x \cdot j \mod p)$ $C_x=1$ otherwise $C_x=0$. Encoding code is $\{0100110100\}$ of prime number $p=3$, cycle $p^2=9$ by port10. If we decoded by port20, it is matched signal. And if we decoded by port11-19, it is unmatched signal.

Figure 2 shows simulation of optical encoding and decoding results. Figure 2 (a), (d) shows power spectrum and waveform of encoded code by port10. Figure 2 (b), (e) shows power spectrum and waveform of decoded code by port20 (matched). Figure 2 (c), (f) shows power spectrum and waveform of decoded code by port11 (unmatched). In matched case, decoder output peak value of waveform became high. In unmatched case, decoder output peak value of waveform became low. Optical power after decoding by port20 (matched) is -16.6dBm, by port11 (unmatched) is 34.5dBm. Therefore extinction ratio is 17.9dB. We architectured simulation system for optical encoding and decoding to study optical code changing conditions.

**Fig.1 Structure of multi-port OSS.**

**Fig.2 Simulation results of prime code en/decoding.**

(a), (d) Encoding port10
(b), (d) Decoding port20(Matched)
(c), (f) Decoding port11(Unmatched)
4. Multi-wavelength frequency comb generator

Recently, 19x10-GHz electro-optic ultra-flat frequency comb generation using a Mach-Zehnder modulator (MZM) has been demonstrated [4].

Multi-wavelength electro-optic frequency comb generator consists of four laser diode (LD)’s, optical coupler, polarization controller (PC), a LiNbO$_3$ dual-drive MZM, radio-frequency (RF) amplifiers, signal generator (SG), LiNbO$_3$ intensity modulator (LN-IM), and pulse pattern generator (PPG) as shown in Fig. 4. The MZM was large-driven with sinusoidal signals with different amplitudes. The RF sinusoidal signal at a frequency of 10GHz was generated from SG, amplified with microwave booster, divided half with a hybrid coupler, and then fed into each modulation electrode of the modulator [5]. Central wavelengths of four LDs are 1556.96nm (λ1), 1554.18nm (λ2), 1552.34nm (λ3), and 1550.69nm (λ4). Four different central wavelengths tuned in to central wavelength of each diffraction order of AWG. Figure 4 shows simultaneously generated frequency comb with four different central wavelengths. Therefore, combination of multi-port OSS and multi-wavelength frequency comb generator realize large scale OC-label processing.

5. Multiple OC-label generation and processing

Figure 3 shows the experimental set-up for multiple OC-label generation and processing. To generate 80 different OC-labels at the OSS outputs, pulses from the comb generator is fed into one of the OSS input port. Figure 3 (a)-(d) and (i)-(iv) are encoded waveforms and spectrum with each central wavelength (λ1, λ2, λ3, λ4) by port 1. Figure 3 (e)-(h) and (v)-(viii) are sample of encoded waveforms and spectrum with λ1 by port 1-20. We assigned ports 1 to 10 for encoding and ports 11 to 20 for decoding instead of two OSS’s for en/decoding. The generated OC-labels by port 1-10 were decoded by port 11-20. Figure 3 (i)-(l) and (ix)-(xii) are matched decoding waveforms and spectrum by port 11 of λ1, 15 (λ2, λ3), and 20(λ4). Figure 3 (m)-(p) and (xiii)-(xvi) are decoded waveforms and spectrum with λ1 of the OC-labels forwarded from port 10 to port 11-20. Only matched signal is extracted and unmatched signals are well suppressed. 40 parallel matched-filtering has been performed by an OSS.

6. Conclusions

We have proposed and developed a new multi-port OSS with VOA and VOPS. 80 different optical-code labels have been generated simultaneously using an OSS and multi-wavelength electro-optic frequency comb generator. 40 parallel matched-filtering has been performed by an OSS. A multi-port OSS works as large number of spectral domain matched-filtering devices. The optical CDMA application is also expressed in the presentation. The feasibility is confirmed by simulation.

References

Transmission Characteristics of loss and fiber nonlinearity tolerance on Coherent Time-Spreading OCDM with Optical Time-Gatings

Naoki Minato, Satoko Kutsuzawa, Shuko Kobayashi, and Kensuke Sasaki

Corporate Research & Development Center, Oki Electric Industry Co., Ltd., 550-5, Higashiasakawa-cho, Hachioji-shi, Tokyo 193-8550, Japan (e-mail : minato242@oki.com)

Abstract
Transmission characteristics of loss budget and nonlinear fiber effect are evaluated in OCDM with optical time-gating in the receiver, and the results show that OCDM has a potential to extend service area in PON systems.

1. Introduction
Passive optical network (PON) has grown into commercial service with over gigabit total bandwidth. Recently, it has been required to extend area of PON and guarantee high quality of service. Optical code division multiplexing (OCDM) has achieved multiplexing users comparable to conventional PON by adopting coherent scheme [1, 2] and expected a candidate as a technique for the next generation PON. For this requirement, one of the key issues to limit service area in a PON system is transmission characteristic of the OCDM signal. To extend the number of users, optical splitting loss increases in the transmission line. Also, nonlinearity of transmission fiber is another key parameter to extend the transmission distance between an office and a user, since transmitting signal composes of much shorter optical pulses than the inverse of transmission rate. In this paper, we will report experimental and simulation results of such transmission characteristics, and show the tolerance for transmission and splitting loss.

2. Experimental results on receiver sensitivity
Figure 1 shows an experimental setup to evaluate receiver sensitivity. In the sending end, an optical RZ-formatted signal was generated with the pulse width of 13ps and the wavelength of 1549.32nm at the bit rate of 622Mbs, using an optical pulse generator and an electro-absorption modulator (EAM). The signal was encoded with coherent time-spreading scheme with the code length of 128 and the spreading-time of 800ps by a fiber-Bragg-grating (FBG)-based encoder [3]. In the receiving end, signal at point A in Fig. 1, whose power was adjusted to P1 by a variable optical attenuator (VOA), was amplified by an erbium-doped fiber amplifier (EDFA) and decoded according to the conjugated code as encoder’s one. The decoded signal went through an optical time-gating

with the gate-width of 50ps at the repetition rate of 622MHz, which eliminates interferers in the case of multiplexing. The time-gated signal, whose power was adjusted to P2, was amplified into the optimum power within a receiving range of a 3R-receiver. The 3R-receiver converted the optical signal to electrical signal to recover the original information signal and clock signal.

Figure 2 shows experimental results of measured bit error rate (BER) with the parameter of the received signal power (P2). The horizontal axis is optical signal power of P2 and the vertical axis is the measured BER. Solid and dashed lines are the results in the case of using the optical time-gating and no optical time-gatings, respectively. From this figure, the minimum value of P2, which is satisfied with power penalty of less than 1dB at the BER of 10−3, is estimated at −37dBm in the case of using the optical time-gating. This value is 3dB lower than minimum power estimated for the case of using no optical time-gatings. This difference is considered due to ASE noise reduction through the optical time-gating. Therefore, received power tolerance is expected to increase still more by optimizing extinction ratio and gate width of the optical time-gating.

![Figure 1 Experimental Setup (PPG; pulse pattern generator, EAM; electro-absorption modulator, EDFA; erbium-doped fiber amplifier, VOA; variable optical attenuator, FBG; fiber Bragg grating, ERD; error detector)](image-url)
3. Simulation results of nonlinear fiber transmission

Figure 3 shows a simulation model to evaluate transmission characteristics through nonlinear fiber. An optical RZ-formatted signal with the pulse width of $\Delta t$ at the repetition rate of 625 Mb/s was encoded by multiplying reflectivity of an FBG-based encoder in the wavelength domain. The optical signal-to-noise ratio of the encoded signal set to 30 dB. Multiple access interference (MAI), which was composed of sum of interferers, was added to the encoded signal. The interferers had the same power but a different code per channel from the encoded signal. The multiplexed signal was propagated through a standard single mode fiber (SMF) of 30 km with the dispersion parameter of 17 ps/nm/km and the nonlinear parameter of 1.3 W$^{-1}$/km, and a dispersion compensating fiber (DCF) of 10 km with the dispersion parameter of 51 ps/nm/km and the nonlinear parameter of 5.2 W$^{-1}$/km, so that accumulated dispersion of the entire fiber was zero. The propagation was calculated by split-step Fourier method from nonlinear Schroedinger equation [4]. The propagated signal was decoded and time-gated. The gate-width was optimized from the viewpoint of output signal power and OSNR, and was the same as chip pulse width. Nonlinear fiber transmission penalty was evaluated from calculated Q-parameter of the received signal at a receiver with 468.75 MHz bandwidth.

Figure 4 shows the simulation results of the calculated power penalty for the case of back-to-back transmission. In the case of multiplexing, power penalty increased drastically at 10 dBm lower transmitting power than single channel case. But, the upper transmitting power was not largely decreased when increasing the number of multiplexing. In the case of 8-channel multiplexing, transmitting power to obtain power penalty of less than 1 dB was ~4 to +3 dBm in the range of signal pulse width $\Delta t$ of 6 to 22 ps. From these results and Sec. 2, this OCDM-PON system tolerates transmission and splitting loss of 33 to 40 dB in the case of 8-channel simultaneously accessed users under the condition of 30 km-long transmission.
Security performance of optical multicoding transmission using a single multiport encoder/decoder

Gabriella Cincotti¹, Gianluca Manzacca¹, Valentina Sacchieri¹, Naoya Wada², Xu Wang² and Ken-ichi Kitayama³

¹: Department of Applied Electronics University Roma Tre, via della Vasca Navale 84, I-00146 Rome, Italy, phone: ++39 0655177399, fax: ++39 0655177026, email: g.cincotti@uniroma3.it
²: National Institute of Information and Communications Technology, 4-2-1, Koganei, Tokyo 184-8795, Japan.
³: Osaka University, 2-1 Yamadaoka, Suita, Osaka 565-0871, Japan.

Abstract
We investigate the confidentiality of an OCDMA P2P transmission, using a single-chip multi-port optical encoder/decoder, as a function of the number of device ports. We consider symmetric bit and block cryptosystems, and investigate the security against brute-force searching attacks.

Introduction
Triple-play services drive an increasing demand for broadband, high-speed access networks, and passive optical networks (PONs) are excellent solutions to meet future demand. A PON has a point to multi point (P2MP) topology, between an optical line terminal (OLT) located at the central office (CO), and N optical network units (ONUs) that serve single or multiple residential or business users. Downstream is broadcast and select (B&S), on a single wavelength using time division multiplexing (TDM); destination and source addresses are usually in plain and Ethernet-based traffic frames do not always include encryption of payload. Upstream traffic is scheduled in time division multiple access (TDMA) that does not guarantee data confidentiality. Optical code division multiple access (OCDMA) is receiving increasing attention thanks to its potential to enhance data confidentiality, spectral efficiency and flexibility in asynchronous multiple access networks. In an OCDMA transmission, the mark bits from each user are encoded by a different code; encoded signals can be overlapped both in frequency and time domains and are recognized only by a matched decoder.

Although multiple access interference (MAI) hides transmitted data from each individual user, an accurate analysis of OCDMA network security should consider a point-to-point (P2P) transmission, because an unauthorized user could be able to tap an isolated users signal (see Fig. 1). Secure spectral phased encoding OCDMA have been investigated in literature [1, 2], according to the Kerckhoff’s principle, that assumes that all the details of the transmission (bit rate, wavelength, etc.) are of public knowledge, and the code is the only secret key.

We investigate the confidentiality of a P2P OCDMA transmission using a single chip multiport encoder/decoder (E/D), in an arrayed waveguide grating (AWG) configuration [3]. The device has N input/output ports and it is able to generate/process N phase shifted keyed (PSK) codes simultaneously. The code cardinality N is not large enough to guarantee a secure transmission, and for this reason, we consider a n-dimensional (n-D) configuration, where n coherent laser pulses are sent to n different input ports. A n-D code is recognized
by the same device, detecting \( n \) autocorrelation peaks (ACPs). On-off keying (OOK) OCDMA is a non-secure transmission, because it can be broken by power detection; for this reason, we consider 2- or multiple-code keying systems.

**Bit cipher cryptographic systems**

The key space of \( n \)-D codes is very large, and a device with \( N=100 \) ports can generate/process more than \( 10^{29} \) different codes; but an eavesdropper that possesses a matched decoder can easily break the code. For this reason, we add a new degree of freedom, inserting phase shifters at the encoder input ports, as illustrated in Fig. 2a; this coding scheme corresponds to a spread-spectrum OCDMA, and the private key is the pseudo-random binary phase code. Both mark and space are encoded, so that balanced detection is also possible, to enhance the OCDMA transmission performance. The system security is shown in Fig.3: we consider a brute-force code searching attack and evaluate the number of years required to break a code, assuming that an adversary is able to test \( 10^7 \) codes per second.

**Block cipher cryptographic systems**

In this scheme, a block of \( n \) bits from each user is encoded by a different codeword, sending \( n \) laser input pulses to different inputs of a multi-port E/D. Figure 2b shows the OLT and ONU architectures: the message from each user is mapped into a \( n \)-bit ciphertext, using an electronic cryptographic system, that drives a switch to generate an optical \( n \)-D code; \( n \) ACPs detected at outputs of the multi-port E/D at ONU can be seen as \( n \) marks of the ciphertext than can decrypted into the original message. The electronic encrypting and decrypting systems can have an internal state: in that case they are named stream ciphers. The key space potentially comprises all the permutations over \( n \) bits that are \( 2^{n!} \); for instance, if \( n=10 \), all the possible keys are \( 5.42\times10^{269} \). Symmetric cryptosystems use only some of all the possible permutations, combined with substitution to provide confusion (to make the relation between key and ciphertext complex) and diffusion (to expand the influence of a single plaintext bit over many ciphertext bits).

Data encryption standard (DES) developed by IBM in 1970 encrypts blocks of \( n=64 \) bits, with 16 multiple iteration, using a key space of \( 2^{56} \) elements. Unfortunately in 1998 Deep Crack was able to break the system in 4.5 days, and cryptosystems with longer keys have been developed to increase the system confidentiality. According to Shannon’s theorem, a perfectly secure encryption system requires that entropy of the key space is larger than that of the message space. The system of Fig. 2b corresponds to an \( n \)-ary transmission, where \( n \) bits from each user are transmitted simultaneously.

**Conclusions**

Planar architectures to encode data messages from a multiple access user have been presented, and the P2P transmission confidentiality has been investigated. We presented both bit and block cipher cryptosystems, and compare their performance against brute-force code searching attacks. To increase data confidentiality, it would be necessary to use time-varying encoding keys; using the Vernam’s one-time pad scheme, a perfectly secure system can be obtained when the private key is changed every bit or every message block.

**References**

Ultrafast Transmission Technology

Reinhold Ludwig,
FhG Heinrich-Hertz-Institute, Einsteinufer 37, 10587 Berlin, Germany

The expected increase of transmission capacity in optical fiber networks presents us with many technology challenges. One of these challenges is an optimized combination of wavelength division multiplexing (WDM) and time division-multiplexing (TDM). TDM may be realized by electrical multiplexing (ETDM) or by optical multiplexing (OTDM) to a high speed data signal, i.e. by electrical signal processing or optical signal processing. Presently, the first 40 Gb/s systems based on ETDM have been installed and in laboratories the first 100 Gb/s ETDM-experiments have been performed. On the other hand at the same data rates, OTDM transmission experiments have been carried out more than 10 years earlier. For instance, the first 100 Gb/s OTDM transmission experiment over a 36 km fiber link was reported by NTT already in 1993. Since then, OTDM transmission technology has been made a lot of progress towards much higher bit rates and much longer transmission links. Recently, OTDM transmission technology succeeded in the transmission of a TDM data rate of 160 Gb/s over a record fiber length of 4320 km and of a TDM data rate of 2.46 Tb/s over a fiber link length of 160 km.

The past has seen that OTDM will be replaced by ETDM as soon as electrical signal processing becomes available at the required TDM data rate. Therefore, OTDM transmission technology is often considered to be an interim technique with which to investigate the feasibility of ultra-high speed data transmission in fibers. Today, the ultimate limits of ETDM technology are not known. An ETDM data rate of 160 Gb/s is very likely in the future. However, the performance of the ETDM terminal equipment may eventually be worse than the OTDM terminal equipment. OTDM demultiplexers perform better than ETDM receivers already at data rates of 80 Gb/s. State of the art OTDM terminal equipment for 160 Gb/s TDM data transmission provides already very stable operation conditions. The performance suggest that 160 Gb/s OTDM transmission systems can be used in deployed systems and can be operated error-free for years. ETDM technology must work hard to compete with these systems for instance as regards receiver sensitivity. Moreover, it is not sure that ETDM terminal equipment for 160 Gb/s will be less expensive and less energy consuming than the corresponding OTDM terminal equipment.

The ultimate limits of OTDM transmission technology are not given by the terminal equipment but by the transmission properties of the fiber link including all repeater or amplifier stages. With higher TDM bit rate data transmission in fiber is stronger affected by chromatic dispersion (CD), polarization-mode dispersion (PMD), fiber non-linearity, and the limited bandwidth of repeaters or amplifiers in the transmission link. This is independent of the signal processing in the terminal equipment whether it is based on ETDM or OTDM technology. At present therefore, a main task of OTDM technology is to explore the ultimate capacity for fiber transmission in a single wavelength channel. The most challenging view as regards OTDM technology is that optical networks will evolve into "photonic networks", in which ultra-fast optical signals of any bit rate and modulation format will be transmitted and processed from end to end without optical-electrical-optical (O/E/O) conversion. This "photonic network" is a target for the distant future, and it presents us with the present challenge of investigating and developing high-speed optical signal processing and exploring the ultimate capacity for fiber transmission in a single wavelength channel.

In this tutorial, we review the components, techniques and characteristics of recent ultrahigh speed transmission systems.
Reinhold Ludwig was born in 1952 in Lahnstein, GERMANY. He received the Ing.grad. degree from the Fachhochschule Kobenz in 1974 and the Dipl.-Ing. and Dr.-Ing. degrees from the Technical University Berlin in 1985 and 1993.

In 1985 he joined the Heinrich Hertz Institute (HHI) Berlin, where he is involved in research on photonic components and systems.

He worked as a visiting scientist at Nippon Telephone & Telegraph Co.(NTT),Japan in 1991 and at Bell Labs, USA in 1993. Since 1985 he authored and coauthored more than 300 scientific papers and holds several patents.

In 1996 he founded the first HHI spin-off company, LKF Advanced Optics GmbH, and served as CEO until the merger of LKF and u2t Innovative Optoelectronic Components GmbH to the u2t Photonics AG in 2001.

In 1999 his group received the Philip-Morris-Science Award and he was nominated for the Innovation Award of the German Bundespraesident.

Dr. Ludwig is a member of the VDE.
Abstract: Broadband millimeter-wave wireless access with optical backbone is an ideal candidate for the provision of future broadband services. We provide an overview of our research activities in this area.

Introduction

Fixed wireless access networks operating at millimeter wave (mm-wave) frequencies with an optical fiber backbone offer the capacity to deliver future broadband services. Such networks require a pico-cellular architecture which in turn requires the deployment of a large number of antenna base stations (BSs). With a drastic increase in the total number of BSs, it is essential to simplify them and reduce their cost. This can be achieved by migrating all of the signal processing, routing and switching functionalities to the optical headend or central office (CO) where they can be shared among all the BSs. Fig. 1 shows a typical hybrid radio-on-fiber (RoF) network layout. The CO of the RoF network acts as a gateway to the optical wavelength-division-multiplexed (WDM) backbone while serving a large number of widely distributed antenna BSs. It has been shown that transporting the radio signals at the designated mm-wave frequency over fiber (RF-over-fiber transport) has the potential to reduce the BS complexity by enabling a centralized control architecture.

Although an RF-over-fiber transport scheme potentially simplifies the antenna BS design, there are optical impairments associated with this approach that need to be addressed. Fig. 2 illustrates a point-to-point link connecting a CO to a BS that incorporates RF-over-fiber transport and summarizes all the optical impairments that exist. The conversion of the radio signals to optical signals and vice versa in both the CO and BS is a non-linear process that introduces intermodulation distortion (IMD) when multiple carriers are present. In addition to the nonlinear optical interface, the transport of the radio signals at mm-wave frequencies significantly compromises the overall link budget and performance. The transport of mm-wave modulated optical signals also suffers from severe fiber chromatic dispersion penalties [1] and inefficient use of optical bandwidth.

In this paper, we review and discuss some of our research activities focused on addressing these optical impairments. We have previously proposed and demonstrated techniques to improve optical spectral efficiency and link budget, as well as a linearization technique to reduce third-order IMD for RF-over-fiber transport.

Wavelength-Interleaving Technique

It has been well-established that optical single sideband with carrier (OSSB+C) modulation overcomes the penalty associated with fiber dispersion.
for the distribution of mm-wave modulated optical signals [2]. It also improves the optical spectral usage by 50%. To further improve the spectral usage, a wavelength-interleaving technique was introduced [3,4] where it was shown that a potential threefold improvement in optical bandwidth usage using OSSB+C could be achieved [5]. This technique enables multiple mm-wave radio signals to be multiplexed in such a way that the optical channel spacing between adjacent channels is less than the mm-wave radio frequency as shown in the schematic in Fig. 3. We have successfully demonstrated this technique and have also implemented the essential optical interfaces for the interleaved channels (multiplexer and demultiplexer) using arrayed-waveguide gratings, optical circulators and fiber Bragg gratings (FBGs) [6].

Improvement in Link Performance

The OSSB+C signal is generated using a dual-electrode Mach-Zehnder modulator (DE-MZM), however due to the poor modulation efficiency at these frequencies the optical modulation depth (OMD) is typically very low. As a result, the modulated mm-wave optical sideband power can be more than 20 dB below that of the optical carrier. To improve the link performance, the optical power of the signals can be increased by using a high power optical source or an optical amplifier, however this may lead to increased IMD at the receiver (or even damage the receiver) due to too large an optical power incident on the detector.

We have presented experimental, analytical and numerical studies on the impact of optical modulation efficiency of the mm-wave signals on the overall fiber-radio link performance [7]. The OMD of the mm-wave modulated signals was varied by removing a portion (via a FBG) of the optical carrier from the OSSB+C signal before detection. Shown in Fig 4a is a measured optical spectrum of an OSSB+C signal with and without an 80% reflective FBG that reduces the optical carrier by ~7.5 dB. The corresponding BER curves are shown in Fig. 4b with a 3 dB improvement in the sensitivity at a BER = 10⁻⁶ when the optical carrier was decreased by 7.5 dB (with 80% reflectivity FBG). We have also shown that the sensitivity performance of the RoF link is dependent on the optical carrier-to-sideband ratio (CSR) of the OSSB+C with an optimal performance occurring at CSR of 0 dB [7].

Linearization Technique

In a multi-carrier environment, system linearity is vital to maintaining the required dynamic range. It has been shown that the nonlinearity of the optical frontend of a RoF link limits the overall system dynamic range [8]. Recently we have proposed a linearization technique for dispersion-tolerant OSSB+C signals to reduce the third-order IMD contributions. We have quantified the IMD generation with respect to the optical components generated due to the nonlinear characteristics of the OSSB+C modulator and have shown that the major contributor to the overall IMD are the optical components located in the vicinity of the optical carrier, i.e. components at ω₁−ω₂ and ω₁+ω₂ if the OSSB+C were driven by two RF tones (f₁ and f₂) [9]. Our proposed linearization technique was based on the removal of these components. We have experimentally demonstrated the technique and have shown an improvement of 9 dB in the resulting carrier-to-interference ratio [9].

Conclusions

We have presented a brief overview of some of our research work in the area of RoF. The main focus of these research activities has been to overcome the impairments for transporting mm-wave modulated optical signals. We have introduced wavelength-interleaving the technique to improve optical spectral usage, quantified RoF link performance and demonstrated a linearization technique to combat these impairments.

Acknowledgement

This work was supported by the Australian Research Council Discovery Grant DP0452223.

References


Fig. 4 (a) Measured optical spectra and (b) measured BER for OSSB+C with and without FBG
20-GHz 16-QAM Radio-over-Fiber Data Transmission Using a Directly Modulated Laser

Hoon Kim and Ho Chul Ji*

Samsung Electronics, Telecommunication R&D Center, Suwon, Korea 444-742
Phone) +82-31-279-5612, Fax) +82-31-279-5255, Email) hoonkim@ieee.org

*Photonic Network Laboratory, Osaka University, Osaka 565-0871, Japan

Abstract: We demonstrate the transmission of 200-Mb/s data coded with 16-QAM at 20 GHz over 25-km conventional single-mode fiber using a directly modulated laser. To avoid the signal fading by fiber dispersion, we generate single-side band signals with an optical filter.

I. Introduction

Radio and millimeter-wave over fiber systems have gained a lot of attention for video or wireless signal transmission since they transmit the signals without format conversion, and thus minimizing the size and cost of remote base stations (RBSs). In these systems, RBSs can be implemented simply with optical-to-electrical converters, electrical-to-optical converters, RF front-ends, and RF amplifiers. External modulators have been commonly employed for transmitters since the modulators can generate chirp-free signals and provide wider bandwidth than directly modulated lasers [1]. However, the use of external modulators significantly increases the cost of RBSs. In addition, high insertion loss of the modulator compels the system to employ the optical amplifiers, which further increases the systems cost.

Hence, we demonstrate the transmission of 50-Msymbol/s 16-QAM data at 20 GHz using a large current-biased directly modulated laser. Since the directly modulated signals generate double sideband (DSB) at the output, we utilize an optical filter to filter out one of the sidebands and make the signals immune to fiber dispersion. Our cost-effective system exhibits good error vector magnitude (EVM) performance after 25-km transmission over conventional single-mode fiber (SMF).

II. Experimental Setup and Results

Fig. 1 Experimental setup

Fig. 2. E/O frequency response of the DFB-LD measured with a network analyzer
We first measured the RF power degradation due to fiber dispersion as a function of frequency. The DSB signals have been found to be susceptible to fiber dispersion since each sideband experiences different phase shift as fiber dispersion or modulation frequency increases [2]. This induces RF fading at the receiver but it can be overcome by using SSB transmission. Fig. 3 shows the measured frequency response of the 25-km link for DSB and SSB modulation. The DSB signals suffer from RF fading of >10 dB at modulation frequencies of 18.2 and 25 GHz. Taking into account the RF amplitude increase by large chirp of the directly modulated DFB-LD (at around 13, 22, and 28 GHz), the RF power variation is measured to be as large as 30 dB [3]. However, the RF power variation of the SSB signals is less than 3 dB throughout the frequency.

Fig. 3. Measured RF power degradation due to dispersion.

Next, we measured the signal constellation together with EVM. The EVM, defined as the root-mean-square (RMS) error vector magnitude referred to the peak symbol magnitude, is related to the symbol-error ratio through the average signal-to-noise ratio per symbol. Fig. 4 shows the EVM performance versus the RF input power into the LD when the signals are carried at 20 GHz. The results show typical bathtub curves: when the signal power is low, noise is a major factor limiting performance whereas it is nonlinearity when the signal power is high. In our experiment, clipping distortion starts to limit the system performance as the signal input power is higher than 0 dBm, which corresponds to the RMS optical modulation index (OMI) of 23%. Throughout the RF input power range, DSB outperforms SSB due to the higher RF power at the output of the receiver (see Fig. 3 for RF power difference between DSB and SSB at 20 GHz). Fig. 4 also shows the signal constellations of the SSB signal when the RF input powers are -10, 0, and 10 dBm.

We next changed the carrier frequency of the signals to 18.2 GHz where the DSB signals experience severe RF fading. The EVM results are depicted in Fig. 5. Due to the large signal fading of DSB signals, we are able to measure the EVM performance of DSB signals only at the RF input power of 6 dBm. It is measured to be 11%. However, the SSB signals exhibit similar performance to the case with 20-GHz carrier frequency. At the RMS OMI of 23%, we achieve the best EVM performance of ~4% with the SSB signals.

![Figure 3](image.png)

**Fig. 3.** Measured RF power degradation due to dispersion.

![Figure 4](image.png)

**Fig. 4.** Measured EVM performance of 20-GHz RoF link.

![Figure 5](image.png)

**Fig. 5.** Measured EVM performance of 18.2-GHz RoF link

### III. Conclusion

We have demonstrated the 20-GHz radio-over-fiber transmission of 16-QAM signals over 25 km of SMF using a directly modulated laser. Thanks to the high bias current of the laser diode and optical sideband filtering, our cost-effective system exhibits good EVM performance after transmission.

### References

UWB monocycle pulse generation by optical polarization time delay method

Minghua Chen, Hongwei Chen, Jian Zhang and Shizhong Xie
Department of Electronic Engineering, Tsinghua University, Beijing, 100084 P.R China.
Tel: 8610-6278-4784; Fax: 8610-6278-8161; Email: chenmh@tsinghua.edu.cn

Abstract
A novel method to generate ultrawideband (UWB) monocycle pulse based on birefringence time delay in polarization-maintaining fiber (PMF) is proposed and experimentally demonstrated. Two polarity-reverse pulses are generated by polarization modulation using a single phase modulator.

1 Introduction
Ultrawideband (UWB) is emerging as a solution for future wideband personal access network (PAN) recently [1]. It has more advantages than traditional wireless communication technologies, such as low power consumption, high bit rate, immunity to multipath fading [2]. In UWB system, one of the most used modulation techniques is impulse radio (IR) direct-sequence code-division multiplexing. And the carrier-free impulse modulation attracts more attention which not only does not need the complicated frequency mixer, intermediate frequency, and filter circuits, but also has good pass-through performance due to the base-band transmission. Gaussian monocycle pulse has better bit error rate and multipath performance and wider bandwidth than the other impulse signals [3]. With the radio-over-fiber technology improvement, UWB over fiber can be a candidate solution for future wideband access networks [4]. So there are many schemes to be carried out to optically generate and distribute the monocycle pulses. In ref [5] and [6], optical pulses are transmitted by fiber and monocycle pulses are obtained by microwave differentiator in electrical domain. Recently, Yao’s group gives some attractive methods to optically generate monocycle pulses. One is to generate monocycle pulse signal by cross-phase modulation in nonlinear fiber and an optical frequency discriminator [7]. The other utilized cross-gain modulation in semiconductor optical amplifier (SOA) and a pair of fiber Bragg gratings (FBG) to generate polarity-reversed optical pulses and combined them together [8]. However, these two approaches both use two laser sources which may increase the system complexity.

2 Principle and Experiment

The principle of proposed scheme is shown in Fig. 1. Electrical pulse is used to drive an optical phase modulator (PM) and keep the peak level to $V_p$ of PM. For a linearly-polarized light launching at 45° relative to a principal axis of PM, the phase shift between the two principal axes will depend on the driving voltage. If the driving voltage is set to $V_p$, the signal polarization will rotate to the orthogonal direction at the output of PM [11]. So at the input orientation, there is a negative pulse. While at the orthogonal orientation, there is a positive pulse. Then the optical signal is sent into a differential group delay (DGD) component with two polarization orientation along the principle axes. Thanks to the birefringence time delay, the two inverse polarity pulses can be delayed properly. After O/E converter, monocycle pulse can be obtained by this method.
The experiment setup is shown in Fig.2. Optical signal from laser is tuned by polarization controller (PC) and then fed into an optical phase modulator driven by electrical pulse generator. Electrical pulse is generated by a BER tester (BERT, Advantest D3186). The electrical signal is fixed pattern “1000 0000 0000 0000” (one “1” every 16 bits) with bit rate of 8.2GHz, so the repetition rate of pulse pattern is about 512MHz. The electrical pulse FWHM is about 122ps which is shown in Fig. 3. The output orientation of phase modulator is adjusted by a PC and fed into a section of polarization-maintaining fiber with length of 90m. The beat length of PMF is 3.8mm and the time delay of two principle state of polarization (PSP) is about 122ps. Then the optical signal is amplified to the O/E converter and measured by a digital sampling oscillator (DSO, Tektronix TDS8200).

Figure 4 shows the generated monocycle pulse. The pulse width is almost the same as the input electrical pulse. And there is no power fluctuation at the DC level compared with result of ref [9]. For the input optical signal orientation can be easily tuned, the two polarity-reverse pulses can pass along different principle axes of the PMF. So it is very easy to get an inverted monocycle pulse as shown in Fig. 5 which can be used in biphase modulation format for UWB systems.

The spectrum of generated monocycle pulse is measured by electrical spectrum analyzer (ESA, Agilent E4446A) and shown in Fig.6. The repetition rate of electrical pulse sequence from BERT is about 512MHz which is equal to the interval of discrete frequency parts in spectrum. And the central frequency is about 4.1GHz and -10dB bandwidth is nearly 6.15GHz which corresponds to the fractional bandwidth of 150% due to the definition of UWB. The discrete spectrum envelope is just like the spectrum of a single monocycle pulse.

3 References

[7]. F. Zeng, Q. Wang and J. Yao, IEEE International Topical Meeting on Microwave Photonics, MWP’06, P 13.
Multi-channel Ultrawideband Monocycle Pulse Generation via Cross Phase Modulation and Spectral Filtering

Bill P. P. Kuo, P. C. Chui and Kenneth K. Y. Wong
(Photonics Systems Research Laboratory, Department of Electrical and Electronic Engineering, The University of Hong Kong, Pokfulam Road, Hong Kong. Tel: (852)2859-2698; Email: ppkuo@eee.hku.hk)

Abstract
We demonstrate a novel technique for multi-channel ultrawideband monocycle pulse generation by using cross phase modulation and spectral filtering. High quality monocycle pulse on three WDM channels has been successfully generated.

1. Introduction
Ultrawideband (UWB) technology, which has been demonstrated to be able to provide high speed data link with low power consumption [1], has received great interest as a wireless replacement for short distance wired connections. Traditionally, the RF sources for UWB devices, which are in form of monocycle or doublet pulse train, are generated and transmitted in electrical domain. Until recently, UWB-over-fiber technology has been introduced to extend the reach of UWB radio service by taking advantage of low loss transmission over optical fiber. Therefore UWB pulse generation in optical domain would be an attractive alternative over electrical means for UWB-over-fiber service. Previously single channel optical UWB pulse generation has been demonstrated using specialized modulator [2], cross gain modulation (XGM) in optical parametric amplifier (OPA) [3], and phase-modulation to intensity modulation conversion through dispersion in single mode fiber and fiber Bragg grating [4-5]. However, simultaneous multiple channel UWB pulse generation has never been realized to the best of our knowledge. In this paper, multiple channel UWB impulse generation by using cross phase modulation (XPM) in optical fiber and spectral filtering in arrayed waveguide grating (AWG) is demonstrated.

2. Principle
The principle of multi-channel monocycle pulse generation is shown in figure 1. When a pulsed pump propagates with continuous-wave signals in a nonlinear medium with third order susceptibility (e.g. highly nonlinear fiber), the signal waves will be phase modulated according to the instantaneous intensity of pump due to XPM effect. Such phase modulation then causes the frequency of the signal waves to down-shift on the rising edge and up-shift on the falling edge of pump pulse. Mathematically, when dispersion and fiber loss are ignored, the frequency shift can be expressed as:

$$\delta f = -\gamma L \frac{d}{dt} \left( P_p(t) \right)$$

where $\delta f$ is the frequency shift of the signal waves due to XPM, $\gamma$ and $L$ are the nonlinear coefficient and length of the medium, and $P_p$ is the power of pump pulse. Thus if the pulse shape of the pump follows a Gaussian function, the frequency variation of the signals will follow the shape of a Gaussian monocycle pulse, which is the first order derivative of a Gaussian function.

---

![Fig. 1. Schematic diagram showing the principle of monocycle pulse generation.](image-url)
3. Experimental Setup

The experimental setup for multi-channel monocycle pulse generation is shown in figure 2. The nonlinear medium used for XPM was a spool of 1km highly nonlinear dispersion shifted fiber (HNLSDF) with nonlinear coefficient $\gamma \approx 14\text{W}^{-1}\text{km}^{-1}$ and zero-dispersion wavelength $\lambda_0 \approx 1560\text{nm}$. A DFB laser (DFB1) with emission wavelength $\lambda_{\text{pump}}$ at 1557.1nm was used as pump source. The pump wave was intensity modulated with 3GHz electrical pulse train at 25% duty cycle through an intensity modulator MZM. The pump was then amplified to 17dBm by two stages of EDFAs (EDFA1 and EDFA2) with a tunable band-pass filter TBPF1 inserted in between to reduce ASE noise level at EDFA2 input. On the other hand, three tunable laser sources TLS1-3 with wavelengths $\lambda_{\text{ch}1}$, $\lambda_{\text{ch}2}$ and $\lambda_{\text{ch}3}$ at 1569.4nm, 1571.0nm and 1572.6nm respectively were served as signal sources and subsequently boosted to total power of 14dBm by EDFA3. The pump and signals were then combined using a WDM band combiner WDMC1 and launched into the HNL-DSF. After propagating through the HNL-DSF, the output waves were attenuated by a variable optical attenuator (VOA) and passed through WDMC2 to decouple signal waves from pump pulse. The L-band signal waves were launched into an AWG with 0.8nm channel spacing for spectral filtering of signal waves. The pass-bands center wavelengths $\lambda_{\text{ch}1}$, $\lambda_{\text{ch}2}$, $\lambda_{\text{ch}3}$ and $\lambda_{\text{ch}4}$ for output ports 1-4 were adjusted according to the allocation scheme shown in figure 3 so that only one channel was present at each output port. The polarization controllers PC2-4 were adjusted to obtain best pulse quality. The signals at AWG output ports were monitored by a photodetector (PD) and a digital communication analyzer (DCA).

4. Results and Discussions

The waveforms of monocycle pulse generated in three different channels at output ports 1, 3 and 4 are shown in figure 4. As seen from the diagrams, the waveforms resembled an ideal Gaussian monocycle pulse well, except that slight overshoot was observed at the trailing edge of the pulse. This was due to non-ideal electrical response of the modulator and driving electronics which gave rise to the overshoot at the trailing edge of pump pulse. Besides, the polarity of pulse generated in channel 3 was inversed with respect to channel 1 and 2 as the signal wavelength of channel 3 was on the red edge of pass-band of output port 4, which the frequency-loss relationship was reversed in this case. This also shows the possibility of pulse polarity modulation (PPM) by placing signal at the blue edge and red edge arbitrarily or switching between adjacent output ports when the signal is in between the pass-bands of the ports. Moreover, although in this work simultaneous pulse generation was demonstrated only on three channels, the maximum number of channel that the proposed method could support was only limited by the number of output port of the AWG used, and so could be increased by using AWGs with more output ports.

5. Conclusion

We have demonstrated a multi-channel monocycle pulse generation technique based on XPM and spectral filtering. This technique would be helpful in providing cost-effective solution for generation and distribution of UWB signals to multiple access points.

6. Acknowledgement

The work described in this paper was partially support by a grant from the Research Grants Council of the Hong Kong Special Administrative Region, China (Project No. HKU 7179/06E). The authors would also like to acknowledge Sumitomo Electric Industries for providing the HNL-DSF.

7. Reference

High-quality electrical Gaussian-monocycle pulse generation by electrical-optical hybrid signal processing

Masanori Hanawa¹, Kazuhiko Nakamura¹, Koji Nonaka²

¹University of Yamanashi, 4-3-11 Takeda, Kofu, Yamanashi 400-8511, Japan
²Kochi University of Technology, 185 Tosayamada, Kami, Kochi 781-5103, Japan
¹Tel/Fax: +81-55-220-8683, E-mail: hanawa@yamanashi.ac.jp

Abstract
High-quality pulse generation for ultra-wideband impulse-radio systems is experimentally demonstrated. The generated pulses have the repetition rate, the duration and the timing jitter of 1.5013GHz, from 224ps to 443ps, and <6ps, respectively.

1 Introduction
In ultra-wideband impulse-radio (UWB-IR) systems, short electrical pulses, with the duration of several hundreds of picoseconds and without any carrier wave, are used [1]. Such short pulses have rather unique properties and can provide certain advantages, for example, high data rate for wireless communications, high-resolution for radars, and so on.

The quality of the pulses used in the UWB-IR systems is very important. So far, several pulse generation methods based on electrical signal processing techniques have been proposed [2,3]. Such electrical approaches enable the integration and miniaturization of the system; meanwhile, they have some weakness such as pulse ringing or an intrinsic error in the pulse shape due to their operating speed limitations.

Recent investigations have revealed that it is possible to generate such the UWB-IR pulses precisely using optical signal processing techniques [4]. Although the method proposed in the Ref. [4] is applicable to any pulse shaping, it is, however, a little bit complicated for the simple Gaussian-monocycle pulse (the basic pulse for the UWB-IR systems) generation. In this paper, the high-quality Gaussian-monocycle pulse generation is experimentally demonstrated using a simple experimental setup based on the electrical-optical hybrid signal processing technique using the timing-jitter-suppressed gain-switched pulse laser and the balanced photo detector.

2 UWB-IR pulse shaping method

Fig. 1 explains the basic concept of the proposed Gaussian-monocycle pulse shaping. The optical pulse train is spread out in time using an optical dispersive media like the standard single mode fiber (SMF). The pulse train is split in two with an optical 3dB splitter and one of the split signal is delayed by a tunable optical delay line. After the O/E conversion, two electrical signals are electrically subtracted, thus resulting in the desirable Gaussian-monocycle pulses. The resulting signal \( g(t) \) is given as follows:

\[
g(t) = |y(t - \Delta t)|^2 - |y(t)|^2, \quad (1)
\]

where \( y(t) \) is the output signal from the dispersive media and \( \Delta t \) is the relative time delay between two signals launched into the O/E converters. When the light source generates the clear Gaussian pulse train and the time delay \( \Delta t \) is suitable for the pulse width of the output from the dispersive media, the differentiated signal of two Gaussian pulse trains, the Gaussian-monocycle pulse train, is generated.

3 Experiment
The proof-in-principle experiments were carried out using the setup shown in Fig.2. In the experiment, the gain-switched pulse laser diode (LD) module with very low timing jitter [5] was used. This repetition-rate-flexible pulse LD module was tuned to output a quasi-Gaussian pulse train with the full-width at half-maximum and the repetition rate of about 70ps and 1.5013GHz, respectively. The quasi-Gaussian pulses were spread out using the SMFs of the length from 2.2km to 20km and subsequently split in two paths using the 3dB optical coupler. In the one path, the variable optical attenuator and the fixed optical delay line (FDL) was inserted to make the signal power in two paths even and to align the relative time delay of the signal roughly. The variable optical delay line (VDL) in...
the other path was used to give a precise relative time delay $\Delta t$ to the pulse trains.

Fig. 3 shows the Gaussian-mono-cycle pulse observed by the digital communication analyzer (DCA). In this case, the repetition rate of the source pulse, the SMF length and the relative time delay $\Delta t$ were 1.5013GHz, 20km and 200ps, respectively. The generated pulse train was highly stable and the timing jitter, defined as the standard deviation of the histogram for the narrow horizontal slice of the waveforms measured by the DCA, was only 5.95ps. As the timing jitter was measured by the histogram analysis function of the DCA, it consists of the instrument jitter; therefore, the exact value should be smaller. Anyway, it is only 1.4 percent of the pulse duration (full width at tenth maximum (FWTM)) of 443ps and it is quite small. Fig. 4 shows the comparison between the measured and ideal frequency spectra. Here, the measured spectrum was obtained using not an electrical spectrum analyzer but the fast Fourier transform function of the DCA. The two spectra have good agreement within a range of 6GHz. The small timing jitter and the good agreement of the spectra clearly show the quite high quality of the generated Gaussian-mono-cycle pulse train.

The duration of the generated Gaussian-mono-cycle pulse can be controlled by changing the SMF length and the relative time delay $\Delta t$. Fig. 5 shows the duration vs. the SMF length. Here, $\Delta t$ was set to almost a half of the FWTM of the SMF output. The timing jitter values are also shown in the same figure and those were less than 6ps for all conditions. From the results it is found that the FWTM of the Gaussian-mono-cycle pulses could be successfully adjusted from 224 to 443ps without the significant degradation of the timing jitter.

4 Conclusion
A high-quality Gaussian-mono-cycle pulse generation for UWB-IR system was experimentally demonstrated. The generated pulses have the almost ideal Gaussian-mono-cycle shape and the quite low timing jitter less than 6ps. With the proposed method, the FWTM of the Gaussian-mono-cycle pulse was successfully adjusted with a range from 224ps to 443ps. The use of the dispersion compensation fiber instead of the SMF makes the narrower duration possible. Although the repetition rate was fixed at 1.5013GHz in this report, it also can be adjusted from 1.0GHz to 2.4GHz. From these high quality and flexibility, the proposed scheme is suitable for the UWB-IR pulse simulator for further developments of the UWB-IR systems.

Acknowledgement
This work was supported by Grant-in-Aid for Scientific Research (B), No. 18360180, of Japan Society for the Promotion of Science (JSPS).

References
We discuss a digital coherent receiver, which can demodulate multi-level coded optical signals. The carrier phase is estimated with digital signal processing, and the optical complex amplitude is entirely restored without the optical phase-locked loop.

I. INTRODUCTION

Recently, with the advent of high capacity WDM transmission technologies, the spectral efficiency has become one of the main concerns of researchers. Coherent optical receivers are very attractive for increasing the spectral efficiency because any kind of multi-level modulation format can be introduced by using such receivers.

We have recently demonstrated a digital coherent receiver [1]. The in-phase and quadrature components of an optical signal are retrieved using a homodyne phase-diversity receiver without locking the phase of the local oscillator (LO). The carrier phase is recovered after homodyne detection by means of digital signal processing (DSP). While an optical phase-locked loop (PLL) that locks the LO phase to the signal phase is still difficult to achieve, DSP circuits become increasingly faster and provide us with simple and efficient means for estimating the carrier phase. In addition, electrical signal processing such as group-velocity dispersion (GVD) compensation is achievable in the homodyne receiver, making the coherent scheme more attractive. This paper describes the principle of operation of our DSP-based homodyne phase-diversity receiver and its applications.

II. PRINCIPLE OF OPERATION OF THE COHERENT RECEIVER

The optical phase-diversity homodyne receiver used in our experiments is shown in Fig.1 [1]. Orthogonal states of polarization for LO and the incoming signal create the 90° hybrid necessary for phase diversity. With the λ/4 waveplate (QWP), the polarization of the LO becomes circular, while the signal remains linearly polarized and its polarization angle is 45° with respect to the principle axis of polarization beam splitters (PBSs). After passing through the half mirror (HM), the polarization beam splitters separate the two polarization components of the LO and signal while two balanced photodiodes PD1 and PD2 detect the beat between the LO and signal in each polarization.

Let the complex amplitude of the signal be represented by

\[ E(t) = E_s(t) \exp \left[ j(\theta_s(t) + \theta_n(t)) \right], \]

where \( E_s(t) \) is the signal amplitude, \( \theta_s(t) \) the phase modulation and \( \theta_n(t) \) the carrier phase in reference to the LO phase. Then, the currents from PD1 and PD2 are expressed as

\[ I_{PD1}(t) = RE_{LO}E_s(t)\cos[\theta_s(t) + \theta_n(t)], \]
\[ I_{PD2}(t) = RE_{LO}E_s(t)\sin[\theta_s(t) + \theta_n(t)], \]

where \( R \) is the responsivity of the photodiodes and \( E_{LO} \) the amplitude of LO. The electrical signals \( I_{PD1} \) and \( I_{PD2} \) contain information on the in-phase and quadrature components of the complex amplitude of the optical signal.

The signals \( I_{PD1}(t) \) and \( I_{PD2}(t) \) are simultaneously sampled once every symbol period \( T \) with analog-to-digital converters (ADCs). We can thus reconstruct the signal complex amplitude, ignoring unimportant constants, as

\[ E_s(t) = I_{PD1}(t) + jI_{PD2}(t), \]

where \( T \) denotes the sampling time interval, and \( i \) the number of samples.

Since the linewidth of semiconductor DFB lasers \( \Delta f \) used as the transmitter and LO typically ranges from 100 kHz to 10 MHz, the optical carrier phase \( \theta_s(t) \) varies much more slowly than the phase modulation, whose symbol rate is 10 Gsymbol/s in our experiments. Therefore, by averaging the carrier phase over many symbol intervals, it is possible to obtain an accurate phase estimate. In the case of \( M \)-ary PSK signals, we take the \( M \)-th power of \( E_s(t) \) because the phase modulation is removed from \( E_s(t) \). Averaging \( E_s(t)^M \) over 2\( k + 1 \) samples constitutes a phase estimate as

\[ \theta_s(t) = \arg \left( \sum_{j=1}^M E_j((i+j)T)^M \right) / M. \]

The phase modulation \( \theta_s(t) \) is determined by subtracting \( \theta_s(t) \) from the measured phase of \( \theta_s(t) \).

![Fig.1 Construction of the homodyne phase-diversity receiver.](image-url)
III. BIT-ERROR RATE MEASUREMENTS OF M-ARY PSK SIGNALS

The back-to-back BER of the BPSK, QPSK, and 8-PSK signals was measured to access the receiver performance [2]. The DFB laser output was modulated through LiNbO$_3$ phase modulators to generate PSK signals at a symbol rate of 10 Gsymbol/s. The received signal was amplified with an erbium-doped fiber amplifier (EDFA) to $-10$ dBm before it was detected with the coherent receiver. The linewidth of the transmitter and the LO was about 150 kHz and frequency drifts of the lasers were kept below 10 MHz. The signals $I_{PD1}$ and $I_{PD2}$ were simultaneously sampled at a rate of 20 Gsample/s with analog-to-digital converters (ADCs). The collected samples were resampled to keep only one point per symbol and combined to form a 100-ksymbol stream. The signal was demodulated through the digital phase estimation process described in Sec.II, and the number of bit errors was counted through off-line measurements. Fig.2 shows BERs measured as a function of the received power. Fig.3 represents the constellation map of the 8-PSK signal obtained in the error-free state.

![Fig.2 Back-to-back BER curves for BPSK, QPSK, and 8-PSK signals.](image)

![Fig.3 Constellation map for the 8-PSK signal.](image)

IV. POST-PROCESSING FUNCTION

Coherent detection can linearly recover the amplitude and phase information of optical signals. Therefore, post-processing of the received signal allows fully electronic compensation for chromatic dispersion, which is a linear transfer function operating on the optical complex amplitude. In this section, we demonstrate unrepeated transmission of 20-Gbit/s optical QPSK signals over a 200-km standard single-mode fiber (SMF), where after homodyne phase-diversity detection, digital signal processing is employed for carrier phase estimation as well as dispersion compensation [3]. A total dispersion of up to 4,000 ps/nm is compensated effectively through a simple transversal digital filter implemented in our coherent receiver.

Fig. 4 shows measured BER of the 20-Gbit/s QPSK signals, when the fiber dispersion is compensated with various filters. For comparison, the back-to-back BER is also shown in Fig. 4. We find that even a small number of taps provides significant improvement of the BER performance. 39 taps are sufficient to compensate for dispersion value of 4,000 ps/nm. Although chromatic dispersion is thus compensated to significant degree, power penalties of about 5 dB still remain after transmission of 200-km distance. These penalties may stem from the nonlinear distortion due to the relatively large input power and non-ideal filter coefficients.

![Fig.4 BERs measured as a function of the received power after transmission through a 200-km SMF.](image)

V. CONCLUSION

We have investigated the homodyne phase-diversity receiver, where the carrier phase is estimated with digital signal processing, alleviating locking the phase of the local oscillator to the carrier phase. We have demonstrated demodulation of M-ary PSK signals ($M=2, 4, 8$) at the symbol rate of 10 Gsymbol/s, and post-compensation for chromatic dispersion of 4,000 ps/nm with such receiver.

REFERENCES


Impairment mitigation in high speed optical communications using digital signal processing

R. I. Killey, P. M. Watts, S. J. Savory, R. Waegemanns, Y. Benlachtar, V. Mikhailov, M. Glick*, and P. Bayvel
Optical Networks Group, University College London, Torrington Place, London, WC1E 7JE, UK, r.killey@ee.ucl.ac.uk
* Intel Research Pittsburgh, 4720 Forbes Avenue, Suite 410, Pittsburgh, USA

Abstract We present results of recent studies on transmission impairment mitigation techniques using digital signal processing. The methods investigated include electronic predistortion, and coherent polarisation-multiplexed QPSK transmission with receiver-based equalization.

Introduction
The use of digital signal processing (DSP) in the next generation of optical transmission systems, to compensate for chromatic dispersion, fibre nonlinearity and optical filtering, offers the advantages of low cost and size, and adaptive operation. Ongoing increases in the speed and reductions in the power consumption of CMOS make this technology suitable for signal processing at optical line-rates of 10 Gb/s and above. Recent demonstrations of A/D and D/A converters operating at 20 Gs/s confirm the feasibility of these components for applications in high bit-rate optical communications.

A key question concerns the fundamental limits to the performance that can be achieved with DSP. Many studies to date have focused on direct (square law) detection schemes in which electronic dispersion compensation (EDC) is implemented at the receiver. However, in comparison to conventional optical compensation methods, the performance of such schemes is limited due to the loss of the optical phase information following detection. In contrast, signal processing techniques which operate directly on the optical field, either using electronic predistortion (EPD) [1,2] or coherent detection [3], offer the possibility to overcome these limitations.

With EPD, the optical amplitude and phase are predistorted at the transmitter to achieve the desired signal waveform at the receiver, allowing direct-detection to be used, and avoiding the requirement for local oscillators. We have carried out studies, assessing the transmission performance that can be achieved by electronic compensation of dispersion and intra-channel nonlinearities. In addition, we are developing EPD transceiver designs, based on field programmable gate array (FPGA) devices, to operate at up to 10 Gb/s.

In our research on coherent systems, the approach we adopt is based on a phase- and polarization-diverse coherent receiver combined with DSP to compensate for transmission impairments and to recover the data. By using polarization multiplexed QPSK as the modulation format, since each symbol carries 4 bits of information, a capacity of 40 Gbit/s can be achieved at 10 Gbaud, allowing the use of 20 Gs/s DSP. Using this technique, we have recently reported transmission over 6480 km of standard fiber without optical dispersion compensation [4]. In this paper we describe work carried out at UCL, assessing DSP-based EDC. Firstly, results of numerical studies of electronic predistortion are presented. Following this, the use of FPGAs in the implementation of digital EDC schemes is discussed. Finally, recirculating loop transmission studies on 42.8 Gb/s polarization-multiplexed QPSK are described.

Electronic predistortion
The EPD transmitter generates predistorted signals such that the chromatic dispersion (CD) and/or nonlinearity of the fibre reverse the distortion during transmission, resulting in the desired signal waveform at the receiver. Fig. 1 shows the design of an EPD transmitter, based on a Cartesian Mach-Zehnder modulator (MZM).

Transmission simulations were carried out with predistorted 10 Gb/s NRZ-OOK signals over 1200 km of standard SMF, with no optical compensation. Details of the simulations are presented in [5]. Fig. 2 shows the required OSNR plotted for the case of chromatic dispersion compensation alone, and for combined dispersion and SPM compensation. In the latter case, the penalty at 0 dBm launch power was reduced from 3.7 dB to 0.2 dB. The upper limit on launch power is governed by the sampling rate of the modulator drive signals, in this case 20 Gs/s (2 samples per bit). Experimental confirmation of the effectiveness of combined CD and SPM compensation is presented in [6]. An experimental demonstration of 72 channel WDM EPD signal transmission over 1600 km of standard SMF is described in [7], while [8] presents a detailed numerical assessment of multi-channel nonlinear effects.
**FPGA-based EPD transmitter design**

For the implementation of digital EDC schemes, while custom integrated circuit designs offer optimum performance, the field programmable gate array (FPGA) is a useful alternative tool for experimental work, as it is low cost and reprogrammable. In the latest FGPA, high speed serial interfaces are increasingly being used. Chips with up to 10 Gb/s input/output capabilities are in production or under development, allowing the latest FGPA to be used for real-time DSP operating on signals at standard optical line rates. We are currently assessing the use of FGPA to implement experimental EPD transceivers.

![Fig. 3 FPGA-based EPD transmitter design](image)

**Coherent polarization multiplexed QPSK**

In our work on coherent systems, we carried out polarization multiplexed QPSK recirculating loop transmission experiments at 10.7 Gbaud, giving a channel capacity of 42.8 Gb/s. At the receiver a local oscillator with linewidth of 100 kHz was combined with the signal via two asymmetric 3 fiber couplers, with coupling ratios 1:2:2, to detect the in-phase and quadrature components of the two polarizations. On conversion into the electrical domain, the signal was digitized at 20Gs/s using a Tektronix TDS6154C digital storage oscilloscope with the waveforms then processed off-line. Further details of the experiment are given in [4].

![Fig. 4 Contour plot of launch power vs distance for BER=3×10^{-3}](image)

Transmission over distances up to 6480km over standard fiber with no optical dispersion compensation was demonstrated (Fig. 4). A total dispersion of 107,232ps/nm was compensated using digital signal processing, with an OSNR penalty of 1.2dB.

**References**

5. R.I. Killey et al, Proc. ECOC 2005, paper Tu4.2.1
7. M. Birk et al, Proc. ECOC 2006, paper Th2.5.6

143
Advanced Modulation/Demodulation Technologies for High-Speed Optical Transmission Systems

Itsuro Morita and Sander L. Jansen
KDDI R&D Laboratories, 2-1-15 Ohara Fujimino Saitama, 356-8502, Japan
Tel: +81-492-78-7865, Fax: +81-492-78-7821, Email: morita@kddilabs.jp

Abstract
This paper reviews advanced modulation and demodulation technologies to increase the channel bit rate in optical transmission systems.

Introduction
Practically all commercial optical transmission systems with bit rates up to 10 Gbit/s are realized with intensity modulation direct detection (IM-DD). The main advantage of IM-DD is that the least components are required at transmitter and receiver. For bit rates of 40 Gbit/s and higher, however, a large signal-to-noise ratio (OSNR) is required for error-free operation and expensive electrical and optical components with high bandwidth are needed at the transmitter and receiver. In addition, accurate compensation for chromatic dispersion and polarization-mode-dispersion (PMD) is required for long-haul transmission systems with a bit rate of 40 Gbit/s and higher.

Most of these difficulties can be solved by reducing the symbol rate of the high speed signal by using an advanced modulation format. In this paper, promising advanced modulation formats for next generation transmission systems are reviewed.

DQPSK
Phase-shift keying (PSK) in combination with balanced detection is an attractive technology to improve the sensitivity of high speed transmission systems with respect to IM-DD. Recently, differential quadrature phase-shift keying (DQPSK) has been intensely studied [1]. DQPSK carries two bits information per symbol and therefore has a narrower spectral bandwidth at the same data rate (shown in Fig.1). The reduction of the symbol rate brings other benefits as well: The bandwidth requirement of the components in the transmitter and receiver is reduced by a factor of two and the tolerance towards chromatic dispersion and PMD is enhanced. The narrow spectral width of DQPSK enables WDM transmission with a high spectral efficiency. With 80-Gbit/s RZ-DQPSK, 1.6 bit/s/Hz [2] and 3.2 bit/s/Hz [3] spectral efficiency have been achieved using polarization interleaving and multiplexing, respectively. In [3], the record capacity of 25.6 Tbit/s has been demonstrated with C- and L-band EDFA amplifiers.

The use of DQPSK signals is also effective to increase the channel bit rate to 100 Gbit/s and higher. With a combination of RZ-DQPSK signals and optical time-division-multiplexing (OTDM) technologies, 2.56 Tbit/s transmission has been demonstrated [4]. This is the highest single-channel transmission experiment demonstrated so far. However, OTDM systems are not cost effective as they require many components and complex optical signal processing. Without OTDM, the highest bit rate ever achieved is 200 Gbit/s. To investigate the applicability of DQPSK in 100-Gbit/s transmission without OTDM, we conducted a DQPSK transmission experiment over 2-km of standard SMF (SSMF) without chromatic dispersion compensation [5]. In this experiment, a BER of $1 \times 10^{-9}$ was obtained with no sign of an error floor. After 2 km transmission only a 6-dB penalty was observed for the 37-ps/nm uncompensated chromatic dispersion. Long-haul 100 Gbit/s-based WDM transmission experiments have been also demonstrated with DQPSK [6, 7]. In [6], 200-GHz spaced 10 x 100-Gbit/s RZ-DQPSK signals have been transmitted over 2000 km. These results indicates that the DQPSK modulation format is a promising candidate both for short reach uncompensated transmission and long-haul transmission at 100 Gbit/s.
DSP-aided technologies

Coherent detection using digital signal processing (DSP) [8, 9] is an attractive method to increase the number of levels in multi-level (M-ary) signals. In such a system, the carrier phase is recovered after homodyne or heterodyne detection by means of DSP. Various experiments have been reported with multi-level (M-ary) signals. 8-PSK signal transmission with a symbol rate of 10 GSymbol/s (bit rate of 30 Gbit/s) has been demonstrated [10]. DSP has also enabled 32-level signal transmission with incoherent detection using delayed interferometers [11]. In this demonstration, a combination of quaternary amplitude-shift keying (QASK) and 8-PSK has been utilized to obtain 50-Gbit/s signal at a symbol rate of 10-Gsymbol/s.

DSP aided coherent detection can be employed as well to increase the robustness of the transmission system. By using DQPSK and polarization multiplexing in combination with a polarization diverse receiver, the symbol rate can be reduced by a factor of four. Recently, this technology has been employed to demonstrate 40-Gbit/s and 100-Gbit/s long-haul transmission [12-15]. In these systems digital equalization is used to significantly increase the chromatic dispersion and PMD tolerance.

Orthogonal frequency division multiplexing (OFDM) is a frequency efficient form of subcarrier multiplexing (SCM) in which subcarriers partly overlap. Recently OFDM, which is widely used in wireless communication, has been proposed for high speed optical transmission [16] and we have conducted OFDM transmission experiment at a bit rate of 20 Gbit/s [17]. The configuration of the transmitter and receiver, used in the demonstration, is shown in Fig. 2. In this case, DSP is used in both transmitter and receiver. The 20-Gbit/s OFDM signal was transmitted over 4160 km SMF without dispersion compensation. The main advantage of OFDM is that unlike pre-distortion [18], the signal is continuously detectable along the whole transmission link. Such a large dispersion tolerance is attractive for high speed transmission systems as it eliminates the necessity of inline dispersion compensation even in dynamically reconfigurable networks.

Conclusion

Advanced modulation and demodulation technologies to increase the channel bit rate are discussed. One of the main goals for advanced modulation formats is to reduce the symbol rate, since this relaxes the requirements for the bandwidth of optical and electrical components and enhances the robustness against the chromatic dispersion and PMD. The use of DQPSK is considered to be the current most promising method. To increase the robustness, DSP-aided technologies are attractive and further progress is expected.

Acknowledgement

The authors wish to thank Dr. S. Akiba, Dr. M. Suzuki, Dr. M. Usami and Dr. H. Tanaka of KDDI R&D Laboratories for their continued encouragement and helpful discussions. This work was partly supported by a project of the National Institute of Information and Communications Technology of Japan

References

Electronic mitigation techniques for 40 Gbit/s optical fiber transmission systems

Henning Bülow, Bernd Franz, Fred Buchali, Axel Klekamp, Alcatel-Lucent Research and Innovation
Holderaeckerstr. 35, 70499 Stuttgart, Germany, Dept. ZFZ/OT, henning.buelow@alcatel-lucent.de

Abstract
Different aspects of electronic distortion equalization (EDE) at 40Gbit/s are highlighted: from the experimental assessment of the adaptation of analog equalizer circuits, over the numerical simulation of digital MLSE for spectrally narrowed transmission channels, to the estimation of the digital processing (DSP) effort in various schemes.

Introduction
Today’s upgrade scenario for existing 10Gb/s WDM transmission links presumes a gradual replacement of 10Gb/s terminal equipment by 40Gb/s line cards while keeping the optical path designed for 10Gb/s operation, i.e. parameters and tolerances such as optical bandwidth given by WDM mux/demux for 100GHz or even 50GHz channel spacing, residual chromatic dispersion (e.g. up to 50km SMF) and PMD (e.g. up to 8ps). This demands robustness and distortion tolerance of the 40Gb/s transmission equipment. Today this is met by optical dispersion compensators (tunable grating CFBG) and optical PMD compensator. With the sustaining efforts to reduce size and complexity in future releases of 40Gb/s receiver boards, analog or digital (DSP) electronic signal processing schemes are investigated for electronic distortion equalization (EDE).

Adaptive analog distortion equalization
Lab devices of analog operating feed forward (FFE) and decision feedback equalizer (DFE) in SiGe or InP HBT technology have been realized and tested for 40Gb/s EDE application [1-3]. Though these devices of typically 5 forward taps and 1 feedback tap increase the tolerance of the system to many kinds of distortions by only some tens of percent, this might already be sufficient for links whose parameters are not so far away from 40Gb/s requirements.

For automatic adaptation of the equalizer consecutive dithering of the FFE taps are commonly used. The FEC error count provided by an FEC circuit might serve as feedback signal. The adaptation speed is mainly determined by the error rate since each dither step must wait for the collection of a sufficiently high number of errors (e.g. 1000) in order to obtain a stable (noiseless) feedback signal. With a BER around 10⁻⁴, an update of the feedback signal can be expected within a few milliseconds.

Alternatively also an eye monitor can be incorporated which provides a parameter which is related to the eye opening or the Q-factor of the eye diagram of the equalized signal. For eye monitoring at 40Gb/s a novel IC has been fabricated in a preproduction state-of-the-art SiGe bipolar technology with two independently operating decision gates for data decision and monitoring, respectively. Opposite to FEC error count the eye monitor gate can inspect the eye closer to the rails at higher BER around 10⁻². Thus a sufficient number of errors can be collected in a shorter time enabling an approximately 100 times faster adaptation speed.

In the following experiment the reduction of the number of the automatically adapted taps of a 5 tap FFE has been investigated. The error count served as feedback signal. A 42.7Gb/s ASK-NRZ signal was distorted by a tunable DGD emulator. First the 1st and 5th tap have been optimised at 0ps DGD and then only the remaining three centre taps have been automatically adapted to the actual distortion (blue boxes in Fig. 1). In a further experiment the first three taps were automatically adapted and the last two were fixed (red triangles). Finally all 5 taps were adjusted (green dots). Fig. 1 exhibits that there is no difference visible between 3 and 5 adapted taps up to a DGD of 15 ps. This behaviour can be used to reduce the adaptation time.

Digital signal processing (DSP)
Complex digital 10Gb/s equalizers such as the maximum likelihood sequence estimator (MLSE) are already introduced as product. These MLSEs are suitable for distortion corresponding to an inter-symbol-interference (ISI) of up to 3 bits (4 state MLSE) [4] or 5 bits (16 states). In a numerical simulation (Monte-Carlo sim. for noise, BER = 10⁻⁴) we compared the CD tolerance of an MLSE (4 states) in optical bandwidth limited channels. Fig. 2 shows the tolerance vs. optical filter bandwidth without equalizer (solid, RX) and with equalizer (dashed, MLSE) for the two modulation formats NRZ ASK and DPSK, respectively.

It is obvious that in the range of filter bandwidths of 50GHz down to 30GHz as they might be seen in ROADM cascades or in a 50GHz channel grid, the MLSE helps to keep the CD tolerance high.

Beside performance assessment based on dispersion or PMD tolerance analysis, in some presentations the realization constraints are already discussed by analyzing the penalty introduced by a finite analog-to-digital converter (ADC) resolution. But so far little attention has been paid to the complexity of the DSP processor itself.
The coherent detection scheme also referred to as intradyne detection, can be considered as the ultimate equalizer solution since it maps the phase and polarization information of the signal completely into the electrical domain [8,9]. If no further PMD compensation is applied and only one optical hybrid for phase diversity with two ADCs is realized, a complex 65 tap FIR filter processing the I and the Q signal samples is sufficient. It can further be reduced to 2 real 65 tap FIRs whose summation points are added, leading to 517 OpT, the same value as for the pre-compensation.

Optical OFDM [10,11] relies on fixed electronic DSP processing in both transmitter and receiver by fast and efficient realization of (inverse) Fourier transformers (IFFT/FFT) circuits and parallel but very simple adaptive processing of the individual subcarriers at the FFT output of the OFDM receiver. For a coarse DSP complexity assessment we determine the number of operations within the FFT and IFFT processor only. When splitting the 40Gb/s signal among 128 binary OFDM subcarriers the OFDM symbol has a length of roughly 3 ns. The IFFT processor at the transmitter with N=256 input samples (half of which are 0) provide a complex output signal for the I-Q modulator. An (inverse) DFT radix-2-algorithm requires about 5N log2(N) real numerical operations per OFDM symbol (3 ns) in both transmitter and receiver, which finally leads to 160 OpT.

All the operations per bit-period are summarized in the table Tab. 1. It is obvious that with 28000 OpT a MLSE for the distortion of 1000 ps/nm might be out of scope for realization. On the other hand optical OFDM with 160 OpT appears to be comparable to 16 state MLSE with the difference, that the processing is split among transmitter (IFFT) and receiver (FFT). Also electronic precompensation and intradyne detection are comparable from the signal processing effort (517 OpT).

![Fig. 2: Simulated chrom. dispersion tolerance vs. optical bandwidth for ASK (blue) and DPSK (purple) with MLSE (dashed lines) and w/o. equalizer (solid lines).](image)

Here we introduce the parameter “Operations per bit-period T” (OpT) which quantifies the average number of calculation (real addition, multiplication, or metric extraction by RAM look-up table) applied to the digitized signal samples within the DSP processor. Further processing for adaptation, carrier or phase recovery as well as demultiplexing are not considered here. In the following we will discuss the processing effort of different DSP schemes for 1000 ps/nm dispersion compensation (60 km standard SMF).

The first DSP scheme we are looking at is the electronic pre-compensation [6]. It is based on an optical field synthesizer with electrical DSP calculation of the complex transmitter field. After digital-to-analog conversion (DAC) the electrical representatives of I and Q field components are converted into the optical domain by an external modulator. The inter symbol interference (ISI) of roughly 32 bits induced by 1000 ps/nm necessitates two FIR filters with 65 tap leading to 516 OpT (65 tap weight multiplications and 64 additions per filter and per half bit).

The core operation of the MLSE is concentrated the add-compare-select (ACS) unit which is positioned at each state of the trellis and governs 7 numerical operations per T. An extrapolation of 10Gbit/s simulations [7] to 40Gbit/s indicates that at least 4096 states are needed leading to a high processing effort of 28000 OpT.

The coherent detection scheme also referred to as intradyne detection, can be considered as the ultimate equalizer solution since it maps the phase and polarization information of the signal completely into the electrical domain [8,9]. If no further PMD compensation is applied and only one optical hybrid for phase diversity with two ADCs is realized, a complex 65 tap FIR filter processing the I and the Q signal samples is sufficient. It can further be reduced to 2 real 65 tap FIRs whose summation points are added, leading to 517 OpT, the same value as for the pre-compensation.

Optical OFDM [10,11] relies on fixed electronic DSP processing in both transmitter and receiver by fast and efficient realization of (inverse) Fourier transformers (IFFT/FFT) circuits and parallel but very simple adaptive processing of the individual subcarriers at the FFT output of the OFDM receiver. For a coarse DSP complexity assessment we determine the number of operations within the FFT and IFFT processor only. When splitting the 40Gb/s signal among 128 binary OFDM subcarriers the OFDM symbol has a length of roughly 3 ns. The IFFT processor at the transmitter with N=256 input samples (half of which are 0) provide a complex output signal for the I-Q modulator. An (inverse) DFT radix-2-algorithm requires about 5N log2(N) real numerical operations per OFDM symbol (3 ns) in both transmitter and receiver, which finally leads to 160 OpT.

All the operations per bit-period are summarized in the table Tab. 1. It is obvious that with 28000 OpT a MLSE for the distortion of 1000 ps/nm might be out of scope for realization. On the other hand optical OFDM with 160 OpT appears to be comparable to 16 state MLSE with the difference, that the processing is split among transmitter (IFFT) and receiver (FFT). Also electronic precompensation and intradyne detection are comparable from the signal processing effort (517 OpT).

<table>
<thead>
<tr>
<th>DSP @ 40Gb/s</th>
<th>ops/T: OpT</th>
<th>CD @40Gb/s</th>
</tr>
</thead>
<tbody>
<tr>
<td>MLSE 1.6 states</td>
<td>150 @ 200ps/nm</td>
<td>1.6 states 28000</td>
</tr>
<tr>
<td>El. precompensation 2x65t. FIR</td>
<td>517 @ 1000ps/nm</td>
<td>Intradyne equalizer 2x65t. FIR</td>
</tr>
<tr>
<td>Opt. OFDM, Ts=3ns [IFFT+FFT]</td>
<td>160 @ 1000ps/nm</td>
<td></td>
</tr>
</tbody>
</table>

Tab. 1: DSP processing effort (pure equalization effort) of different DSP-based equalization schemes.: Operations per bit period (T) for CD mitigation at 40 Gb/s.

Conclusion

At 40Gbit/s electronic distortion equalization performance and adaptation dynamics of analog electronic equalizer circuits are already investigated in lab experiments. Due to the higher complexity various kinds of DSP based equalization schemes are in the infancy for 40Gbit/s application ranging from MLSE to intradyne equalization. Numerical evaluation indicates that the MLSE facilitates or even enables transmission at high distortion tolerances even over existing 10Gb/s infrastructure (WDM filters) with low bandwidth (e.g. ROADM cascade). For the assessment of DSP signal processing complexity the parameter “operations per bit-period” (OpT) has been introduced and applied to chromatic dispersion mitigation (1000 ps/nm). Optical OFDM has the potential of the least complexity, followed by intradyne detection and electronic pre-compensation. MLSE is more than two orders of magnitude more complex.

Part of this work has been supported by the European Commission within the IST project NOBEL phase 2.

References

[2] H. Jiang et al., OFC 2006, OWB4,OTuE1
[7] P.Poggio et al., ECOC 2006, PDP, Th4.4.6
[8] S. Tsukamoto et al., ECOC 2006, Mo4.2.1
[9] S.J. Savory et al., ECOC 2006, Th2.5.5

Conclusion

At 40Gbit/s electronic distortion equalization performance and adaptation dynamics of analog electronic equalizer circuits are already investigated in lab experiments. Due to the higher complexity various kinds of DSP based equalization schemes are in the infancy for 40Gbit/s application ranging from MLSE to intradyne equalization. Numerical evaluation indicates that the MLSE facilitates or even enables transmission at high distortion tolerances even over existing 10Gb/s infrastructure (WDM filters) with low bandwidth (e.g. ROADM cascade). For the assessment of DSP signal processing complexity the parameter “operations per bit-period” (OpT) has been introduced and applied to chromatic dispersion mitigation (1000 ps/nm). Optical OFDM has the potential of the least complexity, followed by intradyne detection and electronic pre-compensation. MLSE is more than two orders of magnitude more complex.

Part of this work has been supported by the European Commission within the IST project NOBEL phase 2.

References

[2] H. Jiang et al., OFC 2006, OWB4,OTuE1
[7] P.Poggio et al., ECOC 2006, PDP, Th4.4.6
[8] S. Tsukamoto et al., ECOC 2006, Mo4.2.1
[9] S.J. Savory et al., ECOC 2006, Th2.5.5

Conclusion

At 40Gbit/s electronic distortion equalization performance and adaptation dynamics of analog electronic equalizer circuits are already investigated in lab experiments. Due to the higher complexity various kinds of DSP based equalization schemes are in the infancy for 40Gbit/s application ranging from MLSE to intradyne equalization. Numerical evaluation indicates that the MLSE facilitates or even enables transmission at high distortion tolerances even over existing 10Gb/s infrastructure (WDM filters) with low bandwidth (e.g. ROADM cascade). For the assessment of DSP signal processing complexity the parameter “operations per bit-period” (OpT) has been introduced and applied to chromatic dispersion mitigation (1000 ps/nm). Optical OFDM has the potential of the least complexity, followed by intradyne detection and electronic pre-compensation. MLSE is more than two orders of magnitude more complex.
Solid Photonic Crystal Fibres
R. Goto, S. Matsuo, K. Himeno
Fujikura Ltd.
1440, Mutsuzaki, Sakura, Chiba, 285-8550, Japan
Tel: +81-43-484-2197, Fax: +81-43-481-1210, E-mail: ryugoto@lab.fujikura.co.jp

Abstract
All-silica solid photonic band gap fibres (PBGF) are reviewed focusing on our work. A 2-D PBGF, which has a low loss of 3.6 dB/km at 1520 nm and a zero-dispersion wavelength below 1300 nm, is introduced.

1. Introduction
Photonic crystal fibres (PCFs) have been of great interest since they have unique optical properties such as endlessly single-mode operation [1] and light guidance in air [2]. Conventional PCFs have holes in their glass structures to obtain a large refractive index difference. However, these “holey” structures suffer from several difficulties, such as pressure control of the holes during fibre drawing and hole collapse during fusion splicing. Therefore, an all-glass fibre structure is advantageous in terms of fabrication and fusion splicing. Solid photonic crystal fibres, especially solid photonic band gap fibres (solid PBGFs) [3][9] have been attracting much attention ever since photonic band gap guidance in an all-silica glass structure has been demonstrated [5]. Photonic band gap guidance enables various unique optical properties such as wavelength filtering or dispersion property. An all-glass structure enables easier fabrication and fusion splicing. The fibre core can be doped with rare earth metals and combined with photonic band gap effects. Furthermore, an all-silica structure has a potential toward low loss operation, and the lowest loss of 6 dB/km at 1550 nm has been reported [7].

Table 1 shows two categories of solid PBGFs. 1-D structure PBGFs, also called Bragg fibres, are made by the MCVD method and hence the fabrication process is relatively simple. The lowest loss of ~3 dB/km has been reported [9]. However, the effective bandwidth is narrowed since, at certain wavelengths, the core mode couples to leaky modes which propagate in the cladding rings. In contrast, 2-D structures do not exhibit core-cladding coupling and the whole band gap can be used as a transmission band. Although 2-D solid PBGFs are fabricated by the stack-and-draw method and the fabrication is more complicated compared to the MCVD method, they offer a wider variety of structures and design flexibility. In this paper, we focus on our work on a 2-D solid PBGF and present its properties hereafter.

2. Structure and properties of 2-D solid PBGF
Figure 1 shows a microscopic picture of the solid PBGF we fabricated [8]. The cladding structure is the triangular lattice structure and composed of seven periodic layers of Ge-doped high index rods around the core. The core is formed by removing three layers from the periodic structure and replacing them with silica rods. The pitch between two neighbouring rods, \( \Lambda \), is 4.2 \( \mu \)m. The diameter of the high index rods, \( d \), is 2.1 \( \mu \)m. The fibre diameter is 125 \( \mu \)m. We used a large relative refractive index difference (\( \Delta \)) of 2.8% between the pure silica cladding and the high index rods. A large refractive index difference is effective to reduce the fibre attenuation, since the attenuation of solid PBGFs is limited by macro bending loss [7] and high index contrast helps reduce the bending loss.

![Fig. 1 Microscopic picture of solid PBGF](image.jpg)

Figure 2 shows the band diagram of the cladding structure. The plane wave expansion method is used for the band diagram calculation [10]. The refractive index of silica is set to 1.45 in the calculation. The first band gap exists over \( \lambda \geq 1.1 \) \( \mu \)m-1.7 \( \mu \)m, and the second band gap exists over \( \lambda \geq 0.65 \) \( \mu \)m-0.9 \( \mu \)m. Guided modes exist in these wavelength ranges.

![Fig. 2 Band diagram of cladding structure](image.jpg)
We measured the near-field patterns of the guided modes at 1550 nm. A conventional single-mode fibre is used as a launch fibre. Figure 3 shows the near-field patterns of (a) LP01-like and (b) LP11-like modes after 1 km of propagation. The LP11-like mode is observed only when the fibre is bent in a small diameter smaller than 10 mm. The LP01-like mode usually does not couple to higher order modes in this fibre. No other higher order modes were observed after 1 km.

![LP01-like mode](image1)

(a) LP01-like mode

![LP11-like mode](image2)

(b) LP11-like mode

Fig. 3 Near-field patterns at 1550 nm

The OTDR traces at 1.55 μm using three different launch fibres are shown in Figure 4. The mode field diameters (MFDs) of the launch fibres at 1550 nm are 6.8 μm, 10.5 μm and 22 μm and all the fibres are single-mode at 1550 nm. We can see the slopes of the OTDR traces are different in the first 200 m. However, these slopes become the same after 200 m. Since only LP0x-like modes are excited by the launch fibres and the slopes of the OTDR traces change as we change the MFDs of the launch fibres, the slope differences are explained by the existence of a LP02-like mode with high attenuation. But after 200 m, all the slopes become the same. This means that the LP02-like mode has disappeared and only the LP01-like mode is guided in the fibre.

![OTDR traces of solid PBGF](image3)

Fig. 4 OTDR traces of solid PBGF

The attenuation spectrum of the fibre is shown in figure 5. To avoid the LP02-like mode being taken into account in the measurement, the fibre was cut back from 380 m to 150 m. The lowest attenuation is 3.6 dB/km at 1520 nm. This is the lowest attenuation ever obtained by all-silica solid PBGFs. However, we can see two attenuation peaks due to OH impurity absorption at 1240 nm and 1383 nm. Reduction of OH will possibly reduce the attenuation of the fibre. We next measured the chromatic dispersion of the fundamental mode of this fibre by selectively exciting the fundamental mode using a single-mode fibre as a launch fibre. The length of the fibre was 413 m, which was long enough for the LP02-like mode to disappear. Figure 6 shows the chromatic dispersion of the fibre. The chromatic dispersion of the fibre at 1310 nm and 1550 nm are 6.6 ps/nm/km and 32.6 ps/nm/km, respectively. The dashed line in figure 6 is an interpolation, since a light source was not available. The zero-dispersion wavelength is 1261.8 nm. For comparison, typical chromatic dispersion of a conventional single-mode fibre is also shown in the figure. These chromatic dispersion properties are not obtainable by conventional all-silica fibres and hence these results show one of the unique optical properties of solid PBGFs.

![Attenuation spectrum](image4)

Fig. 5 Attenuation spectrum

![Chromatic dispersion](image5)

Fig. 6 Chromatic dispersion

3. Conclusion

All-silica solid photonic band gap fibres (PBGFs) are reviewed. Optical properties of a 2-D PBGF we fabricated have been introduced. The fibre has demonstrated a low loss of 3.6 dB/km and a zero dispersion wavelength below 1300 nm. This result implies that 2-D PBGF is a practical solution for realising the unique characteristics of PBGFs.

References

Designs and Fabrications of Photonic Crystal Fiber Couplers with Air Hole Collapsed Taper Regions

Hirohis YOKOTA, Hiroki KAWASHIRI, Hirotomo YASHIMA, and Yutaka SASAKI
Graduate School of Science and Engineering, Ibaraki University
4-12-1 Nakanarusawa, Hitachi, Ibaraki 316-8511, Japan
Phone & FAX: +81-294-38-5215
E-mail: hirohis@mx.ibaraki.ac.jp

Abstract
We theoretically clarified that the excess loss of the air hole collapsed photonic crystal fiber coupler (PCFC) is lower as the air hole collapsed elongation ratio is smaller. A prototype of the PCFC was fabricated using CO₂ laser irradiation.

1. Introduction
Since photonic crystal fibers (PCFs) have unique properties that cannot be obtained in conventional fibers, PCFs are expected to the applications as functional fibers [1]. Recently, low loss PCFs had been developed [2] and realization of optical fiber systems that consist of PCFs are expected. To construct optical fiber systems using PCFs, introducing of PCF devices are desired.

An optical fiber coupler is one of the basic components in an optical fiber system and a development PCF coupler (PCFC) for PCF system is desired. Recently, PCFC fabrication using CO₂ laser irradiation technique had been reported [3]. In this fabrication method, it can be selected whether air holes in taper region of the PCFC are remaining or collapsed by controlling irradiated laser power. Theoretical investigations of PCFC were also reported; where coupling characteristics of PCFCs with air hole remaining tapers for PCF structures were clarified [3]. However, theoretical investigations of PCFCs with air hole collapsed taper regions have not been reported.

In this paper, we present numerical examples of air hole collapsed PCFC designs, where extinction ratios and losses for air hole collapsed elongation ratio are clarified. A PCFC with air hole collapsed taper region is fabricated and its characteristics are presented.

2. Theoretical analysis of air hole collapsed PCFC
In the theoretical investigation of air hole collapsed PCFC characteristics, we calculated light propagations in PCFC. The three dimensional refractive index profile of the PCFC was replaced to an equivalent slab waveguide coupler using effective index method [4], where the air hole remaining coupler cross section was approximated to equivalent step index fiber coupler cross section similar to the approximation in PCF analysis [5] and the air hole collapsed coupler taper cross section was treated as a dumbbell-like channel waveguide cross section. The light propagation in the slab coupler was calculated using finite difference beam propagation method (FD-BPM). We can obtain extinction ratios at the output ports and excess loss from calculated field distribution at the end of coupler taper.

In the numerical calculations, the air hole diameter d and the air hole pitch A were chosen as 2.65μm and 5μm, respectively (i.e. d/A = 0.53). The degree of fusion [4] of PCFC was chosen as 0.95. Parameters of tapered shape of the air hole collapsed PCFC are shown in Fig. 1. The elongation ratio τ that is defined as τ = c_{min}/c_{0} was chosen as 0.25 and 0.3. The air hole collapsed elongation ratio τ_{col} is defined as τ_{col} = c_{col}/c_{0}. The tapered shape of the PCFC was assumed to be SIN²-type [6], and the taper length L_{c} was set as 18mm. The extinction ratios at ports 3 and 4 and the excess loss of the PCFC, in which the light with 1550nm-wavelength is launched from port 1, were calculated.

Figure 2 shows the extinction ratios and excess loss for air hole collapsed elongation ratio τ_{col}. The excess loss is lower as τ_{col} is smaller. For τ = 0.25, the extinction ratios at port 3 and port 4 are almost constant for τ_{col}. For τ = 0.3, the extinction ratio at port 3 is increased and that at port 4 is decreased as τ_{col} is larger in the range of τ_{col} > 0.45.
Fig. 2 Extinction ratios and excess loss for air hole collapsed elongation ratio.

3. Fabrication of air hole collapsed PCFC

A prototype of air hole collapsed PCFC was fabricated using CO$_2$ laser irradiation technique. A photograph of the cross section of PCF for PCFC fabrication is shown in Fig. 3. An air hole diameter and an air hole pitch of the PCF were 1.5μm and 3.0μm, respectively.

Fig. 3 Cross section of the PCF for PCFC fabrication.

In PCFC fabrication, the time for fusion before elongation was 75s and the time for elongation was 225s. The elongation speed was set to 60μm/s. Photographs of fabricated PCFC side view are shown in Fig. 4. It is found that air holes are collapsed at taper transition region. It was clarified that air hole collapsed position in coupler taper i.e. air hole collapsed elongation ratio $r_{col}$ can be controlled by changing laser power in the elongation process.

![Taper transition region]

(a) Taper transition region

![Taper waist]

(b) Taper waist

Fig. 4 Air hole collapsed PCFC side view.

Characteristics of the fabricated PCFC at 1554nm-wavelength were measured. The extinction ratios at port 3 (straight port) and port 4 (cross port) were 59% and 41%, respectively. The excess loss was ~5dB, and reducing of it is a future subject.

4. Conclusions

In this paper, we present numerical examples of air hole collapsed PCFC designs. Characteristics of the PCFC for air hole collapsed elongation ratio $r_{col}$ were theoretically obtained. It was clarified that the excess loss is lower as $r_{col}$ is smaller. It was also clarified that the extinction ratios for $r_{col}$ depend on the elongation ratio $r$. A prototype of the air hole collapsed PCFC was fabricated using CO$_2$ laser irradiation technique. The extinction ratios at port 3 (straight port) and port 4 (cross port) were 59% and 41%, respectively. The excess loss was ~5dB, and reducing of it is a future subject.

References

Photonic crystal fiber interferometer composed of a long-period grating and one point collapsing

Hae Young Choi, Myoung Jin Kim, Seok Han Kim, and Byeong Ha Lee
Department of Information and Communications, Gwangju Institute of Science and Technology
1 Oryong-dong, Buk-gu, Gwangju, 500-712, Korea
Tel: +82-62-970-3153, Fax: +82-62-970-2204
E-mail: leebh@gist.ac.kr

Abstract

We present all-fiber interferometer based on a long-period fiber grating and collapsing of the air holes in PCF. As a potential application of the proposed all fiber interferometer, strain sensing is experimentally demonstrated.

1. Introduction

A photonic crystal fiber (PCF), which is generally composed of a single material with an array of air holes running along its longitudinal axis, has been widely studied over the last decade [1]. Since the structure of the air holes is adjustable, its unique characteristics such as single mode operation over a wide wavelength range and tailoring chromatic dispersion have been obtained. These properties make the PCF an attractive alternative for conventional single mode fibers. Therefore, the recent applications of the PCF have surged from the traditional telecommunication field to the field of sensing. The optical fiber sensors based on the PCF interferometer utilizing long period fiber gratings (LPGs) pair and tapering technique have been reported [2, 3]. They have demonstrated that PCF was very attractive for sensing applications such as high temperature and strain.

In this article, we present the PCF interferometer that is composed of a single LPG and one point collapsing region. The LPG was imprinted in pure silica PCF by using the electric arc discharge of a commercial fusion splicer. The air-hole collapsing was made by the same fusion splicer also. With this interferometer, all-PCF strain sensor is experimentally demonstrated.

2. Fabrication of LPG in PCF

The LPG was imprinted in pure silica PCF by using a commercial fusion splicer (S183PM, FITEL co.). A piece of PCF was unjacketed and fixed on a motor-driven translation stage. An LPG was made by discharging the electric arc repeatedly while moving the stage. This process was repeated several times until a required attenuation loss peak was obtained. The transmission spectrum of the fabricated PCF-LPG was measured by using a broadband LED source and an optical spectrum analyzer (OSA).

Figure 1 shows the transmission spectrum of the fabricated PCF-LPG having a grating period of 480 μm and 20 grating periods. The resonant wavelength of the LPG was located around 1.31 μm.

![Transmission spectrum of the PCF-LPG](image)

Fig. 1. Transmission spectrum of the PCF-LPG. The grating period was 480 μm, and the number of periods was 20.

3. One point collapsing of the air holes in PCF

The PCF (Crystal Fiber Co.) used in our experiment has 4 layers of hexagonally structured air holes around a solid core of diameter 11 μm as shown in Fig. 2 (a), which is designed to have endless single mode property.

![Cross sectional views of the PCF](image)

Fig. 2. (a) The cross sectional view of the PCF before arc discharge, (b) the side view and (c) cross sectional view of the PCF after strong arc discharge.

When the PCF was heated by strong electric arc discharges, the air holes in the cladding region were collapsed. Figure 2 (b) and 2 (c) show the side and cross sectional views of the collapsed region, respectively. We can see that the air holes were fully collapsed.

At the collapsed region, the PCF is no more single mode fiber since the fiber has no cladding part as shown in Fig. 2 (c). Thus, the beam that has been originally in the core mode can be partly directed into cladding region. Therefore, the air hole collapsing can be used to excite the cladding modes of PCF.
4. All fiber interferometer and its application

Figure 3 shows schematic of the proposed technique for an all fiber interferometer.

Fig. 3. Schematic of the interferometer based on the LPG and one point collapsing in a single piece PCF.

A part of the core mode was coupled to a cladding mode by an LPG. This coupling was wavelength selective due to the LPG. The coupled cladding mode was re-coupled to the core mode at the collapsing region. This re-coupling was not wavelength selective. It was just the mode coupling from a multimode waveguide to a single mode waveguide.

Since the re-coupled core mode makes interference with the mode that has not been coupled at both elements, an all-fiber interferometer can be simply implemented. The distance from the center of the LPG to the center of the collapsing region corresponds to the physical length of the interferometer.

An all-fiber interferometer having a physical length of 8 cm was fabricated, and the variation of its transmission spectrum was monitored while applying longitudinal strain on it.

Fig. 4. Transmission spectra of the interferometer measured at 0 and 2200 με strain.

Figure 4 shows the spectra measured without strain (solid line) and with 2200 με (dash line) strain. Interestingly, the spectrum moved toward the short wavelength direction. The strain-induced variations of the interference fringes were measured with small steps and plotted in Fig. 5. Interference peaks were linearly blue shifted with a strain sensitivity of ~1.8 pm/με. Such a behavior is opposite to that of the strain sensor based on conventional single mode fiber [4].

In the in-line fiber interferometer, the phase difference induced by the propagation along the fiber is given as \( k(n_{\text{core}} - n_{\text{cl}}) L \). When the involved modes are non-dispersive or suffer small dispersion, the \( n_{\text{core}} - n_{\text{cl}} \) term is nearly constant, thus the wavelength becomes proportional to the interferometer length \( L \). Therefore, peaks of fringe patterns are shifted to the long wavelength direction (red shift). It usually happens with standard single mode fibers.

However, because the cladding region of the PCF is very dispersive, the \( n_{\text{core}} - n_{\text{cl}} \) term cannot be constant anymore. In our experiment, it is thought that, \( n_{\text{core}} - n_{\text{cl}} \) decreases greatly with applied strain and causes the blue shift. The similar behavior was observed in sensor with single long period fiber grating inscribed in PCF [5].

Fig. 5. The strain response of the interference peak centered around 1.31 μm.

5. Conclusions

We have demonstrated an all-fiber interferometer that was implemented with pure silica PCF by combining an LPG and air-hole collapsing. Its potential application as a strain sensor has been experimentally verified. The proposed PCF interferometer will be very powerful as a high temperature sensor because the PCF itself is composed of only fused silica, which can overcome the core diffusion problem of a conventional fiber.

This work was supported in part by the BK-21 and APRI projects at GIST.

References

A Simple and Practical Design Approach to Realize Band-Pass Photonic Crystal Fiber Filters

Shailendra K. Varshney, Kunimasa Saito, Nikolaos J. Florous, and Masanori Koshiba
Division of Media and Network Technologies, Hokkaido University, Sapporo 060-0814, Japan
E-mail: skvarshney_10@yahoo.co.uk

A new in-line photonic crystal fiber band-pass filter based on a simple design approach with more than 76% transmission, centered at 1136 nm, is numerically investigated. A beam-propagation-method was used to verify the operation of filter.

1. Introduction

Wavelength selective devices such as optical band-pass or band-rejection filters are in great demand for wavelength division multiplexing (WDM) or dense-WDM (DWDM) optical fiber communication systems. Such devices can serve as building blocks for all-optical channel selection. Several types of all-fiber spectral filters have been reported during the last decade [1-2]. Among them, filters constructed with fused-taper couplers, dissimilar fiber couplers, and dual-core fibers have received much attention. On the other hand, photonic crystal fibers (PCFs) have been a subject of intensive research due to their novel optical properties such as ultra-wide band single mode operation [3]. PCFs can provide elegant ways to act as functional devices such as, narrow band pass filters based on solid and hollow core photonic crystal fibers and polarization independent couplers [4-6].

Most of the PCF designs were based on dual core separated by a resonator. The biggest disadvantage of such functional devices is their practical realization as they may show difficulty in the fabrication due to complex geometry in the cladding. Therefore, to realize a feasible structure, we have adopted a simpler design approach to achieve PCF band-pass filter. Our design procedure is based on the fusion of two different physical mechanisms [7] that is short-wavelength cut-off (long-wavelength band-pass filter) and long wavelength cut-off (short wavelength band-pass filter) of the fundamental mode. The short wavelength cut-off of the fundamental mode is achieved by doping the core with fluorine material of lower refractive index [8], while the long wavelength cut-off is obtained by increasing the size of the air-holes in the first ring. This will impart a band-pass filter behavior to PCFs. The fiber’s geometrical parameters as well as the doping concentration and doping size play an important role in determining the optimum performance of the device.

2. Fiber’s profile and design principle

Figure 1 shows the transverse cross-section of the proposed fiber filter. The doping region (red circle) is characterized by doping size \( D \) and fluorine concentration \( \Delta \) (%). The air-holes surrounding the doped region have diameter \( d' \) while the rest of air-holes have hole-diameters \( d \). The design parameters are \( d = 0.90 \mu m \) and \( \Delta = 3.2 \mu m \) and the index of silica was assumed 1.458 in calculations. This PCF presents an anti-guiding wavelength [8] which is simply a short-wavelength cut-off and thus the fiber allows all longer wavelengths larger than the anti-guiding wavelength and hence the device can acts as a long wavelength-pass filter. By increasing the size of the air-holes fencing the fluorine-doped region, a long-wavelength cut-off can be achieved where the index of the guided mode becomes smaller than the index of the cladding mode and thereof the fiber can work as a short-wavelength pass filter. Therefore, the proposed PCF device can operate as a band-pass filter. The 3-dB bandwidth can be controlled by manipulating the geometrical parameters. We have employed a full-vectorial finite element method and beam propagation method (BPM) [9] to simulate the modal characteristics and transmission response of the PCF based band-pass filter.

For the present analysis, we have considered two sets of simulation parameters where the total number of air-hole rings \( N \), size of doping region, and doping concentration were varied, while other design parameters such as pitch, hole diameters \( d \) and \( d' \) are fixed. Figure 2(a) depicts the spectral variation of the absolute difference between the index of the fundamental guided core mode \( n_{eff} \) and the effective cladding index \( n_{cl} \), while the inset graph shows the index variation of these two, where \( d' = 1.24 \mu m, D = 3.2 \mu m, \) and \( \Delta = 0.16 \) %. It can be seen from the graph that the index of the guided mode matches at two different wavelengths, that is short and longer wavelengths, respectively, as presented by arrows in Fig. 2(a). This establishes the band-pass filtering operation of the proposed PCF. Both short and long-
wavelength cut-off can be tailored by changing the hole diameter \( d \) and \( d' \). In addition, if we increase \( \Delta \), the fiber may not guide the fundamental mode and it becomes leaky, whereas if we decrease \( \Delta \), the short-wavelength cut-off shifts to shorter side and long-wavelength cut-off shifts towards longer wavelength for a particular set of geometrical parameters. Next, we run the BPM solver for the design parameters as mentioned in Table 1. Figure 2(b) shows the transmission characteristics of the device for two sets of parameters. The solid red-curve corresponds to the case where the doping radius is half of the pitch, fluorine concentration is 0.16 % \((n_f = n_{atm} - \Delta)\), and total number of air-hole rings are eight. It is seen that the insertion loss (IL) is 2.2 dB while the 3-dB bandwidth is 385 nm when the PCF is 5 cm long. The 3-dB bandwidth decreases when the length of fiber increases at the cost of higher insertion loss. This is due to the fact that the operation of the device relies on the leakage loss characteristics of the fiber. The fiber is very leaky near the cut-off wavelengths, thus we observe a trade-off between the 3-dB bandwidth and the length of the fiber. Next, we analyze the scenario where the doping concentration is increased and the doping radius is reduced to 1 \( \mu \)m in order to compensate for the reduced doping index (solid blue curve). An insertion loss of 2.8 dB and 307 nm of 3-dB bandwidth can be obtained when the fiber length was 8 cm. The insertion loss comes from the mismatch of the mode-fields that can be understood as an optical mode field from the unperturbed PCF (when \( d' = d \)) without fluorine doping is launched to the proposed PCF filter.

Fig. 1 Schematic representation of the proposed PCF band-pass filter, the core is doped with fluorine material as shown by red circle with doping size \( D \), the size of the first six air-holes is increased to \( d' \) from \( d \), and the lattice constant is \( \Lambda \).

3. Summary

The band-pass filtering characteristics of solid core PCFs have been obtained through a simple and easy to fabricate design approach which was based on a W-type index profile with a depressed core. The basic physical insight in the operation of the device was the control of the leakage loss property of the fundamental mode. It has been found that the controlled variation of the fiber’s structural parameters, namely doping radius, doping level, and first ring’s hole-diameters, the band pass characteristics of PCFs can be tuned and such study as well as the thermo-optical properties of the device is currently under investigation and will be reported in the conference.

References


Table 1. Fiber design parameters

<table>
<thead>
<tr>
<th>Parameters</th>
<th>( d'/\Delta )</th>
<th>( d/\Delta )</th>
<th>( t_g )</th>
<th>( \Delta )</th>
<th>( N )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Set-1</td>
<td>0.387</td>
<td>0.28</td>
<td>1.6 mm</td>
<td>0.16 %</td>
<td>8</td>
</tr>
<tr>
<td>Set-2</td>
<td>0.387</td>
<td>0.28</td>
<td>1.0 mm</td>
<td>0.35 %</td>
<td>11</td>
</tr>
</tbody>
</table>
Wavelength-dependent core-mode blocker using micro air bubble in the hollow optical fiber

J. Jeon$^1$, J. Kim$^2$, Y. Jung$^2$, W. Shin$^3$, D. Ko$^3$, J. Lee$^3$ and K. Oh$^1$

1. Institute of physics and applied physics, Yonsei University 134 Shinchon-dong, Seodaemun-gu, Seoul, 120-749, Republic of Korea, Tel: 82-2-2123-7657, Fax: 82-2-365-7657, Email: koh@yonsei.ac.kr
2. Department of Information and Communications, GIST, Buk-gu, Gwangju, 500-712, Republic of Korea
3. Advanced Photonics Research Institute, GIST, Buk-gu, Gwangju 500-712, Republic of Korea

Abstract
We have experimentally demonstrated a new wavelength dependent core-mode blocker using micro air bubble. Distinctive collapsing of air hole in hollow optical fiber generated various size of air bubble in optimal fusion splicing conditions.

1 Introduction
To date, several all-fiber core-mode blocker have been proposed and experimentally demonstrated for applications in all-fiber bandpass filters. For example, the local fiber damage in the core with high power laser$^{(1)}$ and a short hollow optical fiber (HOF)$^{(2)}$ segment have been introduced between long period grating pairs. The former scheme induced significant scattering loss and lacked the fabrication reproducibility. The later scheme took advantages in effective core-mode to radiation-mode conversion, and minimal perturbation of cladding mode but required high precision in cleaving a short piece of fiber.

In this paper, we introduce a new-type of core-mode blocker using micro air bubble in trapped using a distinctive collapsing process of the air hole in HOF. We report propose a simple and control the degree of the air-hole collapse to realize all-fiber core-mode blocker by arc discharges and its spectral response.

2 Experiment
In prior papers$^{(3,4)}$, the authors reported adiabatic collapse of HOF to splice with single mode fiber (SMF) seamlessly. In this study, contrary to adiabatic lossless splice, we discovered that micro-bubble can be deliberately formed in the splicing zone with an appropriate arc condition.

The schematic of the process of bubble formation is shown in Fig.1.

Similar to the conventional fusion splice method of solid core fiber, the HOF was prepared for fusion (stripped, cleaved, cleaned and placed in splicer, Fig.1a). Then the melting and collapsing process started near the midpoint of the splicing zone (Fig.1b). The inner air is pushed out to the central point and capsulized in the fiber to form a micro-bubble (Fig.1c).

Typical parameters of the arc condition for a HOF with the hole diameter of 6μm are; arc duration time of 0.55 second, pre-fusion time of 0.04 second and arc power level of 110.(In our research, we have used
the Furukawa S175 splicer whose default values are arc
duration time of 0.75 second, pre-fusion time of 0.24 second
and arc power level of 95 in SMF splicing mode) The actual
photographs of the micro-bubble entrapped between HOF
splice is shown in Fig. 2, for two types of HOF, hole
diameter of 4 and 6 μm.

In Fig. 2a, the bubble is almost symmetric about the fiber
axis, so its appearance is elliptical at the side view and
circular at the cross section. The long axis and short axis of
air hole were about 59, 34 μm respectively.

Varying arc conditions, we were also able to make a
spherical air bubble whose diameter size is about that of the
core as in Fig. 2b, using HOF with hole diameter of 4μm
with optimal electric arc conditions. Here the bubble
diameter was about 7 μm

Through the experiments, we could adjust the air-bubble
size by controlling arc duration time and arc power level.
After several trials and errors, we discovered that there are
critical arc condition for formation of air hole and
excess/lack of arc power and duration time decrease the size
of air bubble and finally collapse them. The fact that bubble
size can be tunable encouraged us to control the
transmission spectra.

Fig.3. shows transmission spectra of small size air bubble
that is as big as diameter of core, and large size air bubble
(as listed in Fig.2). According to this data, small size bubble
blocker has 8dB loss and large one has 13dB loss in short
wavelength range (1250–1450nm). Rejection
efficiency was 14dB at small one and 25dB at large
one in long wavelength range (1550–1650nm).

It is noteworthy that the transmission of the
micro-bubble mode blocker is highly wavelength
dependent, which could be applied to wavelength
selective devices that separates O band from C/L
band.

3 Conclusion
Micro air bubble in HOF core was demonstrated as
a core-mode blocker, which determine amplitude of
light transmission. Moreover, the range of blocked
wavelength is tunable according to bubble size which
can be controlled by splicing condition.

4 Acknowledgement
This work was supported in part by the
KOSEF(Program Nos. R01-2006-000-11277-0,
R15-2004-024-00000-0), and the Science and
Technological Cooperation program between and
from MOST

5 References
Vol.17, No.1
Advanced optical fiber and fiber device technologies for ultrahigh-speed optical transmission

Masataka Nakazawa
Research Institute of Electrical Communication, Tohoku University
2-1-1 Katahira, Aoba-ku, Sendai-shi, Miyagi-ken
980-8577, Japan

With the huge growth of communication traffic on the Internet, which ranges from simple data to voice, photographs, and video, it has become increasingly important to realize a high-speed and high capacity network to support the daily needs of a modern society. Ultrahigh-speed photonics and communication technologies are fundamental components of such an advanced information society.

The recent maturity of WDM technology has also proved that high-speed TDM techniques are increasingly important in terms of achieving a simple high-speed network. In fact, there has been a lot of research focusing on speeds of above 40 Gbit/s, and now 100 Gbit/s Ethernet appears to be feasible. Furthermore, it can be said that channel speeds of 160 Gb/s and above clearly reveal the advantage of the light beam as a communication medium, and has triggered the current global trends in optical communication technology as regards handling ultrashort pulse trains for high-speed signal processing and for high-capacity communication.

Optical fibers are excellent materials for high-speed optical devices because of their ultrafast relaxation time in the femtosecond regime and excellent figure of merit due to their low-loss characteristics. This tutorial reviews advanced functional optical fiber devices for ultrafast pulse generation, propagation, and signal processing, which are indispensable for ultra-high speed optical transmission. Some advanced non-telecom fiber technology is also described.
Biography
Masataka Nakazawa

After receiving his Ph.D. from the Tokyo Institute of Technology in 1980, Dr. Nakazawa joined the Electrical Communication Laboratory of Nippon Telegraph & Telephone Public Corporation. He was a visiting scientist at MIT from 1984 to 1985. From 1989 he led the High-speed Optical Transmission Research Group, and he became the first Distinguished Technical Member of NTT Laboratories in 1994 and the first NTT R&D Fellow in 1999. He was appointed a professor at the Research Institute of Electrical Communication of Tohoku University in 2001. His current interests are ultrahigh-speed optical transmission, coherent transmission, EDFAs, fiber devices, and lasers.

Dr. Nakazawa is a Fellow of the IEEE, OSA, NTT and the IEICE. Recently, he served as the president of Electronics Society of the IEICE. He has published 350 papers and holds more than 100 patents. He has received many awards including the D. E. Noble Award 2002, the R. W. Wood prize 2005, and was a Thomson Scientific Laureate in 2006.
The Optical Access Technologies and Applications in FTTH

(Invited Paper)
Yuefeng Ji (jyf@bupt.edu.cn), Yongmei Sun (ymsun@bupt.edu.cn), Rikun Liao (lyfen@sina.com)
(Key Laboratory of Optical Communication and Lightwave Technologies, Ministry of Education, Beijing University of Posts and Telecommunications, P. R. China)

Abstract
The paper presents the current R&D activities and prospects of optical access networks in China. Main technologies, applications and trends in FTTH are analyzed from the technical and market viewpoint.

1 Introduction
The rapid growth of broadband access subscriber and network traffic leads to urgent requirement for bandwidth in access network, therefore optical access technologies have been widely studied and developed in recent years. As shown in Fig.1, the number of China's broadband subscribes has already reached 137,000,000 in 2006 which 23.4% higher than that of in 2005, and it is expected to increase with the growth rate of more 15% per year next few years [1]. To fulfill the growing user’s bandwidth requirements, R&D activities on broadband access technologies have been greatly promoted. Among them, FTTH (Fiber to the Home) has been drawing more attention of both research and industry.

Fig.1. The number of China broadband subscribers

This paper will discuss the current R&D activities in China and analyzes key technologies in FTTH, including passive optical network (PON) and point to point (P2P). The FTTH related field test, applications and trends in China are also given.

2 Challenges and motivation of China’s FTTH
At present, the challenges of FTTH mainly come from the relatively high cost and the some regulation. But the service demand and the end users’ needs, the biggest drive factors of FTTH deployment, have greatly motivated FTTH development. More and more consumers have the need for lager and more stable broadband as broadband services like P2P TV, IPTV, online game, VOIP, VPN, etc..

Therefore, promoted by the increasing applications and bandwidth requirements, China has made a great progress in FTTH with the effort from research, industry and government in the last few years. Fig.2 shows China’s FTTH roadmap for R&D, field trial, and commercial deployment.

In China, some National Programs, such as O-Time (Optical Technology for IP with Multi-wavelength Environment) program and 3T-net (Tbps-level router, Tbps-level optical transmission system and Tbps-level switching capacity equipment) project, have made great contributions to FTTx. As a part of National 863 Program in Tenth Five-Year Plan, one target of O-Time (2001-2005) is to solve the high efficiency multi-service access problem for the next generation Internet, with main focus on EPON. Some key problems have been studied, such as dynamic bandwidth allocation, related standardization, and construction of application system. The research has been put to use in more than 20 cities and areas. As for 3T-net project (2002-2006), it has been built to form a large-scale interactive TV testing network, which is oriented towards large-scale concurrent
DTV/HDTV broadband stream video and Peer to Peer service application.

3 Key technologies in FTTH

While the service drives the new technology emerged, the technologies are also rely on the users’ up-to-date needs. Hereinto, PON and P2P are widely studied and implemented which can best satisfy the multi-services demand in FTTH. For example, EPON or GPON is a high-speed platform fully compatible existing standard and represents the development direction.

Table 1 gives the comparison between xPON and P2P which are complimentary rather than conflicting. At present, among various xPON technologies, EPON is mainly used for FTTH in high population density area, while GPON in corporations that need a large number of TDM leased line services. P2P can be used for both business cases and residential cases. For business case, P2P will be used when the required bandwidth is more than that a PON can provide. For residential case, P2P is preferred when the residential homes are scarcely located. Until now, most local vendors focus on GEPON system. GPON system is under consideration by some vendors. WDMPON and OCDMA are still under research or experiment.

Table 1. xPON vs. P2P access solutions

<table>
<thead>
<tr>
<th></th>
<th>xPON</th>
<th>P2P</th>
</tr>
</thead>
<tbody>
<tr>
<td>Subscribers</td>
<td>closely located</td>
<td>sparsely located</td>
</tr>
<tr>
<td>Central office</td>
<td>limited</td>
<td>enough</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>low</td>
<td>high</td>
</tr>
<tr>
<td>Solutions</td>
<td>B/E/GPON</td>
<td>P2P SDH/Ethernet</td>
</tr>
</tbody>
</table>

From the viewpoint of market, EPON can achieve more economical access while GPON fits for business circumstance such as VoIP, IPTV and HDTV. Since the cost of fiber, Ethernet card, passive optical component, etc, keeps dropping, FTTH will be economically acceptable in the coming year.

4 The applications and trends in FTTH

China’s 137 million broadband users are potential FTTH subscribers because they have experienced what broadband can bring to them. In the commercial deployment of FTTH, GEPON is widely used and triple-play can be provided as illustrated in Fig.3.

Although the applications of online music/films/TV, Internet games are developed widely, HDTV, P2P and the coming tele-presence services are undoubtedly considered to be the killer applications of FTTH, because they need much more bandwidth than existing ones.

Fig.3. FTTH multi-services based on GEPON

Due to the emerging of various broadband applications, FTTH becomes a hotspot in government, universities and enterprises. In China, by the end of 2006, there are about 60K commercial FTTH subscribers and another 20K in field trials. The total number is expected to reach 160K in 2007 [2]. Beijing, Wuhan, Shanghai are major cities with FTTH deployment, and more than 20 cities are involved. The biggest FTTH network has about 1000 subscribers. Almost all the deployments now are using GEPON.

5 Conclusion

Broadband access and service-driven with dynamic flexibility are typical characteristics and developing trend of future optical access network. Facing the high-bandwidth-required services and applications emerged in endlessly, FTTH will definitely be the most promising solution and will be dominating in future access market.

Acknowledgments

This research was supported by the National Science Fund for Distinguished Young Scholars (No.60325104), NSFC(No.60572021),863 Program(No.2006AA01Z243), the PCSIRT Project of MOE(No.IRT0609), International Cooperation Project of MOST(2006DFA11040), P. R. China.

References

Technologies for FTTH system in Korea
Euy Don Park, Kwanhee Han, and Woon Jin Jung
FTTH team, Seed Business Group, LS cable
19F, ASEM Tower, 159, Samsung-dong, Gangnam-gu, Seoul 135-090, Korea
(phone) +82-2-2189-9614, (fax) +82-2-2189-9259, (email) edpark@lscable.com

Abstract
We review the current FTTH technologies in Korea and report the relevant technical progresses at LS cable.

1. Introduction
The optical access network has been introduced in the 1980s, developed through the 1990s, and commercially deployed in real field. For example, in Japan, it has been reported that the total number of FTTH subscribers exceeded 2.5M in 2006. In US, there have been numerous announcements of FTTH deployments by both municipalities and operating companies including RBOCs. The home connection via FTTH technology became more than 2.65M in 2006. In Europe, currently 0.65M subscribers were reported, 80% of which reside in Italy, Sweden, Denmark. From MIC, Korea has more than 3.3M subscribers for FTTB solution and 0.12M subscriber for FTTH solution. The replacement trend from DSL to FTTH is rapidly spreading all over the world. From the forecast of heavy reading, the FTTH subscriber will reach up to 76M by 2011. In this paper, we review the current issues in FTTH system and report the relevant technical progresses achieved at LS cable.

2. FTTH technology in Korea
In Korea, various kind of access network technology has been proposed and deployed. Since 60% of household live multi-dwelling unit such as apartment and the 95% of subscriber located within 4km from central office, digital subscriber line (DSL) was dominant technology. However, the market share (M/S) of DSL was decreasing from 54% in Dec. 2005 to 39% for last 1 year. One of challenging technology is Apt. LAN, one of popular FTTB solution, the M/S of which was reached up to 23%. This is because it is simple and cost-effective solution especially for Korean apartment environment. Conceptually it has double star architecture so that low-cost layer-2 switch with 100Base-Tx interface is installed at underground room of apartment and UTP cable is deployed from switch to each home directly.

Recently several FTTH solutions are commercially available. First one is active optical network (AON) which uses the Layer-2 switch with 100Base-Fx interface. It has similar architecture with Apt. LAN. As Korean government announced the FTTH Emblem mark for promotion of FTTH, it was started to deploy from 2004 due to its cost-effectiveness and technology maturity. From plan of KT, the estimated number of AON subscriber reaches up to 40K.

The most popular technology in Korean market is Gigabit Ethernet passive optical network (GEPON), which connecting from optical line terminal (OLT) to multiple optical network terminals (ONT) through the passive splitter located at manhole or electric pole. From KT, the number of GEPON subscriber will reach up to 1M subscribers by end of 2007. However, as the bandwidth requirement for triple play service increase, the evolution of network technology is essential for the future.

One of emerging technology is Gigabit-capable PON (GPON) and another is WDM PON which is beginning to attract significant attention in Korea. Even though WDM PON has long been considered as an ultimate solution for the access networks due to its large capacity, easy management, network security, and upgradeability, but it has the critical bottleneck due to high cost. In case of GPON, it has lower cost than WDM-PON and several advantages in comparison of GEPON. For example, it has twice bandwidth and better transmission efficiency. It also supports more than 64 splitting ratio and advanced security. In addition, GPON system can support....
various services including broadband internet, voice over IP (VoIP), IPTV and leased line service using a single platform. Fig. 1 shows the detailed service architecture using GPON system. It can also support conventional broadcasting service by using RF overlay technology. Therefore GPON will be major technology over the world in the near future.

3. GPON system

Fig. 1 shows the general network architecture using GPON system. It is composed of the active equipment such as OLT and several ONTs and ONUs and the outside plant (OSP) component including the optical cable and splitter located in remote node. The system is fully compatible with ITU-T G.984 standard and supports the triple play service over single platform. The transmission speed is 2.5Gbps and 1.25Gbps for downstream and upstream, respectively, and the maximum splitting ratio is 64. The maximum distance is 60km logically but physically about 20km when the class B+ optical transceiver was used.

Table 1 shows the detailed specification of OLT system. It has open architecture with modular design. When the PON card was fully employed, one OLT support more than 48 PON port. It means that it can support more than 3072 subscribers simultaneously when the splitting ratio at RN is 64. To support full bandwidth for each subscriber, it supports maximum 320Gbps switch capacity and 48Gbps upstream interface. In addition, it has the redundancy system for carrier-class reliability.

<table>
<thead>
<tr>
<th>Item</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>GPON</td>
<td>• ITU-T G.984.x GPON Fully Compliant</td>
</tr>
<tr>
<td></td>
<td>• Max. 48 GPON (1:64 Split Ratio)</td>
</tr>
<tr>
<td></td>
<td>• 20km Distance at 1:64 Split Ratio</td>
</tr>
<tr>
<td></td>
<td>• DS 2.5G, 1490nm</td>
</tr>
<tr>
<td></td>
<td>• US 1.25G, 1310nm</td>
</tr>
<tr>
<td></td>
<td>• (Optional RF Video) 1550nm</td>
</tr>
<tr>
<td>IP</td>
<td>• 24 GbE Interface, 2 10G Interface</td>
</tr>
<tr>
<td></td>
<td>• L3 Protocol: Static, OSPF, RIP</td>
</tr>
<tr>
<td></td>
<td>• L2 Protocol: VLAN, STP, LACP</td>
</tr>
<tr>
<td></td>
<td>• Switch Fabric: 320Gbps, Load Sharing</td>
</tr>
<tr>
<td></td>
<td>• Multicasting: IGMP Snooping, PIM-SM/DM</td>
</tr>
<tr>
<td></td>
<td>• IPv6: IPv4/IPv6 Support at the Same Platform</td>
</tr>
<tr>
<td></td>
<td>• QoS: Marking/Remarking, Rate Limiting, Classification, IEEE802.1p/DSCP</td>
</tr>
<tr>
<td></td>
<td>• Mapping, Shaping, etc.</td>
</tr>
<tr>
<td>Console</td>
<td>• RS-232C</td>
</tr>
<tr>
<td></td>
<td>• Fast Ethernet</td>
</tr>
</tbody>
</table>

The ONU was developed for business subscriber and supports more than 24 Ethernet port or 12 POTS port or 12 E1 port. By adapting dual architecture, it can provide the PON redundancy as recommend in G.984.1. Two type of ONT was developed for indoor and outdoor application. By using these ONT, the subscriber can receive the VoIP service as well as high-speed internet.

4. OSP solution

For single dwelling unit, the optical splitters are installed at the remote node, which located at the manhole or electric pole, and the feeder cable is deployed from central office to remote node. Usually drop cable is installed from the remote node to home. For multi-dwelling unit, the optical splitter is installed under the apartment so that it is not easy to install the optical cable through the original duct. To overcome these problems, the air-blown fiber (ABF) solution was releases. It is a concept for the installation of fiber-optic networks utilizing viscous air flow to propel blown fiber unit through pre-installed blown tube cable. Blown fiber unit could be proceeding by pushing force of drive wheel and compressed air. The installation speed and blowing distance are limited by the friction force between micro duct and blown fiber unit. By optimizing the fiber structure inside of micro duct and outside coating of blown fiber unit, the maximum blowing distance could be achievable.

5. Conclusion

As the traffic requirement is growing in Korea, a variety of access solution is tried to deploy and competing each other. The Apt. LAN and GEPON solution is the prevailing technology in current situation. To provide the triple play service cost-effectively, GPON and WDM-PON will gain more attention by virtue of high bandwidth and guaranteed quality of service. For the OSP, the drop cable and ABF was introduced and widely deployed in Korea.
Optical fibres and cables for the access networks;
Standardization & FTTH development in Europe

Gerard Kuyt, Piet Matthijsse*
Draka Comteq Optical Fibre, Zwaanstraat 1, 5651 CA Eindhoven, The Netherlands,
e-mail: gerard.kuyt@draka.com

Abstract The development of ITU-T Recommendations for fibre cable used in access networks is highlighted, emphasising the new Rec. G.657 for bend insensitive single mode fibre. In addition developments of FTTH in Europe will be reported.

Introduction
The International Telecommunication Union (ITU-T) has started activities to serve broadband services in the access network, e.g. Access Network Transport Standards and Optical Systems for Fibre Access Networks. In November 2006 the first recommendation on bend insensitive single mode fibres was consented, under the designation G.657. Further work continues on application of bend loss optimized graded-multimode fibres in optical access networks for multi-tenant buildings.

New activity in ITU-T
ITU-T activities are organized in Study Groups [1]; subjects related to (cabled) optical fibres are assigned to Study Group 15 (SG15): “Optical and other transport network infrastructures”. Sub-activities are organised by means of “Questions”. The regular work on cabled optical fibre is taken care of in Question 5 of SG15.

Late 2003 a new sub-activity was created on fibre cable designed for access networks, organised in Question 107 of SG15. The motivation to start this work is the growing demand for broadband services (multimedia, high-speed internet, HDTV) in buildings and homes requiring high-capacity transmission media into the local network. Optical fibre is viewed to be an important option for any broadband media mix created for this purpose.

The first year dealt with organization of work and – avoiding duplication – gathering existing information concerning fibres and cables for access networks. In November 2004 the first technical contributions were offered [2], containing two proposals:

1). Small bending radius optimized single mode fibres (SMF). Due to the many obstacles and the restricted space available, stringent low bend radius requirements may arise, for the fibre to and in the building. In addition, access networks require increased resistance to incidental bends originating from improper fiber deployment. Although this should be prevented in general, the employment of lower skilled installation crews, the installation in various types of premises and the large scale use of patch panels in telecom offices and street cabinets will make the occurrence of incidental bends inevitable. Modern SMF preferably has to answer the requirement for an improved resistance to this maltreatment.

2). Graded-index multimode fibres, used in optimised optical access networks in multi-tenant building environments (e.g. in Europe and Asia) with a separate access switch in the basement. Due to the potentially high connection density and the short lengths of the optical distribution cable, investment cost per connection can be relatively low in this particular sub-segment of access networks.

These two proposals have been the subject of study in the following period, with G.smx as draft recommendation for bend insensitive single mode fibre and G.mmxx as draft recommendation for multimode fibre.

Due to large interest in the market (operators and industry) Rec. G.smx was already consented in November 2006, more than a half year earlier as originally planned. This official recommendation for bend insensitive fibre has been published as Rec. G.657 [3].

Rec. G.657: Bend insensitive single mode fibre
Various contributions from multiple delegations lead to a split of the G.657 single mode fibre recommendation into two categories:
Category A fibres, suitable to be used in the O, E, S, C and L-band (i.e. 1260 to 1625 nm range). These fibres are a subset of G.652.D fibres (are therefore fully backwards compatible) and have the same transmission and interconnection properties. The main improvements are in reduced bending loss and tighter dimensional specifications. Both improvements reduce connectivity losses.

Category B fibres, suitable for transmission at 1310, 1550, and 1625 nm for restricted distances that are associated with in-building transport of signals. These fibres may have different splicing and connection properties than G.652 fibres, but are capable at very low values of bend radius.

Table 1 shows the main characteristics of both categories.

<table>
<thead>
<tr>
<th>Table 1: Main characteristics of Rec. G.657A and G.657B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Attributes</td>
</tr>
<tr>
<td>Mode field diameter 1310 nm</td>
</tr>
<tr>
<td>Nominal range</td>
</tr>
<tr>
<td>Tolerance</td>
</tr>
<tr>
<td>Macrobending loss</td>
</tr>
<tr>
<td>Radius (mm)</td>
</tr>
<tr>
<td>Number of turns</td>
</tr>
<tr>
<td>Max. at 1550 nm (dB)</td>
</tr>
<tr>
<td>Max. at 1625 nm (dB)</td>
</tr>
<tr>
<td>Main transmission attributes (a) PMD / Chrom. Dispersion)</td>
</tr>
</tbody>
</table>

Lifetime of bent fibre
An important discussion in Q10/15 concerned the lifetime aspects of bent fibres. Based on common lifetime modelling of fibres stored in reduced size cassettes of fibre management systems, it was shown that the current minimum proof stress value of 0.69 GPa as recommended in G.652 suffices to maintain a guaranteed operational lifetime of 20 years [4, 5]. See also Figure 1.
FTTH development in Europe

FTTx connections:
By June 2006, 28 countries in Europe (EU-28) counted around 820,000 FTTx subscribers (+32% vs June 2005), with 2.74 millions homes passed (+13% vs June 2005) and a growing penetration rate of 29.8% in 2006 (2005: 25.6%) [6]. This level of subscribers is far below that in Asia, in particular in Japan (April 2006: 5,642,000 FTTx subscribers), but not too far behind the USA with around 4M homes passed and 1M connected mid 2006 [7].

One reason for the lower level of FTTx connections in Europe may be the relative good copper and cable-TV networks, allowing high DSL speeds, specially in Scandinavia, France and the Benelux. For example Belgium is leading this bandwidth league with 91.5% of DSL subscribers connected to at least 2 Mb/s, followed by Iceland (45.7%), Sweden (41.9%), Norway (30.7%), Netherlands (30%) and France (28.8%).

Other factors may be a lower commitment to FTTH by most incumbent operators, compared to e.g. some Asian countries. In the USA the decision of the FCC to lift the unbundling requirements for optical fibre access networks in 2003 lead to a strong impulse of investments in FTTx networks. In Europe incumbent operators criticize the present unclear regulatory framework as not being helpful in investing new FTTx networks.

96% of the EU-28 FTTx subscribers are concentrated in 5 countries: Denmark, Netherlands, Sweden, Italy and Norway, which shows the large difference between the European countries. Municipalities and power utilities still lead the build out of FTTx networks in Europe; yet, Incumbent and alternative operators are warming up for the deployment of very high speed networks.

Positive demand indicators can be mentioned: strong growth of ownership of digital camera’s (with increasing file sizes), breakthrough of large-screen HD-ready displays, new next-gen games machines and new HD-DVD formats, all pushing for increased demand of higher speeds in the access networks [8].

Technical architecture of European FTTx deployments:
In Europe still P2P Ethernet is outnumbering PON [6], also in new significant projects like in Paris (Ilad/Freet) and Amsterdam. On the other hand, PON has been selected for several major projects since 2005:
- in Spain, the government of Asturias: GPON;
- EnergiMidt, a Danish power utility: BPON;
- France Telecom since June 2006: GPON.
One of the considerations for ILECs may be the more difficult unbundling practice when using GPON.

Open Network Models are often chosen by Municipalities.

Future important FTTH deployments [6,8]:
- France Telecom deployment confirmed:
  - Deployment began in 2006 based on GPON. Plan to connect 10 major French cities by end 2008. 1 M homes passed and up to 200,000 subscribers by end 2008.
  - Investment estimated to 270 M EUR in two years.
- Free (Ilad) FTTH deployment launched in France:
  - Commercial launch of fibre-to-the-home (FTTH) local loop network in Paris during the first half of 2007 based on Ethernet. 4 M home passed through 2012 (15% of penetration at least expected). Plans on investing 1 Billion EUR in the project through 2012.
  - The Amsterdam Municipality CityNet: Based on an Ethernet access network; deployment started in Oct. 2006. 25,000 homes connected by the end of 2007; 450,000 total by 2013.
  - Dong Energy (Denmark): 900 000 homes.
  - Vienna (Austria): 900 000 by 2011.
  - Reykjavik (Iceland): 200 000 by 2011.

And FTTN+VDSL...:
- Ericsson, network, in Ireland: Work started to upgrade national telecommunication network with FTTC equipment.
  - 500,000 homes to be connected by the end of 2007. Ericsson plans to invest 1 Billion EUR to complete this build out.
  - Deutsche Telekom has planned to cover 2.9 millions homes in 10 major cities in Germany with FTTN and VDSL2 (end of 2006).
  - KPN and Swisscom will also deployed widely VDSL platforms in 2007.

Conclusions
Preceded by its use in Japan, the bend-insensitive single mode fibre proofs its added value worldwide in the labour intensive and scarcely spaced access network environment. The ITU-T has acknowledged this and developed its new G.657 Recommendation.
In many European FttH plans, with progress hampered by the lack of a sufficiently clear regulatory framework, the use of G.657 compliant single mode fibres is expected to offer great added value for the many FttH networks still to be constructed.

References
[1] G. Bonaventura et al., 54th IWCS
11C3-4 (Invited)

Development of optical fiber cable and wiring techniques in full-scale FTTH era

Mitsuo KAMA, Kazuyuki SHIRAKI, Hiroshi AOYAMA, Osamu INOUYE, and Shigekazu MATSUI

NTT Access Network Service Systems Laboratories, NTT Corporation
1-7-1 Hanabatake, Tsukuba, Ibaraki, JAPAN

Abstract
This paper describes the current state of broadband services in Japan and the conditions required for optical wiring techniques in the access network as we approach the full-scale FTTH era. Furthermore, key techniques and recent development examples for mass construction are introduced.

1. Trend of broadband services
Recently, the “full-scale FTTH era” has already arrived in Japan with the commercial introduction and spread of optical services.
As we see from the recent state of broadband services in Japan, the net increase in the number of broadband subscribers per year is approaching 3 million, while ADSL subscribers have started to decrease since the first quarter of 2006. (Figure 1) Therefore, the FTTH service is clearly becoming a leading component of Internet provision.

![Figure 1. Transition of broadband subscribers in Japan](image)

The spread rate of FTTH services to households now exceeds 15% in urban areas. Even in rural areas, the rate is between 2.5% and 5% as efforts are made to eliminate the digital divide. Therefore, it is predicted that the spread rate will continue to grow rapidly not just in urban areas but also throughout the whole of Japan.

When we look at the state of broadband services worldwide, we see that many countries are promoting their introduction. In terms of FTTH services, Japan is becoming the world’s No. 1 in service charge and number of subscribers. Therefore our FTTH experiences and technologies could assist other countries.

The “u-Japan” strategy was proposed as a national policy in 2006. The main targets to be achieved by 2010 are shown below.
1. Total broadband coverage
2. More than 90% of households equipped with broadband

And NTT plans for a further acceleration of the expansion of FTTH services and aims to increase the subscriber number to 30 million by fiscal 2010

2. Requirements for cable wiring and R&D View
The conditions required for full-scale FTTH service deployment are regulated based on the situation in Japan. Previously, the main target was to reduce the cost of service construction for a fixed level of demand. By contrast, as we move towards the full-scale FTTH era, we need both cost reduction and a quick response to demands for construction. With regard to access network development, we need to construct facilities with long-term durability and develop techniques that enable economical and efficient large-scale construction.

Three examples are provided below.

(1) Quality assurance
This is the basis of all services. Although problems are commonplace, their affect in large-scale facilities is very serious and we must prepare for many predictable issues at the construction stage. Facilities must offer long-term durability to such environmental factors as wind pressure, vibration, temperature change, the ultraviolet rays of the sun, and the harmful effects of birds and insects. A technique for avoiding human errors in construction is also needed. We have to follow previous specifications and handle new tasks.

(2) Mass construction
Facilities and techniques are required for unprecedented large-scale construction and maintenance. The amount of construction will have doubled by 2010, and the number of service changes and amount of re-construction will increase corresponding to the number of facilities.
It is difficult to achieve the above target with present methods because human resources themselves are limited. Simple and efficient work is needed that reduces the total cost. This will also contribute to a reduction of construction time and improved customer satisfaction.

(3) Flexible network
To correspond to the provision of large-scale services, the total access network from the central office to user equipment must be efficient as regards investment and maintenance. The facilities must provide a flexible response to service demand with the fewest facilities

3. Key techniques and recent developments
As specific approaches to meet the above requirements, an easy construction technique that does not require skilled workers, an easy post-construction operation technique, and a flexible response network technique have been developed. The use of connectors and the elimination of bare fiber storage are effective ways to achieve these goals. Some examples are introduced here.

(1) House wiring technique in user houses
With conventional house wiring, a hard indoor optical fiber is installed and bare fiber must be spliced and stored at the lead-in part and connecting part in an ONU. This takes considerable time to accomplish. So, a free bending optical fiber cord has been developed that can be handled like copper cable. Figure 2 shows the application of free bending optical fiber cord. By providing a connector point at an optical rosette, easy connection is achieved and this is effective for mass construction. It is also allows the ONU made smaller by eliminating the need for bare fiber storage, reducing the fiber breakage that sometimes occurs in house wiring, and eliminating unattractive fixed wiring.

Figure 2. Application of free bending optical fiber cord

(2) Aerial optical closure technique
Conventionally, new construction and fiber transfer always required the splicing and storage of bare fiber. This was complicated skilled work that took much time. To simplify this, we have developed stub connectors and a splitter module. This means that fiber storage work is no longer needed, and it is sufficient to assemble connectors and then connect them. The use of a cable sheath holding stub connector facilitates post-construction transfer work in a closure.

Figure 3. New aerial optical closure

(3) Aerial wiring technique
Conventional cable has a simple structure to keep costs low, so optical loss increases when fibers are extracted. As a result, all drop closures must be set up at the same time as the cable is installed. Therefore, an 8-fiber distribution cable and a corresponding aerial optical closure have been developed. This cable enables optical fibers to be easily extracted with less loss change while maintaining a simple structure. Therefore, it is possible to set up a closure according to service demand. Connection work is simplified by using connectors.

Figure 4. DF cable and AOF closure

(4) Underground wiring technique
We are developing a multiple cable installation technique as an effective way of using existing conduits. Conventionally, a maximum of 2000 fibers can be accommodated with a straight-line distance limitation in a φ 75 mm conduit. Figure 5 shows multiple cable installation in a conduit by using a small diameter inner pipe with a thinner wall. The adoption of a small diameter and a thinner wall is expected to reduce the required pulling force. And this will lead to a relaxation of the distance limitation and the ability to accommodate up to 3000 fibers. This approach controls investment without the need for civil construction and also shortens the construction time.

Figure 5. Multiple cable installation in a conduit

4. Conclusion
The broadband access market for mass users is expanding and starting to shift from ADSL to FTTH. If we are to achieve our targets by 2010, it is very important that we realize prompt installation, and establish easy operating techniques for optical access networks. The keys are efficient construction and cost reduction.
InGaAs/AlAs/AlAsSb intersubband transition all-optical switch for ultrafast all-optical signal processing

Hiroshi Ishikawa, Hidemi Tsuchida, Takasi Simoyama, Cheng Guan Lim, Masanori Nagase, Teruo Mozume, Ryoichi Akimoto and Toshifumi Hasama
Ultrafast Photonic Devices Laboratory,
National Institute of Advanced Industrial Science and Technology (AIST)
Tsukuba-Central 2-1, 1-1-1 Umezono, Tsukuba, Ibaraki, 305-8568 Japan
TEL: +81-298-3202, FAX: +81-298-5255, email: hiroshi-ishikawa@aist.go.jp

Deep phase modulation on TE probe light was observed in InGaAs/AlAs/AlAsSb intersubband transition (ISBT) switch when pumped by TM pulse. This enables us to realize ultrafast and low-loss all-optical signal processing devices.

So far, ISBT all-optical switch was developed as intensity modulation switches, where the change in transmittance by absorption saturation was used [1]. However, this caused large insertion loss. Recently we found a new all-optical phase modulation in InGaAs/AlAs/AlAsSb ISBT switch [2][3]. The TE probe light is phase modulated when the device is pumped by TM gate pulse. As the intersubband transition takes place only for TM mode, the probe TE mode does not suffer the absorption loss. This enables us realize low-insertion loss devices for ultrafast all-optical signal processing. Here, we review this phase modulation effect and its novel mechanism, and then discuss possible application of this new effect for signal processing.

Figure 1 shows the phase modulation spectrum of the TE probe light when pumped by 10Gb/s-repetition TM pulses with pulse width of 2.6 ps. We used ISBT switch module with pigtailed fibers. The pump energy to the fiber was 2.07 pJ. We can confirm fairly deep phase modulation. The amount of the phase shift is evaluated by converting the phase modulation to intensity modulation using interferometer with differential group delay (DGD) of 25 ps. The evaluated phase shift is shown in Fig. 2. The largest phase shift was 1.88 rad for the fiber input energy of 4 pJ. The response speed also was evaluated to be of the order of pico-second. The change in refractive index causing the phase modulation was confirmed to be positive side by the interference fringe measurement.

According to the Kramers-Kronig relation, the change in absorption coefficient for TM light by TM pumping does not cause the refractive...
index change for TE light. So, to explain the observed phenomena, new model have to be set up. We have shown that the refractive index change is due to the change in plasma dispersion associated with the effective mass difference among subbands [3]. Figure 3 shows the calculated refractive index change for 1ps full-width at half-maximum TM pulse pumping. Pumping intensity was adjusted to give 50% reduction in the absorption coefficient. Because of the coupled-double quantum well used in our device, there are four subbands. The plasma dispersion at the first and second subband increases the refractive index because of the electron depletion. While that of the third and fourth subband reduces the refractive index because of the increased electron density. If the effective masses at each subband are equal, there arises no refractive index change. However, as the effective masses at each subband are different, there arises positive refractive index change when we sum up the effect of all subbands.

The newly observed phase shift and the new mechanism are highly attractive for signal processing. Because of high speed, low insertion loss and wide spectral range for the probe (signal) light. Figure 4 shows the example of the application to wavelength conversion [2]. In this experiment, TE-CW light (1559.90 nm) was phase modulated by 10Gb/s TM pump signal (1550.09 nm) with 2.6ps pulse width. The phase modulation was converted into the intensity modulation by interferometer with DGD of 25ps. Because of DGD we obtain two pulses. The two pulses were combined to one pulse with wider width using band pass filter. Then error rate was evaluated. As can be seen form Fig.4, error free wavelength conversion is achieved.

As demonstrated by wavelength conversion experiment, this effect is quite useful for ultrafast signal processing [2]. For example, intensity modulation switch can be realized by putting the device at one arm of the Mach-Zehnder interferometer. The device can also be applied to the conversion from on-off keying to phase shift keying. Optimization of device design for larger index change with smaller pump energy will realize highly attractive ultrafast signal processing devices.

References


Abstract
We have successfully demonstrated the wavelength switching in the straight arrayed waveguide with linearly varying refractive index distribution by changing the refractive index using thermo-optic effect. We have obtained the wavelength demultiplexing and its changing of the output ports by refractive index change in the waveguide.

Introduction
In WDM networks, key components are the wavelength multi-/demultiplexers, which combine/separate wavelength channels. Various kind of demultiplexing principles have been proposed and already commercially available, such as arrayed-waveguide gratings (AWG). In the latest paper, a novel wavelength demultiplexer using a GaInAs/InP multiple quantum well (MQW) arrayed waveguide has been proposed in which refractive index varies linearly across the array [1,2]. The conventional AWGs are designed so that phase differences between adjacent waveguides are obtained by gradually varying waveguide length. In the proposed design, however, phase differences between adjacent waveguides are achieved by varying the waveguide thickness which is the refractive indices of the waveguides. Therefore, it is possible to be realized a straight waveguide type wavelength demultiplexer [2,5]. Furthermore, the device could be applied to wavelength switching since the refractive indices of the waveguides in the array can be controlled dynamically by QCSE or TO effect. In this report, we show the successful wavelength switching in the arrayed waveguides with linearly varying refractive index distribution by controlling the refractive index using TO effect.

Device Structure
Fig. 1 shows the schematic design of the wavelength switch. The arrayed waveguide region was fabricated by selective metal-organic vapor phase epitaxy (MOVPE) growth using the asymmetric SiO$_2$ mask pattern on both sides of the array [3,4].

The number of arrayed waveguides was $N=16$, waveguide width $w=3$ μm and waveguide spacing was narrowed from $d=3$ μm to 2 μm toward the star coupler, and the length of arrayed waveguides were 10 mm. Four output waveguides spaced by 6 μm were located at the end of output star coupler. In the designed star coupler, incident lights are distributed to array with equal phase front and output lights from array are focused to the output waveguides based on Rowland circle geometry. All the arrayed waveguide between input and output star couplers has equal length though the straight arrayed waveguides and the star couplers were joined with curved waveguides. Additionally, the positions of the input and output star couplers, which are not placed on straight axis, are effective to reduce the recoupling of the scattered light. TO control the refractive index in the...
waveguides, Ti/Au electrode was placed on the center of the arrayed waveguide region with 1000 μm length.

Wavelength Switching

The successful wavelength switching was demonstrated in the device mentioned above. Fig. 2 shows the three port switching characteristics of the $\lambda=1540$ nm wavelength light which is the dependence of output intensity on electric power. When there was no input electric power, the light was outputted from Port 2. By increasing the input electric power, the output port was changed to Port 3 and Port 4. The extinction ratio was 8.7 dB at Port 2, 6.2 dB at Port 3, and 17.3 dB at Port 4. From these results, the estimated refractive index change in the waveguide was 0.13 % for one-port switching.

Fig. 2. Three port switching characteristics of the $\lambda=1540$ nm wavelength light. The dependence of output intensity on electric power.

Further development of electric-field control of refractive index would also be expected to yield high speed wavelength switching.

Fig. 3. Switching characteristics of other wavelength light of (a) $\lambda=1521$ nm and (b) $\lambda=1562$ nm

Conclusion

We have demonstrated the wavelength switching in the arrayed waveguide with linearly varying refractive index distribution. We have obtained the three port switching by the refractive index change using the thermo-optic effect.

References

Abstract—An active integrated optical bandpass filter consisting of two parallel waveguides including two gratings with slanted grating lines is presented. The filter concept enables the combination of filtering and amplification within one device.

I. INTRODUCTION

In DWDM systems optical filters are required to access individual channels. In this paper an active integrated bandpass filter based on the InP/InGaAsP system is presented formed by two parallel 2D-waveguides upon a slab waveguide. The 2D-waveguides are coupled by two gratings within the 2D-waveguides with slanted grating lines. The multichannel signal initially guided in the input waveguide is converted to a wave in the slab waveguide by the first grating. This wave is incident on the second grating. Coupling of this wave to a guided wave of the output waveguide is strongly wavelength sensitive. By properly choosing the grating slant angles a bandpass filter is obtained.

II. DEVICE STRUCTURE

A schematic view of the device is shown in Fig. 1. The 3-layer slab waveguide consists of an n-doped InP substrate, an n-doped InGaAsP film and a p-doped InGaAsP cladding. The cores of the two 2D-waveguides are the two undoped stripes of InGaAsP on top of the film, which are the active regions. The waveguides are designed to support only the fundamental mode. Optical gain is provided by forward biasing of the two heterostructures. The material compositions have been chosen to ensure waveguiding and carrier confinement. Below the stripes are the slanted gratings. The centre wavelength of the device is determined by the grating period $L_g$. Each 2D-waveguide together with the respective grating forms a Distributed Bragg Deflector (DBD) [1] which converts a guided mode of the 2D-waveguide to a guided mode of the slab waveguide.

III. FREQUENCY RESPONSE

The dielectric profile $\varepsilon_0\varepsilon_r(\mathbf{r})$ of the 2D-waveguide is perturbed by the grating. Consequently, the incident fundamental mode $E_{fm}$, which is the quasi-TM mode, induces a perturbation polarization in the grating lines that excites the field $E^s$ scattered by the grating. For the calculation of the scattered field it is assumed as in [1], that the field within the waveguide in the grating region

$$E^s(r) = E_{fm}(y,z)e^{j\varphi(y,z)/2-j\beta_{fm}x}$$

is given by the field of the unperturbed waveguide attenuated by the grating. The net gain $\gamma_M$ is the effective gain of the unperturbed waveguide reduced by $\alpha_G$, the attenuation due to the grating. The perturbation polarization is then given by $\delta\varepsilon E^s$ where $\delta\varepsilon$ is the difference between the dielectric profile of the perturbed and the unperturbed waveguide. The scattered field excited by the perturbation polarization can be calculated from the Hertzian potential which is obtained with the aid of the Hertzian potential Green’s dyad $G(r,r')$ [2] to ($y_g$ is the grating volume)

$$\Pi_e(r) = \int_{y_g} G(r,r') \frac{\delta\varepsilon(r')}{\varepsilon_0\varepsilon_r(r')} E^s(r')dy'$$

The scattered field is obtained from $E^s = (\nabla \times + j\omega \varepsilon_r(\mathbf{r}))\Pi_e$. Only the fraction of the scattered field guided in the slab waveguide affects the output waveguide and takes part in the filter process. This fraction of the scattered field has been calculated using the Hertzian potential Green’s dyad of the slab waveguide. The slant angle $\psi$ is adjusted such that a single grating line radiates in the slab waveguide in the $y$-direction. Then, for a large number of grating lines $N$, the total field in the interspace between the two waveguides forms a TM-polarized mode of the slab waveguide. The angle between the direction of propagation of the wave in the slab waveguide and the $y$-direction is given by

$$\tan \gamma = \frac{2\pi - \beta_{fm}L_g}{\beta_M L_g}$$

where $\beta_M$ is propagation constant of the mode in the slab waveguide. The dielectric profile of the slab waveguide is perturbed by the core of the output waveguide and the respective grating. Accordingly, a perturbation polarization $P^s$ is induced in the volume of the perturbation, that is determined by the product of the incident field in the slab waveguide and the difference between the dielectric profile of the perturbed and the unperturbed slab waveguide. This polarization acts as a source in the output waveguide which

Fig. 1: (a) Top view (b) cross sectional view
excites the quasi-TM mode with the amplitude $A$. Since the amplitude of the mode in the input waveguide has been chosen to 1, $A$ equals the frequency response and $|A|^2$ is the power transfer function. For the use as bandpass filter, coupling to the output waveguide has to be maximized at centre wavelength. Coupling can be achieved either to the mode propagating in the $-x$- or in the $+x$-direction. In each case a different angle $\vartheta$ is obtained. The amplitude of the mode propagating in the $-x$-direction is obtained from [3]

$$A = -\int_{V_{g2}} j\omega P^s(r) E_m(y,z) e^{j(g_N/2 - j\beta_{fm}) z} dV$$

$$\left/ \int_{CS} E_m(y,z) \times H_m(y,z) e_j dy dz \right. \frac{2}{(4)}$$

where $V_{g2}$ is the volume of the second grating and $CS$ is the waveguide cross section. The necessary grating slant angle is obtained to

$$\vartheta = \pi - \arctan \frac{\beta_{fm}}{\beta_sl} \frac{}{}$$

Finally, one obtains the modal amplitude to

$$A = Ce^{j(\frac{2\pi}{\lambda} - j\beta N_m)(N+1)L_g} \frac{\sin((\beta_{fm} + j\frac{2\pi}{\lambda} - j\beta N_m)N L_g)}{\sin((\beta_{fm} + j\frac{2\pi}{\lambda})L_g)} \frac{}{}$$

The factor $C$ can be treated as wavelength independent in the wavelength region of interest. The proper bias point of the device is when $g_N = 0$. In this case a linear phase filter is obtained and (6) becomes

$$A(f) = Ce^{-j2\pi f_m (N-1)} \frac{\sin(2\pi f_m N)}{\sin(2\pi f_m)} \frac{}{}$$

where $\beta_{fm} L_g = 2\pi f_m$ has been used. The full width of half maximum (FWHM) of the power transfer function is approximately given by

$$FWHM = 2.783 \frac{\lambda}{2\pi N f_m} \frac{}{}$$

and is inversely proportional to the number of grating lines $N$. The device gain at centre wavelength is obtained to

$$|A(f_m)|^2 = |C|^2 N^2 \frac{}{}$$

IV. NUMERICAL EXAMPLE

A filter with a centre wavelength of 1.55 $\mu m$ and a FWHM of 0.2$\mu m$ has been designed. Each grating consists of 3427 grating lines. The dimensions of the waveguides and the gratings as well as the dielectric profile at 1.55$\mu m$ are summarized in Fig. 1. The calculated frequency response is shown in Fig. 2. A device gain of 12.4$\mu m$ at centre wavelength has been obtained and a side-lobe suppression of 13.2$\mu m$. For a fixed waveguide and grating geometry, the only design parameter is $N$. The FWHM and the device gain as a function of $N$ is shown in Fig. 3.

Fig. 2: Calculated frequency response

Fig. 3: FWHM and gain at centre wavelength

V. CONCLUSION

A bandpass filter consisting of two 2D-waveguides coupled by two slanted gratings has been presented. In contrast to filters based upon the coupling of forward and backward propagating modes by a grating with grating lines perpendicular to the waveguide axis, for example Fibre Bragg Gratings, coupling occurs only between forward propagating modes. One advantage of this approach is the spatial separation of the device input and output. A circulator to separate the input signal from the reflected signal as in the case of the Fibre Bragg Grating is not necessary. Due to the coupling between forward propagating modes only, a linear phase filter can be obtained and no feedback path exists. The absence of a feedback path allows the introduction of optical gain and the combination of filtering and amplification of a channel.

REFERENCES

Analysis and Design of Wavelength Selective Switches Based on MMI Assisted Microring Resonators

Thanh Trung Le and Laurence W. Cahill
Department of Electronic Engineering, La Trobe University, Melbourne, Vic 3086, Australia
E-mail: t118le@students.latrobe.edu.au; Phone: +61 3 9479 3731

Abstract—The paper proposes a novel structure of wavelength selective electro-optic switches based on MMI assisted microring resonators (MRs). The analysis and design are based on the scattering matrix theory. The device parameters are to be considered and optimized to obtain the best performance with a high FSR and a low switching power.

I. INTRODUCTION

An optical switch is a basic component in optical communication systems. Many types of optical switches based on various mechanisms and materials have been designed and fabricated in recent years. Of these, the optical switch based on microring resonators is one of the most promising devices due to its advantages of high wavelength selectivity, large cavity enhancement of the electric field, and small size [1-2].

Moreover, Multimode Interference (MMI) devices have been a growing interest in compact photonic circuits due to their advantages, such as compact size, low loss, stable splitting ratio, low cross-talk and imbalance, insensitivity to polarization, ease of production and good fabrication tolerance [3]. Therefore, in this paper, we would like to propose analysis and design approaches to an optical switching structure based on MMI assisted microring resonators. The device parameters are then optimized to have the desired characteristics.

II. ANALYSIS AND DESIGN

Fig. 1 shows the configuration of the proposed optical switch. It consists of two MMI couplers along with a racetrack resonator. When a voltage is applied to the electrode on the top of the outer microring resonator, the switching is obtained.

In Fig. 1, the radii of the two rings are \( R_1 \) and \( R_2 \), and the coupling coefficients of the two couplers are \( \kappa_1 \) and \( \kappa_2 \). The relationship between the complex field amplitudes of the input and output ports of the couplers can be expressed by

\[
\begin{align*}
\begin{bmatrix} h_1 \\ b_2 \\ \end{bmatrix} &= \begin{bmatrix} \tau_1 & \kappa_1 \\ -\kappa_1^* & \tau_1 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \\
\begin{bmatrix} d_2 \\ c_1 \\ \end{bmatrix} &= \begin{bmatrix} \tau_2 & \kappa_2 \\ -\kappa_2^* & \tau_2 \end{bmatrix} \begin{bmatrix} c_2 \\ c_1 \end{bmatrix}
\end{align*}
\]

where \( \tau_1, \tau_2 (|\kappa_i|^2 + |\tau_i|^2 = 1, i = 1, 2) \) are the transmission coefficients of the couplers. The couplers are assumed to be lossless, for simplicity. If the couplers are 3dB MMI couplers, the transmission coefficients can be expressed by

\[
\begin{align*}
\tau_1 &= \tau_2 = 1/\sqrt{2} \quad \text{[4-5]}. \quad a_i, \quad b_i, \quad c_i \quad \text{and} \quad d_i \quad (i = 1, 2)
\end{align*}
\]

are complex field amplitudes satisfying the relations:

\[
\begin{align*}
c_1 &= \alpha_2 \exp(j\phi_2) b_1 \\
c_2 &= \alpha_1 \exp(j\phi_1) b_2 \\
\end{align*}
\]

where \( \alpha_1, \alpha_2 \) are notations calculated by

\[
\begin{align*}
\alpha_1 &= \exp(-\alpha_0 2\pi R_1) \\
\alpha_2 &= \exp(-\alpha_0 2\pi R_2) \\
\end{align*}
\]

\( \alpha_0 \) (dB/cm) is the loss coefficient in the core of the optical waveguides. It is assumed that the optical waveguides support only one mode which is the fundamental one, then \( \phi_1 = \beta_1 L_1 \) and \( \phi_2 = \beta_2 L_2 + \pi V / V_x \) are the phases accumulated over the waveguides with propagation constants \( \beta_i \), where \( \beta_1 = 2\pi n_{eff} / \lambda \), \( L_i = \pi R_i \), \( n_{eff} \) and \( \lambda \) are effective refractive index of the waveguide core and optical wavelength, respectively. \( V \) is the voltage applied to the electrode on top of the outer microring, and \( V_x \) is the voltage causing a differential phase shift \( \Delta \phi = \pi \) [6].

Combining these relationships results in expressions for the transmitted optical intensity \( T \) as follows:

\[
T = \left| \frac{d_1}{a_1} \right|^2 = \left| y_1 + y_2 \right|^2
\]

where \( x_1, x_2, y_1, y_2 \), and \( y \) are notations calculated by

\[
\begin{align*}
x_1 &= \alpha_2^* \tau_2 \exp(j\phi_2) \\
x_2 &= \alpha_1 \alpha_2^* \kappa_2 \exp(j\phi_1) \\
y_1 &= \alpha_2 \kappa_2^* \tau_1 \exp(j\phi_1) + \alpha_2 \tau_2 \kappa_1 \exp(j\phi_2) \\
y_2 &= -\alpha_1 \kappa_2 \tau_1 \exp(j\phi_1) + \alpha_2 \tau_2 \kappa_1 \exp(j\phi_2) \\
y &= -\alpha_1 \kappa_1 + x_1 \tau_1 \\
   &\quad + x_2 \tau_1 \left( 1 - (x_1 \tau_2 + x_2 \kappa_1) \right)
\end{align*}
\]

It is shown in the next section that resonant wavelength and extinction ratio can be tuned by tuning the applied voltage. By using this characteristic, one can realize switching functions using the proposed structure.

III. SIMULATION RESULTS AND DISCUSSION

In this section, the performance of the proposed switch with the assumption that the device is designed on the polymer material is considered. The transmission characteristics of proposed switching structure are shown in Fig. 2, in which the parameters used in the simulations are as follows: radii of the
racetrack resonators are $R_1 = 25\mu m$, $R_2 = 50\mu m$, optical waveguides have a width of 2 $\mu m$ and thickness of 1.8 $\mu m$ to support only the fundamental mode and have an effective index of 1.57 calculated by the effective index method. It is assumed that the couplers used in these simulations are 3dB MMI couplers. The MMI size is $8 \times 90\mu m^2$ to obtain a 3dB MMI coupler. The attenuation coefficient of the optical waveguides is $\alpha_0 = 5dB / cm$. The simulations showed that at a resonant wavelength $\lambda = 1550.8nm$, the switching can be obtained with a very high extinction ratio and low applied voltages. At off resonance, the transmission is unity, so the input signal can go through the device and does not be affected by the applied voltages. Using this mechanism, a wavelength selective switch is obtained.

The transmission characteristics depend on the applied voltage at resonant wavelength $\lambda = 1550.8nm$ as shown in Fig. 3. To make the device design issues closer to practical applications, we have taken into account propagation losses in our analyses and simulations. The results showed that the performance of the switch depends on the propagation loss of the optical waveguides. Therefore, it is not like the previous analyses, in which the optical losses are neglected, the analyses presented in this paper took into account the propagation loss and considered its effects on the performance of the device.

By using MMI tunable couplers [7], the coupling coefficients of the couplers and the performance of the device can be tuned to desired characteristics. The transmission of the device at different coupling coefficient values is shown in Fig. 4. When the coupling coefficient $\tau = 0.8$, an optical switch with highest extinction ratio can be obtained.

The performance characteristics of the device depend on the applied voltage and loss coefficients, so the desired spectral characteristics can be obtained by changing the applied voltage. The spectral characteristics at variety of applied voltages are shown in Fig. 5. The simulation results showed that by tuning the applied voltage to a proper value, the device acts as a switch or acts only as a optical device with unity transmission at all wavelengths. For example, at the applied voltage $V/V_\pi = 0.5$, the transmission is almost unit at all wavelengths, so input signal goes through the device. However, at an normalized applied voltage of $V/V_\pi = 0.8$, the device acts as an optical switch with an extinction ratio of -13dB and a high FSR of 20nm. This value of FSR is the almost highest one of known switching structures based on a single microring resonator up-to-date.

IV. CONCLUSION

We have presented a analysis and design of a novel optical switch structure based on MMI assisted microring resonators. Compared to previous switch designs based on electro-optic effect, very low switching voltages can be obtained. The design of the device are given in detail. Moreover, we have demonstrated that by choosing a proper value of applied voltage, a wavelength selective switch with very high FSRs and a low switching power can be achieved.

V. REFERENCES

All-Optical Transistor Operation Based on Bistability Principle in Nonlinear DFB GaInAsP-InP Waveguide: A Transient Perspective

Yosia¹, Yoichi Akano², Kazuhiko Tamura², Tetsuya Mizumoto², and Shum Ping¹

¹Network Technology Research Centre, Department of Electrical and Electronic Engineering, Nanyang Technological University, Research Technoplaza, 50 Nanyang Drive, 4th Floor, XFrontiers Block, Singapore 637553
²Mizumoto Laboratory, Department of Electrical and Electronic Engineering, Tokyo Institute of Technology, 2-12-1 Ookayama, Meguro-ku, Tokyo 152-8552, Japan
Email: yosi0001@ntu.edu.sg

We demonstrated all-optical transistor based on bistability principle in nonlinear DFB GaInAsP-InP waveguide experimentally. The unstable state is shown to play a crucial role in distinguishing the probe transmission transient artifacts between switching and transistor mode.

1. Introduction

In the vision to realize the future all-optical technology, there is a need to do signal processing purely in optical domain without electronics. Nonlinear periodic structures such as fiber Bragg grating (FBGs), photonic crystal (PhC), and DFB GaInAsP-InP waveguide are capable of producing plethora of nonlinear phenomena [1-4], in which some of them have inherent optical signal processing capabilities. One of the most useful nonlinear phenomena in periodic structures for optical signal processing purpose is optical bistability, based on which various all-optical signal processing capabilities have been demonstrated theoretically and experimentally [5-9].

Building on the research accomplishments above, we report the experimental demonstration of all-optical transistor operation based on bistability principle in nonlinear DFB GaInAsP-InP waveguide. By increasing the pump power above a certain limit, it is experimentally shown that the probe transmission in the switching mode is turned into transistor mode. By numerical simulations, the unstable state is shown to play a crucial role in distinguishing the probe transmission transient artifacts between switching and transistor mode. Finally, the existence of unstable state is verified in the experiments by taking the snapshots of probe transmission while it is operating in unstable mode.

2. Experimental Demonstrations of All-Optical Transistor Operation

The strong probe and strong pump are propagating simultaneously inside nonlinear DFB GaInAsP-InP waveguide. The probe wavelength λ is set to be around the edge of the band-gap and its power is set in the bistable regime, whereas the pump wavelength is set far from the band-gap hence unperturbed by the grating and its power is varied to investigate various probe transmission modes. The probe is set at 135 ns long, while the pump is set at 50 ns long and overlapped around the middle of the probe as shown in Fig 1. Both the probe and pump output is monitored by oscilloscope in the time averaging mode to suppress influence of the noise.

![Normalized Input Probe (solid) and Pump (dotted)](image)

Fig 1. Input probe (solid) and pump (dotted)

As the pump power is increased to be in the high regime, the probe transmission is turned from switching to transistor mode. The term of all-optical transistor operation
shown in Fig 2 is derived based on the fact that the relatively lower pump power when it is “on” can produce high probe transmission as it switches from low to high state of bistable regime. The best gain (G) of the transistor operation was measured at around 6 dB. However, when the pump is off, some of the probe transmission remains as it operates in the low state of bistable regime. This is regarded as noise. The highest SNR was measured at 12.7 dB.

3. A Transient Perspective of All-Optical Transistor by Numerical Simulations and Experiments

To understand the physical reason behind all-optical transistor operation when the pump is high, we performed numerical simulations by Implicit 4th Order Runge-Kutta Method [9] to solve the set of Nonlinear Coupled Mode Equations under strong probe and strong pump conditions as shown in Fig 3.

When the pump power is in the low regime, the probe transmission exhibits switching operation (dashed). However as the pump power is increased, it turns the probe transmission from switching to transistor mode (dotted) as the cross phase modulation (XPM) is stronger for the later case. The unstable state is shown as a boundary to distinguish the two transient artifacts of probe transmission.

Lastly, the existence of unstable state is verified in the experiment by taking the snapshots of probe transmission as shown in Fig 4.

4. References
Invited Talk

Nano- and Micro-photonic devices for Inter- and Intra-Board Level Optical Interconnects

Ray T. Chen
Cullen Trust endowed professor
Microelectronics Research Center
The University of Texas, Austin
chen@ece.utexas.edu
512-471-7035

Abstract
The role of polymer-based and silicon-based Nano- and Micro-photonic passive and active devices will be presented under the consideration of plausible applications in intra- and inter-board level interconnects beyond 10 Gbit/sec with acceptable bit error rates.

Summary
The speed and complexity of integrated circuits are increasing rapidly as integrated circuit technology advances from very-large-scale integrated (VLSI) circuits to ultra-large-scale integrated (ULSI) circuits. As the number of devices per chip, the number of chips per board, the modulation speed, and the degree of integration continue to increase, electrical interconnects are facing their fundamental bottlenecks, such as speed, packaging, fan-out, and power dissipation. In the quest for high-density packaging of electronic circuits, the construction of multichip modules (MCM), which decrease the surface area by removing package walls between chips, improved signal integrity by shortening interconnection distances and removing impedance problems and capacitances. The employment of copper and materials with lower dielectric constant materials can release the bottleneck in a chip level for the next several years. The International Technology Roadmap for Semiconductors (ITRS) expects that on-chip local clock speed will constantly increase to 10 GHz by the year 2011. Electrical interconnects operating at a high-frequency region have many problems to be solved, such as crosstalk, impedance matching, power dissipation, skew, and packing density. Optical interconnection has several advantages, such as immunity to the electromagnetic interference, independence to impedance mismatch, less power consumption, and high-speed operation. Although the optical interconnects have great advantages compared with the copper/low K interconnection, they still have some difficulties regarding packaging, multilayer technology, signal tapping, and reworkability[1-7]. In this presentation, the progress of optical interconnects for intra and inter-board levels will be presented with both nano and micro photonic passive and active components suitable for system integration. These include thin film planar waveguides, vertical cavity surface emitting lasers (VCSELs), PIN photodiode array (figure 1) and silicon nano-photonic crystal waveguide modulators and 3D photonic devices for laser beam steering. Optical bandwidth of 150GHz has been experimentally confirmed for a 51 cm long polymer waveguide as shown in Fig.2 [2].

Figure 1 Schematic of board level fully embedded optical interconnects

References
2. X. Wang, L. Wang, W. Jiang, and R. T. Chen, “Hard-molded 51 cm long waveguide array with a
150GHz bandwidth for board level optical interconnects,” will be published in Optics Letters, Vol.32, no.7, April 2007

The author

Dr. Chen is the Cullen Trust endowed professor at UT Austin. He received his BS degree in Physics from National Tsing-Hua University in 1980 in Taiwan and his MS degree in physics in 1983 and his PhD degree in Electrical Engineering in 1988, both from the University of California. He joined UT Austin as a faculty to start optical interconnect research program in the ECE Department in 1992. Prior to his UT’s professorship, Chen was working as a research scientist, manager and director of the Department of Electrooptic Engineering in Physical Optics Corporation in Torrance, California from 1988 to 1992.

Chen also served as the CTO/founder and chairman of the board of Radiant Research from 2000 to 2001 where he raised 18 million dollars A-Round funding to commercialize polymer-based photonic devices. His research work has been awarded with more than 84 research grants and contracts from such sponsors as DOD, NSF, DOE, NASA, the State of Texas, and private industry. Chen’s group at UT Austin has reported its research findings in more than 430 published papers including over 60 invited papers. He holds 14 issued patents. He has chaired or been a program-committee member for more than 60 domestic and international conferences organized by IEEE, SPIE (The International Society of Optical Engineering), OSA, and PSC. He has served as an editor or co-editor for eighteen conference proceedings. Chen has also served as a consultant for various federal agencies and private companies and delivered numerous invited talks to professional societies. Dr. Chen is a Fellow of IEEE, OSA and SPIE. He was the recipient 1987 UC Regent’s dissertation fellowship and of 1999 UT Engineering Foundation Faculty Award for his contributions in research, teaching and services. Back to his undergraduate years in National Tsing-Hua University, he led a university debate team in 1979 which received the national championship of national debate contest in Taiwan.

There are 30 students received the EE PhD degree in Chen’s research group at UT Austin.
Repetition Rate and Central wavelength Tunable Terahertz Optical Clock Generation Using Variable Bandwidth Spectrum Shaper

S. Anzai¹, Y. Komai¹, M. Mieno¹, N. Wada₂, T. Yoda³, T. Miyazaki², K. Kodate¹
1. Japan Women’s University, 2-8-1, Mejirodai, Bunkyo-ku, Tokyo 112-8681, Japan, awg@fourier.jwu.ac.jp
2. National Institute of Information and Communications Technology, 4-2-1, Nukukita, Koganei-shi, Tokyo 184-8795, Japan
3. Optoquest Co., Ltd. 1335, Haraiichi, Ageo-shi, Saitama 362-0021, Japan

Abstract: Tunable terahertz optical clock generation from 1.28 to 4.0THz is successfully demonstrated experimentally, using a variable bandwidth spectrum shaper and a supercontinuum light source, without using any high speed electronics.

1. Introduction
Recently, a number of researches have been carried out to develop new technologies of ultra high-speed repetition rate with terahertz (THz) order optical clock generation based on ultrafast and ultra broadband property of light. Applications of this THz repetition rate optical clock are broad, including ultra high-precision measurement, development of new material, optical and biomedical analysis and generation of standard clock in ultrafast photonic network and so on[1-3]. Currently, optical clock generation techniques based on compression by high nonlinear fiber and optical time division multiplexer, Fourier synthesis of multiple wavelengths and mode locking distributed-Bragg-reflector are available[4-7]. Ultrafast clock with repetition rate of 10 THz have been realized by applying molecular vibration[8]. However, the tunability of repetition rate and center wavelength of optical clock has not been achieved yet by those technologies. We have proposed repetition rate and central frequency tunable THz order optical clock generator[9].

In this paper, we demonstrated THz order optical clock generation using a variable bandwidth spectrum shaper (VBS), a 48-channel, 100GHz-channel-spacing spectrum equalizer (SE) and supercontinuum (SC) generator. Optical clocks with sharp spectrum components of 1.0-4.0THz mode spacing are demonstrated, respectively.

2. Spectrum signal processing based on Fourier transform
Figure 1 shows the principle of pulse processing in the spectral region. The pulse width \( \Delta t \) is proportional to the inverse of the spectrum width. The pulse repetition rate corresponds to the inverse of frequency spacing of spectrum mode components. The mode spacing of ultrafast optical pulse train is so broad that it could easily be processed in the spectral region. Ultrafast optical clock is generated by suppressing the frequency components of the optical pulse. We focus on this principle of pulse processing in the spectral region as a method of generating ultrafast optical clock.

3. Repetition rate tunable THz optical clock generation
Hz order optical clock is generated by a VBS, a 48-channel, 100GHz-channel-spacing spectrum equalizer (SE) and supercontinuum (SC) generator. Optical clocks with sharp spectrum components of 1.0-4.0THz mode spacing are demonstrated, respectively.

Figure 2 shows the experimental set-up for generation of optical clock and spectrum of SC light.
Dispersion Flattened Fiber (DFF), Polarization Controller (PC), polarizer, circulator, VBS, SE, and interferometer as shown in Fig.2. VBS consists of collimating lens, grating as the functional Fourier transforming device, lens and Spatial Light Modulator (SLM). The amplitude and phase of each spectrum components are controlled by SLM. Parameter of VBS is 10GHz channel-spacing, insertion loss is within 6dB and controlled for whole C band.

The 10GHz SC light generated by DFF is based on 2ps, 10GHz pulse train from the MLLD as shown in Fig.2. The Spectrum of 10GHz SC light is synthesized by VBS.

Figure 3 (a), (c), (e) and (g) shows measured spectrum of 1.28THz-4.0THz mode spacing. The central wavelength of these optical clocks is 1548.0nm. Figure 3 (b), (d), (f) and (h) show autocorrelations of generated optical clocks with repetition rate of 1.28THz-4.0THz measured by Michelson interferometer. And peak spacing of autocorrelations is corresponding to 1.28THz-4.0THz. The power of main- and side-lobes is virtually the same. It means the waveforms of generated optical clocks are much stable. In addition, changing spectrum peaks of optical clock by VBS, it is able to tune central wavelength of generated optical clock. Figure 4 (k), (l) shows spectrum and autocorrelation of 1.28THz and 2.56THz repetition rate optical clock with central wavelength of 1550nm.

4. Conclusions
A new generation technique of THz repetition rate optical clock has been proposed. 1.28THz, 2.56THz, 3.0THz and 4.0THz optical clocks have been generated by 10GHz-spacing, attenuation and phase tuneable VBS and a spectrum equalizer with supercontinuum generation. Optical clocks with 2 and 5 sharp spectrum components of 1.28THz, 2.56THz, 3.0THz and 4.0THz mode spacing within 10dB intensity range have been demonstrated, respectively. This technique has capabilities to generate repetition rate and central wavelength widely tunable optical clock.

References
Abstract:

Continuous-wave (CW) terahertz waves were generated via excitation of phonon polaritons in GaP. We investigated the phase matched angle frequency dependence of THz wave output power. We studied frequency-tunability of the CW THz wave.

1. Introduction

The Terahertz (THz) frequency region (100GHz to 10 THz) lies between radio and infrared regions and it has attracted significant interest due to number of potential applications. However pulsed THz waves have limited applications. CW THz waves can be used for carriers and modulators in telecommunications, and CW THz waves pumped using semiconductor lasers have a narrow line width of a MHz order. CW THz waves also have applications in high-resolution and high-speed THz spectroscopy, and imaging technologies, but only few CW THz sources have been developed so far. We have developed a widely frequency-tunable CW THz signal generator with laser diode (LD) pumping from GaP crystal and obtained high output power.

In 1963, Nishizawa predicted the generation of THz-waves from compound crystals via the excitation of phonons or molecular vibrations [1,2]. An electromagnetic wave with a frequency of 12.1 THz was generated from GaAs pumped by GaP Raman laser, at a power of 3 W [3,4]. Recently, wide-frequency-tunable high-power THz wave signals have been generated from GaP with Q-switched pulse pumping [5,6]. THz waves were generated using difference frequency generation (DFG) via the excitation of phonon-polaritons in GaP, which converts energy very efficiently, and the THz wave had a pulsed energy of 9 nJ/pulse (peak power of 1.5 W). Frequency-tunable CW THz-wave generation from GaP should be possible by enhancing the power density of the incident beams. This study describes the CW THz waves generation from GaP based on LD pumping as well as the frequency tunability.

2. Experimental

The pump and signal lasers are an external cavity laser diode (ECLD) and laser diode pumped Nd: YAG (1064 nm) laser respectively. The wavelength of ECLD can be varied from 1012 nm to 1080nm. The intensity of signal laser is amplified by using ytterbium-doped fiber amplifier (FA). The power from the ECLD and from FA was 0.24 and 3.6 W respectively. The incident beams were focused to spot sizes near the wavelength of THz waves on a GaP crystal. From the results of pulse pumping, the THz wave generated under the small angle phase-matching condition at each THz frequency should be tunable. Generated THz waves are detected by using Si bolometer cooled at 4.2 K. The bolometer signal is measured with a lock-in-amplifier.

3. Results and Discussion

Figure 1 shows the phase matching condition comparing that of theoretical and pulse pumping at 1 μm. At present the output power is in pW order. It is investigated that the THz power is likely proportional to square of the GaP crystal length ξ, when the coherence length ξcoh is sufficiently larger than the crystal length [7].

![Fig. 1 Phase matching condition in GaP THz wave signal generator.](image-url)
We investigated the phase matched angle dependence of the THz wave output power at various THz frequencies for different lengths of the GaP crystal. THz waves were generated over a range from 0.68 to 2.75 THz. CW THz power spectrum is obtained by piezo tuning the ECLD, in a wide range of THz frequencies.

4. Conclusion

Frequency-tunable CW THz wave generated with LD pumping based on the difference frequency generation via excitation of phonon polaritons in GaP. By tuning a small angle between pump and signal beam directions, a CW THz wave can generated in widely frequency range. The THz frequency of 0.2 to 7.5 THz can be swept widely with external cavity tuning and CW THz waves have very narrow line width of MHz. The generated output power can be enhanced by employing a longer GaP crystal with lower carrier density. CW THz power spectrum is studied by tuning ECLD, in a wide range of frequencies.

References

Anti-Shake Mechanism for Mobile Optical Wireless Communication

Masaharu Hattori, Tomonori Yazaki, and Hideaki Tanaka
KDDI R&D Laboratories Inc., 2-1-15 Ohara, Fujimino-shi, Saitama 356-8502, Japan
Tel: +81-49-278-7654 Fax: +81-49-278-7821 E-mail: m-hlattori@kddilabs.jp

Abstract
We implement and demonstrate an anti-shake mechanism to suppress the involuntary hand movements for mobile optical wireless communication. The demonstration suggests that the effectiveness of the anti-shake mechanism to usual use.

Introduction
The memory volume of mobile devices such as mobile phones and digital cameras has been increasing rapidly in recent years. Some users want to exchange large data between mobile devices and fixed devices such as personal computers and information appliances by using high-speed interfaces. Optical wireless interfaces are very attractive among the various wireless interfaces because of high potential for high-speed communication and high security. The standard of an infrared light communication at a speed of 100 Mbit/s with a transmission distance of 100 cm have been defined [1]. Furthermore, demonstration of the 1 Gbit/s data transfer between mobile devices with the transmission distance of 10 cm by using infrared light has been reported [2]. The transmission speed increase results in decreases of transmission distance and/or receivable area because the minimum sensitivity decreases as the transmission speed increases. In order to enhance the usability, it is desirable to extend the communication distance, but more reduction of receivable area is inevitable. The involuntary hand movements may result in frequent “link down” states due to misalignment in case of high-speed communication. In order to expand the receivable area, an automatic optical axis alignment system was reported so far [3]. The posture of the receiver was adjusted by monitoring the direction of the data beam. However, the technology was suitable for communication between fixed or semi-fixed devices, but wasn’t seemed to be considered for compensation of the involuntary hand movements. We have already proposed an anti-shake mechanism for mobile optical wireless communication [4]. Beam fluctuation can be compensated by a monitor of beam position and an adjustment of the photodiode position.

In this paper, we described the implementation of the anti-shake mechanism for mobile optical wireless communication in detail. We have also demonstrated the anti-shake mechanism.

System Configuration
In the transmitter, two laser diodes at wavelengths of 850 nm and 658 nm are utilized for data and guide respectively. The data and guide beams are collimated and radiated along the same axis.

In the receiver, the guide beam is eliminated by an optical filter in front of the front lens as shown in Fig. 1. After that, the data beam is split into two beams for a communication and a position sensing with the ratio of 9:1 by a beam splitter. A position sensing beam is focused on a position sensitive detector (PSD) with an active area of 45 mm x 45 mm square. A communication beam is injected to a multimode fiber with a core diameter of 1 mm and is guided to a 200 μm - diameter photodiode through the multimode fiber. The PSD and one of the ends of the multimode fiber are fixed on the 2-dimentional motorized stage in order to avoid the mount of RF circuits with the photodiode on the motorized stage. The stage is controlled by monitoring the PSD signals so that the communication beam is coupled into the multimode fiber.

Evaluation of anti-shake mechanism
Firstly roughly estimation of the involuntary hand movements was carried out. A beam position fluctuation on the optical filter due to the involuntary hand movements was measured by the PSD which was set behind the optical filter. A distance between a transmitter and a receiver was set to 2 m. Note that the transmitter was handheld with a stand and the beams were aimed to the center of the optical filter. Typical fluctuation of examples which were one-dimensionally measured was shown in Fig. 2 (a). The position was measured every 0.02 seconds for 5 seconds. The maximum position fluctuation was within 20 mm. Figure 2 (c) shows the frequency distribution calculated...
Amplitude, it was measured under the same condition set to 20 mm and 2 Hz, respectively. As for the movements, the applied amplitude and frequency were estimated to be approximately 0.1 mm. Considering the multimode fiber on the motorized stage was the result of the analysis of the involuntary hand movements. The distance between the transmitter and the receiver was also set to 2 m. The radiated hand movements. The distance between the transmitter and the receiver was set on the motorized rotary stage to emulate the involuntary hand movements. The evaluation was carried out in one-dimension for the purpose of simplification. The transmitter was set on the motorized stage to emulate the involuntary hand movements. The distance between the transmitter and the receiver was also set to 2 m. The radiated power and the diameter of data beam were set to 1.26 mW (+1 dBm) and 1.4 mm, respectively. The power and the diameter of data beam were set to 1.26 mW (+1 dBm) and 1.4 mm, respectively. The radiated power was drastically fluctuated in ranging between 0 and 40 µW without anti-shake mechanism. The link state was seemed to be changed alternately between the “link up” and “link down” states. On the contrary, the received power was stable with a variation of 5 µW with anti-shake mechanism. As the power was kept higher than the minimum receiver sensitivity, it was expected that the link was always established.

Next, the performance of the anti-shake mechanism was evaluated as shown in Fig. 3. Although the anti-shake mechanism was designed to provide effectiveness to two-dimensional beam fluctuation, the evaluation was carried out in one-dimension for the purpose of simplification. The transmitter was set on the motorized rotary stage to emulate the involuntary hand movements. The distance between the transmitter and the receiver was also set to 2 m. The radiated power and the diameter of data beam were set to 1.26 mW (+1 dBm) and 1.4 mm, respectively. The communication beam diameter measured at the end of the multimode fiber on the motorized stage was estimated to be approximately 0.1 mm. Considering of the result of the analysis of the involuntary hand movements, the applied amplitude and frequency were set to 20 mm and 2 Hz, respectively. As for the amplitude, it was measured under the same condition for the evaluation of the involuntary hand movements. Figure 4 shows the received power fluctuation calculated by the photodiode current. The dot-and-dash line shows a minimum receiver sensitivity of -17 dBm (20 µW) which is defined in the standardized specification of Gigabit Ethernet at the wavelength of 850 nm. The received power was drastically fluctuated in ranging between 0 and 40 µW without anti-shake mechanism. The link state was seemed to be changed alternately between the “link up” and “link down” states. On the contrary, the received power was stable with a variation of 5 µW with anti-shake mechanism. As the power was kept higher than the minimum receiver sensitivity, it was expected that the link was always established.

**Conclusions**
The anti-shake mechanism against the involuntary hand movements has been implemented and demonstrated for mobile optical wireless communication. We have successfully confirmed the effectiveness of the anti-shake mechanism to suppress the involuntary hand movements.

**Acknowledgement**
The authors wish to thank Dr. S. Akiba, Dr. M. Suzuki and Dr. M. Usami for their continuous encouragement.

**References**
Optoelectronic Package using Optical Waveguide Hole for Chip-to-chip interconnection

Takeshi Ohno, Yutaka Takagi, Toshikazu Horio, Toshifumi Kojima, Toshikatsu Takada, Kazushige Obayashi and Masahiko Okuyama
R&D Center, NGK Spark Plug Co., Ltd.
2808 Iwasaki, Komaki, Aichi 485-8510 Japan
Phone: +81-568-76-5262, Fax: +81-568-76-1295
E-mail: ta-ono@mg.ngktk.co.jp

Abstract
We have proposed an optoelectronic package structure having optical waveguide holes through a package for chip-to-chip interconnections. This can be coupled optical device to a fiber array by passive alignment from LSI mounting plane.

Introduction
Today fiber optic interconnections have been extending their field to the short reach region within computer systems because it is said that it has become the bottleneck of electrical transmission due to increase LSI performance. So, optical interconnection modules that can be coupled to optical fiber array within servers or routers have been studied and progressed at many organizations [1,2], but the optical interconnection within computer systems such as CPU (Central Processing Unit) to memory, the so-called chip-to-chip interconnection, has not been enough studied thus far.

Therefore, we have been studying an optoelectronic package for chip-to-chip interconnections, which can be coupled to an optical fiber array. Unlike long-haul optical transmission, a package used in consumer-computer is required to be the low-cost optical coupling structure that can be used the established assembly process in the very short region such as chip-to-chip interconnections. Additionally, it needs to design the optimal electrical transmission lines for higher bandwidth. We have been considering taking following steps against above issues, as follows:

1. Using a fiber array as an optical transmission media at the portion of higher data transmission, and utilizing MT coupling techniques between media and package.
2. The package having optical waveguide holes to transmit optical signals between the flip chip optical device and the optical fiber connector.
3. The package having optimal electrical transmission structure for higher bandwidth because a processor, a controller IC and an optical device are located to be near on the same plane to ease the impedance matching.

Concepts of Optoelectronic Package
Our proposed package has the shortest possible electrical paths that utilize only surface lines and standard Flip Chip bonding pads without difficult impedance matching such as wire bondings, through holes and vias. Fig.1 shows the conceptual image of optoelectronic package structure. The established socket is connected to the motherboard, and then the LSI (Large Scale Integration) package is set in the socket with the PGA (Pin Grid Array). In this structure, electrical power or low-speed signals are supplied through the PGA socket. However, high-speed data signals can be transmitted through the optical paths and not the PGA socket. The processor, controller IC and optical device are mounted on the same surface plane of the package by flip chip connection. The package with guide holes [3] can be simply coupled to the passive optical connector with guide pins on the backside. The passive optical connector has the 45-degree mirror for picking up optical signals.

Proposed package can be assembled by all passive alignment, because flip chip mounting of active devices and fabrication of guide holes to incorporate guide pins are handled with respect to the same alignment patterns on the surface plane. Optical devices have active faces and electrodes on the same plane. So when optical devices are mounted by flip chip in
common use, these work forward package. Therefore, the optical through holes that are fabricated with guide holes at the same time are required for transmission through the package. Consequently, we have developed the simple and efficient optical transmissions through the package.

**Optical waveguide Hole**

If output beam from a VCSEL (Vertical Cavity Surface Emitting Laser) is transmitted in free space, optical receiving power significantly decreases due to the beam divergence. Additionally, if it is transmitted in the drilled holes, this will lead to lower transmission quality due to the diffuse reflection at the rough sidewalk. To get sufficient quality of optical signals cannot be expected by both ways. In order to overcome these issues, it is reported that optical paths by a filling transparent resin into the prepared holes availed against power loss and low transmission quality [4].

We would like to propose the optical waveguide hole with core and clad structure for more efficient transmission. Table.1 shows ray-trace simulation results of free space and optical waveguide hole transmission, respectively. In spite of only 0.4mm free space transmission, the optical coupling loss was -13.3dB. Therefore, it is difficult to establish the optical transmission system for transmitting and receiving by free space transmission. However the optical waveguide hole with the same size as an optical fiber can get sufficiently low loss for optical transmission systems.

Fig.2 shows the experimental field patterns of 0.4mm free space and optical waveguide hole transmission, respectively. Due to the optical waveguide hole, we can now have the higher optical power and the smaller spot size to be able to combine with the optical fiber rather than free space transmission.

<table>
<thead>
<tr>
<th>Table.1 Simulation result of optical coupling loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmission media</td>
</tr>
<tr>
<td>Free space transmission</td>
</tr>
<tr>
<td>Optical waveguide hole</td>
</tr>
</tbody>
</table>

**Table.2 Properties of optical waveguide hole**

<table>
<thead>
<tr>
<th>Material properties</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Difference in reflective index</td>
<td>8.3%</td>
</tr>
<tr>
<td>Insertion loss</td>
<td>-1.7dB</td>
</tr>
<tr>
<td>Jitter (p-p)</td>
<td>25ps (Δ:&lt;2ps)</td>
</tr>
</tbody>
</table>

**Conclusions**

We have proposed the optoelectronic package that can couple the optical fiber array to LSIs for chip-to-chip optical signal transmission. We have demonstrated that the optical waveguide holes through a package have excellent optical transmission against insertion loss and jitter characteristics. We will verify the usefulness of our proposed package with guide holes for passive alignment and the optical waveguide holes for optical paths, which can be coupled to an optical fiber connector.

**References**

Very Low Noise AlInAs Avalanche Photodiodes with Gain-Bandwidth product of 140 GHz
M. Achouche, A. Rouvié, J. Decobert, N. Lagay, F. Pommereau, D. Carpentier
Alcatel Thales III-V Lab, Route de Nozay, 91460 Marcoussis, France
mohand.achouche@3-5lab.fr

Abstract:
We demonstrate a new planar junction AlInAs APD with carbon-doped charge layer achieving low dark current of 13nA (M=10) and very low excess noise factor of 3 (M=10). These characteristics are simultaneously achieved with a responsivity of 9.5 A/W (M=10) and a gain x bandwidth of 140 GHz.

Introduction:
Recent demand for high sensitivity receivers for 10Gb/s metro networks requires high performance avalanche photodiodes (APDs) because their internal gain improves sensitivity (>10dB) over traditional PIN photodiodes. In addition, avalanche photodiodes constitute a cost effective solution due to their compactness and low power consumption [1].

SAGM (Separate Absorption Grading and Multiplication) APD structures using bulk AlInAs avalanche material appear today as a good alternative to InP avalanche layer and achieve similar performances as AlInAs/AlGaInAs MQW APDs, the epitaxial growth of which is somehow more complicated [2], [3]. Owing to a high ionization coefficients ratio of 3-4, AlInAs APDs demonstrated a high gain-bandwidth product of 120 GHz and a receiver sensitivity of −29.8 dBm at 10Gb/s and BER of 10⁻³, using either front or back illuminated structures [4], [5]. These results have been demonstrated using a simple mesa avalanche diode to define the absorption window which requires however an InP-regrowth step acting as a guard-ring layer to improve device reliability. Recently, a front-side illuminated APD has been realized using the well established planar junction technology using Zn-diffusion process and has demonstrated a low dark current of 160 nA at 0.9xVb (Vb is the breakdown voltage) and a gain-bandwidth product of 120 GHz [5]. However, as the diffusion process occurs at high temperatures (>500°C), a stable charge doping layer, which reduces the electric field in the absorption layer, is required.

In this paper, we report on a new planar junction AlInAs APD with carbon-doped charge layer. The flip-chip mounted APD primary responsivity reaches 0.95 A/W at 1.55 µm with a gain-bandwidth product of 140 GHz. The stability of the APD structure is displayed through a very low dark current figure with less than 20 nA dark current (at 0.9xVb) for avalanche gains up to 20. A very low excess noise factor of 3 at gain of 10 and an ionization coefficients ratio of 5-7 are additionally demonstrated.

APD Design and Fabrication:
The epitaxial structure of the SAGM APD is grown by MOVPE (Metallic Organo Vapor Phase Epitaxy) on an n-doped InP substrate. The vertical structure includes from the top an unintentionally doped InP window layer (1 µm) followed by an unintentionally doped InGaAs absorption layer (1.2 µm) and an unintentionally doped AlInAs avalanche layer (0.2 µm). An InGaAlAs grading layer (0.1 µm) is inserted between absorption and avalanche layers to avoid carriers trapping during transport between both layers. In addition, a stable and accurately defined charge doping layer made of C-AlInAs (4x10⁻¹² cm⁻²) is used to avoid high electric field in the absorption layer.

The process flow of our back-side illuminated APDs fabrication is made using optical lithography and is mainly based on dry etching process for mesa realization. Highly reliable planar junction is carried out by Zn diffusion to obtain p-InP region in the undoped InP window layer. This operation is made in an MOVPE reactor using DMZn as source doping material to improve process control by immediate switch on-off of doping gas and uniformity by substrate rotation. A PtAu and a TiPtAu layers which are patterned using lift-off are used for P- and N-contacts, respectively. After wafer thinning and anti-reflection coating on the back-side of the wafer, the APDs are diced and mounted using flip-chip on a high frequency carrier.

Figure 1 shows the equipotential lines calculated using a drift diffusion model. The simulation demonstrates a gradual evolution of the potential at the periphery of the junction as well as in the avalanche layer which avoids possible edge breakdown.

Fig.1: Equipotential lines simulation of the APD (starting from the substrate).
**AlInAs APD Photodiode characteristics:**

Figure 2 shows typical dark current, photocurrent and multiplication gain versus reverse bias voltage characteristics for a 30 µm diameter flip-chip mounted APD. The dark current remains low and stable (~ 10 nA) over a wide range of applied bias (from punchthrough 11-12 V up to gain M=10) and is less than 30 nA for avalanche multiplication factors of 30. This low dark current characteristics even at high avalanche gain indicates the good passivation process of the planar junction as well as the excellent stability of electric field distribution along the structure achieved using our newly developed Carbon-doped field control layer (charge layer). A high primary responsivity of 0.95 A/W is measured at 1.55 µm owing to the flip-chip assembly with no additional parasitic capacitances (depletion capacitance of 140 fF). The breakdown voltage is around 32 V allowing possible available multiplication gain of >100.

![Fig.2: APD dark current, photocurrent and gain characteristics versus reverse bias voltage](image)

Figure 3 shows the excess noise factor versus multiplication gain deduced from the noise spectral density measurements. Very low excess noise factors of 3 at M=10 are deduced corresponding to an ionization coefficient ratio k=5-7 (Teich model [6]). The frequency response of the fabricated APD was measured from 130 MHz to 20 GHz at 1.55 µm. The bandwidth (3-dB) dependence on multiplication factor is displayed in Figure 4 where a gain-bandwidth (G.B) product of 140 GHz is achieved allowing a large maximum achievable gain at 10 Gb/s operation. This is to our knowledge the first time that a back-side illuminated APD displays such a large G.B product together with a primary responsivity of 0.95 A/W (M=1) and a low noise and dark current. In addition, to illustrate systems benefit of the new AlInAs APD for 10 Gb/s operation, calculations have been conducted on two avalanche receivers (TIA equivalent input noise is 16 pA/(Hz)^{1/2}): the first APD is characterized with an ionization ratio (k) of 3.3 and a dark current (Idark) of 1 µA (as shown in InP-APD) and the second (similar to the present new APD) with a k of 5.3 and Idark of 0.01 µA. A clear improvement (~ 2.5 dB) of the minimum received power is achieved at 10 Gb/s using the later APD with a calculated sensitivity of – 32 dBm.

![Fig.3: Excess noise factor vs multiplication factor (Teich noise model displayed for determination of k)](image)

![Fig.4: Frequency response versus multiplication factor of a 30µm AlInAs APD](image)

**Conclusions**

A flip-chip back-illuminated AlInAs APD with a planar junction is proposed. Accurate control of the charge layer allows to achieve very low dark current even at high multiplication gain (Idark < 30 nA for M=30) and a multiplication process that is strictly concentrated in the AlInAs avalanche layer without unwanted multiplication in InGaAs absorption layer. Therefore a high ionization coefficients ratio of 5-7 is deduced from noise measurements (F=3 at M=10). High primary responsivity of 0.95 A/W at 1.55 µm was achieved on a 30 µm diameter APD together with a G.B of 140 GHz. These performances associated to a robust planar junction technology make those C-AlInAs APDs the choice candidates for 10 Gb/s receivers.

**References**

1. J.C. Campbell OFC’06, 2006, OFI4
2. M. Itzler et al. OFC’00, 2000, FG5
4. E. Yagyu et al. ECOC’05, 2005, We.3.6.1
5. S. Tanaka et al. ECOC’02, 2002, 10.5.3
Wide Temperature Range 1.25-Gb/s Burst-Mode Receiver Chip Set for G-PON Optical Receiver

Y. Tsunoda*1, S. Ide*1, T. Yamabana*1, Y. Hattori*2, S. Sakuramoto*3, T. Matsuyama*3, K. Mori*1

*1 Fujitsu Laboratories Ltd. *2 Fujitsu VLSI Ltd. *3 Fujitsu Limited

Abstract We developed a 1.25-Gb/s optical burst-mode receiver chip set with a new ATC stabilizing method, which supports a wide operating temperature range. This new ATC method suppresses sensitivity degradation and output waveform duty fluctuation and pattern jitter caused by fluctuations in the threshold level. Using this method, we achieved high sensitivity, a wide dynamic range, and a fast settling time over the full -40°C to 85°C temperature range. These performance results comply with the specifications defined in ITU-T Recommendation G.984.2 Class B+.

1. Introduction

Recently, the use of high-speed access networks for broadband communications has been spreading widely throughout the world. A passive optical network (PON) system is one of the most promising technologies to implement access networks because of its advantages in communication speed and cost. In particular, a gigabit-capable passive optical network (G-PON) that meets ITU-T Recommendation G.984.2 [1] is demanded in North America because it is consistent with SONET/SDH systems.

For G-PON up-stream transmission, a 1.25-Gb/s burst-mode optical receiver is a key technology for high-speed decision threshold of the received packets, whose power varies widely due to different optical path lengths. An automatic threshold control (ATC) circuit is an optimal solution for a burst-mode receiver, which detects the threshold levels of the received packets using peak and bottom hold circuits. To reduce ATC settling time, it is necessary to use small hold capacitances in the peak/bottom detector.

However, if small hold capacitances are used, a fluctuation in the threshold level occurs, which is caused by leakage of electric charge. Threshold level fluctuation leads to changes in the duty ratio and pattern jitters of the output signals. In that case, it not only degrades the sensitivity but also makes it difficult for clock data recovery (CDR) circuits to recover the clock [2,3]. Because the leakage of charge occurs through the transistor, which is followed by small hold capacitance at high temperature, it is important to suppress threshold level fluctuation by this leak to achieve an optical receiver that operates in a wide temperature range.

In this paper, we developed a 1.25-Gb/s burst-mode optical receiver chip set using a new ATC method that suppresses the threshold level fluctuation to reduce the change in the duty ratio. Using this method, the pattern jitters were reduced, and a receiver that had this chip set achieved high sensitivity of -34.4 dBm and a wide dynamic range of over 30 dB over the full -40°C to 85°C temperature range.

2. Structure

Figure 1 depicts a block diagram of our burst-mode optical receiver. The optical receiver consists of a photodiode (PD), a transimpedance amplifier (TIA) with variable-transimpedance, and a burst-mode limiting amplifier with the ATC circuit.

To reduce the threshold level fluctuation, we devised the “ATC Stabilizer Method” as shown in Figure 2. The ATC circuit, which consists of the peak hold circuit, bottom hold circuit, and a 1/2 circuit, detects the middle level of the signals as the threshold level during preamble time. The ATC stabilizer fixes the peak and bottom hold levels after detecting the preamble.

Figure 3 plots the simulation results of the pulse width of the output signals at 85°C, assuming a worst-case data stream with long successive 1’s or 0’s. As shown in Figure 3(a), the pulse width fluctuation of the output signals was ±300 ps when the ATC Stabilizer was not
used. However, the pulse width fluctuation of the output signals was reduced to ±150 ps when the ATC Stabilizer was used, as shown in Figure 3(b). This result indicates that the ATC stabilizer reduced the pulse width fluctuation by half.

An avalanche photodiode (APD) was used as the PD in this experiment, and the pseudo random binary sequence (PRBS) part of the packet1 as high as -8 dBm. The receiver exhibited sensitivity of -34.4 dBm and overload of -4 dBm at a BER of $10^{-10}$ over the full -40°C to 85°C temperature range.

The burst-mode optical receiver with our developed chip set can support a wide operating temperature range, and has high sensitivity, a wide dynamic range, and fast settling time, resulting in a G-PON OLT receiver defined in ITU-T Recommendation G.984.2 Class B+.

4. Conclusion

We developed a 1.25-Gb/s burst-mode optical receiver chip set for G-PON OLT with an ATC Stabilizer. This chip set drastically reduced threshold level fluctuation and pattern jitter. The optical burst-mode receiver incorporating our chip set achieved high sensitivity and a wide dynamic range, which complies with specifications defined in ITU-T Recommendation G.984.2 Class B+, over a wide operating temperature range of -40°C to 85°C.

References

[1] ITU-T Recommendation G.984.2

Fig. 3 Output waveform pulse width of receiver

Fig. 4 Output waveform of receiver

Fig. 5 Output waveform of the optical burst

Fig. 6 BER measurement result
Characterization of Metal/Semiconductor/Metal Photo Diode for Optical Interface of Single-Flux-Quantum Circuit

Satoshi Shinada, Hirotaka Terai, Naoya Wada, Zhen Wang, Tetsuya Miyazaki
National Institute of Information and Communications Technology,
4-2-1 Nukui-Kitamachi, Koganei, Tokyo, 184-8795 Japan
E-mail : sshinada@nict.go.jp

Abstract  We fabricated and characterized a metal/semiconductor/metal photo diode for optical interface of a single flux quantum circuit, which could be a high-speed and low-power-consumption buffer memory in the optical packet switch.

1. Introduction

An optical packet switch [1] is an important technique for a future optical packet network. However, a buffer memory which can avoid the packet collision is one of components to be improved for the practical use. A conventional buffering technique using a fiber delay line can make a system huge with increasing a number of ports. Also, a semiconductor memory such as a SRAM can be used, but the physical size and power consumption required for a serial parallel conversion might be problems as the system scale increases.

We have proposed to use a single-flux-quantum (SFQ) circuit [2] as the buffer memory [3]. This circuit can operate as logic and/or memory circuits over 100 GHz clock frequency with extremely low power. By applying the SFQ circuit to the buffer memory, optical packet can be stored without reducing its data rate. The most serious problem in the SFQ technology is the interface between room temperature and cryogenic environment. In spite of the high internal clock, the input/output bit rate is limited to 10 Gbps due to a large thermal load through a metal coaxial cable. Therefore, the optical interface with smaller thermal load is desirable for higher data transmission. Figure 1 shows a schematic of the SFQ circuit with optical input and output fibers. A mutual conversion between optical and SFQ signals is an important issue.

In this paper, we fabricated and characterized a metal/semiconductor/metal photo diode for the optical input interface of the SFQ circuit.

2. Implementation of SFQ circuit and photo diode

We proposed to use the multi-chip module (MCM) technique for an implementation of a SFQ circuit and a photo diode (PD) as shown in fig. 2. The SFQ chip and the PD chip are connected by using a flip chip bonding technique on same wafer. The transmission through the solder bump is over 100 GHz and a low loss transmission between the chips is realized by superconducting micro strip lines in the cryogenic environment. We used a metal/semiconductor/metal photo diode (MSM-PD) as a detector for the input interface. The MSM-PD has a simple interdigitated finger electrodes and a flat structure, therefore implementation with a SFQ circuit can be easy. For exciting the SFQ pulse by the communication wavelength band optical signal (1550 nm band), we fabricated a MSM-PD on InGaAs which was grown on the InP substrate, which could become the MCM wafer.

3. Fabrication and characterization of MSM-PD

Figure 3 shows a schematic structure of a MSM-PD fabricated on the InGaAs/InP wafer. The top layer of InAlAs was for suppressing the dark current [4]. The interdigitated finger was fabricated by using an electron-beam lithography and a metal lift off process. A photo current of MSM-PD was conducted through a coplanar waveguide.
Figure 4 shows current and voltage characteristics of MSM-PD with 0.3 µm wide lines and 0.9 µm wide spaces as a function of the input power (CW laser). The sensitivity of 0.6 A/W was obtained at a bias voltage of 0.5 V and was sufficient to excite a SFQ even by the power using in a conventional fiber communication without an amplifier.

For evaluating the time response of MSM-PD, we used an experimental setup as shown in fig. 5. An optical pulse with FWHM of 1.5 psec was generated by a mode-locked laser diode (MLLD) and irradiated to the MSM-PD, which was mounted in a liquid helium flow type cryostat with an optical access window. The photo current was output from the electrical output port. Figure 6 shows the pulse response of MSM-PD with 0.3 µm wide lines and 1.2 µm wide spaces. The FWHM of 30 psec and the decay time of less than 100 psec were obtained at both room temperature and cryogenic environment, and the estimated bandwidth is about 10 GHz. The higher response would be realized by a MSM-PD with narrower space.

4. Conclusions

We have proposed to use a SFQ circuit as an optical buffer memory in the optical packet switch and developed an optical input interface for exploiting the high speed throughput of SFQ circuit. The MCM technique could realize the implementation of some functional SFQ circuits and optical modules on one wafer. We fabricated the MSM-PD, which was suitable for MCM wafer, and characterized it. The sufficient sensitivity for exciting a SFQ was obtained in the cryogenic environment. The bandwidth of 10 GHz would be increased by a MSM-PD with narrower spaces.

References
Novel Optical Interconnection for Silicon-on-insulator Waveguide Tap Monitor

Shih-Hsiang Hsu
Department of Electronic Engineering
National Taiwan University of Science and Technology
No. 43, Sec. 4, Keelung Rd., Taipei, Taiwan
TEL: +886-2-2737-6399
FAX: +886-2-2737-6424
E-Mail: shhsu@mail.ntust.edu.tw

Abstract: A photodetector can be applied onto the silicon-on-insulator waveguide tap to monitor light signals on waveguide. The reflective silver metal layer integrated with 54.7-degree angles from silicon wet-etching demonstrated low polarization-dependent-loss of 0.2 dB.

Introduction
One or more tap waveguides, implemented in the optical components for monitoring, is usually utilized to extract a portion of the light signal traveling along a primary waveguide. Furthermore, a photodetector can be positioned onto tap waveguides to receive the light signals on the tap waveguide. Since a portion of the light signal from the primary waveguide is carried by the tap waveguide, the output of the photodetector can be utilized to indicate characteristics on the primary waveguide. The tap waveguide typically ends at the edge of an optical component. The photodetector is then positioned at the edge of the optical component over the end of the tap waveguide and receives the light signals directly from the tap waveguide. As the complexity of optical circuits formed on optical chips increases, many tap waveguides cannot be employed and terminated at an edge of an optical component. Hence, there is a need for a waveguide tap that is suitable for use with complex optical circuits and can be positioned on a wafer base.

Recently silicon-on-insulator (SOI) is a critical platform for integrated optoelectronic circuits since it offers the potential of monolithic integration for photonic and electronic functions on a single substrate [1,2]. Integrating photonics functions on a silicon platform will be a low cost solution if integrated optoelectronics circuits are feasible.

In a rib SOI waveguide, the interface roughness from core/cladding and etched silicon surfaces mainly plays the role of the scattering centers to contribute the optical propagation loss. If the SOI waveguide design is appropriate, the polarization dependent loss (PDL) can be controlled within 0.1 dB. 5 µm thick SOI waveguide was a good choice to demonstrate low birefringence and efficient interconnection [3]. Nevertheless many tap waveguide monitors/photodetectors are associated with an undesirably high level of PDL, which means that the detector will sense a result of significantly different polarization modes. Because the tap and primary SOI waveguides are usually showing very small PDL (<0.2 dB), the distribution of different polarity modes received by the photodetector cannot represent and monitor the polarization modes in the primary waveguide of an optical component. Accordingly, there is a need for a tap waveguide arrangement where the output of the photodetector can represent the conditions in the primary waveguide.

Optical Interconnection for Waveguide Tap Monitoring
A novel optical interconnection was demonstrated by a total internal reflection (TIR) mirror on the GaAs based platform. The TIR mirror was etched by chemically assisted ion beam etching (CAIBE) to couple the optical signal from the waveguide into the detector for monolithically optical interconnections [4]. This approach requires a special CAIBE etcher and a complex dry etch processing development. Another simple optical interconnection is utilizing the anisotropic wet etch on silicon, which can be implemented by an alkaline solution, KOH, and will be stopped at (111) planes slanted 54.7° for (100) silicon wafer, shown in Fig.1. The angles of 35.3°, less than the angle for total internal reflection, can be utilized as the input angle from SOI waveguide to polyimide medium. The silver metal on top of SiO2 in the direction-changing-region is reflecting the optical signals, traveling along the tap waveguide, away from the base toward a photodetector via an input angle of 81.08°.

Fig. 1(a). Top view of an optical component having a primary waveguide and a tap waveguide positioned on a base. (b) A cross section of the optical component, taken at the dashed line labeled in (a).

In Fig. 2(a), the ray optics simulation shows that the PDL is around 0.1 dB at the input angle of 35.3° for the transmission from silicon to polyimide layers. The positive PDL value means that the optic loss of transverse-magnetic (TM) polarization is higher than transverse-electric (TE). In Fig. 2(b), an aluminum metal, frequently utilized in silicon processing, was taken to calculate the PDL of around 0.75 dB, contributed by the reflection from polyimide medium to aluminum metal at the input angle of 81.08°. The net PDL through silicon, polyimide, and aluminum is -0.7 dB. If the reflective light was simulated for different kinds of reflective metals with the variation of the real and imaginary parts of refractive index from 0.1 to 5, the PDL contributed by the polyimide and metals at 81.08° is shown
The usual metals utilized in the processing laboratory are listed in Table 1. The corresponding PDL performance can be demonstrated between -0.3 and -4.1 dB, which means that the propagation loss from TE polarization mode is always higher than TM. A silver (Ag) reflective metal integrated with SOI waveguide and polyimide medium is a suitable layer in directing light signal up to the photodetector, excluding SOI tap waveguide in the PDL budget, to achieve the lowest net PDL of -0.2 dB, +0.1 dB coming from the transmission from silicon to polyimide layers and -0.3 dB coming from the reflection of silver layer.

### Table 1. Refractive indices of different metals

<table>
<thead>
<tr>
<th>Metal</th>
<th>$n$</th>
<th>$k$</th>
<th>PDL between polyimide and metal</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cu</td>
<td>0.606</td>
<td>8.26</td>
<td>$-0.4$</td>
</tr>
<tr>
<td>Ag</td>
<td>0.469</td>
<td>9.32</td>
<td>$-0.3$</td>
</tr>
<tr>
<td>Al</td>
<td>0.559</td>
<td>0.81</td>
<td>$-0.6$</td>
</tr>
<tr>
<td>W</td>
<td>2.22</td>
<td>4.85</td>
<td>$-1.6$</td>
</tr>
<tr>
<td>Ni</td>
<td>3.38</td>
<td>0.82</td>
<td>$-2.0$</td>
</tr>
<tr>
<td>Ti</td>
<td>4.04</td>
<td>3.82</td>
<td>$-3.1$</td>
</tr>
<tr>
<td>Cr</td>
<td>4.24</td>
<td>4.81</td>
<td>$-3.3$</td>
</tr>
<tr>
<td>Pt</td>
<td>5.11</td>
<td>7.04</td>
<td>$-4.1$</td>
</tr>
</tbody>
</table>

PDL $< 0$ means that TE loss is higher than TM.

It is also clear that as the waveguide is increased in size, the TE and TM modes become similar and the polarization dependence is reduced [5]. A beam propagation method employed by BeamPROP shows that the PDL for SOI waveguides will be increased and the propagation loss from TM polarization is higher than TE when the waveguide dimensions are getting compatibly small with the order of the operating wavelength. In that way, the PDL from the tap SOI waveguides of thinner core will not be as low as 5 µm thickness. Platinum (Pt) and chrome (Cr) are then the suitable layers to compensate this kind of higher PDL from the small core of tap SOI waveguides.

If the polyimide medium is replaced by the air, the incident angle for the reflection on the metal layer is changed to 74.5° and the PDL will be different in the direction-changing-region area. For example, the PDL of -0.7 dB for an aluminum in the polyimide medium will be reduced to -0.3 dB in the air medium. Under this situation, more than one metal will be qualified to be the reflective layer in the tap waveguide monitoring applications.

**Summary**

A novel optical interconnection was demonstrated via a SOI waveguide tap monitor using the silver metal as the reflective layer under the detector to get the lowest PDL. The different metals as reflectors and filling media for reflection can be resources to compensate the significant PDL contributed by other optical components utilized in the waveguide tap monitoring system.

**References**

High-Speed (>55 GHz) Electro-Absorption Modulator Based on Two-Step Undercut Active Region Waveguide

T. H. Wu, F. Z. Lin, D.R. Lee and Y. J. Chiu
No 70, Lien Hai Rd., Kaohsiung, Taiwan R.O.C.
Institute of Electro-Optical Engineering, NSYSU
Email: d943050003@student.nsysu.edu.tw
Tel: +86-7-525-2000 ext 4472, Fax: +86-7-525-4499

Abstract

Higher than 55GHz electro-optical (EO) response was measured in electroabsorption modulator (EAM) fabricated by using the two-step-undercut-etching-active-region-waveguide (TSUEARM) method. It is shown that the high-speed EO performance can be obtained by simultaneously reducing cladding resistance, parasitic capacitance and the electrical insertion loss.

1. Introduction

High-speed electro-absorption modulators (EAM’s) are getting more and more interest for use in both digital and analog high-speed fiber-optic links. A 50-GHz bandwidth lumped EAM (L-EAM) has been reported with the active waveguide length shortened to 63μm [2]. However, reducing the waveguide length results in smaller modulation efficiency due to both shorter modulation length and smaller power handling ability. To overcome the bandwidth limit without severely sacrificing modulation efficiency, the traveling-wave EAM (TW-EAM) has been proposed and experimentally investigated by several authors [3]. The TW-EAM speed is still restricted by the microwave properties inside the p-i-n waveguide, including of impedance mismatch, velocity mismatch and microwave propagation loss. We have previously proposed undercut-weting-active-region (UEARW) structure to enhance the microwave propagation loss by reducing the cladding layer resistance to [1]. However, the capacitance is higher attributed to the parasitic capacitance of the undercut region between p- and n-type cladding layers (as shown in figure 1). In order to reduce this parasitic capacitance, two-step-undercut-etching method was performed to increase the gap between the cladding layers at the undercut region (as shown in figure 1). As compared with the UEARW structure, there is about 62.4% reduction in resistance-capacitance (RC) product leading to >55-GHz electro-optical (EO) response in TSUEAW EAM.

Fig. 1: Schematic structure of the device.

Fig. 2: The optical transmission against voltage for EAM.

2. Fabrication and Measurement

The semiconductor material is grown on semi-insulated InP by metal-organic chemical vapor deposition (MOCVD) method. The InGaAsP MQW consisted of ten tensile-strained QWs (10.4 nm) and barriers (7.6 nm). The tensile-strained QW was used to reduce the polarization dependence of the electroabsorption effect. The top p-cladding layer is 1.9 μm. A 6-μm wide, 1.5-μm deep (0.4 μm above...
MQW) P-InP ridge was first defined by HBr based solution. After covering the ridge by photoresistor to protect the side wall, the HCl : H3PO4 solution was used to selectively undercut etch the InP. About 2.8-μm width of MQW was formed by using H3PO4:H2O2:H2O solution. The p-type (Ti-Pt-Au) and n-type (Ni-AuGe-Ni-Au) metallization are deposited by an e-beam evaporator. PMGI (MicroChem Inc.) is used for waveguide passivation, planarization and bridging the coplanar waveguide (CPW) electrodes for the input and output microwave feed lines. The schematic plot of the device is also shown in Figure 1.

Light at wavelength 1550 nm was launched into the 300-μm-long EAM for the optical transmission against bias measurement. The coupling fibers are lensed fibers with 3-μm full-width-half-maximum (FWHM) mode size. The fiber-to-fiber insertion loss is 15 dB. As illustrated in Figure 2, over 20-dB/volt modulation efficiencies for both TE and TM polarization were obtained in this device. A high-speed vector network analyzer (Anritsu) is used to measure the device microwave S-parameters and EO response. Less than -7dB of S11 (<40GHz) and over 20 GHz 3-dB microwave transmission (S21) is obtained in TSUEARW EAM (Figure 3) showing the better microwave properties as compared with the UEARW EAM [6]. The measured EO response is given in Figure 4, which show the 3-dB response can exceed 55 GHz. The high-speed performance was analyzed by using the equivalent circuit for the TSUEARW EAM transmission line. It is found that the high EO response is due to the small cladding layer resistance associated with small intrinsic capacitance of the TSUEARW structure.

3. Conclusion

Further reduction in parasitic capacitance of UEARW EAM was achieved by using TSUEARW method. Over 55 GHz EO response was observed in TSUEARW EAM. Polarization insensitivity in optical transmission can be found resulting from the use of the tensile QWs. The modulation efficiency is 25dB/v.

4. Acknowledgements

The authors would like to thank the financial supports from the National Science Council, Taiwan (NSC 93-2215-E-110-018) and (NSC 93-2215-E-110-013), Technology Development Program for Academia (92-EC-17-A-07-S1-025) and “Aim for the Top University Plan Taiwan”.

Fig. 3: Electrical transmission property.

Fig. 4: High-speed EO response of EAM.

5. References

Polymeric Integrated Optical Devices for ROADM Systems
Polymer-Based Waveguide Grating Devices

Kin Seng Chiang
Department of Electronic Engineering, City University of Hong Kong, 83 Tat Chee Avenue, Hong Kong, China
* Tel: +852-2788-9605, Fax: +852-2788-7791, Email: eeksc@cityu.edu.hk

Abstract
This paper reviews the development of widely tunable long-period grating devices fabricated with polymer waveguides, including band-rejection filters, bandpass filters, variable attenuators, and add/drop multiplexers.

1 Introduction
Long-period waveguide gratings (LPWGs) [1] have been proposed with the purpose of relaxing the geometry and material constraints of long-period fiber gratings (LPFGs) for the realization of optical filters. Because the transmission characteristics of a LPWG depend critically on the physical dimensions of the waveguide [1,2], which can be controlled accurately with the microfabrication technology, there exists much flexibility in the design of LPWG devices. Over the last few years, there has been significant progress in the development of LPWG devices with different waveguide structures and different materials. Polymer-based LPWGs have attracted special attention because of their excellent thermal tunability and other desirable features, such as easy fabrication and strong photosensitivity. This paper provides a brief review of the development of polymer LPWG devices.

2 Fabrication Techniques
LPWGs have been demonstrated with polymer-clad ion-exchanged glass waveguides [3], all-polymer channel waveguides [4,5], and rib waveguides [6]. In these studies, the grating is formed on the surface of the core of the waveguide by the conventional photolithography and reactive ion etching (RIE) techniques. While this conventional fabrication process allows an accurate control of the waveguide dimensions, it is slow and expensive. To take advantage of the molding property of polymer, a simple and low-cost nano-imprint lithography technique has been demonstrated for the fabrication of a LPWG in a polymer rib waveguide [7]. This technique has the potential of being further developed for mass production of LPWGs. Meanwhile, a direct ultra-violet (UV) writing technique based on a 248-nm KrF excimer laser has been demonstrated for the fabrication of LPWGs in polymer ridge waveguides [8]. This technique is practically the same as the UV-irradiation technique for the fabrication of LPFGs. It allows in-situ monitoring of the growth of the transmission spectrum of an LPWG and thus relaxes the fabrication tolerances significantly.

3 Devices
A LPWG, which couples light between the guided mode and the cladding mode of a waveguide, is inherently a band-rejection filter. For the implementation of a thermally tunable filter, the temperature dependence of the center wavelength of the rejection band of a LPWG has been studied both theoretically and experimentally for different waveguide structures [5,8]. The temperature sensitivity of a LPWG is found to depend mainly on the dispersion properties of the coupling modes and the difference of the thermo-optic coefficients of the core and the cladding and can be controlled over a wide range by choosing suitable waveguide dimensions and materials. The temperature sensitivity of a LPWG is particularly sensitive to the cladding thickness for a ridge waveguide [8], while it is more sensitive to the core size for a channel waveguide [5]. Thanks to the large difference (~25 times) between the thermo-optic coefficients of polymer and glass, a LPWG formed in a polymer-clad glass waveguide can provide wavelength tuning over the entire (C+L)-band with a temperature control of only ~10 °C [3], which corresponds to a sensitivity of ~9 nm/°C. By relaxing the stress in an all-polymer LPWG through etching of the cladding width, a polarization-insensitive center wavelength tunable over the (C+L)-band with a temperature control of ~8 °C has been achieved [4]. On the other hand, the principle of realizing
a temperature-insensitive polymer LPWG by controlling the cladding thickness has been demonstrated recently [9]. Post-processing of the cladding profile along a uniform LPWG has also been demonstrated for the generation of more sophisticated transmission characteristics by using the conventional etching technique [10] or the UV-irradiation technique [11]. In addition, a widely tunable polymer LPWG bandpass filter has been fabricated, which consists of two identical gratings placed in series and separated by a gap to block the light at off-resonance wavelengths from passing through [12]. The filter shows a peak transmission efficiency of ~60% and the passband can be tuned over the (C+L)-band with a temperature control of ~30 °C.

The large thermo-optic effect in polymer material has been explored for the realization of dynamic LPWGs [13-15], where the grating is induced thermally by applying current to an electrode grating deposited on a polymer waveguide. The transmission spectrum of such a filter can be adjusted dynamically by controlling the electric powers applied to the electrode grating. In one device, the contrast of the rejection band can be tuned by ~28 dB with 45-mW electric power applied to the metal grating and the center wavelength can be tuned over the (C+L)-band with a temperature control of ~40 °C [15]. Dynamic LPWGs may find applications as tunable notch filters, polarization-dependent loss compensators, or variable attenuators.

To further increase the functionality of LPWGs, the structure of two parallel LPWGs for the realization of a broadband add/drop multiplexer has been proposed and analyzed in detail [16]. A peak coupling efficiency of ~34% and a wavelength tuning sensitivity of 4.7 nm/°C has been demonstrated experimentally [17]. The use of parallel LPWGs has the clear advantage over parallel LPFGs in device packaging. It is easy to put more waveguide gratings on the same substrate to form multi-port couplers and add/drop multiplexers [17,18].

4 Conclusion
The potential of developing new devices through the manipulation of the cladding modes with LPWGs is large. Polymer LPWGs have the additional advantage of providing excellent thermal tunability. LPWGs share many of the features of LPFGs, yet enjoy the flexibility offered by the optical waveguide technology. It is envisaged that the activities in the development of LPWG devices to fulfill more functions will continue to increase.

This research was supported by a grant from the Research Grants Council of the Hong Kong Special Administrative Region, China, under Project CityU 112005.

5 References
A Planar-waveguide 8-channel Arrayed Wavelength Division Multiplexing Coupler using Directional Coupler with Higher-order Mode cutting Filter for Waveguide Amplifier

Takara Sugimoto¹, Yukari Deki²,³, Takeshi Takeuchi¹,², Sekizen Takaesu²,³ and Yutaka Urino²,⁴

¹System Platform Research Laboratories, NEC Corporation
²Optoelectronic Industry and Technology Development Association (OITDA)
³R&D Support Center, NEC Corporation
⁴Fundamental and Environmental Research Laboratories, NEC Corporation

34, Miyukigaoka, Tsukuba, Ibaraki, 305-8501, Japan
E-mail: t-sugimoto@dc.jp.nec.com, Phone: +81-298-50-2619, Fax: +81-298-50-1106

Abstract
We have developed a planar-waveguide 8-channel arrayed WDM coupler for EDWA application. The coupler is compact and can be directly connected to an EDWA chip. The insertion loss is less than 2.0 dB over the full C-band.

Introduction
The erbium-doped fiber amplifier (EDFA) has played a key role in the development of broadband fiber networks, which now span the globe. The challenge now is to reduce their size and cost. A candidate for meeting this challenge is the erbium-doped waveguide amplifier (EDWA) integrated with a wavelength division multiplexing (WDM) coupler [1,2]. However, it has two drawbacks: a large propagation loss for C-band signals and large crosstalk from the pump light to the signal light during demultiplexing, which degrades the noise figure (NF). We have thus taken a different approach and have developed an arrayed WDM coupler with high-β SiON waveguides, which are compact, have low loss and low crosstalk, and can be directly connected to an EDWA chip.

Design
As shown in Figs. 1 (a) and (b), our WDM coupler consists of a spot-size converter (SSC) for Hi-NA fiber, a higher-order mode cutting filter for the pump light, a directional coupler (DC), a pitch converter, and another SSC for the EDWA. The higher-order mode cutting filter consists of a tapered waveguide and an S-bend waveguide, as shown in the inset of Fig. 1 (b). Simulation results for the directional coupler during demultiplexing with and without the mode filter are shown in Fig. 2. The directional coupler was designed to demultiplex C-band signals and pumping light at 980 nm to cross and bar ports, as shown in Figs. 2 (a) and (b), respectively.

Fig. 1: Design of WDM coupler: (a) structure of the 8-channel arrayed WDM coupler, (b) one of the 8-channel arrayed WDM coupler.

Single-mode propagation is essential for low crosstalk. Because the wavelength of the pump light is much shorter than that of the signal light, if there is no mode filter, higher order modes of pump light can easily propagate along the directional coupler and then couple to the cross port, as shown in Fig.2 (d). This crosstalk degrades the NF of EDWA systems. Therefore, we added a higher-order mode cutting filter to suppress the crosstalk. In the wider portion of the tapered waveguide, both fundamental mode light and higher-order mode pump light can propagate. At the end of the waveguide, the fundamental modes of the signal and pump lights are still strongly confined in
the core, as shown in Figs. 2 (a) and (b), respectively. The higher-order mode lights are weakly confined and radiate at the S-bend waveguide, as shown in Fig. 2 (c). Simulation showed that, with the mode filter, the excess loss for the fundamental mode signal light was less than 0.1 dB, and the crosstalk was reduced by more than 24 dB.

![Mode filter](image)

**Fig. 2:** Simulated operation of directional coupler during demultiplexing: (a) fundamental mode signal light with filter, (b) fundamental mode pump light with filter, (c) 1st order mode pump light with filter, and (d) 1st order mode pump light without filter

**Performance**

We fabricated an 8-channel arrayed WDM coupler with a higher-order mode cutting filter. The chip had a footprint of 11 × 5 mm.

As shown in Fig. 3, the insertion loss of the coupler was less than 2.0 dB, and the polarization-dependent loss was better than 0.2 dB over the full C-band. The estimated fiber-to-PLC coupling loss was about 0.7 dB/facet, and the estimated excess loss of the mode filter and directional coupler was about 0.5 dB. As shown in Fig. 4, the inter-channel loss uniformity was better than 0.3 dB. The crosstalk between the pump and signal lights was less than –30 dB. These characteristics are suitable for EDWA application.

![Insertion loss vs Wavelength](image)

**Fig. 3:** Measured wavelength-dependent loss of WDM coupler

![Insertion loss vs Channel](image)

**Fig. 4:** Inter-channel loss uniformity of WDM coupler @1550 nm

**Conclusion**

We have developed a planar-waveguide 8-channel arrayed WDM coupler for multiplexing and demultiplexing C-band signals and 980-nm pump light for EDWA applications. It uses a directional coupler with a higher-order mode cutting filter. The WDM coupler chip is compact and can be directly connected to an arrayed EDWA chip. The insertion loss is less than 2.0 dB over the full C-band, and the inter-channel loss uniformity is better than 0.3 dB. The crosstalk between the pump and signal lights is less than –30 dB. This coupler is thus suitable for EDWA application.

**Acknowledgements**

A part of this work was performed under the management of OITD supported by NEDO.

**References**


Refractive-index difference dependence of loss reduction in Y-branch waveguides designed by wavefront matching method


NTT Photonics Laboratories, NTT Corporation, 3-1 Morinosato Wakamiya, Atsugi, Kanagawa, 243-0198 Japan,
Tel: +81-46-240-4093, Fax: +81-46-240-4528, Email: sakamaki@aecl.ntt.co.jp

Abstract

We designed low-loss Y-branch waveguides by using our developed wavefront matching method. We demonstrated the loss reduction principle experimentally and revealed the possibility of improving the performance of waveguide devices with Y-branches.

1. Introduction

Y-branch waveguides play important roles as optical signal dividers/combiners in optical waveguide devices. A Y-branch made with planar lightwave circuit (PLC) technology has such advantages as low excess loss, low wavelength-dependent loss and low polarization-dependent loss. As a result silica-based 1 × N splitters consisting of cascaded Y-branches [1] are now widely used for optical signal distribution in passive optical networks. In addition, Y-branches are becoming increasingly important in functional PLC devices such as the differential quadrature phase-shift-keying modulator [2], because of recent developments in the integration of PLC devices.

To improve the performance of waveguide devices with Y-branches, it is very important to reduce the Y-branch excess loss. The excess loss of the conventional Y-branch shown in Fig. 1 (a) results from the mode mismatch at the branching point, namely the interface between a tapered waveguide and two output waveguides, due to the gap caused by the limitations of the fabrication process. Several approaches have been proposed to overcome this problem, and they include the introduction of segmented structures [3], vertical tapered structures [4] and a multimode interference waveguide [5].

In this paper, we compare experimental results obtained using samples with different refractive-index differences (Δ), and show that this method is very efficient for realizing higher Δ waveguides suitable for highly integrated PLC devices.

2. Design

If we are to reduce the excess loss of a Y-branch, we must optimize the tapered waveguide pattern to convert the fundamental mode of the input waveguide φ(x, 0) into the fundamental local normal mode of two output waveguides ψ(x, L), which has a double-peak intensity profile, where L is the tapered waveguide length. We designed such a Y-branch by using the WFM method, which can synthesize the optimum waveguide pattern from the input and desired output fields [6].

In this study, we used φ(x, 0) as the input field and ψ(x, L) as the desired output field. The principle of the WFM method is described simply below. The WFM method is based on the fact that the coupling efficiency between φ(x, 0) and ψ(x, L) equals the overlap integral between the forward-propagating input field φ(x, z) and the backward-propagating desired output field ψ(x, z).

The overlap integral at z can be increased by matching the wavefronts of φ(x, z) and ψ(x, z). Therefore, if we modify the refractive-index distribution at z so as to match the two wavefronts, the output field φ(x, L) becomes closer to ψ(x, L), which leads to a loss reduction. The optimum waveguide pattern is obtained by scanning the optimization point z from 0 to L.

As previously described, the excess loss of a Y-branch results from the mode mismatch caused by the gap between the two output waveguides. In particular, the mode mismatch becomes more significant for higher Δ waveguides. Thus, we designed Y-branches for 0.45%, 0.75%, 1.5% and 2.5%−Δ waveguides to compare the effectiveness of the WFM design. The length of the optimized region and the core size of the input and output waveguides are shown in Table 1. To clarify the
potency of the WFM method, we set the minimum gap width to 3 μm, which is larger than that of the commercially available PLC splitters. Fig. 1 (b) shows the Y-branch for a 2.5%-Δ waveguide designed using the WFM method. As shown in Fig. 1 (b), the WFM-designed Y-branch has an periodically modulated core width. Figs. 2 (a) and (b) show the waveform transients of the optical power calculated using the beam propagation method (BPM), at a wavelength of 1550 nm. As shown in Fig. 2 (b), the WFM-designed waveguide generates a double-peak profile at z=L and resolves the problem of mode mismatch.

3. Results

The designed Y-branches were fabricated using silica-based PLC technology including flame hydrolysis deposition, photolithography and reactive ion etching.

Fig. 3 shows the measured excess losses of the fabricated Y-branches, as determined relative to the reference waveguide loss and intrinsic 3 dB splitting loss, at a wavelength of 1550 nm. In Fig. 3, the dotted lines show the loss calculated by BPM for the conventional and WFM-designed Y-branches. It is clearly shown that the WFM-designed Y-branches have a much lower loss for every Δ waveguide. As Δ increases, the advantage of the WFM method becomes more obvious. Fig. 4 shows the measured excess loss of 2.5%-Δ Y-branches as a function of wavelength. We found that the loss of the WFM-designed Y-branch is lower than that of the conventional Y-branch across the entire wavelength range. These results show that the WFM method enables us to improve the performance of Y-branches, particularly for higher Δ waveguides suitable for highly integrated PLC devices. Note that the minimum gap width was set at 3 μm in this study. The conventional PLC process can provide a narrower gap, and the absolute insertion loss values of commercial Y-branch splitters are lower than that in this report. Of course, the WFM method has a substantial loss reduction effect for narrow gap values [7].

4. Conclusion

We demonstrated low-loss Y-branches designed with the WFM method for waveguides with several different Δ values. We believe that the WFM designed Y-branches will enable us to improve the performance of splitters and integrated PLC devices consisting of Y-branches.

5. References

Two-dimensional photonic crystal devices are characterized by a largely periodic patterning of the dielectric, through nanofabrication techniques, on a scale that is shorter than the optical wavelength at which the device operates. Patterning on this length scale allows us, in principle, to engineer the electromagnetic properties of photonic devices in microscopic detail. This patterning often results in structures for which no analytical solutions of Maxwell’s equations are known. As a result, it is a serious challenge to understand how to utilize this freedom to improve device performance. Nevertheless, a great deal of progress in photonic crystal device development has been made in the past few years.

This tutorial presentation will discuss basic photonic crystal design and analysis techniques and illustrate design approaches with examples of device results. The presentation will describe the optical loss mechanisms in photonic crystal resonant cavities and I will discuss photonic crystal lasers with particular emphasis on devices capable of room temperature CW operation. The presentation will also address device issues associated with passive photonic crystal components such as optical loss, waveguide dispersion, and the design of waveguide junctions.
Dr. O’Brien received the B.S. degree from Iowa State University in Electrical Engineering in 1991 and the M.S. and Ph.D. degrees in Applied Physics from the California Institute of Technology in 1993 and 1996, respectively.

In 1997 he joined the Department of Electrical Engineering at the University of Southern California as an Assistant Professor. In 1999 he received the Presidential Early Career Award for Scientists and Engineers, and in 2000 he was awarded an NSF Career award. In 2003 he became an Associate Professor, and he was promoted to Professor of Electrical Engineering in 2006. His research interests are in nanophotonics and photonic crystal devices.

40-Gbit/s operation of dispersion compensator based on multiple 1D coupled-defect-type photonic crystals

Misuzu Sagawa¹2,3, Shigeo Goto¹2,3, Kazuhiko Hosomi¹2,3, Toshiki Sugawara²3, Toshio Katsuyama¹3 and Yasuhiko Arakawa¹

¹Nanoelectronics Collaborative Research Center, Univ. of Tokyo,
²Central Research Laboratory, Hitachi Ltd.,
³Optoelectronic Industry and Technology Development Association

1-280 Higashi-koigakubo, Kokubunji, Tokyo 185-8601, Japan
Tel: +81-42-323-1111 Fax: +81-42-323-7673 email: misuzu.sagawa.wx@hitachi.com

Abstract: We have demonstrated 40-Gbit/s dispersion compensation operation over a 10-km-long standard single-mode fiber using a dispersion compensator based on multiple one-dimensional coupled-defect-type photonic crystals.

1. Introduction
Chromatic dispersion compensators are strongly required for the next-generation fiber network systems with the channel rates of over 40 Gbit/s because of their low dispersion tolerance. Accordingly, many types of dispersion compensators have been developed and demonstrated[1-7]. Multiple one dimensional (1D) coupled-defect-type photonic crystals is another promising candidate of a dispersion compensator because of its large optical group velocity dependence on the wavelength. To exploit these characteristics, we proposed a dispersion compensator based on multiple 1D coupled-defect-type photonic crystals and preliminarily demonstrated its dispersion compensation operation[8,9]. In this paper, we have successfully demonstrated 40-Gbit/s dispersion compensation operation over a 10-km-long standard single-mode fiber.

2. Device structure and optical property
The structure of one dimensional coupled-defect-type photonic crystals for the dispersion compensator is shown in Fig. 1. It consists of three sets of 13 pairs of SiO₂/TiO₂ distributed Bragg reflector (DBR)-layers and two SiO₂ defect layers (3λ₀) sandwiched between two other defect layers (0.5λ₀) and 7 pairs of SiO₂/TiO₂ DBR layers. It was designed for the 1.55-µm, 40-Gbit/s optical communication system. The thin film structure is substrate-free, and this enables the compensator to be as small as a 1.4 mm edge cube. To obtain a large group velocity difference, 60 substrate-free films are stacked. In addition, 10 and 30 layer substrate-free film stacks were fabricated for tuning operation.

Figure 2 shows the structure of the dispersion compensator module. All three stacks were mounted in the module. The tuning operation was carried out by sliding the device mechanically as shown in the figure. The size of the module is 22 x 40 x 24.5 mm.

Group delay spectra for 10, 30 and 60 layer stacks are shown in Fig. 3. The passband was about 2
nm. The group delay time-differences within the band were almost proportional to the number of the film stacks, which indicates that the module could operate as a tunable dispersion compensator. For 60 layer stacks, the delay time-difference was more than 100 psec.

4. Transmission experiment
We carried out a 40 Gbit/s non-return-to-zero (NRZ) optical transmission experiment. The experimental set-up is shown in Fig. 4. NRZ pseudorandom binary sequence (PRBS) data ($2^{15}$-1 word-length) at 40 Gbit/s was directly generated by pulse pattern generator. Data were encoded into a data pulse train with a LiNbO$_3$ Mach-Zehnder modulator. The wavelength of the CW laser was 1549 nm and the transmission link was a 10-km standard single mode fiber.

The eye diagrams obtained after the 10-km transmission are shown in Fig. 5. The 60-layer-stack compensator was used in this experiment. Well-defined eye opening was observed when the dispersion compensator was used, while serious degradation was seen in the eye diagram without the compensator. The corresponding dispersion was 170 psec/nm.

5. Conclusion
We have successfully demonstrated 40-Gbit/s dispersion compensation operations over a 10-km-long standard single-mode fiber by a dispersion compensator based on multiple 1D coupled-defect-type photonic crystals.

Acknowledgement
A part of this work was financially supported by OITDA under contract with NEDO and also by the IT Program of MEXT.

References
A Study on Photonic Crystal Waveguide Based on Absolute Photonic Band Gap

Yuki MORITA Yasuhide TSUJI Koichi HIRAYAMA

Department of Electrical and Electronic Engineering, Kitami Institute of Technology
address : 165 Koen-cho, Kitami-shi, Hokkaido, 090-8507 Japan
Tel : +81-157-26-9279 E-mail : mel06012@std.kitami-it.ac.jp

Abstract: We propose a novel photonic crystal waveguide with absolute photonic band gap which realizes single-mode operation for both TE and TM modes. 60 degree bend and polarization splitter based on proposed waveguide are also demonstrated.

1. Introduction

On two-dimensional photonic crystals (PCs), it is known that absolute photonic band gaps (PBGs) which forbid both TE and TM waves propagation are realized by reducing structural symmetry[1]. We had been studying about compound-type PC of honeycomb and triangular lattices and also have proposed PC waveguide created by line defect with elliptical air holes[2]. In this structure, it was revealed that single-mode operation for both TE and TM waves is realized. On the other hand, when considering to construct planar lightwave circuits, we also need to study about some basic waveguide components, such as waveguide bends. However, it seems to be difficult to create bending waveguide based on the waveguide structure proposed in [2] because of its complexity. In this paper, we propose a novel PC waveguide which can realize single-mode operation and efficient transmission through 60 degree bend. In order to calculate dispersion relations and propagating properties, finite element method (FEM)[3] is used. As a device application of the proposed waveguide, a compact polarization splitter is also demonstrated.

2. Absolute PBG and PC waveguide

Fig. 1(a) shows a compound-type PC structure, where the background relative permittivity is assumed to be 10.0[4], a is a lattice constant, the diameters of air holes, \(d_1\) and \(d_2\) are 0.55a and 0.32a, respectively. These diameters are selected to maximize an absolute PBG bandwidth. In these structural parameters, the dispersion relations calculated by FEM are shown in Fig. 1(b). We can see that the absolute PBG exists between \(a/\lambda = 0.602\) and 0.645.

Now then, we consider PC waveguides which can realize single-mode operation. In [2], PC waveguide which propagation direction corresponds to \(\Gamma-K\) direction has been proposed. On the other hand, we propose a PC waveguide which propagation direction corresponds to \(\Gamma-M\) direction. This waveguide is created by changing the size of air holes along \(\Gamma-M\) direction. This structure is simpler than the conventional waveguide[2], and it seems to be easy to create waveguide bend. Here, we assume the diameter of air holes within line defect is \(d_2\). Fig. 2(b) shows the dispersion relations of the proposed PC waveguide. Here, \(a' = \sqrt{3}a\) is a period of PC waveguide. As shown in Fig. 2(b), a single-mode operation for both TE and TM modes can be achieved between \(a/\lambda = 0.607\) and 0.619. Fig. 3 shows the propagating fields at \(a/\lambda = 0.613\) for both TE and TM modes. These are calculated by FEM with perfectly matched layer (PML)[3]. The calculated normalized transmitted power for TE wave is 0.98, and that for TM wave is 0.54.

3. Polarization splitter

As an example of device application of the proposed PC waveguide, we consider a polarization splitting device based on PC directional coupler. Fig. 4(a) shows the PC coupler, where waveguide separation, \(s\), is set to be 3.5a. Fig. 4(b) shows the dispersion relations.
of even and odd modes of TE and TM supermodes. Fig. 5 shows coupling length of the PC coupler, $L_c$. From Fig. 5, the coupling length for TE mode is much longer than that of TM mode, therefore the PC coupler can be used as a polarization splitter by choosing coupling length for TM mode as a device length. Fig. 6 shows the propagating fields at $a/\lambda = 0.613$ for TE and TM waves in the polarization splitter, where TE or TM fundamental wave is inputted into the lower waveguide. As shown in Fig. 6, TE and TM waves are separated and output to the lower and upper ports, respectively.

4. Conclusion
We proposed the novel PC waveguide which propagation direction corresponds to $\Gamma$-$M$ direction. We showed that single-mode operation for both TE and TM modes is realized in the proposed PC waveguide. And also, it was shown that efficient transmission waveguide bend is realized for TM wave. The application of this PC waveguide to the polarization splitter was also demonstrated. Optimization in order to improve transmission property for TM mode and other device applications with this PC waveguide are now under consideration.

Reference
Bonded photonic crystal components and circuits: Toward 2.5-D micro-nano-photonics


INL (Lyon Nanotechnology Institute), UMR CNRS 5270, Ecole Centrale de Lyon, 36 avenue Guy de Collongue, F-69134 Ecully Cedex, France

Tel: +33 4 72 18 60 81, Fax: +33 4 78 43 35 93, e-mail: Xavier.Letartre@ec-lyon.fr

Abstract: We will present various surface addressable active devices and surface emitting microlasers. These 2.5D structures combine planar photonic crystal resonators and vertical stacks. They can be fabricated using InP/Silicon wafer bonding, micro-nanopatterning and MOEMS technology.

Introduction:
Planar photonic crystal (PC) structures fabricated on semiconductor membranes are among the best candidates to build ultra-compact non linear optical devices including microlasers, modulators or switches. Various designs including high quality factor PC microcavities were proposed and could be used as basic building blocks for highly integrated photonic integrated circuits. In order to exhibit a low operation threshold, PC cavities should be designed in such a way to minimize the vertical optical losses, while maintaining a limited modal volume.

A radically different approach consists of the exploitation of slow light optical modes that stand over the light line of PC structures. The goal is then not to inhibit, but to control and to use the interplay between “in-plane” resonances in PCs and free space modes and “vertical” resonances. Enabling manipulation of light in the three dimensions of space by using a combination of intrinsically planar 1D or 2D PC structures is referred to as a “2.5D micro-nanophotonics” approach.

In order to stack such membranes, we have developed two technological strategies. The first one is based on InP wafer bonding onto a silicon host wafer, that may include a Si/SiO\(_2\) Bragg reflector. The second one, that combines micro-nanopatterning of holes or slits and sacrificial layer etching, consists in the fabrication of a PC structure on one or more superimposed InP membranes.

Different examples of micro-nanostructures and devices based on this approach will be presented in this communication.

PC-based surface emitting microlasers:
A first activity concerns the control of high-Q slow light modes in PCs in order to achieve low threshold photonic crystal-based surface emitting lasers, surface addressable all-optical modulators and bistable devices. In particular, by using a resonant mode around the Γ-point (i.e., that may interact with free-space modes along the vertical direction), Q-factors and laser threshold can be controlled by combining such InP-based planar PC resonators and vertical Si/SiO\(_2\) Bragg reflectors [1], see Fig. 1.

![Fig. 1: Schematic view of a photonic crystal surface emitting laser assisted by a SiO\(_2\)/Si Bragg reflector (a), and L-L characteristics of such lasers with different values of d (distance between the PC and the reflector). These characteristics are compared to those of a reference photonic crystal laser, with no Bragg reflector.](image-url)
Using this approach, the lasing threshold could be reduced down to ~200µW using an InAsP/InP multi-quantum well, and CW room temperature laser operation was achieved. Moreover, on similar structures, laser emission could be achieved at room temperature using a single layer of InAs/InP quantum dots as the gain material [2].

PC-assisted VCSELs:
Another key feature of PC low group velocity modes is their ability to reflect light very efficiently. Indeed, by using and combining such resonances, and provided their Q-factor is limited, PC membrane may exhibit a very high reflectivity (possibly over 99.9%), on an extended wavelength range (typically some 100nm). We have proposed to combine such PC reflectors, processed using an SOI waveguide, together with an InP multi-quantum well heterostructure, and a standard top Bragg reflector (see Fig. 2). This constitutes an alternative to classical VCSELs where the PC membrane may control the lateral size of the resonant mode, and the polarization of the emitted light.

![Fig. 2: Schematic cross-section view of the PC-VCSEL, and mapping of the electromagnetic field at resonance, as calculated by FDTD.](image)

Another technological approach that may be used to reach the same goal is based on the combination of a PC membrane with a vertical Fabry-Perot-like cavity that both include the gain material and an air-gap. Corresponding devices were fabricated using a similar design as in the latter case, and vertical emission laser was achieved [3], see Fig. 3. These structures operate under optical pumping, exhibit a threshold of 15mW and a high directivity of the emission (+/- 4.5°).

![Fig. 3: Cross-section schematic view of a photonic-crystal assisted VCSEL (a), L-L characteristics, and spectra of the laser mode, over and under the laser threshold](image)

The combination of slow light modes in PC membranes, vertical stacks and Fabry-Perot resonances could enable much more structure and devices development, including the integration of microlasers and passive integrated waveguides and circuits, 3D photonic integrated circuits, as well as selective and tunable optical PC-MOEMS filters. Some of the perspectives will be discussed in the communication.

Abstract
We discuss several interesting applications of ultrahigh-Q and ultrasmall nanocavities recently realized in silicon photonic-crystal slabs: 1) All-dielectric slow-light media, 2) All-optical bistable switching elements towards all-optical logic chips, 3) Adiabatic wavelength conversion, 4) Super-efficient optomechanical energy converter.

1. Ultrahigh-Q and ultra-small nanocavities based on photonic crystals
Recently, there has been rapid progress in the quality factor (Q) of wavelength-sized nanocavities based on semiconductor photonic crystal slabs owing to improvements in design and fabrication resolution. [1,2,3] We have recently proposed a novel design of photonic-crystal cavity, namely width-modulated line-defect cavity, and realized Q of 1.2 million with the mode volume of 1.5(λ/n)^3 by detailed spectral and time-domain measurements. We observed a long photon lifetime (~1 nsec) for this high-Q cavity. [3,4,5]

2. Slow-light media
A cavity having long photon lifetime should exhibit a substantial group delay. If the cavity size is small, we can expect significantly small group velocity. We have measured the speed of light passing through photonic-crystal nanocavities and found that it is reduced down to c/50,000, which is the slowest value reported for all-dielectric slow-light media.[3,5]. To increase its operation bandwidth, we have recently realized large-scale coupled-resonator waveguides employing these nanocavities. We observed significantly high transmission up to N=200 (N is the number of cavities). [6]

3. On-chip all-optical logic processing
One of the most important aspects for high-Q nanocavities is the fact that intrinsically weak light-matter interaction can be greatly enhanced by employing high-Q nanocavities. Such enhancement is expected for various optical phenomena, such as light emission, optical nonlinearity, etc. We have recently employed these nanocavities for all-optical switching based on optical nonlinearity in materials.[7,8,9,10] The device is based on high-Q nanocavities having two resonant modes coupled to input and output waveguides implemented in silicon photonic crystals. We have observed all-optical bistable switching operation with significantly small switching energy (<100 fJ) and fast switching speed (~100 ps). This switching operation is based on two-photon absorption in silicon and subsequent carrier-plasma dispersion effect. Although silicon is not an ideal material for nonlinear application in comparison with III/V materials, we observed fairly low-power switching operation. This is because the switching power should be scaled as V/Q (the switching energy should be ~V^2/Q).[11] Although the loaded Q is finally determined by the required operation speed, it is always advantageous to have smaller V and larger unloaded Q, which is the case for photonic-crystal nanocavities.

This tiny bistable switching element has some important aspects for future application. 1) It requires very small switching energy, 2) Bistable transistor-like operation is possible, 3) It is fundamentally suited for large-scale optical integration, 4) It is based on silicon. Although there had been extensive studies for all-optical logic using optical nonlinearity, we believe that this element can potentially solve many of problems in previous works towards on-chip integration of all-optical logic processing elements. Recently, we have designed all-optical flip-flop processor consisting of doubly-coupled nanocavities, by which we demonstrate all-optical retiming circuit operation numerically.[12] Although this is just one of examples how photonic-crystal based integrated circuits can perform, we believe much complex logic function may be possible by integrating a large number of these elements.

4. Adiabatic wavelength conversion
In the previous section, we discuss the possibility of manipulating light propagation in a cavity via optical nonlinearity of materials. Here we show a fundamentally different way of the manipulation of light using nanocavities. If we have a sufficiently small and high-Q optical cavity, it becomes possible to change the property of this cavity within its photon lifetime. Recently, we have clarified that this dynamic tuning[13] leads to very interesting optical phenomenon. Suppose that an optical pulse is stored in a cavity and we change the resonant wavelength of the cavity. Then, what will happen? We numerically computed this phenomenon by the finite-difference time-domain technique, and we found that the wavelength of light captured in a cavity is...
coupled-nanocavity-based optical MEMS

If we apply the previous adiabatic conversion process to a certain type of coupled cavities, very interesting opto-mechanical system can be realized. We have examined a double-layer cavity consisting of ultrahigh-Q cavities studies in the first section, as shown in Fig. 3. We calculated an optical force generated by an optical pulse captured in this coupled cavity by using two different methods (Maxwell’s stress tensor calculation and \(-dU/dz\) calculation), and round that the it can generate very large optical force (~1 \(\mu\)N/PJ). In addition, this large optical force can do mechanical work upon the slab during the system’s long photon lifetime. Consequently, this system can work as a very efficient optical to mechanical energy converter. The estimated energy conversion efficiency is close to 10%, although the system is in a non-relativistic regime. Conventional opto-mechanical systems, such as optical tweezers, suffer from their poor energy conversion efficiency, which fundamentally due to mass-less nature of light. However, we clarified that we can extremely enhance this efficiency using nanocavity-based optomechanical systems. In fact, this energy conversion is a reverse process of wavelength conversion in the previous section, and thus the converged energy corresponds to the shifted wavelength. In other words, this system is also very effective wavelength converter. If we move one of the slabs faster than the photon lifetime of the cavity, very large wavelength conversion is indeed possible. We have numerically confirmed that \(\Delta \lambda = \lambda \Delta \phi\) can be 20%.

References

Prospect of Optical Devices with One-dimensional Photonic Crystal Structure

Toshio Katsuyama\textsuperscript{1,2}, Kazuhiko Hosomi\textsuperscript{1,2,3}, Misuzu Sagawa\textsuperscript{1,2,3}, and Yasuhiko Arakawa\textsuperscript{1}

\textsuperscript{1}Nanoelectronics Collaborative Research Center, Institute of Industrial Science, The University of Tokyo, 4-6-1, Komaba, Meguro-ku, Tokyo 153-8505, Japan
\textsuperscript{2}Optoelectronic Industry and Technology Development Association, Sumitomo Edogawabashi-kikai Bldg. 7F, 20-10, Sekiguchi 1-Chome, Bunkyo-ku, Tokyo 112-0014, Japan
\textsuperscript{3}Central Research Laboratory, Hitachi Ltd., 1-280, Higashi-Koigakubo, Kokubunji-shi, Tokyo 185-8601, Japan

Abstract

Optical devices constructed with one-dimensional photonic crystals are reviewed. In particular, dielectric thin-film and deeply etched Si photonic crystals are discussed from the viewpoint of their application to optical transmission networks and optical sensing systems. Here, we discuss the present status of research on 1D PhCs, particularly dielectric thin-film and deeply etched Si PhCs, which can be used in optical transmission networks and optical sensing systems.

1. Introduction

Photonic crystals (PhCs) can be classified into one-, two-, and three-dimensional structures. Two- and three-dimensional PhCs are a particularly popular focus of the research because they have unique features such as an ultimate light-confining effect due to their full photonic band gap. One-dimensional (1D) PhCs, on the other hand, have the advantages of being relatively easy to fabricate and being non-polarization dependent. Therefore, 1D PhCs have also attracted a lot of attention in various fields.

Here, we discuss the present status of research on 1D PhCs, particularly dielectric thin-film and deeply etched Si PhCs, which can be used in optical transmission networks and optical sensing systems.

2. Photonic crystals composed of dielectric thin films

One of the functions of PhCs is to control group velocity and its dispersion. Coupled-defect-type photonic crystals are the most promising candidates for such control, which enables extremely small dispersion compensators to be used in optical fiber transmission systems [1]. The structure optimized for obtaining a large group-velocity dispersion was formed with SiO\textsubscript{2}/Ta\textsubscript{2}O\textsubscript{5} thin films, as shown in Fig. 1 [2]. It was designed for the 1.55-μm, 40-Gbit/s optical communication system. The thin-film structure is substrate-free, which means the device chip can be as small as a 1.4-mm edge cube. To obtain a large group velocity difference, 60 pieces of substrate-free film were stacked. The obtained optical properties are shown in Fig. 2. The hatched area in the figure was used for the 40Gb/s modulation band. The loss was about 10 dB, and the group velocity difference in the band reached 100 ps, which enables fabrication of realistic modules.

The obtained fiber-to-fiber type module is as small as 22 x 40 x 24.5 mm, as shown in Fig.3. We carried out a 40 Gb/s non-return-to-zero (NRZ) optical transmission experiment. The wavelength of the laser emission was 1549 nm and the transmission link was a 10-km standard single mode fiber. The eye diagrams after 10-km transmission are shown in Fig. 4. A clear eye opening was observed when the dispersion compensator was in use, and there was serious degradation in the eye diagram when it was not in use. The corresponding dispersion is 170 ps/nm. We have also recently succeeded in fabricating an extremely compact dispersion compensator module integrated with a photo receiver. The dispersion compensation in this case was also excellent.
3. Deeply etched Si photonic crystals

Another type of 1D PhCs is composed of periodically aligned semiconductor (typically silicon) walls, thus forming 1D PhCs, in which a light beam propagates parallel to the substrate plane. Si-based 1D PhCs were formed using a cryogenic etching process [3,4]. The etching depth reached about 20 μm, and the etched surface was extremely smooth compared with those produced by conventional etching processes such as the Bosch process, as shown in Fig. 5. A clear transmission band based on defect formation was obtained with the deeply etched Si structure embedded in the polymer medium. The peak position can be artificially controlled by changing the temperature of the PhC, as shown in Fig. 6. Thus, this etching technique enables the formation of various optical elements such as the optical wavelength-tuning filters used in optical transmission networks and optical sensing systems.

4. Summary

Using dielectric thin-film 1D PhCs, we successfully demonstrated 40-Gb/s dispersion compensation, which can be used in the metro network systems. Deeply etched 1D Si PhCs have also been fabricated using a cryogenic etching process. The Si-based PhCs make it possible to monolithically integrate PhCs with other optical elements, such as the optical waveguides on the substrates. This could lead to realization of the integrated Si-based photonic circuits concept.

Although the structural variations in the 1D PhCs are not that large compared with those of 2D and 3D PhCs, they may still have potential as applications in various optical fields.

Acknowledgements

A part of this work was financially supported by OITDA under contract with NEDO and also by the IT Program of MEXT.

References

Add-drop multiplexing in WDM signal transmission link using silicon photonic crystal R-OADM devices
Shigeru Nakamura1), Tao Chu1), Akiko Gomyo1), Jun Ushida1), Hirohito Yamada2), Satomi Ishida3), and Yasuhiro Arakawa3)
1) Fundamental & Environmental Res. Labs., NEC Corporation
34 Miyukigaoka, Tsukuba, Ibaraki 305-8501, Japan
Phone: +81-29-850-1191 Fax: +81-29-856-6139 E-mail: s-nakamura@dy.jp.nec.com
2) Department of Electrical and Communication Engineering, Tohoku Univ.
6-6-05 Aramaki-Aza-Aoba, Aoba-ku, Sendai, Miyagi, 980-8579 Japan
3) NCRC, RCAST, and IIS, Univ. of Tokyo
Komaba, Meguro-ku, Tokyo, 153-8505 Japan

Abstract: We present add-drop multiplexing of WDM optical signal using silicon photonic crystal R-OADM modules in the fiber transmission setup. Wavelength tunable add-drop multiplexing of 4-channel WDM signal with each channel bit rate of 10 Gb/s has been achieved with very compact, integrated R-OADM devices.

1. Introduction
For developing ultra-small photonic devices and integrating them, Si photonic device technology attracts considerable attention [1-3]. Taking advantage of sophisticated Si electronic device manufacturing technology, Si photonic devices utilizing various micro-/nano-structure can be realized cost-effectively. Such devices are highly expected for photonic network nodes. Currently, the concept of photonic networks is being realized by installing reconfigurable optical add-drop multiplexers (R-OADMs) consisting of discrete optical components. For extending photonic networks in terms of functions and areas [4], newly developed, compact and integrated photonic devices are required. We previously demonstrated an R-OADM device in which a tunable filter and a 2 x 2 switch were integrated on a Si-on-insulator (SOI) substrate using photonic crystal (PhC) structure and Si wire waveguides [3,5-12]. In this report, we present the add-drop multiplexing of wavelength-division-multiplexing (WDM) optical signal using moduled R-OADM devices in a fiber transmission setup. Wavelength tunable add-drop multiplexing of 4-channel WDM signal with each channel bit rate of 10 Gb/s has been achieved with very compact, integrated R-OADM devices.

2. Si Phonic crystal R-OADM device and module
Our R-OADM device consisting of a tunable filter and a switch is shown in Fig. 1. Waveguides defined by line defects in 2-dimensional photonic crystal (2D-PhC) slab were used as Bragg reflectors placed in a Mach-Zehnder interferometer (MZI) composing the tunable filter and also phase shifters placed in another MZI composing the switch [5-8]. The 2D-PhC structure was formed by making holes arranged in a hexagonal lattice through a 250-nm thick Si layer and then being buried with SiO2. The hole diameter was about 250 nm and the lattice constant was about 420 nm. The Bragg reflector was designed to reflect one wavelength channel among input WDM channels. A refractive index change based on a thermo-optical effect on Si was used for tuning the reflection wavelength in the filter and for inducing a phase shift in the switch. These PhC waveguides were connected with Si wire waveguides through low-loss trapezoidal interfaces [9-10]. Sharp bending and very short directional couplers [11] provided by Si wire waveguides also contributed to device miniaturization. The net device footprint of the R-OADM device was about 500 µm x 140 µm [12].

To make a fiber-pigtailed module, optical coupling of two beams between the Si wire waveguides and the arrayed fibers on each device facet was done using two lenses. The photograph and the schematic of the module are shown in Fig. 2.
3. Add-drop multiplexing in transmission setup

We did the experiment on the transmission and the add-drop multiplexing of WDM optical signal using R-OADM modules. The setup is shown in Fig. 3. The WDM optical signal consists of four-channel 10 Gb/s optical signals with a wavelength spacing of 5 nm. The first R-OADM module on the transmitter side was used to add the channel at 1557 nm to the other three channels at 1552, 1562, and 1567 nm. In this R-OADM device, the 1557-nm signal was propagated through the switch and then reflected with the Bragg reflector of the filter. The optical spectrum of the four wavelength channels output from the THROUGH port is shown in Fig. 4(c). The four-channel WDM optical signal was transmitted through 80-km single mode fiber with dispersion compensation fiber inserted on the receiver side. The second R-OADM module on the receiver side dropped the channel at 1557 nm from the four wavelength channels. The optical spectrum shown in Fig. 4(d) was obtained from the DROP OUT port. Optical signal from the DROP OUT port shows clear eye opening, as shown in Fig. 4(b). Measured BERs of this channel at 1557 nm shown in Fig. 4(g) indicate error-free operation with low power penalty.

The tunability of the wavelength to be added or dropped enables add-drop multiplexing of a different wavelength channel. Fig. 4(c) and (f) show optical spectra when the channel at 1562 nm was added and then dropped. In the R-OADM devices, 5-nm shift in the Bragg reflection wavelength was induced with an electric power of about 500 mW applied to the heater of the R-OADM device. As shown in Fig. 4(g), good performance was also confirmed at this wavelength.

4. Conclusions

In conclusion, we have demonstrated add-drop multiplexing of WDM optical signal using silicon photonic crystal R-OADM modules in the fiber transmission setup. Wavelength tunable add-drop multiplexing of 4-channel WDM signal with each channel bit rate of 10 Gb/s has been achieved with very compact, integrated R-OADM devices.

References


Fig. 3: Transmission setup with Si PhC R-OAMDs

Fig. 4: Experimental results
(a)-(b) Eye diagrams before add and after drop
(c)-(d) Optical spectra after add and after drop at 1557 nm
(e)-(f) Optical spectra after add and after drop at 1562 nm
(g) Measured BERs
Development of directional coupler switch with ultra-short switching length based on flat-band photonic crystal structure

Jun-ichiro Sugisaka\(^{(1,2)}\), Noritsugu Yamamoto\(^{(1)}\), Makoto Okano\(^{(1)}\), Kazuhiro Komori\(^{(1)}\), Masahide Itoh\(^{(2)}\) and Toyohiko Yatagai\(^{(2)}\)

\(^{(1)}\): AIST Photonics, Umezono 1-1-1 AIST Central 2, Tsukuba, Ibaraki, Japan
Tel:+81-29-853-5042, Fax:+81-29-853-5205
\(^{(2)}\): Inst. of Appl. Phys. Univ. of Tsukuba, Tennoudai 1-1-1 Tsukuba, Ibaraki, Japan
Tel:+81-29-861-5201, Fax:+81-29-861-5602
jun-ichiro.sugisaka@aist.go.jp

Abstract
We have proposed a directional coupler (DC) switch with short switching length and wide bandwidth, which have a flat-band in even mode. We fabricated the DC and demonstrated that the DC had a flat-band.

1 Design of DC for short switching length and wide bandwidth

The DC switch we proposed is composed of a two-dimensional photonic crystal slab of air bridge type. This consists of GaAs substrate (thickness of 190nm) and air cylinders (circular holes). The lattice constant \((a)\) is 390 nm, and the circular holes whose radii are 0.29\(a\) are arranged in the triangular lattice.

The coupled waveguide, which is a pair of parallel waveguides (shown in Fig. 1), is separated into two region A and B. The switching operation is realized by changing the refractive index of region A (the length of region A in the direction of waveguide is called "switching length").

![Diagram of coupled waveguide](image)

**Fig. 1:** The structure (top view) of coupled waveguide. The entire length is set to integral multiple of coupling length \(L_c\). Switching operation is realized by changing the refractive index of region A.

A DC switch requires short switching length and wide bandwidth. However they are a relation of trade-off for conventional DC. This relation can be dissolved using a DC which have a flat-band in even mode\(^{(1,2)}\).

![Dispersion curves](image)

**Fig. 2:** Dispersion curves for conventional DC. The open circles denote the dispersion relation for even mode, and the rectangles denote that for odd mode.

The dispersion curves of conventional DC switch are shown in Fig. 2. Although conventional DC has no flat-band, dispersion curves can be generally transformed by structural modulation of photonic crystal. We designed the structure of DC so that the DC has a flat-band in even mode. The structure and dispersion curves of structure-modulated DC switch are shown in Fig. 3. By numerical analysis with two-dimensional (2D) FDTD method, it was found that the switching length was \(11a\) (\(a\) is a lattice constant), which is 2.86\% of conventional one. The bandwidth is 5.01nm, which is 35.0\% of conventional one. The bandwidth becomes narrower than that of conventional DC, however, considering the shortening rate of the switching length, it is very wide.

2 Fabrication of DC

Next, the designed DC was lithographically fabricated with electron beam. The SEM images are shown in Fig. 4. In order to eliminate the reflection at bent waveguides, adiabatic bend waveguides (Fig. 4(c)) were introduced to all bent points. In addition, in order to inhibit the reflection at the edge of the coupled waveguide, and to improve the coupling efficiency of propagation mode between single waveguide and coupled waveguide, adiabatic branch waveguides (Fig. 4(a))
were introduced at both ends of the coupled waveguide.

3 Measurement of wavenumber difference between even and odd modes

In order to confirm whether the fabricated DC has a flat-band in even mode, the difference of wavenumbers ($\Delta k \equiv k_e - k_o$) between even and odd mode were measured with the following method. The $\Delta k$ cannot be measured directly, then the coupling length ($L_c$) for every frequency was measured, and $\Delta k$ was calculated by the following equation

$$\Delta k = \pi / L_c.$$  

(1)

The coupling length is obtained by measuring the power spectra of DCs for different length of coupled waveguide, then the difference of length of the coupled waveguide between bar condition and cross condition is equal to coupling length.

By using the measured power spectra, we obtained the $\Delta k$, which are shown in Fig. 5 with open circles. Figure 5(a) is the result for conventional DC and (b) is that for structure-modulated DC. The solid lines in Fig. 5 are theoretical values obtained by band calculation with 2D-FDTD method. Experimental values well agree with theoretical one. The $\Delta k$ for conventional DC gently changes for the frequency of incident light, this corresponds to the band diagram in Fig. 2.

4 Conclusion

A novel structure of DC switch was designed and fabricated. The power spectra were measured correctly by introducing adiabatic structures. Next the difference of wavenumbers between even and odd modes were calculated from the measuring result of power spectra. From the result, it was shown that the fabricated DC has a flat-band in even mode.

Reference

RF Output Power Improvement Using Heterodyne Technique for Radio-on-Fiber Down-link Transmission System Eliminating Electric Power Supply at Base Station

Motoharu Matsuura, Taichiro Okabe, Naoto Kishi, and Tetsuya Miki
University of Electro-Communications, 1-5-1 Chofugaoka, Chofu, Tokyo 182-8585, JAPAN
Tel: +81-424-435206, Fax: +81-424-907096, E-mail: matsuura@ice.uec.ac.jp

Abstract - RF output power improvement using heterodyne technique for radio-on-fiber down-link transmission system at base station is proposed. We have successfully improved the RF power of about 7.5 dB without electric power supply facilities.

Introduction

In future access network, radio-on-fiber (ROF) transmission system is one of the important techniques which deliver broadband radio-frequency (RF) signals for wireless communications [1, 2]. In this system, more base stations (BSs) are needed due to higher transmission loss as the transmission speed becomes higher. Thus, in order to implement ROF system, simple and cost-effective configuration must be required at each BS. So far, various types of simplified BS scheme have already been proposed [3, 4]. In particular, the electric power supply facilities in each BS are a big obstacle for the construction and maintenance of the BSs. To overcome this problem, we proposed a novel ROF system, which eliminates the electric power supply facilities at the BS by means of the power delivery, which feeds optical power over optical fiber from central station (CS) [5]. However, further performance improvement is required to implement the proposed scheme more effectively. Heterodyne technique is a practical approach to improve the transmission performance for ROF systems. Previously, the elimination of dispersion effect in fiber [6] and 60-GHz millimeter wave transmission [7, 8] were reported using such technique.

In this paper, we present RF output power improvement at the BS for ROF down-link transmission system by means of heterodyne technique for the optical transmission signal. Using the proposed scheme, the RF output power is improved approximately 7.5 dB without adding to electric power supply facilities such as low noise amplifier at the BS.

Operation principle and experimental setup

The experimental setup of ROF down-link transmission system is depicted in Fig. 1. The 2.45-GHz data signal is generated from a signal generator (SG) with WLAN, 54 Mbps, and IEEE 802.11 format, which uses orthogonal frequency division modulation (OFDM) each modulated with 64 quadrature amplitude modulation (64QAM). In the conventional scheme as shown in Fig. 1(a), the optical signal at the wavelength of 1550 nm is directly modulated by an electroabsorption modulator (EAM) biased at 17 dBm. After the modulation by the EAM, the optical signal is amplified by an erbium-doped fiber amplifier (EDFA). The amplified optical signal is mixed down to the 0.45-GHz IF signal by applying the 2.0-GHz local oscillator (LO) to the EAM. Then, the modulated optical signal is delivered to the base station (BS) by using an optical fiber. At the BS, the modulated optical signal is mixed by the LO signal and the 0.45-GHz IF signal is down-converted to the 2.0-GHz IF signal. The 2.0-GHz IF signal is detected by the photodetector (PD) and then amplifed by the LO signal and the 0.45-GHz IF signal is down-converted to the 2.0-GHz IF signal. The 2.0-GHz IF signal is detected by the photodetector (PD) and then amplified by the RF amplifier (RF-Amp) and then amplified by the RF power amplifier (RF-Pwr) and then amplified by the RF power amplifier (RF-Pwr) and then amplified by the RF power amplifier (RF-Pwr)

The RF output power improvement using heterodyne technique for radio-on-fiber down-link transmission system at base station is proposed. We have successfully improved the RF power of about 7.5 dB without electric power supply facilities.
Experiment and results

Figure 2: RF output spectra for the cases of conventional and proposed schemes.

We believe that the power penalty is due to the noise of the added EDFA and the characteristics of the LNM. On the other hand, the excess power penalty of approximately 1.5 dB around the lowest BER is much smaller than the power margin of 7.5 dB, which is improved using the proposed scheme. Thus, these results have proved that the proposed scheme is effective to improve the transmission performance for the ROF down-link system.

Conclusion

We have successfully improved the RF output power of the electrical signal at the BS for ROF down-link transmission system by means of heterodyne technique. Using this system, the RF output power has been increased up to about 7.5 dB without electric power supply facilities at the BS. The proposed scheme will be also attractive for future ROF system using higher frequency millimeter-wave.

References

A Study of Interference Reduction in Radio over Fiber System Applying MIMO Technique

Yasuhiro Kanaoka¹, Ikuo Yamashita¹, Satoru Kashimura², Kinya Asano² and Satoru Shimizu²

¹ Kansai Electric Power Co., Inc., Power Engineering R&D Center, 3-11-20 Nakoji, Amagasaki-shi, Hyogo, 661-0974 Japan
Tel: +81-6-6494-9704, Fax: +81-6-6498-766, e-mail: kanaoka.yasuhiro@d3.kepco.co.jp
² Oki Electric Industry Co., Ltd. Wireless Technology R&D Division 3-4 Hikarino-oka, Yokosuka-shi, Kanagawa, 239-0847 Japan

Abstract
Multi zone ROF wireless LAN system using MIMO was proposed and its characteristic was experimentally examined. By applying MIMO, improvement of the throughput in the interference area was confirmed.

1. Introduction
Radio over fiber technique (ROF) is attractive method for wireless communication. Because the number of wireless equipment can be reduced by dividing and combining of optical signals on the optical fiber and major equipment can be located in a central station (CS)¹. Especially for multi zone system, these features of ROF become effective. However in the case transmission characteristics might be degraded by inter-symbol interference in the area where radio waves come from different base stations (BS). In order to obtain stable throughput with whole area, interference reduction scheme is needed.

For this purpose we propose multi zone ROF system applying MIMO technique which realize high speed transmission by utilizing multi path propagation.² Multi path condition is formed in the interference area by allocating two or more of MIMO antenna output to each adjacent BS, and transmission characteristic at the area is improved applying the effect of MIMO transmission.

In this report, the transmission characteristic was examined by the laboratory experiment about this method.

2. System configuration
Figure 1 shows the system configuration in which MIMO is applied to ROF system. A MIMO access point (AP) which has two antenna outputs is set in a CS. The adjoining BS is optically connected to different antenna of the AP and non adjoining BS is connected to the same antenna output by dividing and combining of optical signals at the transmission line. In this configuration, wireless communication area is classified into two. The area where station (ST) communicates to only one BS (One BS area) and the area where the waves from two adjoining BS are reached (Multi BS area). In One BS area, transmission characteristics is not influenced by applying MIMO, on the contrary, in Multi BS area, as the effect of multi path propagation, characteristic is improved by applying MIMO.

3. Transmission experiment
Transmission characteristics measurement was performed using ROF system designed for wireless LAN (W-LAN)³ band and 2.4GHz W-LAN-AP(corega CG-WLBARGM-P). This AP has two antenna outputs and works both IEEE802.11g and 2×2 MIMO mode by changing the equipment setting. Table 1 shows the characteristics of this W-LAN. In the case of direct connection between the AP and the ST, obtained throughputs are 39.4Mbps in MIMO mode and 24.9Mbps in IEEE802.11g mode respectively. The throughput between the AP and the ST with inserting the ROF equipment are also shown in the table. The result shows the influence of the ROF system on the performance is a little. And it is found that the transmission characteristic in the case of one antenna in MIMO is almost the same as in the case of IEEE802.11g.
Then using a fading simulator, performance of multi zone ROF system applying MIMO is examined. Experimental set up is shown in Figure 2. Propagation models set at the fading simulator are shown in Table 2. The direct path (delay=0µsec) was set to be changed according to Rice rule, and other paths were to Rayleigh fading.

![Experimental setup](image)

<table>
<thead>
<tr>
<th>Model</th>
<th>Path 1</th>
<th>Path 2</th>
<th>Path 3</th>
<th>Path 4</th>
<th>Path 5</th>
<th>Path 6</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.004</td>
<td>0.01</td>
<td>2.2</td>
<td>0.01</td>
<td>0.19</td>
<td>1.7</td>
<td>19.0</td>
</tr>
<tr>
<td>2.006</td>
<td>1.7</td>
<td>0.01</td>
<td>2.2</td>
<td>0.4</td>
<td>1.7</td>
<td>23.0</td>
</tr>
<tr>
<td>3.042</td>
<td>0.1</td>
<td>2.2</td>
<td>0.4</td>
<td>1.7</td>
<td>25.0</td>
<td></td>
</tr>
<tr>
<td>4.241</td>
<td>1.7</td>
<td>0.1</td>
<td>2.2</td>
<td>15.0</td>
<td></td>
<td></td>
</tr>
<tr>
<td>5.395</td>
<td>0.1</td>
<td>2.2</td>
<td>15.0</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Average Delay (µsec)</th>
<th>Delay Time (µsec)</th>
<th>Path Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Measured throughput between the AP and the ST is shown in Figure 3. The vertical axis is the throughput normalized by that of the static propagation in One BS area (via ROF case at Table 1). In Multi BS area characteristic without MIMO is always deteriorated to One BS area under the influence of interference. On other hand, improved characteristic is always observed in MIMO case. In One BS area with and without MIMO is basically the same condition. The slight difference appeared in the figure seems to be caused by internal processing in the AP.

The characteristic is greatly deteriorated where average delay of more than 0.24µsec, because the influence of large delay waves that exceed the guard interval of IEEE802.11g (=0.8µsec). In these cases, throughput in Multi BS area without MIMO is degraded to less than 3% compared with non fading case. On the other hand, by applying MIMO, 30% performance is still obtained even in the worst case and over 100% performance is seen in some cases.

In the case of the multi path cannot be reduced in One BS area, it is considerable that by setting new BS in the One BS area and making new Multi BS area, transmission characteristics is to be improved.

![Normalized throughput](image)

4. Conclusion

We proposed multi zone ROF system applying MIMO technique for the purpose of reducing the interference caused by multi waves from different BS's. Though the transmission characteristic of without MIMO case was severely degraded in Multi BS area, it was found that the throughput was improved by applying MIMO.

Reference


Cycle Attack by MAI Noise Propagation and its Online Detection in OCDM-based Transparent Optical Networks

Shaowei Huang (1), Ken-ichi Baba (2), Masayuki Murata (3), Ken-ichi Kitayama (1)

(1) Graduate School of Engineering, Suita city, Osaka University, Osaka, Japan, 565-0871; (2) Cybermedia Center, Osaka University, Ibaraki city, Osaka, Japan, 565-0047; (3) Graduate School of Information Science and Technology, Osaka University, Suita city, Osaka, Japan, 565-0871

Abstract — In OCDM-based networks, MAI noise propagation can possibly build up unintended all-optical cycles, where MAI noise oscillation corrupts all lightpaths. A depth-first-search (DFS)-based algorithm is proposed to detect such cycle attacks.

I. INTRODUCTION

Transparent WDM all-optical networks enable ultra-high speed data transmission in the physical layer without any optical-electrical (O/E) and electrical-optical (E/O) conversions. However, due to the present immature all-optical regeneration technologies, unintended all-optical cycles may possibly be created due to Erbium-Doped Fiber Amplifiers (EDFAs) and the crosstalk oscillation in lightpath establishment [1].

An OCDM-based network architecture was proposed in [3]-[4], where channels with finer granularity than a wavelength, encoded by distinct optical codes (OCs) can be multiplexed onto the same wavelength and thus improve the wavelength utilization effectively. The multiple-access interference (MAI) noise is the dominant factor degrading the signal performance in OCDMA systems as well as the OCDM-based networks. As the MAI noise propagates and accumulates along the path, the cycle attack created in OCDM-based transparent optical networks may have unlimited attacking capability, which hinders from establishing an OCDM path along with the existing lightpaths.

The remainder of this paper is organized as follows. In the Section 2, we firstly formulate the cycle attack problem due to MAI noise oscillation. In Section 3, a depth-first-search (DFS)-based algorithm is proposed to detect such cycle attacks in the dynamic lightpath establishment. In Section 4, the numerical results are shown. Finally, we conclude this study briefly.

II. CYCLE ATTACK IN OCDM-WDM NETWORKS

In this section, the model of MAI noise propagation and the cycle attack due to the MAI noise oscillation in OCDM-based networks are first defined by us.

A. MAI noise propagation mechanism

As presented in [3]-[4], different channels (OCDM-LSPs) coded by distinct optical codes can be multiplexed onto the same wavelength. As shown in Fig. 1, OCDM-LSPs 1 and 2 are discriminated at Node A (shown in Fig. 1) by the optical correlation. In the ideal case, only the decoded peak signal of OCDM-LSP 2 will enter an encoder and be encoded by another code before being forwarded to the next node. But in the realistic case, since MAI noise introduced by OCDM-LSP 1 cannot be eliminated completely, it will also be encoded and transmitted along with OCDM-LSP 2.

There exists the possibility that OCDM-LSP 2 and OCDM-LSP 3 (using the same wavelength) meet at Node B (shown in Fig. 1) and are switched to the same output as illustrated in Fig. 1. Consequently, the OCDM-LSP 3 will have an energy leakage not only from OCDM-LSP 2, but also from OCDM-LSP 1. Due to this mechanism, the signal degradation by MAI noise will be spread in the network.

B. Cycle attack due to MAI noise propagation

Based upon the above MAI noise propagation mechanism, a phenomenon named cycle attack is first defined by us as follows. As illustrated in Fig. 2, OCDM-LSPs 1 (solid line) and 2 (dotted line) are existing lightpaths using different OCs but on the same wavelength in the network, and OCDM-LSP 3 (dashed line) is a new lightpath supposed to take a route along 1 → 2 → 3 → 5 → 6. As described above, since OCDM-LSP 2 and OCDM-LSP 3 are sharing the same link L2,2 and wavelength, and due to imperfect regeneration, MAI noise from OCDM-LSP 2 will be propagated along links L1,3 and L2,5. Furthermore, since OCDM-LSP 3 and OCDM-LSP 1 are sharing the same link L5,6 and same wavelength, MAI noise from OCDM-LSP 3 and the propagated MAI noise from OCDM-LSP 2 will be leaked to OCDM-LSP 1. Unfortunately, OCDM-LSP 1 and OCDM-LSP 2 are also sharing link L6,1, all the noises from OCDM-LSP 1 will be fed back to OCDM-LSP 2, and this causes the MAI noise oscillation along a cycle.
1 → 2 → 3 → 5 → 6 → 1, which is shown on the right-hand side in Fig. 2.

Consequently, OCDM-LSP 1, 2 and 3 are broken down because of this MAI noise oscillation. This is identified as the most distinguishing characteristic in OCDM-based networks, and is considered very different from WDM networks. Therefore, detecting and avoiding such cycle attacks become essential while dealing with traffics in OCDM-based networks.

III. DFS-BASED CYCLE-ATTACK DETECTION ALGORITHM

In this section, we will focus on developing an algorithm able to detect such cycle attacks explained above.

Depth-first-search (DFS) is a common algorithm for traversing a graph. In DFS, a visited node is stacked, and is avoided being visited twice in the next search, and finally all the nodes can be found in a graph with a recursive manner. Based upon this mechanism, we propose a cycle-attack detection (CAD) algorithm. The main implementation of CAD can be described as, 1) a network topology will be firstly created, which is based upon the information of existing lightpaths (including the new lightpath) distributed in the network instead of the physical topology. Intuitively, the set of edges is determined by whether they are attacked by the MAI noise propagation described above. For example, as illustrated in Fig. 2, at Node 2, since there is not any lightpath traversing link 2-6, Node 6 is not considered as the adjacent node of Node 2 even though they are physically connected; 2) DFS is performed to traverse the topology created in step 1). If a node is visited twice, the algorithm will return a value announcing there exist a cycle, and block the new lightpath establishment request; otherwise, the new lightpath will be established.

IV. SIMULATION RESULTS

Simulations are conducted in a single-fiber bi-directional 6-node-based network. Each fiber has $L=5$ wavelengths, and $n$ OCDM channels (varying from 1 to 8). Taking the capacity of a wavelength as 10Gbps, each OCDM channel owns the capacity of $10/n$ Gbps.

In this paper, the cost function per link for calculating the route is defined as the ratio of total channels ($5n$) and free OCDM channels. Therefore, a link with more free channels is tended to be selected. In OCDM-based networks considering cycle attacks, wavelength assignment schemes are considered much more important, because a cycle attack can only occur among the lightpaths sharing the same wavelength. Here, we propose an advanced first-fit (AFF) wavelength assignment scheme, which means only the wavelength with low sequence number among all the available wavelengths not only satisfying the wavelength continuity constraint but also not causing any cycle attack will be selected.

Fig. 3. Single-fiber-based 6-node network, 5 wavelength per fiber, network offered load = 220; no wavelength conversion, but OC conversion.

As illustrated in Fig. 3, to the ideal case, first-fit, random and AFF wavelength assignment schemes obtain the same performance with an increase of OC channels per wavelength. While to the cycle-attack case, as not taking the cycle attack into account, the first-fit and random schemes suffer much higher blocking comparing to the ideal case. However, our proposed AFF wavelength assignment can still achieve the performance very close to the ideal even cycle attacks exist in the network.

VII. Conclusion

Cycle attacks by MAI noise propagation have been firstly defined. A DFS-based algorithm for detecting such cycle attacks and an advanced first-fit (AFF) wavelength assignment scheme for reducing blocking effectively have been proposed.

ACKNOWLEDGEMENT

S. Huang would like to acknowledge the ICOM Electronic Communication Engineering Promotion Foundation and Japan Society for the Promotion Science (JSPS).

REFERENCES

Optical CDMA Networks Using Extended Hamming Code for Interference Elimination

Chao-Chin Yang and Larn-Ying Yeh
Dept. of Electronic Engineering, Kun Shan University, Taiwan.
E-mail: ccyang@mail.ksut.edu.tw
E-mail: yehyu@mail.ksu.edu.tw

Abstract
One spectral-amplitude-coding scheme employing extended hamming code for optical code division multiple access networks is present. By using this scheme, each user can adopt complementary coding for information encoding with lower bit error rate.

1 Introduction
Optical code-division multiple-access (OCDMA) techniques are promising solutions for optical access and local area networks. Among these techniques, the spectral-amplitude-coding (SAC) scheme has received more and more attention due to its excellent interference elimination ability [1]. Besides, its complementary coding ability is also suitable for the applications of other OCDMA schemes such as time-spreading/wavelength hopping OCDMA with enlarged cardinality and improved performance[2].

In [3], one SAC scheme cooperating with wavelength division multiplexing for coder simplification was proposed. However, this scheme doesn’t have complementary coding ability. The following scheme is proposed to alleviate this drawback.

2 System description
The proposed scheme uses the well-known Extended Hamming (EH) code for codeword construction[4] [5].

The codewords of extended [8,4] Hamming code \(X_{m,n}\) are listed in Table I (\(m=0,\ldots, M-1\) and \(n=0,1\).) Note that these codewords are generated from the generator matrix of [7,4] Hamming code and added by an overall parity-check bit. The Hamming distance between these extended [8,4] Hamming code is

\[
d(X_{m,n}, X_{q,r}) = \begin{cases} 0, & m = q, n = r, \\ 8, & m = q, n \neq r, \\ 4, & m \neq q. 
\end{cases}
\]

Thus

\[
X_{m,n} \odot X_{q,r} = \begin{cases} 4, & m = q, n = r, \\ 0, & m = q, n \neq r, \\ 2, & m \neq q, 
\end{cases}
\]

where \(\odot\) is the dot-product of two vectors[6]. Note that the above property is not valid for \(X_{0,1}\), thus \(X_{0,0}\) and \(X_{0,1}\) are not used in the proposed scheme. By using the method of code construction in [3], WS-EH codewords \(A_{\alpha,m,n}\) can be obtained from \(X_{m,n}\), where \(u\) is the group number( \(u = 0, 1, \ldots, N-1\) ). \(A_{\alpha,m,0}\) and \(A_{\alpha,m,1}\) are assigned to user \(u(m, m)\) for the encoding of “1” and “0” bits, respectively. As an example, the encoder of user \(u(0, 1)\) using \(A_{0,1,0}=1 0 0 0 0 1 1 1 0 0 0 0 0 0 0 0\) and \(A_{0,1,1}=0 1 1 1 1 0 0 0 0 0 0 0 0 0 0 0\) as signature sequences is shown in Fig. 1 for \(N=2\). The encoder contains light source, 1*2 optical switch, circulators and
fiber Bragg gratings with different center wavelengths [3]. According to the information bits, the wavelength signals from the light source are directed to one of the upper or lower circulator inputs in Fig. 1(a) by the 1*2 optical switch, and spectrally encoded by one of the fiber Bragg grating arrays. The resulting signals are directed by the same circulator again and appear at the output port of the encoder.

The decoding scheme for user \(u, m\) to eliminate interference from user \(v, q\) is

\[
A_{u,m,i} \odot A_{v,q,j} = A_{u,m,i} \odot \delta_{v,q,i}
\]

\[
= \begin{cases} 
4, & u = v, m = q, n = 0, \\
-4, & u = v, m = q, n = 1, \\
0, & \text{otherwise},
\end{cases}
\]

and it can be implemented simply using optical components. One example is the FBG-based decoder shown in Fig. 1(b).

WS-EH codes proposed here, the results for BIBD codes and WS-BIBD codes in [3] are also shown for comparison and the code lengths of these code families of Fig. 2 are about 183 or 273. It is found that when the code lengths \(M*N\) is fixed and the number of active users is smaller, the WS-EH codes obtains lower BER as compared to that of the other two code families. When the number of active users become larger, the BERs of these code families are getting closer. In addition, it should be noticed that the encoder of the WS-EH codes is more complex than that of WS-BIBD codes. Thus the selection of suitable code families should be careful.

4 Conclusion

We propose one code family using EH codes for SAC OCDMA networks. These codewords can be used for complementary coding and the BERs are lower than that of previous proposed code families when the number of active users is relatively small.

5 References

Comparison of Physical Network Configurations for Next Generation Home Network

Yu Kakishima*, Kohei Okada and Kimio Oguchi**
Information Networking Lab., Graduate School of Engineering, SEIKEI University
3-3-1 Kichijoji-Kitamachi, Musashino, 180-8633 Japan
Tel: +81-422-37-3732, Fax +81-422-37-3871, Email: *kakishimaru@yaho.co.jp, **oguchi@st.seikei.ac.jp

Abstract - Physical network configurations for the next generation home network are compared in terms of total cable length by considering the general house model. Longest distance between the router and optical wall socket is also discussed by using typical numerical values of Japanese houses.

1. Introduction
The next generation home network will accommodate several different kinds of terminal or appliances with large capacity links [1]. One of the heaviest bandwidth consumers is the audio visual (AV) terminal; its traffic is heavy and bursty. Therefore, the transmission medium used must offer very wide bandwidth. Optical fiber is the most promising candidate. However, actual installation guidelines of optical fiber cables have not been well discussed.

The authors have already presented the basic house model for evaluating cable length in a home [2].

This paper extends the basic model to yield a more realistic solution; Cable wiring routes that replicate those in actual homes are considered. Four optical wall sockets are also considered. This paper also describes total cable length and the longest length for three network topologies.

2. Cable wiring route in a home
Network topologies considered here are single star, double star, and ring. Wiring routes are considered using the housing model shown in Fig. 1 and Fig. 2 where a house consists of multiple boxes stacked with each size of a x b x c [2]. Four optical wall sockets locate each at the four corner of each room.

Figure 3 shows the wiring route of each network topology; (a) single star, (b) double star, (c) a ring. It is also assumed that each room has a branching filter and ADM (Add/Drop Multiplexer) in the ceiling (left corner). Moreover, the double star and the ring utilize WDM (Wavelength Division Multiplexing) technology for signal multiplexing. The ring type is structured as one ring per floor.
3. Total cable length needed for each network topology

The assumed room size is as shown in Fig. 2, a = x-axis, b = y-axis and c = z-axis. The house has F floors.

Total wiring length $L$ is given by $L=L_1+L_2+...+L_F$ where $L_f$ means total cable length on the $f$th floor with number of rooms $N_{x,f}$ on the x-axis and $N_{y,f}$ on the y-axis.

### 3.1 Total wiring length formula for calculation

i) $L_f$ for the single star is given by

$$L_f = \sum_{i=1}^{N_{x,f}} (2a + 4a(i-1)) \times N_{x,f} + \sum_{j=1}^{N_{y,f}} (2b + 4b(j-1)) \times N_{y,f} + 8N_{f} + 4(1-N_{f})c$$

### 3.2 Comparison of total cable length for the network topologies

For comparing the three topologies in terms of total cable length, several parameters are set such as; $a=1$, $b=1$, $c=1$, $N_x=3$, $N_y>1$, and $F=2$. Figure 4 shows total cable length vs. $N_y$ with parameters set before for each topology. It is observed that the ring has shortest length and the single star has about twice the length of the others.

4 Calculation of longest transmission distance for the network topologies

The longest transmission distance in each network topology is derived by using the mean values of houses in Japan [3]. These values are;

- size of room (Refer to Fig. 2); $a=4.5m$, $b=4.5m$, $c=2.4m$,
- number of floors: $F=2$; two floors, and
- number of rooms on one floor; four rooms, $N_{x}=2$, $N_{y}=2$.

The longest distance from the router to an optical wall socket in the single star is 25.2m. Here, six bends and six connectors are assumed.

The longest distance from the double star type router to an optical wall socket is 25.2m. This value is the same as that of single star; basic configuration is the same except that the double star has a branching filter between the router and the optical wall socket. Thus four bends and eight connectors are assumed. Signal passes through a branching filter in this configuration.

The longest distance from the ring type router to the optical wall socket is 29.7m. Thus four bends and fourteen connectors are assumed. In this configuration, the signal passes through four ADMs.

As a result, the ring type has the longest distance. Moreover, the transmission loss might be large because many connectors and ADMs are needed.

5. Conclusion

This paper examined the wiring routes likely for the next generation home network. Total wiring length of three basic network topologies was compared. The longest distance in each network topology between a router and optical wall socket was elucidated by using the average value of Japanese houses. We found that the double star appears to be the optimal configuration for the next generation home network.

Cable installation trials in actual houses are needed determine real-world optical budgets.

References


QoS Management in a Next Generation Home Network

Shingo Yamakawa*, Kunio Tojo, Shohei Terada and Kimio Oguchi**
Graduate School of Engineering, SEIKEI University, 3-3-1 Kichijoji-Kitamachi,
Musashino-shi, Tokyo, 180-8633 JAPAN,
Tel: +81-422-37-3732, Fax: +81-422-37-3871, E-mail: *shingo5_86@yahoo.co.jp, **oguchi@st.seikei.ac.jp

Abstract: We propose to classification the sensing data in the next generation home network according to the level of importance. Priority control for the classified data is described, together with some simulation results. The proposed QoS management scheme is shown to control the data correctly.

1. Introduction
The number of broadband access users, which includes FTTH (Fiber-To-The-Home), is rapidly increasing. In the next generation home network (NgHN), home terminals/apppliances that are not now connected to each other will be networked by using IP technology. Therefore, the amount of traffic data in the HN will increase a lot. Video terminals are the major bandwidth consumers and their traffic is bursty [1].

When all kinds of sensors will be accommodated in the NgHN, and information regarding the daily life of the user passed across the NgHN, new such systems in the NgHN as a home automation, remote sensing and life support service should be developed [2]. Several technical issues associated with signal transmission quality such as packet loss, delay and jitter still remain to be finalized before we can realize the NgHN. Those issues are important in ensuring the transfer of daily life information, security data, and disaster data.

It is not efficient to assign a different Quality of Service (QoS) to each sensing data stream. This paper classes the sensing data in the NgHN according to level of importance. The priority control based on the classes is described, together with some simulation results.

2. Classification of the sensing data
Table 1 summarizes the proposed sensing data classes. Both category I, which includes medical care, nursing care and health care information, and category II, which includes security and disaster information have the highest level of importance as they impact the Quality of Life (QoL) of humans.

3. Priority control with PQ/CBWFQ
Category I and II use the Priority Queuing (PQ) method, while the other categories use Class-Based Weighted Fair Queuing (CBWFQ) in order to assure minimum bandwidth as shown in Table 1.

The PQ method is used to guarantee the transmission of packets with the highest priority.

CBWFQ is able to guarantee the minimum bandwidth if the bandwidth value is pre-assigned. As categories III to VI may allow packet loss, use of CBWFQ is sufficient for the NgHN.

<table>
<thead>
<tr>
<th>Category</th>
<th>Sensor</th>
<th>Service to become an object</th>
<th>level of importance</th>
<th>Queuing method</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>medical care, nursing care, healthcare</td>
<td>Beat sensor, heat sensor, acceleration sensor</td>
<td>blood pressure, heat, blood glucose level</td>
<td>High</td>
</tr>
<tr>
<td>II</td>
<td>Security, disaster</td>
<td>Infrared sensor, smoke sensor, heat sensor</td>
<td>Object sensing, smoke, heat,</td>
<td>PQ : Lower priority</td>
</tr>
<tr>
<td>III</td>
<td>environment control, saveenergy control</td>
<td>Light sensor, humidity sensor, gas sensor</td>
<td>heat, humidity, brightness, utility</td>
<td>CBWFQ : Large Bandwidth</td>
</tr>
<tr>
<td>IV</td>
<td>location management</td>
<td>IC (integrated circuit) tag</td>
<td>Article of value, important documents</td>
<td>CBWFQ : Middle Bandwidth</td>
</tr>
<tr>
<td>V</td>
<td>livelihood support</td>
<td>Ultrasound sensor, pressure sensor, infrared sensor</td>
<td>Someone’s location, activity</td>
<td>CBWFQ : Middle Bandwidth</td>
</tr>
<tr>
<td>VI</td>
<td>domestic cares support</td>
<td>IC (integrated circuit) tag</td>
<td>location management in frigidaire, freshness date management</td>
<td>CBWFQ : Small Bandwidth</td>
</tr>
</tbody>
</table>

Table 1 sensor, service, priority Queuing for each categories
4. Simulation

4.1. Simulation summary

To verify the effectiveness of PQ/CBWFQ, a simulation was conducted using NS-2 (Network Simulator 2) [3]. The traffic flows allocated to each node and network topology in the simulation are shown in Fig.1. Maximum capacity between any two nodes was 100 Mbps. Source0 sends both category I and II data with PQ; Source1 the VoIP (Voice over IP) data also with PQ; Source2 category III, IV, V and VI data with CBWFQ. Source3 sends IPTV, HD (High Definition video) and Telnet data also with CBWFQ. Source4 sends FTP (File Transfer Protocol) data using the BE (Best Effort) method. All traffic is passed through intermediate nodes that offer PQ/CBWFQ control.

4.2 Simulation results and discussion

Figure 2 shows the bandwidths of each source, or services vs. starting time of each traffic. It is clear that service bandwidth varies in response to the emergence of new traffic. Figure 3 shows a magnified plot of Fig.2, together with traffic of Source0 without PQ/CBWFQ for comparison.

When the intermediate nodes do not support PQ/CBWFQ, all traffic streams suffered packet loss of about 50% because the total amount of traffic sent was about twice the capacity between two Edge nodes.

The application of PQ/CBWFQ ensured the successful transmission of the high priority class. Source0 had throughput of up to 5.0Mbps without any packet loss. Traffic from Source1 was also sent without packet loss. All traffic of Source2 sent was transmitted without any packet loss because the minimum bandwidth was set to 10Mbps, twice the service rate. Source3 traffic suffered packet loss of about 64% as its service rate was set to 35Mbps. This loss value is reasonable as the service rate was 100 Mbps. Source4 traffic experienced packet loss about 38% because it was sent on the BE basis.

5. Conclusions

For efficient management of QoS in the NgHN, we proposed to class the traffic according to importance. The effectiveness of the proposed priority control method (PQ/CBWFQ) was verified by simulation. The results showed that data with higher priority was transmitted correctly.

References


G-WAPS Internet Access Network based on an Active Optical Network

Takayuki Nakata, Takuya Kaminoogou, Hirono Matsuda, and Tadahiko Yasui
Toyama Prefectural University; 9-3-1 Kurokawa, Imizu, Toyama 939-0398, Japan
{nakata, hmatsuda, yasui}@pu-toyama.ac.jp

Abstract
By replacing the access network from a PON to an AON (Active Optical Network), the G-WAPS (WAPS with GMPLS architecture) optical internet access network can save equipment-cost without changing the architecture of the G-WAPS.

1. Introduction
It seems that telecom carriers are reluctant to introduce active devices into access networks and studies on E-PON or G-PON have progressed, and they are standardized. However, there were studies on the active optical network such as VDSL, and in the CATV field, the electronic equipment in the outside plant, as is in the case of HFC, is not so prohibitive.
We propose an AON to be used in the G-WAPS [1] access network. By replacing the PON in the G-WAPS by an AON, the cost saving of the system is achieved greatly without changing the architecture of the system.
In this paper, the role of the terminal communication board (TCB-E) in the G-WAPS and the issues concerning the TCB-E are described, the solutions to these issues are proposed, and the experimental results are shown.

2. G-WAPS configuration
The G-WAPS configuration is shown in Fig.1. The G-WAPS provides a wavelength path between terminals through the OLSR, and enables a broadband end-end communication with QoS fully guaranteed. The G-WAPS implements the edge function of the GMPLS architecture [2] in the client terminals and application softwares in the client terminal can set up wavelength LSPs between terminals in a connection-oriented manner.

The G-WAPS access network is on the PON shown as PON1 and PON2 in Fig.1, and eight CWDM wavelengths (1-8) for the G-WAPS data-plane and one wavelength (9) for control-plane are multiplexed on the feeder fiber of the PON. Each data plane wavelength carries 100Mbit data. The control plane for the access network uses a wavelength (10) on which the Ethernet PON is configured.

The number of client terminals can be larger than eight (the number of the data plane wavelengths), and the coupler works as a traffic concentrator, which increases the traffic capacity of the system.

3. TCB-E Issues
In the G-WAPS, the TCB-E mounted in a PC plays a key role. The TCB-E generates any of the eight wavelengths according to the request from the control plane. Fig.2 shows the block diagram of the TCB-E. The TCB-E has the photonic parts for TRX of the data and electronic parts for the PCI-bus interface. The TCB-E has several issues as follows.

(1) Insertion loss caused by the splitter limits the maximum number of the client terminals accommodated in a coupler. On the other hand, the traffic characteristics improve or the system capacity increases according to the splitting ratio. When it comes to
amplifying the attenuated optical signals, the CWDM wavelengths can not be amplified at once, which increases the system cost.

(2) A wavelength carries 100MbE data. This conforms to the requirement for the typical terminal application. However, 100 Mbps/wavelength is not suitable for DFB laser, even for the CWDM can be used for much higher bit-rate. In this viewpoint, the system is not cost-effective.

(3) E-PON technologies are progressing rapidly. E-PON we used for the G-WAPS needs some modifications (e.g. the upward and downward wavelength of the G-WAPS is the same, but E-PON is not) which are not always allowable, and it may happen that the requirement cannot be implemented.

(4) The cost and the space of the mux/demux in Fig. 2 cannot be neglected. The TCB-E needs to multiplex and de-multiplex the control wavelength (1.3μm), and this is done through the express channel. This additional function increases the cost of the mux/demux. The total cost is about the same as that of the LDs.

4. New proposal using an AON
In order to solve above issues, we propose a new G-WAPS internet access network configured on an AON, in stead of a PON. The basic concept of the new proposal is as follows.
(1)In stead of optical couplers we propose to use L2 switches with broadcast function supported at the same time.
(2)VLAN tag multiplexing is used in stead of wavelength multiplexing. 1GbE interface is used between the L2 switch and the terminal with TCB-Es replaced by ordinary 1GbE NICs.
(3) The control plane is implemented by assigning a specific VLAN tag to control plane wavelength λc, which is commonly used by all the terminals in one coupler. Thus, the E-PON system is eliminated. System configuration of the newly proposed AON is shown in Fig. 3. The PONs in Fig.1 are replaced by the AONs in Fig.3.

5. Experiments
Economy of the system is shown in Fig.4. When the number of terminals is 32, the G-WAPS access network cost is reduced to 5% of the original G-WAPS.

![Fig.4 Economy of the Proposed System](image)

The system throughput is shown in Fig.5(a). This figure is obtained using a free software NetPIPE [3] under the condition; PC; CPU: Xeon 2.8GHz, Memories: 256Mbytes, NIC: Intel Pro/1000 (with three chip sets; i82540EM/i82541GI/i82545GM), OS: NetBSD3.0, and L2 switch: Catalyst 2970 (Cisco). Fig. 5(a) shows that the throughput varies depending on the NIC. If we select an appropriate NIC, we can obtain the same throughput as the original G-WAPS.

(a) Throughput
(b) Latency

![Fig.5 Measured Characteristics](image)

Fig.5 Measured Characteristics about 43 μs with min/max 33μs/81μs.

6. Conclusions
We have newly proposed a G-WAPS internet access system configured on an AON in stead of a PON. We have made experiments, and the feasibility of the proposed system has been shown.

References
[1] T. Yasui et.al, CPT2006, pp112-113
11P-8

40 Gb/s-2 km photonic crystal fiber transmission at 850 nm
with a single-mode VCSEL

Hideaki HASEGAWA, Yosuke OIKAWA, Toshihiko HIROOKA, and Masataka NAKAZAWA
Research Institute of Electrical Communication, Tohoku University, 2-1-1, Katahira, Aoba-ku, Sendai-shi
980-8577, Japan E-mail: hasehide@riec.tohoku.ac.jp

Abstract

40 Gb/s single-channel transmission over 2 km was demonstrated at 850 nm using a PCF and a single-mode VCSEL. The PCF is single-mode and has a low dispersion at 850 nm. Error-free transmission has been successfully achieved.

Introduction

Recently, the 800 nm band has received a lot of attention with respect to short-distance optical communication systems such as local area networks (LAN), since an AlGaAs-based 850 nm oxide-confined vertical cavity surface emitting laser (VCSEL) is now available. A VCSEL has many advantages over edge emitting lasers including a low threshold current, a single longitudinal mode, and high-density integration, due to its two-dimensional structure and short cavity length [1], [2]. However, since conventional fibers are multi-mode and have large normal dispersion in the 800 nm region, the transmission length is limited to 100-300 m.

On the other hand, photonic crystal fibers (PCF), which have a core region surrounded by multiple air holes, are an attractive transmission medium due to their various unique characteristics including the capacity for endlessly single-mode operation [3], high nonlinearity [4], [5], and arbitrary dispersion control [6]. With PCFs, it is possible to control the dispersion characteristics and transverse mode more flexibly. By combining an AlGaAs VCSEL and PCF, we can construct simple and high-speed communication systems such as 10-40 Gb/s Ethernet over a few km. Recently, a 10 Gb/s single-channel transmission over 5 km at 850 nm has been demonstrated using PCF [7], [8]. However, a 40 Gb/s single-channel transmission at 850 nm has yet to be reported.

In this paper, we report the first 40 Gb/s transmission over 2 km at 850 nm thus further increasing the transmission capacity. We also describe the trade-off between dispersion and transmission loss in PCFs.

Single-mode PCF for 40 Gb/s transmission at 850 nm

When applying PCFs to optical communication in the 800 nm region, we need to reduce both the dispersion and transmission loss of PCF in this wavelength region. Figure 1 shows the dispersion and transmission loss of PCFs at 850 nm. In this figure, $d$ and $\Lambda$ represent the air-hole diameter and the pitch between air holes, respectively. The solid curves show the dispersion value of PCF at 850 nm calculated by the beam propagation method. The circles and squares show the measured transmission loss of our fabricated PCFs and those reported in refs. [7], [9], and [10], respectively. In PCFs, as the core diameter $2\Lambda/d$ decreases, the dispersion value becomes smaller but the transmission loss increases due to structural roughness and small particles in the air holes [11]. For example, a zero-dispersion single-mode PCF in the 800 nm region that we fabricated has a core diameter of 2.4 $\mu$m and a transmission loss of more than 29 dB/km, which means it cannot be used as a transmission medium. Therefore, we have to design a PCF with a larger core diameter and higher dispersion that is allowable in terms of the transmission distance and bit rate. In our experiment, we employed a PCF with a large core diameter of 5.6 $\mu$m indicated by the closed circle in Fig. 1, which has a transmission loss of 5.2 dB/km at 850 nm. The air-hole pitch $\Lambda$, the air-hole diameter $d$, and $d/\Lambda$ were 3.4 $\mu$m, 1.2 $\mu$m, and 0.35, respectively. The fabricated PCF has an endlessly single-mode property because $d/\Lambda$ is less than 0.43 [3]. Figure 2 shows the dispersion characteristics of the fabricated PCF and a
conventional step-index fiber (SIF), which were measured by the time-of-flight method using supercontinuum generation in a highly nonlinear PCF [12]. The zero-dispersion wavelength was 1098 nm and the dispersion value was -62.8 ps/nm/km at 850 nm. Compared with a conventional SIF, whose dispersion is -97.5 ps/nm/km at 850 nm, it is possible to reduce the pulse waveform distortion caused by the group velocity dispersion. In a 10 Gb/s transmission at 850 nm over 5 km that we reported previously, the dispersion had negligible influence on the transmission performance, since the dispersion length was as much as 37.4 km [8]. Here the dispersion length is defined as

\[ L_D = \frac{\pi T_r^2}{2 \ln 2.2 |D|} \]  

where \( c \) is the velocity of light in a vacuum, \( \lambda \) is the wavelength, and \( T_r \) is the pulse width given by half the bit interval assuming an NRZ signal. In a 40 Gb/s transmission, however, \( L_D \) is as short as 2.3 km and the dispersion becomes a dominant factor that limits the maximum transmission distance.

40 Gb/s transmission experiment at 850 nm

Figure 3 shows the experimental setup for a 40 Gb/s transmission at 850 nm. We employed an oxide-confined single-mode VCSEL with a side-mode suppression ratio of 30 dB as a light source. The maximum modulation speed of a directly modulated VCSEL at 850 nm was 22 Gb/s [13]. Therefore, we used an external modulation method with a LN (LiNbO\(_3\)) modulator in our 40 Gb/s transmission experiment. Because \( V_T \) is as low as 1.7 V at 850 nm, driver-free modulation can be achieved. The output light from the VCSEL was modulated by a 40 Gb/s NRZ signal with a length of 2\(^{15}\)1 PRBS. The input power to the PCF was amplified to 6 dBm by an erbium-doped fluoride fiber amplifier (EDFFA). The gain of the EDFFA with forward pumping at 975.4 nm was 25 dB, where the pump power was 500 mW and the input power was -25 dBm. The gain peak wavelength and 3 dB gain bandwidth were 852 and 3 nm, respectively. Since a bit rate of 40 Gb/s corresponds to a bandwidth of 0.096 nm, the gain bandwidth of the EDFFA is sufficient to amplify 40 Gb/s signals. The output signal transmitted through a 2 km-long PCF was amplified by a second EDFFA. The amplified signal was detected with a GaAs-PD with a bandwidth of 25 GHz, and the bit error rate (BER) performance was measured.

Figure 4 (a) shows the eye patterns of the input and output signals after a 2 km transmission. The distortion in the eye pattern after the 2 km transmission was caused by the dispersion-induced pulse broadening. Figure 4 (b) shows the BER performance under a back-to-back condition and after 2 km transmission. There was a power penalty of 1.7 dB at a BER of 10\(^{-9}\) caused by the PCF dispersion. However, error-free transmission over 2 km was successfully achieved.

Conclusion

We have demonstrated the first 40 Gb/s transmission at 850 nm realized by using a single-mode PCF and a single-mode VCSEL. An error-free transmission was successfully achieved. The power penalty of 1.7 dB at a BER of 10\(^{-9}\) was caused by the PCF dispersion. This result indicates that inexpensive and high-speed transmissions over short distances appear to be feasible by combining these devices.

References

Sampled FBG based all optical wavelet analyzer for high speed optical communication signals

Masanori Hanawa
University of Yamanashi, 4-3-11, Takeda, Kofu, Yamanashi 400-8511, Japan
Tel/Fax: +81-55-220-8683, E-mail: hanawa@yamanashi.ac.jp

Abstract
All optical wavelet analysis system for high speed fiber-optic communication signals using sampled FBG based optical wavelets is proposed and investigated numerically. With this analyzer, the root cause of signal degradation is easily identified.

1 Introduction
Wavelet analysis is a popular group of the time-frequency analysis methods in signal processing field. They use a finite length or localized function known as ‘mother wavelet’ as the basis function. Although the effectiveness of the wavelet analysis is widely known, it requires convolution integrals between the analyzed signal and the scaled and time-shifted ‘wavelets’, therefore it requires a huge amount of computation.

To the knowledge of the author, there is no real time wavelet analyzer for the high speed fiber-optic communication signals due to above huge computation requirements. In addition, the slow operation frequency of the electrical digital signal processors, compared to the speed of the fiber-optic communication signals, is another reason for the non-existence of the real time wavelet analyzer for high speed optical signals.

The authors have studied the sampled FBG based optical transversal filters (SFBG-OTFs) and its applications to high speed fiber-optic communications. A SFBG-OTF consists of serially placed FBGs with partial reflectivity and a certain interval between adjacent FBGs. The output of the SFBG-OTF is considered as a convolution integral between the input signal and the impulse response of the SFBG-OTF. Therefore, by using the SFBG-OTFs, all-optical wavelet analysis can be realized. In this paper, all-optical wavelet analyzer is proposed and investigated numerically.

2 All-optical wavelet analysis based on SFBG-OTF
The wavelet analysis is done by convolution integral between the input signal and the scaled and time-shifted ‘wavelets’, therefore it requires a huge amount of computation. A SFBG-OTF consists of serially placed FBGs with partial reflectivity and a certain interval between adjacent FBGs. The output of the SFBG-OTF is considered as a convolution integral between the input signal and the impulse response of the SFBG-OTF. Therefore, by using the SFBG-OTFs, all-optical wavelet analysis can be realized. In this paper, all-optical wavelet analyzer is proposed and investigated numerically.

Fig.1 Sampled FBG based all optical wavelet analyzer using SFBG-OTF.

The sampled FBG based all optical wavelet analyzer using SFBG-OTF is realized. In this paper, all-optical wavelet analyzer is proposed and investigated numerically. With this analyzer, the root cause of signal degradation is easily identified.
Simulations, the signal...computer simulations were performed to investigate that how the distorted optical communication signals were observed by the proposed optical wavelet analyzer. Fig. 4 shows the simulation model. For simplicity, the 40Gbit/s NRZ-OOK signal was chosen as the target signal to be analyzed and its transmission through tens kilo-meters of the dispersion shifted fiber was simulated by numerically solving the Schrödinger’s equation. The signals, passed through the SFBG-OTFs, were O/E converted and contour plots were composed from them. In the contour plots, the horizontal and vertical axes are time and wavelength (i.e. frequency), respectively.

Fig. 5 shows the simulation results. The left column is for waveforms and the right one is for their wavelet transformations. Through these simulations, the signal power launched into the fiber was 20dBm and a optical fiber without loss was assumed to enhance the SPM effects in the fiber. Although the waveform degrades according to the transmission, the reason why it degrades is not able to be known directly from the waveforms and also from the optical spectra. On the other hand, by observing the wavelet transformations, it is clearly seen that the leading and trailing edges of the signal got blue and red shifts, respectively. Therefore, it is easily known that the main cause of the signal degradation is the SPM effects in this case.

4 Conclusion

A concept for all-optical wavelet analyzer for high speed fiber-optic communication systems was proposed. The computer simulation results indicate its validity for characterization of the high speed fiber-optic communication signals. The implementation of the optical wavelets by the SFBG-OTFs, and the proof-in-principle experiments of the proposed all-optical wavelet analyzer are subjects in future.

Acknowledgement

This study was supported by Grant-in-Aid for Exploratory Research, No.17656131, the Ministry of Education, Culture, Sports, Science and Technology (MEXT).
Resolution improvement of all-optical ADC using SPM-induced spectral compression

Takashi Nishitani, Tsuyoshi Konishi, and Kazuyoshi Itoh
Graduate School of Engineering, Osaka University, 2-1 Yamadaoka, Suita, Osaka, 565-0871, Japan
Tel: +81-6-6879-4485, Fax: +81-6-6879-4582, E-mail: nishitani@photonics.mls.eng.osaka-u.ac.jp

Abstract

Resolution improvement of our proposed all-optical analog-to-digital conversion using self-phase modulation induced spectral compression is described. From numerical simulation results, the spectral compression ratio 2 to 1, or 3 dB resolution improvement was obtained.

1 Introduction

Analog-to-digital conversion (ADC) has been investigated as a key interface technology to convert an analog signal into a digital one which is manageable for electrical digital signal processing, transmission, storage and so on. Recent tremendous growths of high-speed digital signal processing and optical communication systems have encouraged the demand for a high-speed and high-resolution ADC. Although a 24 Gsps 3 bit electrical ADC has been proposed [1], it would be difficult to realize high-speed and high-resolution electrical ADC over a few tens Gsps due to RC delay bottleneck, the amplitude and timing jitters of electrical sampling pulse and so on. To overcome these electrical limitations, optical ADC has attracted much attention recently [2]. In general, ADC consists of three procedures; sampling, quantization and coding. The optical sampling technique has been proposed and applied to quality evaluation of high-bit rate optical data signals over 500 Gb/s [3]. On the other hand, optical quantization and optical coding have been investigated for various high-speed and high-resolution applications [4]-[6]. Previouly, we have proposed the all-optical ADC composed of optical quantization using soliton self-frequency shift (SSFS) in a fiber [4] and optical coding using optical interconnection based on a binary conversion table [7]. The proposed system realizes the high-speed operation without speed limitation of electrical signal processing. Meanwhile, the resolution of the proposed all-optical ADC depends on the spectral width of the wavelength-shifted signal after SSFS [4]. To improve the resolution of the all-optical ADC, spectral compression of the wavelength-shifted signal is one promising approach. In this paper, we describe the numerical study of self-phase modulation (SPM) induced spectral compression after SSFS for the resolution improvement of the proposed all-optical ADC.

2 Resolution improvement of the proposed all-optical ADC

The schematic diagram of the proposed all-optical ADC is shown in Fig. 1. It realizes optical quantization and optical coding after an optical sampling process. In optical quantization, we use SSFS in a fiber and a dispersion device. Since the amount of the center wavelength shift increases with increasing the peak power of an input pulse, power levels of an input analog signal converts into the amount of the center wavelength shift. After SSFS, each wavelength-shifted signal is promptly output to each different port of a dispersion device. In optical coding, we use the optical interconnection based on a binary conversion table. It allows us to broadcast the output signal after optical quantization to eign output port corresponding to each bit in a multiple-bit binary number. Consequently, each bit in a multiple-bit binary number corresponding to the power levels of an input analog signal can be detected by each binary photo detector.

In general, the resolution of ADC is described by the achievable quantization level $M$. In the proposed all-optical ADC, the achievable quantization level $M$ is described by eq. (1) [4].

$$M = \frac{\Delta \lambda_{FWHM}}{\Delta \lambda_{FWHM}}$$

(1)
where $\lambda_i$ and $\Delta \lambda_{FWHM}$ are the amount of center wavelength shift and the spectral width of wavelength-shifted signal after SSFS, respectively. From eq. (1), the compression of the spectral width $\Delta \lambda_{FWHM}$ enables the improvement of the achievable quantization level $M$ according to the spectral compression ratio. Previously, spectral compression due to the negatively chirped pulse SPM in a fiber has been proposed [8]. To apply this technique to the resolution improvement of the all-optical ADC, we should choose appropriate fiber combination to generate negatively chirped pulse SPM for wavelength shifted signal. To confirm the spectral compression of the wavelength shifted signal, we numerically study the SPM-induced spectral compression of wavelength shifted signal by propagating dispersion shifted fiber (DSF) and high-nonlinear fiber (HNLF).

3. Numerical simulation of SPM-induced spectral compression for wavelength shifted signal

To verify the SPM-induced spectral compression of wavelength shifted signal, we executed the numerical simulation of pulse propagation in fibers using a split step Fourier method. The schematic diagram is shown in Fig. 2. In the simulation, we used three fibers; HNLF1 for SSFS, DSF for generation of a negative dispersed pulse and HNLF2 for SPM. We used a transform limited sech$^2$ pulse as an input sampled analog signal. The center wavelength and the pulse width were 1558 nm and 0.53 ps, respectively. For the generation of SSFS, the input pulse was propagated in a 1 km HNLF1 (dispersion: $D=+7.2$ ps/nm/km, nonlinearity: $\gamma=16$ /W/km) As a result, the center wavelength of the input pulse was shifted to longer wavelength side depending on the peak power of an input pulse. The SSFS signal was propagated in a 50 m DSF ($D=+2.0$ ps/nm/km) for generation of a negative dispersed pulse and 100 m HNLF2 ($D=+0.28$ ps/nm/km, $\gamma = 9.0$ /W/km) for SPM. Figure 3 shows the simulation results of the spectral evolution of an input pulse propagating in fibers in the case of the peak power of an input pulse was 21 W. From these results, we can confirm the SPM-induced spectral compression of the wavelength shifted signal. Figure 4 shows the simulation results of the relationship between the center wavelength after SSFS and the spectral width before and after spectral compression.

4. Conclusion

We have described the resolution improvement of the proposed all-optical ADC using SPM-induced spectral compression. From numerical simulation results, we can confirm that the spectral compression ratio 2 to 1, or 3 dB resolution improvement was obtained. There is the promise of all-optical ADC with greater $M=32$ (5 bits) resolution.

References

Effects of Frequency Allocations and Zero Dispersion Frequencies on FDM Lightwave Transmission Systems

Jun Onishi, Shinya Kojima, and Takahiro Numai
Graduate School of Science and Engineering, Ritsumeikan University, 1-1-1 Noji-Higashi, Kusatsu, Shiga 525-8577, Japan
Tel.: +81-77-561-5161, Fax: +81-77-561-2663, E-mail: numai@se.ritsumei.ac.jp

Abstract
FWM noises decrease with an increase in a separation between signal frequencies and the zero dispersion frequency, and it is found that FWM noises in ERUS and URUS are much lower than those in ES.

1. Introduction
Transmission characteristics in frequency-division-multiplexing (FDM) lightwave transmission systems with low-dispersion optical fibers such as dispersion-shifted fibers are limited by four-wave mixing (FWM) [1].

We focus on the fact that characteristics of FWM are closely related to frequency allocations and a zero dispersion frequency. From the viewpoint of frequency allocations, unequally-spaced (US) allocations [2], repeated unequally-spaced (RUS) allocations [3], and modified RUSs such as equally-spaced RUS (ERUS) and unequally-spaced RUS (URUS) allocations [4] were demonstrated to overcome the problems in equally-spaced (ES) allocation [5]. It was found that RUS, ERUS, and URUS have lower FWM light intensities with signal frequencies than ES and narrower total bandwidths than US. With regard to the zero dispersion frequency, it has been shown that FWM noises decrease with an increase in a separation between the signal frequencies and the zero dispersion frequency in ES [6].

In this work, FWM noises are calculated by changing a separation between the middle frequency of a total bandwidth $f_M$ and the zero dispersion frequency $f_0$ in ES, ERUS, and URUS, and the calculated results are compared with each other. In our calculations, a dispersion shifted fiber (DSF) and a non-zero dispersion shifted fiber (NZDSF) are assumed to have fiber length $L$ of 80 km and a decay rate $\alpha$ of 0.2 dB/km. Moreover, DSF and NZDSF are assumed to have a derivative dispersion coefficient $dD/d\lambda$ of 0.07 ps/km/nm$^2$ and 0.05 ps/km/nm$^2$, respectively. An oscillation wavelength for a light source is assumed to be 1550 nm. The base unit and the channel spaces are common in all frequency allocations which are studied in this work, and the used values are the same as those in Refs. 3 and 4. It is revealed that FWM noises are reduced with an increase in $|f_M - f_0|$, and FWM noises in ERUS and URUS are lower than FWM noises in ES.

2. Calculated Results
2.1 Averaged FWM Efficiency
Figure 1 shows a relation between an averaged FWM light efficiency and a difference in light frequencies $f_M - f_0$ when the number of channels is 19 with DSF and NZDSF. Figures 1 (a) and (b) correspond to DSF and NZDSF, respectively. Open triangles, open circles, and closed circles correspond to ES, ERUS, and URUS, respectively. Here, $f_M$ is a middle frequency of a total bandwidth, and $f_0$ is the zero dispersion frequency.

In DSF, when $f_M - f_0 = 0$ THz, the averaged FWM light efficiencies for ES, ERUS, and URUS are 9 dB, $-3$ dB, and $-5$ dB, respectively. FWM efficiencies decrease with an increase in $f_M - f_0$. When $f_M - f_0 = 10$ THz, the averaged FWM light efficiencies for ES, ERUS, and URUS are $-36$ dB,
−50 dB, and −51 dB, respectively. These results show that FWM efficiencies decrease by 45 dB, 47 dB, and 46 dB for ES, RUS, and URUS, respectively, with an increase in \( f_M - f_0 \) from 0 to 10 THz. In NZDSF, when \( f_M - f_0 = 0 \) THz, the averaged FWM light efficiencies for ES, ERUS, and URUS are 9 dB, −2 dB, and −4 dB, respectively. When \( f_M - f_0 = 10 \) THz, the averaged FWM light efficiencies for ES, ERUS, and URUS are −33 dB, −47 dB, and −48 dB, respectively. These results indicate that FWM efficiencies decrease by 42 dB, 45 dB, and 44 dB for ES, ERUS and URUS, respectively, with an increase in \( f_M - f_0 \) from 0 to 10 THz. When \( f_M - f_0 = 0 \) THz, the averaged FWM light efficiencies for ES, ERUS, and URUS are −33 dB, −47 dB, and −48 dB, respectively. These results indicate that FWM efficiencies decrease by 42 dB, 45 dB, and 44 dB for ES, ERUS and URUS, respectively, with an increase in \( f_M - f_0 \) from 0 to 10 THz. When \( f_M - f_0 = 0 \) THz, the averaged FWM light efficiencies for ES, ERUS, and URUS are −33 dB, −47 dB, and −48 dB, respectively. These results indicate that FWM efficiencies decrease by 42 dB, 45 dB, and 44 dB for ES, ERUS and URUS, respectively, with an increase in \( f_M - f_0 \) from 0 to 10 THz. When \( f_M - f_0 = 0 \) THz, the averaged FWM light efficiencies for ES, ERUS, and URUS are −33 dB, −47 dB, and −48 dB, respectively. These results indicate that FWM efficiencies decrease by 42 dB, 45 dB, and 44 dB for ES, ERUS and URUS, respectively, with an increase in \( f_M - f_0 \) from 0 to 10 THz.

2.2 Bit Error Rate

In Fig. 2, the largest BERs among FDM channels are plotted as a function of a difference in light frequencies \( f_M - f_0 \) for ES, ERUS, and URUS with DSF and NZDSF when the number of channels is 19 and a modulation speed of 10 Gbit/s. Figures 2 (a) and (b) correspond to DSF and NZDSF, respectively. Open triangles, open circles, and closed circles correspond to ES, ERUS, and URUS, respectively. Without FWM noises, receiver sensitivity to achieve a BER of \( 10^{-9} \) is −20.7 dBm. In Fig. 2, BERs are calculated at receiver sensitivity of −20.7 dBm + 0.5 dBm = −20.2 dBm.

In DSF, when \( f_M - f_0 = 0 \) THz, the largest BERs among 19 channels in ES, ERUS, and URUS are 5.8 × 10^{-4}, 2.1 × 10^{-8}, and 1.2 × 10^{-8}, respectively. In NZDSF, the largest BERs among 19 channels in ES, ERUS, and URUS are 1.6 × 10^{-5}, 1.0 × 10^{-9}, and 5.7 × 10^{-10}, respectively. FWM noises in NZDSF are much lower than FWM noises in DSF. When \( f_M - f_0 = 10 \) THz, the largest BER among 19 channels in ES, ERUS, and URUS with DSF and NZDSF is 1.2 × 10^{-11}. Because these differences in BERs are extremely small, BERs for ES, ERUS, and URUS with DSF and NZDSF seem to overlap with an increase in \( f_M - f_0 \) from 4 to 10 THz. As can be seen from Fig. 2, BERs decrease with an increase in \( f_M - f_0 \).

3. Summary

FWM noises were reduced in ES, ERUS, and URUS with an increase in a separation between the middle frequency of a total bandwidth \( f_M \) and the zero dispersion frequency \( f_0 \) in DSF and NZDSF. In DSF, FWM efficiencies were reduced by 45 dB, 47 dB, and 46 dB for ES, RUS, and URUS, respectively, with an increase in \( f_M - f_0 \) from 0 to 10 THz. In NZDSF, FWM efficiencies were reduced by 42 dB, 45 dB, and 44 dB for ES, RUS and URUS, respectively. When \( f_M - f_0 = 10 \) THz, FWM efficiencies in ERUS and URUS were lower than FWM efficiencies in ES by at least 15 dB. These results unveiled that FWM noises in ERUS and URUS became much lower than FWM noises in ES with an increase in \( |f_M - f_0| \). BERs also decreased with an increase in \( |f_M - f_0| \). In DSF and NZDSF, BER of \( 10^{-9} \) was not obtained in ES. On the other hand, when \( f_M - f_0 = 10 \) THz, BER of \( 10^{-9} \) was obtained in ES. By comparing FWM noises in DSF and NZDSF, FWM noises in NZDSF were much lower than FWM noises in DSF.

References

Influence of Modulation Formats on FWM Noises in FDM Optical Fiber Transmission Systems

Yoshitaka Ito, Jun Onishi, Shinya Kojima, and Takahiro Numai
Graduate School of Science and Engineering, Ritsumeikan University, 1-1-1 Noji-Higashi, Kusatsu, Shiga 525-8577, Japan
Tel.: +81-77-561-5161, Fax: +81-77-561-2663, E-mail: numai@se.ritsumei.ac.jp

Abstract

Influence of modulation formats on four wave mixing noises in FDM optical fiber transmission systems is investigated. It is found that a newly proposed intra-bit division RZ format is most suitable to suppress FWM noises.

1. Introduction

Transmission characteristics in frequency-division-multiplexing (FDM) optical fiber transmission systems with low-dispersion optical fibers such as dispersion-shifted fibers are limited by four-wave mixing (FWM).

We focus on the fact that characteristics of FWM are closely related between modulation formats and frequency allocations. From the viewpoint of frequency allocations, unequally-spaced (US) allocations, repeated unequally-spaced (RUS) allocations [1], and modified RUSs such as equally-spaced RUS (ERUS) and unequally-spaced RUS (URUS) allocations [2] were demonstrated to overcome the problems in equally-spaced (ES) allocation [3]. It was found that RUS, ERUS, and URUS have lower FWM light intensities with signal frequencies than ES and narrower total bandwidths than US.

From the viewpoint of modulation formats, non-return-to-zero (NRZ) and return-to-zero (RZ) formats were usually used in optical fiber transmission systems.

In this work, we propose intra-bit division RZ (IBD-RZ) format to reduce FWM noises in FDM optical fiber transmission systems. FWM noises are calculated in ES, RUS, ERUS and URUS allocations with NRZ and IBD-RZ formats, and the calculated results are compared with each other. In our calculations, a dispersion shifted fiber (DSF) is assumed to have an oscillation wavelength for a light source is 1550 nm, fiber length $L$ is 80 km, a decay rate $\alpha$ is 0.2 dB/km, and DSF is assumed to have a derivative dispersion coefficient $dD_c/d\lambda$ of 0.07 ps/km/nm$^2$. The base unit and the channel spaces are common in all frequency allocations which are studied in this work, and the used values are the same as those in Refs. 1 and 2. It is revealed that FWM noises with IBD-RZ format are lower than FWM noises with NRZ format, and FWM noises in URUS are lower than FWM noises in other frequency allocations.

2. Theory

2.1 Frequency Allocations

Figure 1 illustrates frequency allocations of (a) ES, (b) US, (c) RUS, (d) ERUS, and (e) URUS. RUS, ERUS, and URUS are formed by repeating US as a base unit (BU). In RUS, there are not spaces between BUs. In ERUS, there is a common space $\Delta f_0$ between BUs. In URUS, spaces between BUs are not common such as $\Delta f_1$, $\Delta f_2$.

![Fig.1 Frequency allocation (a) ES, (b) RUS, (c) ERUS, and (d) URUS](image)

2.2 Modulation Formats

Figures 2 (a) and (b) show NRZ format and RZ format in a bit time $T$, respectively. These modulation formats were used in conventional optical fiber transmission systems.

![Fig.2 (a) NRZ format, (b) RZ format, and (c) IBD-RZ format](image)

IBD-RZ format, which is proposed in this work, is shown in Fig.2 (c). In IBD-RZ format, two RZ pulses are placed so as not to be overlapped in a bit time. The pulse’s duty rate is set to be $T/2$. Because
pulse A and pulse B are perfectly divided in a time region, IBD-RZ format has a time-division-multiplexing (TDM) transmission character [4]. Therefore, four wave mixing between pulse A and pulse B never happens.

3. Calculation

Averaged FWM efficiency and bit error rate (BER) in ES, RUS, ERUS and URUS allocations with NRZ and IBD-RZ formats are calculated. In IBD-RZ format-I, odd-number channels and even-number channels correspond to pulse A and pulse B in Fig.1(c), respectively. In IBD-RZ format-II, channels with frequencies \( f_1, f_2, \ldots, f_{10} \) and channels with frequencies \( f_{11}, f_{12}, \ldots, f_{19} \) correspond to pulse A and pulse B in Fig.1(c), respectively. In IBD-RZ format-III, channels with frequencies \( f_1, f_2, \ldots, f_9 \) and channels with frequencies \( f_{10}, f_{11}, \ldots, f_{19} \) correspond to pulse A and pulse B in Fig.1(c), respectively.

4. Calculated Results

4.1 Averaged FWM Efficiency

Figure 3 shows dependence of an averaged FWM light efficiency \( \eta \) on modulation formats and frequency allocations when the number of channels is 19.

![Fig. 3 Averaged FWM efficiency](image)

From Fig.3, it is found that IBD-RZ format-III in URUS has the lowest averaged FWM efficiency of \(-32.2\) dB, which is lower than averaged FWM efficiency with conventional NRZ format in ES by \(41.6\) dB.

4.2 Bit Error Rate

In Fig.4, BERs for channel 10 with a frequency \( f_{10} \) which is a midchannel in 19 channels, are plotted as a function of received power for a modulation speed of 10 Gbit/s.

Figures 4 (a) and (b) correspond to NRZ format and IBD-RZ format-III, respectively. A broken line, a dashed line, a dashed and dotted line, and a solid line correspond to ES, RUS, ERUS, and URUS, respectively.

These results indicate that IBD-RZ format-III in URUS allocation has the lowest BER, which means IBD-RZ format in URUS is most suitable in reducing FWM noises.

5. Summary

Influence of modulation formats and frequency allocations on FWM noises in FDM optical fiber transmission systems was investigated. IBD-RZ format was proposed as a new modulation format, and a combination of IBD-RZ format and URUS frequency allocation led to the lowest FWM noises.

![Fig. 4 Bit Error Rate](image)

References

Effects of Frequency Allocations and Polarization Allocations on FDM Lightwave Transmission Systems

Jun Onishi, Shinya Kojima, and Takahiro Numai
Graduate School of Science and Engineering, Ritsumeikan University,
1-1-1 Noji-Higashi, Kusatsu, Shiga 525-8577, Japan
Tel.: +81-77-561-5161, Fax: +81-77-561-2663, E-mail: numai@se.ritsumei.ac.jp

Abstract
Dependence of four-wave mixing (FWM) noises on frequency allocations and polarization allocations is investigated. FWM noises are drastically reduced when light polarizations are perpendicularly crossed at both sides of the zero dispersion frequency.

1. Introduction
Transmission characteristics in frequency-division-multiplexing (FDM) lightwave transmission systems with low-dispersion optical fibers such as dispersion-shifted fibers are limited by four-wave mixing (FWM) [1]

We focus on the fact that characteristics of FWM are closely related to frequency allocations and polarization allocations. From the viewpoint of frequency allocations, unequally-spaced (US) allocations [2], repeated unequally-spaced (RUS) allocations [3], and modified RUSs such as equally-spaced RUS (ERUS) and unequally-spaced RUS (URUS) allocations [4] were demonstrated to overcome the problems in equally-spaced (ES) allocation [5]. It was found that RUS, ERUS, and URUS have lower FWM light intensities with signal frequencies than ES and narrower total bandwidths than US. With regard to polarization allocations, it has been shown that FWM noises are reduced by arranging the polarization states of the channels [6].

In this work, FWM noises for ES, RUS, ERUS, and URUS are calculated using four polarization allocations such as (a) common polarizations, (b) orthogonal polarization states in adjacent channels, (c) orthogonal polarization states in adjacent base units, and (d) orthogonal polarization states in the channels located at both sides of the zero dispersion frequency. In our calculations, a dispersion shifted fiber (DSF) is assumed to have fiber length \( L \) of 80 km, a decay rate \( \alpha \) of 0.2 dB/km, and a derivative dispersion coefficient \( dD/d\lambda \) of 0.07 ps/km/nm². An oscillation wavelength for a light source is assumed to be 1550 nm. The base unit and the channel spaces are common in all frequency allocations which are studied in this paper, and the used values are the same as those in Refs. 3 and 4. It is revealed that FWM noises are drastically reduced when light polarizations are perpendicularly crossed at both sides of the zero dispersion frequency.

2. Calculated Results

2.1 Averaged FWM Efficiency

Figure 1 shows a relation between an averaged FWM light efficiency and polarization allocations in ES, RUS, ERUS, and URUS when the number of channels is 19. In Fig. 1, (a), (b), (c), and (d) show four polarization allocations where polarization directions are indicated by arrows. Open squares, gray squares, dark gray squares, and black squares correspond to ES, RUS, ERUS, and URUS, respectively.

In polarization allocation (a), the averaged FWM light efficiencies for ES, RUS, ERUS, and URUS are 9.4 dB, 1.7 dB, 2.9 dB and 4.9 dB, respectively. In polarization allocation (b), the averaged FWM light efficiencies for ES, RUS, ERUS, and URUS are 5.6 dB, 3.7 dB, 3.6 dB and 11.5 dB, respectively. In polarization allocation (c), the averaged FWM light efficiencies for ES, RUS, ERUS, and URUS are 5.4 dB, 3.6 dB, 14.6 dB and 8.8 dB, respectively. In polarization allocation (d), the averaged FWM light efficiencies for ES, RUS, ERUS, and URUS are 6.3 dB, 1.7 dB, 24.1 dB, and 30.6 dB, respectively. It is found that FWM noises are lowest in polarization allocation (d). This reason is as follows: When light frequencies \( f_1 \) and \( f_2 \) satisfy \( f_1 + f_2 = f_0 \pm \Delta f \) where \( f_0 \) is the zero dispersion frequency, FWM noises grow up significantly. Therefore, FWM noises are decreased when

Fig. 1 Averaged FWM Efficiency
polarization states of the channels are perpendicularly crossed at both sides of the zero dispersion frequency $f_0$.

2.2 Bit Error Rate

Figure 2 shows BERs of Channel 10, which is a midchannel of FDM signals, as a function of received power for ES, RUS, ERUS, and URUS. Here, the Gaussian approximation is used, a modulation speed is 10 Gbit/s, and the number of channels is 19. In Fig. 2, (a) and (d) indicate polarization allocations, and a broken line, a dash-double-dotted line, a dash-dotted line, and a solid line correspond to ES, RUS, ERUS, and URUS, respectively. In polarization allocation (d), BER for URUS overlap BERs for RUS and ERUS.

In polarization allocation (a) for ES, BER of $10^{-9}$ is not obtained. On the other hand, in polarization allocation (d) for ES, BER of $10^{-9}$ is obtained. Also, receiver sensitivities to achieve $10^{-9}$ for ES and that for the others are $-20.3$ dBm and $-20.7$ dBm, respectively. As can be seen in Fig. 2, BERs are efficiently reduced by alternating the polarization states of the channels.

2.3 Power Penalty

Figure 3 shows power penalties of Channel 10, which is a midchannel of FDM signals, as a function of an input power for ES, RUS, ERUS, and URUS, when the number of channels is 19 in polarization allocation (d). A broken line, a dash-double-dotted line, a dash-dotted line, and a solid line correspond to ES, RUS, ERUS, and URUS, respectively. Power penalty for URUS overlap power penalties for RUS and ERUS.

In polarization allocation (a), power penalties at an input power of 1.3 dBm/ch for RUS, ERUS, and URUS are 0.0988 dB, 0.0209 dB, and 0.0062 dB, respectively. On the other hand, in polarization allocation (d), power penalties at an input power of 1.3 dBm/ch for ES, RUS, ERUS, and URUS are 0.2999 dB, 0.0023 dB, 0.0016 dB, and 0.0015 dB, respectively. It is found that power penalties are also diminished by alternating the polarization states of the channels.

3. Summary

FWM noises were reduced in ES, ERUS, and URUS with arranging polarization allocations of the channels. Among four polarization allocations, FWM noises were lowest in polarization allocation (d). By comparing polarization allocation (a) and polarization allocation (d), averaged FWM efficiencies were reduced by 15.7 dB, 25.8 dB, 26.1 dB, and 25.7 dB for ES, RUS, ERUS, and URUS, respectively.

BERs and power penalties were also reduced with alternating polarization allocations of the channels. In polarization allocation (a), BER of $10^{-9}$ was not obtained in ES. On the other hand, in polarization allocation (d), BER of $10^{-9}$ was obtained in ES. Among four polarization allocations, BERs and power penalties in polarization allocation (d) were lowest.

References

Scalability of Code Matching for Optical Time-Series
WDM Encoded Label Using Collinear Acoustooptic Devices

Nobuo Goto\textsuperscript{1} and Yasumitsu Miyazaki\textsuperscript{2}
\textsuperscript{1}Toyohashi University of Technology, 1-1 Hibarigaoka, Tempaku-cho, Toyohashi 441-8580 Japan
\textsuperscript{2}Aichi University of Technology, 50-2 Manori, Nishihazama-cho, Gamagori 443-0047 Japan

Abstract We investigate a label recognition for optical codes encoded in time and spectral domains where time gating is not required. The system can recognize partial label, which make possible to improve the flexibility of routing control.

Introduction
Routing processing of optical packets will be a bottleneck in large-capacity photonic networks. Label routing system has been expected to improve the processing of optical packets. A variety of label processing methods have been proposed to use effectively the potential of optical signal processing.\textsuperscript{[1,2]} The authors have studied on collinear acoustooptic (AO) devices\textsuperscript{[3,4]} and applications to optical label recognition.\textsuperscript{[5,6]}

Since the AO devices can handle wavelength-division-multiplexed (WDM) optical signals, code recognition for optical codes encoded in spectral domain can be realized with a simple processing system. However, encoding in spectral domain limits the WDM number for optical packets. So, we consider combination of encoding in spectral domain and in time domain. Previously proposed systems for recognition in time domain required time gating to generate output signal. In this report, we propose a system that does not require time gating.

Structure of optical label recognition processor
The system consists of $M$ AO processors (AOPs), optical delay lines and electrical multipliers as shown in Fig.1(a). The label structure is shown in Fig.2. A label $m$ consists of $MN$ pulses $C^m = (c_1^m, c_2^m, ..., c_{MN}^m)$ and an identifying (ID) bit. Each pulse $c_i^m = (c_i^{m_1}, c_i^{m_2}, ..., c_i^{m_N})$ of the pulse train consists of $N$ WDM components. Each AOP consists of parallel collinear AO switches, delay waveguides, photodetectors (PDs) and an electrical multiplier as shown in Fig.1(b). Code matching between the optical incident label and a code represented by surface acoustic waves (SAWs) is performed with the AOP. We consider the operation in the $i$-th AOP. When a WDM optical pulse trains representing a label is incident, the pulse train is divided into $N$ pulse trains. The $i$-th pulse train is delayed for the delay time of $(N-1)iM_p$. We suppose that $i$-th part of label $k$ is recognized with the AOP. In the $i$-th AO switch, frequency multiplexed SAWs corresponding code $C_i^{k,i}$ is loaded in advance. The outputs from each AO switch are differentially detected with the PD. The electric pulse trains from the PDs of the parallel AO switches are processed with the electrical multiplier, resulting in a matched signal pulse.

In the whole system, the incident optical pulse train in each AOP is delayed with a delay line. The ID bit pulse is adjusted at the time of all the matched pulses for $M$ partial labels. Therefore, the matched output pulse is obtained when all the partial labels are matched. When the partial matched output for the $i$-th $N$ bits label is required, the matched output is obtained by multiplying the ID bit pulse with the output from the $i$-th AOP.

In the AO switches, matching is performed between the optical incident pulse train and frequency multiplexed SAWs. Although the propagation time of SAWs is slow, fast processing is expected after the SAWs are propagated along the interaction region. The SAWs are not needed to be changed during the matching process. In the previously proposed system, electrical addition or optical addition is required for on-off keying (OOK) code or binary phase shift keying (BPSK) code, respectively in addition to the time gating. So, the proposed system has a potential for high bit-rate labels because only electrical multiplication is required. The partial matching will be useful for layered label structure for QoS control.

Now, we consider the number of labels that can be recognized with each AOP. We assume WDM components of label consists of OOK orthogonal codes. Then the number of codes for $N$ WDM component is approximately $N^2$. The number of codes for $N$, bit WDM pulse train is $(N^2)^{1/2}$. For the whole system of $M$ AOPs, the number reaches $(N^2)^{MN}.M$. For example, when $N = 4, N = 4, M = 4$, the number of recognizable code is $4^{16}$.

Performance simulation of each AOP
Next, we evaluate the code recognition characteristics of an
The interaction length, $l$, in Fig. 3. We assume the SAWs, $(a, a, a)$, and $(a, a, a)$. We verified the proposed AOP by considering wavelength selectivity of the collinear AO switches. As a numerical example, we assume a bandwidth limited optical pulse train with pulse period 12 ps, pulse width 5.9 ps, bandwidth 160 GHz and $N_s = 3$. The interaction length of collinear AO switch is assumed to be $l_{sw} = 16$ mm. We assume an OOK orthogonal code set as given by

$$c^n = \{a_1, a_2, a_3\} = \begin{bmatrix} 1 & 0 & 1 \\ 0 & 1 & 1 \\ 1 & 1 & 0 \end{bmatrix}$$  

(1)

Optical incident pulses $E^n(t)$ at wavelength $\lambda_3$ are shown for the case of $C^n = (a_1, a_1, a_1)$ in Fig. 3. We assume the SAWs in collinear AO switches are set for recognizing codes $C = (a_1, a_1, a_1)$. The optical output is evaluated for the cases of optical inputs $C^n = (a_1, a_1, a_1)$ and $C^n = (a_1, a_2, a_3)$. Time shifted pulse trains are obtained from the parallel AO switches and multiplication of these pulses results in the output as shown in Fig. 4. The output power for all the optical codes are calculated as shown in Fig. 5. It is found that only the matched code results in large output. Fig. 6 shows the outputs corresponding to the matched code and minimum value of unmatched output, that is, the maximum absolute value, for all the codes. It is concluded that all the codes can be distinguished.

Conclusions

Optical code recognition system consisting of collinear AO switch array was investigated. By using electrical multiplication, a simple processing that does not require time gating is performed. A part of the incident label can be recognized by multiplying an ID bit pulse. This will be useful for layer-structured network control. The code recognition characteristics were evaluated for the case of ideal switching and that of actual AO switching having wavelength selectivity. In future plan, the total system for routing will be discussed. We also consider experimental verification of the proposed code recognition.

References

Crossover Reducing Integrated Nested Rings Optical Switch

Nan Xie and Katsuyuki Utaka
Graduate School of Advanced Science and Engineering, Waseda University
3-4-1 Okubo, Shinjuku-ku, Tokyo 169-8555, Japan. Tel/Fax: +81-3-5286-3394
nani@akane.waseda.jp

Abstract
A novel integrated optical switch topology based on nested rings of MMI switching elements. Grid switch architectures can be realized with reduced number of switches and crossovers compared to the conventional Beneš system.

Introduction
Previously, we proposed a novel non-blocking Multimode Interference-based Ring Switch (MIRS) using a ring topology of 2x2 MMI elemental switches [1]. While conventional switch networks are formed by the side-by-side cascading topology, the MIRS is realized by placing rings around a central core. Regarding a scalable NxN switch network, the aim of this device is to reduce the number of elemental switches and crossovers, smaller physical device area, and omni-directional access.

Reduced MIRS Topology
We first digress from the conventional cascade network of elemental 2x2 switches and realize that a ring arrangement of four MMI switches, along with access MMI switches, gives rise to a 4x4 system. This becomes the core of the MIRS, illustrated in Fig. 1(a). The labels A-H indicate the respective elemental 2x2 MMI switches and numbers 1-4 and 5-8 indicate four input and output ports respectively. A closer view of the pattern leads to an extension of the Beneš switch with inter-row cross-switches that facilitate the feedback between individual input ports. Intuitively, the switching permutations of on/off elements A-H are flexible. Assuming the MMI ‘off’ state generates cross-images then straight-images are formed with the ‘on’ state. We can realize any combination of paths except for when port 1 connects to 2 and port 3 to 4 simultaneously. In total, there are 66 optimal feedback combinations.

Abstract
A novel integrated optical switch topology based on nested rings of MMI switching elements. Grid switch architectures can be realized with reduced number of switches and crossovers compared to the conventional Beneš system.

Reduced MIRS Topology
We first digress from the conventional cascade network of elemental 2x2 switches and realize that a ring arrangement of four MMI switches, along with access MMI switches, gives rise to a 4x4 system. This becomes the core of the MIRS, illustrated in Fig. 1(a). The labels A-H indicate the respective elemental 2x2 MMI switches and numbers 1-4 and 5-8 indicate four input and output ports respectively. A closer view of the pattern leads to an extension of the Beneš switch with inter-row cross-switches that facilitate the feedback between individual input ports. Intuitively, the switching permutations of on/off elements A-H are flexible. Assuming the MMI ‘off’ state generates cross-images then straight-images are formed with the ‘on’ state. We can realize any combination of paths except for when port 1 connects to 2 and port 3 to 4 simultaneously. In total, there are 66 optimal feedback combinations.

If feedback is omitted, we can reduce the 4x4 MIRS, shown in Fig. 1(b), with 24 optimal path combinations out of a total of 64. The total number of required switching elements is five, which is one fewer compared to the conventional Beneš configuration.

Fig. 1. (a) 4x4 MIRS in the ring topology and (b) the reduced version

Fig. 2. 8x8 MIRS in the nested ring topology

The MIRS equation for the number of elemental switches required for an NxN network is represented by m.

\[ m(N) = \frac{7N}{8} + m\left(\frac{N}{2}\right) + 3m\left(\frac{N}{4}\right) \quad (N=16, 32, 64...) \]

This switch network configuration can reduce m up to 20% when N=8 and 11% at N=64 with respect to the Beneš topology. If access ports are restricted only in the horizontal directions, we can realize equivalent Beneš
setups. This, however, eliminates the omni-directional nature of the MIRS.

**MMI Behaviors**

MMI simulations using the Eigenmode Expansion Method (EEM) showed low loss and low crosstalk using bi-directional inputs. For the MMI switching operation, the MIPS method is utilized. Fig. 3 shows the behavior of an index-modulated MMI switch. The ‘off’ state output is at the lower output region while the ‘on’ state causes the interfered fields to relocate towards the upper output region. Considering polymer materials, as an example, an MMI switch with a length of 200µm and a Δn of 0.03 produces a crosstalk less than -30dB. For curved MMI structures, a comparison between cross-port power loss of polymer and silicon MMI couplers as well as plain waveguides is illustrated in Fig. 4. Since higher refractive index difference between the core and the cladding is desired for waveguide bending, the same is true for MMI couplers. In this case, for a bending MMI device power loss of 0.1dB, the radius of curvature for silicon is 500µm, 5-8 times less than typical polymers. Typically, polymer waveguide widths are usually larger compared to semiconductor waveguides.

**Grid Architecture**

In the simplest case of multi-directional access, 8x8 MIRS can be arranged in a grid-like fashion shown in Fig. 5. In this case, node-to-node distance is around the length of five MMI elements plus the diameter of curvature of the MIRS outer ring. Suppose the MMI lengths are 100µm with a 0.1dB device loss diameter of 500µm, then the node-to-node distance is around 2mm. The area of a grid network of 40 ports, shown in Fig. 5, is less than 1cm².

**Crossover Reduction**

In an integrated optical switch system, crossovers are difficult to be realized. One possibility of realizing integrated waveguide crossovers is by using the three-dimensional MMI method [4] with polymers. This method also enables the implementation of multilayer optical devices to increase device density. A comparison between the number of crossovers in the MIRS and the Beneš systems is illustrated in Fig. 6. The benefit of the MIRS and the MMI coupler is visible with 34% reductions in the number of crossovers compared to the Beneš system when N=16.

**Conclusion**

We report a new switching network topology to reduce the number of elemental switches compared to the conventional Beneš setup. The MMI devices are included in the MIRS due to their promising dual input isolation, low loss capabilities, switching operation, and curving options. The MIRS benefit for the reduction of waveguide crossovers is evident as well as the ability to maximize device real estate in a grid-like network topology.

**References**


Novel type BOTDA for measuring distributed strain and temperature with cm spatial resolution in km-long fiber

Yahei Koyamada and Satoshi Sotoyama
Department of Media & Telecommunications Engineering, Ibaraki University
4-12-1 Nakanarusawa, Hitachi, 316-8511 Japan
Tel/Fax: +81-294-38-5128, E-mail: koyamada@mx.ibaraki.ac.jp

Abstract: A novel type Brillouin optical time-domain analysis (BOTDA) system, called double-pulse BOTDA, is proposed for measuring distributed strain and temperature with a centimeter spatial resolution in a km-long fiber.

1. Introduction

The Brillouin scattering process in a fiber has been widely investigated as a basis for measuring distributed strain and temperature since the Brillouin frequency shift (BFS) has a linear dependence on both strain and temperature. Brillouin optical time-domain analysis (BOTDA) is a promising method for measuring distributed BFS [1]. However, the spatial resolution of the conventional BOTDA system is 1 m at best, which is insufficient for use in many application areas. This is because the Brillouin gain spectrum (BGS) measured by the BOTDA system broadens rapidly beyond 100 MHz as the pulse width decreases below 10 ns, which corresponds to a 1-m spatial resolution. This makes it difficult to measure BFS accurately.

Recently, a BOTDA system with a sub-meter spatial resolution has been developed, called the PPP-BOTDA system [2]. This system is based on the Brillouin scattering caused by the acoustic wave that is induced before the passage of a pulsed light [3], [4]. However, it has been pointed out by the developers of this system that there may be a considerable error in the measured BFS [5]. Moreover, such an error increases as the pulse width decreases and this makes it difficult to realize centimeter (cm) spatial resolution with this system.

This paper proposes a novel BOTDA system, called the double-pulse BOTDA (DP-BOTDA) system, which allows true cm spatial resolution for a long length of fiber. Our numerical simulation, which includes an estimation of the signal-to-noise ratio of the system, shows that it is possible to measure distributed BFS with a spatial resolution of 4 cm and accuracies of 1-2 MHz for a 5-km long fiber. BFS accuracies of 1-2 MHz correspond to strain accuracies of 20-40 με.

2. Principle of DP-BOTDA

Figure 1 is a schematic diagram of the DP-BOTDA system. A double-pulsed light instead of a conventional single-pulsed light is launched into a sensing fiber from the input end U, and a cw light is launched from the other end V to measure the Brillouin gain along the length of the fiber. The spatial resolution of the Brillouin gain measured with this system depends on the width of the rear pulse. We carried out numerical simulations with respect to a DP-BOTDA system operating at a wavelength of 1.55 μm based on the coupled differential equations for optical pump and probe waves and an acoustic wave [6]. This DP-BOTDA system transmits double-pulsed lights with a front-pulse width of 0.6 ns, a rear-pulse width of 0.3 ns and a front-pulse to rear-pulse time interval of 5 ns.

First we consider an acoustic wave induced by the interaction between a double-pulsed light and a counter-propagating cw light in the fiber. The acoustic wave amplitude $\rho$ at a point along the length of the sensing fiber is shown in Fig. 2 as a function of time $t$ and $\Delta \nu - f_B$, where $\Delta \nu$ is the frequency difference between the double-pulsed light and the cw light, and

![Fig. 1. Schematic diagram of DP-BOTDA system.](image1.png)

![Fig. 2. Induced acoustic wave amplitude.](image2.png)

![Fig. 3. Brillouin gain at a point along the length of fiber.](image3.png)
$f_B$ is the BFS of the fiber. The front and rear pulses of the double-pulsed light pass through this point when $-0.3 \text{ ns} < t < 0.3 \text{ ns}$ and $4.85 \text{ ns} < t < 5.15 \text{ ns}$, respectively. We can see that the acoustic wave grows rapidly during the passage of the front pulse and then decays exponentially with time when there is no pump light. The amplitude $\rho$ induced by the front pulse does not change greatly with $\Delta \nu$. However, the amplitude $\rho$ changes greatly and periodically with $\Delta \nu$ after the passage of the rear pulse. This periodic change is caused by the interference between acoustic waves induced by the front and rear pulses of the double-pulsed light. Figure 3 shows the Brillouin gain $G_B$ that the cw light experiences when passing through the point under consideration. The gain $G_B$ in the $t=0.3$ to 0.3 ns time range is attributed to the front pulse of the double-pulsed light and that in the $t=4.85$ to 5.15 ns range is attributed to the rear pulse. We can see that the gain $G_B$ produced by the rear pulse has a broad but oscillatory frequency spectrum corresponding to the frequency dependence of the induced acoustic wave shown in Fig. 2. The oscillation period in the frequency spectrum is 200 MHz, which corresponds to a front-pulse to rear-pulse time interval of 5 ns. By making use of this oscillatory nature we can determine the Brillouin frequency shift $f_B$ accurately in spite of the very narrow pulse width of 0.3 ns.

3. Simulation of distributed measurement

The 260-cm length of test fiber shown in Fig. 4 is considered here. It consists of seven sections: A (60 cm), B (2 cm), C (68 cm), D (4 cm), E (66 cm), F (8 cm), and G (52 cm). The BFS, $f_B$, is $f_0$ MHz in sections A, C, E, and G and $f_0 + 80$ MHz in sections B, D, and F. A difference of 80 MHz in the BFS corresponds to a 1621- $\mu e$ strain in the fiber. We performed numerical simulations of measuring the BFS along the length of the test fiber with a DP-BOTDA system. Figure 5 shows simulated BGSs at (a) $z=30$ cm and (b) 204 cm that correspond to the centers of fiber sections A and F, respectively. We can see that the peak frequencies of the BGSs are located almost at the BFSs in sections A and F, respectively. Since each BGS has a periodic and steep frequency dependence, we can find the peak frequency, i.e. the BFS, accurately. Figure 6 shows simulated and actual distributions of the BFS along the length of the test fiber. Here, the simulated BFS was determined from the peak frequency of the BGSs at each point. We can see that the simulated curve well traces the actual curve except in the $z=60$-62 cm range, which corresponds to section B. The discrepancies between the simulated and actual BFSs at the centers of sections B, D, and F are 17, 0.6, and 0.3 MHz, respectively. Figure 6 reveals that the error in the simulated BFS caused by the gain induced by the front pulse of the double-pulsed light is no larger than 0.3 MHz.

We numerically estimated the signal-to-noise ratio of the DP-BOTDA system under the following conditions: a double-pulsed light peak power of 300 mW at input end U, a cw light power of 8 mW at input end V, a sensing fiber length of 5 km, and $10^4$ integrations. As a result, we found that the noise-induced error in the measured BFS was 1-2 MHz, which corresponds to a strain error of 20-40 $\mu e$.

4. Conclusion

We proposed a novel BOTDA system that can measure distributed strain and temperature with a cm spatial resolution in a km-long fiber.

References
All-optical Variable Optical Attenuator Based on Nonlinear Optical Fiber with Fiber Bragg Grating

Seongmin Ju, Pramod R. Watekar¹, Yune Hyoun Kim², Taekjung Kim, and Won-Taek Han¹

Research Center for Specialty Optical Fibers, OptoNest
958-6 Daecheon-dong, Buk-gu, Gwangju, 500-470, South Korea

¹Department of Information and Communications, Gwangju Institute of Science and Technology
1 Oryong-dong, Buk-gu, Gwangju, 500-712, South Korea

²Industrial Materials R&D, LG Chem. Ltd.,
Research Park, 104-1 Moonji, Yuseong, Daejeon 305-380, South Korea

Tel: 82-62-970-2215, Fax: 82-62-970-2204, E-mail: wthan@gist.ac.kr

Abstract
A novel all-optical variable optical attenuator based on a nonlinear optical fiber with fiber Bragg grating was developed with more than 35 dB variation controlled by a laser diode.

Introduction
Variable optical attenuator (VOA) in the optical communication system is a key component for gain control of optical amplifiers in dense wavelength division multiplexing (DWDM) systems and for dynamic channel power regulation and equalization in the cross-connected nodes. Also, VOA plays an important role for dynamic optical networks that perform protection and restoration the functions of the high-priced optical communication devices such as repeaters through decreasing power periodically or continuously. However, several types of the VOAs in the market suffer from disadvantages of complicated fabrication, high price, and response time. [1-5]

In the current communication, we have proposed and demonstrated, for the first time, the all-optical variable optical attenuator (AVOA) operating in the 1550 nm wavelength region based on the highly nonlinear optical (NLO) fiber, the Yb³⁺/Al³⁺ co-doped fiber, with the fiber Bragg grating (FBG).

Experimental
A nonlinear optical fiber doped with Yb³⁺/Al³⁺ ions was fabricated by the modified chemical vapor deposition (MCVD) and the doping solution method. The resonant optical nonlinearity of the fiber was measured using the method proposed by our group earlier and it was found to be ~ 7.5 × 10⁻¹⁵ m²/W. [6]

A FBG with ~ 45dB around 1550 nm was written on the hydrogen loaded Yb³⁺/Al³⁺ co-doped fiber by using a phase mask with the KrF excimer laser (248 nm). The maximum attenuation level, center wavelength, grating length, and bandwidth of the fabricated FBG are about 45 dB, 1549.84 nm, 1 cm, and 0.35 nm, respectively.

The experimental setup for all-optical variable attenuator is shown in Fig. 1, where the Yb³⁺/Al³⁺ co-doped nonlinear optical fiber with FBG was spliced between two 980 nm/1550 nm wavelength division multiplexers (WDM). An ASE source was used as a signal at 1550 nm and a 980 nm laser diode (LD) as the pumping source. The total length, L₁, between two WDM couplers was 10 cm, where the length, L₂, of FBG was 1 cm. The attenuation at required wavelength was monitored using the optical spectrum analyzer (OSA, Ando AQ6317B).

Results and discussion
Optical transmission of the Yb³⁺/Al³⁺ co-doped NLO fiber with FBG was measured by increasing the 980 nm LD pump power and the results are shown in Fig. 2. The transmission spectrum at the resonance wavelength formed by the FBG was found to shift towards the longer-wavelength side with the increase of the pump power. Since the transmission near 1550 nm decreased with the increase of the pump power, the maximum attenuation of about 35 dB
It is easily noticed that the usable range of the present AVOA depends on the selected wavelength and the optical attenuation level depends on the LD pump power. The attenuation range of the present AVOA at 1549.75 nm was 33.53 dB from Max. 45.62 dB (0 mW) to Min. 8.32 dB (77 mW) and the attenuation range at 1549.84 nm was 37.30 dB from Max. 45.62 dB (0 mW) to Min. 8.32 dB (77 mW). On the other hand, in the case of the FBG bands located at longer wavelength, the attenuation increased with the increase of the pump power. The attenuation range at 1550.13 nm was 36.93 dB from Min. 8.69 dB (0 mW) to Max. 45.62 dB (77 mW) and that at 1550.21 nm was 38.36 dB from Min. 1.67 dB (0 mW) to Max. 40.03 dB (77 mW).

The proposed AVOA using the NLO fiber with FBG can be extended to use at various wavelength by changing the position of the FBG reflection band. In addition, the control power of the LD can be decreased by increasing the nonlinearity of the optical fiber and by decreasing the bandwidth of the FBG.

**Conclusion**

We have developed the novel AVOA by using the Yb$^{3+}$/Al$^{3+}$ co-doped NLO fiber with the FBG. The attenuation was found to vary by pumping with the LD at 980nm and the total optical attenuation range of about 35dB was obtained near 1550nm pumped from 0 to 77 mW.

**Acknowledgment**

The authors are grateful to OptoNest Corporation, South Korea. This work has been supported by the GIST Technology Initiative (GTI), South Korea.

**References**

Multi-wavelength fiber laser sources with an elliptical core side-hole fiber Sagnac loop filter

Dae Seung Moon1, Bok Hyeon Kim1, Guoyong Sun1, Young-Geun Han2, Won-Taek Han1 and Youngjoo Chung1

1 Department of Information and Communications, Gwangju Institute of Science and Technology (GIST)

1 Oryong-dong, Buk-gu, Gwangju 500-712, Korea

Tel: +82-62-970-2214, Fax: +82-62-970-3137, Email: yehung@gist.ac.kr

2 Hanyang University, Department of Physics

17 Haengdang-dong, Seongdong-gu, Seoul 133-791, Korea

Abstract

We propose and experimentally demonstrate a multi-wavelength fiber laser based on the Sagnac loop filter using elliptical core side-hole fiber with elliptical or circular cladding shape. We could obtain 18 discrete channels with SNR over 30 dB and channel spacing of 0.8 nm.

1 Introduction

Multi-wavelength fiber lasers are cost-effective sources in wavelength division multiplexed (WDM) optical communication systems, fiber sensors, and optical instrument testing. Semiconductor optical amplifier (SOA)-based multi-wavelength fiber lasers exhibit stable operation because the SOA has the property of inhomogeneous broadening and thus can support simultaneous oscillation of many lasing wavelengths. Several techniques have been proposed to achieve multi-wavelength fiber lasers. Among them, the use of a fiber Fabry-Perot filter increases the insertion loss of the cavity, and a Mach-Zehnder filter is sensitive to environmental changes due to the difference in the optical path lengths of the two arms [1]. The Sagnac loop filter incorporating high birefringence fiber has advantages of simple configuration and better stability compared with the filters based on Fabry-Perot and Mach-Zehnder interferometers. In this paper, we propose and experimentally demonstrate a multi-wavelength fiber laser based on Sagnac filter using the elliptical core side-hole fiber with elliptical or circular shape.

2 Experiment

To fabricate the elliptical core side-hole fiber, we used the MCVD process. Figure 1 shows the cross section of the fabricated side-hole fiber with an elliptical core. The elliptic-shape fiber with elliptical core (Fig. 1(a)) was fabricated by over-jacketing (19 X 25 tube) and collapsing after cutting both sides of the preform. The fiber was drawn at the temperature of 1930°C. The relative index difference Δn (peak) was 0.02. The major/minor axes of the core and the side-hole diameter were 12 μm/6.6 μm and 20 ~ 25 μm, respectively [2]. To fabricate the circular-shape fiber with elliptical core (Fig. 1(b)), two holes with the diameter of 5 mm were drilled at both sides of the core in the preform [3]. The core with elliptical shape was made by partially collapsing the holes during fiber drawing at 2000°C. The major/minor axes of the core and the side-hole diameter were 8.6 μm/3.8 μm and 23 μm, respectively. The relative index difference Δn (peak) was 0.018. Generally, the wavelength separation between two transmission peaks of the Sagnac loop filter’s output is given by Δλ=2λ^2/BL (where B is the modal birefringence, L is the length of the Hi-Bi fiber, and λ is the operation wavelength) [4]. From the transmission spectrum of Sagnac loop filter as shown in Fig. 2(a), the modal birefringence of the fabricated elliptical core side-hole fiber was 1.7×10^-4 (elliptical shape) and 1.16×10^-4 (circular shape), respectively.
By proper adjustment of the polarization inside the cavity, we could obtain the multi-wavelength output spectra of the proposed fiber laser when the two SOAs were driven with the injection current of 200 mA. Figure 4(a) shows the output spectrum composed of 10 discrete channels with less than 3 dB power variation among them. The channel spacing was ~1.4 nm and SNR was over 30 dB. Figure 4(b) shows the output spectrum with 18 discrete channels and less than 3 dB power variation. The channel spacing was ~0.8 nm (100 GHz at 1550 nm, i.e. WDM ITU-grid spacing) and the SNR was over 30 dB.

3 Conclusion

In this work, we proposed and experimentally demonstrated a multi-wavelength fiber laser based on the Sagnac loop filter using the elliptical core side-hole fiber with elliptical or circular cladding shape. By properly adjusting the polarization controller, we could obtain 18 discrete channels with SNR over 30 dB and channel spacing of 0.8 nm and 10 discrete channels with SNR over 30 dB and channel spacing of ~1.4 nm. The fiber laser proposed in this work is expected to find various applications as a tunable multi-wavelength light source for optical communication and sensing.

Acknowledgements

This work was partially supported by the Basic Program Project of KOSEF (Grant No. R01-2006-000-11088-0), the Second-Phase of the Brain Korea 21 Project and the Next-Generation Growth Engine Project of MOCIE.

4 References

Investigations on sensitivity of long period gratings modified by gold nanoparticles

Jaw-Luen Tang\(^1\), Jian-Neng Wang\(^2\), and Ping-Chang Ju\(^3\)

\(^1\)Department of Physics, National Chung Cheng University, 168 University road, Chia-Yi 621, Taiwan
TEL: +886-5-272-0586, FAX +886-5-272-0587, Email: phjflf@ccu.edu.tw
\(^2\)Department of Construction Engineering, National Yunlin University of Science and Technology, Yun-Lin 640, Taiwan
TEL: +886-5-534-2601-4723, FAX +886-5-531-2049, Email: wangjn@yuntech.edu.tw
\(^3\)IR Section, Materials and Electro-Optics Research Division, Chung Shan Institute of Science and Technology, Lung-Tan, Taiwan

Abstract:
A simple and effective method is proposed to improve spectral sensitivity and detection limit of long period gratings for refractive index or chemical sensing, where the grating surface is modified by colloidal gold nanoparticles.

Introduction
Recently, long period fiber gratings (LPGs) have been demonstrated as excellent refractive index (RI) sensors for chemical and biochemical sensing. At the same time, there has been an increasing interest in applying the extraordinarily optical properties (e.g., absorbance and resonance wavelength) of noble metal colloidal nanoparticles such as Au or Ag, due to their potential applications for chemical and biological sensing. A number of nanoscale biosensors and chemosensors have been realized through shifts in the localized surface plasmon resonance (LSPR) extinction maximum of gold or silver nanoparticles [1-5]. The wavelength shifts are mainly caused by adsorbate-induced local refractive index changes at the surfaces of nanoparticles. Among the LSPR sensors, we have recently demonstrated an optical fiber biosensor that exploits the LSPR of self-assembled Au colloids (SAGC) on the grating portion of a long period fiber grating (termed CM\(_{Au}\)LPG) [6]. The attenuated total reflection spectrum of SAGC is sensitive to the refractive index of its surrounding medium, which can be used for monitoring the solution bulk and for label-free detection of antigen/antibody binding at the surface of the Au colloids.

Taking sucrose solution and sodium chloride solution as examples, this paper will present the significant enhancement of sensitivity and limit of detection for chemical solution concentration sensing using an LPG sensor with its cladding surface modified by gold nanoparticles.

Principle of Refractive Index Sensing
An LPG is a photo-induced periodic modulation of RI along the core of a single-mode fiber, with a typical index perturbation of 10\(^{-4}\), grating periods between 100 \(\mu\)m-1 mm and length of 2-5 cm. The LPG couples light from a guided fundamental core mode LP\(_{01}\) to different forward-propagating cladding modes HE\(_{m0}\) in an optical fiber. The coupling of the light into the cladding region generates series of resonant bands centered at wavelength \(\lambda_m\) in the transmission spectrum, since a cladding mode is rapidly attenuated in the fiber due to the scattering losses. The center wavelengths \(\lambda_m\) of an attenuation band are solutions of the following phase matched conditions: [7]

\[
\lambda_m = \left[ \pi n_{core}^{01} - \pi n_{cladding}^{1m} \right] / \Lambda
\]

where \(\pi n_{core}^{01}\) is the effective refractive index of the fundamental core mode at the wavelength of \(\lambda_{m}\), which is also dependent on the core refractive index and cladding refractive index. Also \(\pi n_{cladding}^{1m}\) is the effective refractive index of the \(m\)th cladding mode at the wavelength \(\lambda_{m}\), which is also a function of cladding refractive index and the refractive index of the surrounding medium. \(\Lambda\) is the period of grating. When the concentration or the refractive index of the surrounding medium changes, also \(\pi n_{cladding}^{1m}\) changes and a wavelength shift can be obtained. The wavelength shift can be linearly related to the concentrations of the solution under test. LPG can be very sensitive to the changes in temperature and deformations by fiber imperfections and bending. Therefore, temperature changes and strain effects must be compensated or avoided.

Experimental
The LPGs studied in this work were fabricated in Furakawa SM-332 hydrogen-loaded fibers, utilizing an amplitude technique with a pulsed 193nm ArF excimer laser (laser energy of ~135 mJ/cm\(^2\) and total exposure time of ~3 minutes) operating at 10Hz. The fiber was hydrogen loaded at a pressure of 120 bar over a period of two weeks at room temperature. The length of the LPG was 2.3 cm long and the grating period was about 550\(\mu\)m long. After the laser exposure, all the LPGs were annealed at 150 \(^\circ\)C for 24 h to stabilize their transmission spectrum. The preparation of Au colloids can be found in Ref. [2] and the production of colloidal Au-modified LPGs (CM\(_{Au}\)LPG) was followed by the procedure as reported in Ref. [6]. The fiber-optic sensing system used to measure the transmission spectrum of the sensor is also adopted from Ref. [6]. In this work, a number of LPGs an CM\(_{Au}\)LPG sensors associated with various attenuation bands were measured and investigated.
Results and Discussion

For all LPGs used in this work the strain sensitivity were found to be very small (0.09 ± 0.01 pm/µe) and exhibited nearly insensitive to strain changes. The sensitivity of the LPG sensors to temperature was about 0.05 ± 0.02 nm/°C. We kept sample solution at the same temperature (within 0.1°C). Therefore, the results reported here were not influenced by temperature and strain effects. To characterize the LPG or CM₃₄AuLPG as a concentration sensor, measurements with sucrose and sodium chloride aqueous solutions were performed. The surrounding RI was controlled through the use of sucrose solutions with various concentrations [6]. The experiment used to measure refractive index sensitivity was followed by Ref. [8]. Our results for both solutions show that both LPG and CM₃₄AuLPG sensors exhibited a linear decrease in the transmission loss and resonance wavelength shift when the concentration increased, demonstrating that LPG and CM₃₄AuLPG are suitable for chemical sensing.

By comparison, when a LPG and a CM₃₄AuLPG was used to measure the concentration of a sucrose solution, our results show that RI sensitivity of LPG sensor in term of refractive index increased from -33 ± 1.0 nm/RI to -36 ± 1.0 nm/RI in the range of 1.34-1.41, as shown in Fig. 1 and Fig. 2, respectively. Combined with the sensitivity analysis of transmission loss peak this leads to a RI resolution of 8.6×10⁻³ and 2.92×10⁻⁴, respectively. The measured response of sodium chloride solutions by weight concentration from 0% to 20 %, with and without the modification of colloidal gold nanoparticles, indicated that the sensitivity increased from -0.05 ± 0.01 (nm/%) to -0.07 ± 0.01 (nm/%), respectively. When conducting the same performance tests, experimental results showed that the accuracy of concentration measurement with this gold-coated LPG sensor was better than 0.2 ± 0.1 % by weight and limit of detection can be improved from 0.04 ± 0.01 % to 0.02 ± 0.01 %. The limit of detection of 0.0001 % by weight can easily be achieved if an optical fiber analyzer with spectral resolution of 0.1 pm is used. Therefore, results presented in this paper demonstrated that gold nanospheres modified on the fiber grating could increase its sensitivity in detecting solution concentrations significantly.

Conclusion

In this paper, we have demonstrated an optical fiber sensor based on anomalous reflection of self-assembled Au colloids on the grating portion of an LPG that can significantly enhance the sensitivity and detection limit for chemical solutions. This type of sensor has potential applications in medical diagnostics, biochemical sensing, and environmental monitoring. The advantage of this type of the sensor is relatively simple of construction, small, light, robust, low-cost, and ease of use. Moreover, the sensor has the potential capability for on-site, in vivo, and remote sensing, and has the potential use for disposable sensors.

References

Simple pure apodization method for fiber Bragg gratings by sequential UV writing

Kuei-Chu Hsu and Yinchieh Lai
Dept. of Photonics & Inst. of Electro-Optical Engineering, National Chiao Tung Univ., Hsinchu, Taiwan 300
Tel: +886-3-571-2121 ext 56335, Fax: +886-3-571-6631, Email: jessica.eo91g@nctu.edu.tw

Abstract
A simple UV exposure method is proposed to achieve pure apodization for fiber Bragg gratings fabricated by sequential UV writing. Through the exposure phase and/or time control of multiple UV shots, the ac-index can be adjusted independently with the dc-index kept constant.

1 Introduction
Many pure apodization methods for realizing advanced fiber Bragg gratings (FBGs) have been developed in the literature during the past decade [1-3]. Here pure apodization means to keep the average refractive index to be constant through the entire grating length while the ac index modulation can change independently. Previous developed technologies include the double-ultraviolet exposure method, complex design of phase mask, moving-fiber scanning-beam dithering technique, phase mask method with polarization control, and two beam interference with polarization control. However, these methods may suffer from the introduction of additional uncertainties, requiring additional optical elements, limitation of grating length, or only being suitable in either phase mask or holographic approach. In this work, a simple and cost-effective FBG writing method is proposed to realize pure apodization for both the phase mask and holographic sequential UV writing schemes [2,4]. The UV dose exposed on the fiber to form every grating section is divided into two sequentially writing shots instead of one. In this way one gain the freedom to adjust the ac-index independently while keeping the dc index profile fixed. By precisely connecting grating sections with partial overlap, the desire grating profile can be matched while the dc index become constant through the whole grating length.

2 Theory and experiment
For long FBG fabrication by the sequential UV writing scheme, Gaussian-shaped writing beam is equally-spaced and partially-overlapped to sequentially imprint grating sections. The phases of the overlapped grating sections must be controlled precisely to form a long-length grating without phase errors. The average refractive index \( n_{dc} \) has to be constant along the fiber while the ac index modulation \( n_{ac} \) can be locally changed to form specific apodized profiles. In the proposed method we set the exposure location to be \( x_i \), \( i = 1, 2, \ldots, N \), and let the UV dose to be \( 2I_0 \) at every \( x_i \). The total UV dose, \( 2I_0 \), is divided into two shots, shot 1 and shot 2, with the beam amplitudes \( I_1 \) and \( I_2 \), respectively, and the phases of the two UV fringes are \( \theta_1 \) and \( \theta_2 \), respectively, as shown in Fig. 1(a) and Fig. 1(b). The index modulation amplitude and phase of every grating section are determined by the superposition of the two shot profiles, as shown in Eq. 1.

\[
I(x) = I_1 e^{i(kx+\theta_1)} + I_2 e^{i(kx+\theta_2)} = n_{ac}(x) + n_{dc}(x) \tag{1}
\]

\[
I(x) = 2I_0 e^{i\Delta \theta} \tag{2}
\]

\[
I(x) = 2I_0 e^{i\Delta \theta} (m-1) \tag{3}
\]

Fig. 1. (a) Illustration of UV sequential writing. (b) Illustration of intensities and phases of the two shots.
To achieve pure apodization, two exposure configurations can be used. In configuration 1, the total UV dose is $2I_0$, and the two shots have equal intensity $I_0$. The ac index modulation is adjusted by symmetrically changing the fringe phase of the two shots to be $\Delta \phi$ and $-\Delta \phi$, resulting in zero phase-shift at $x_i$ due to amplitude adjustment. Equation 2 in the previous page shows the superposed UV fringe distribution, which amplitude can be determined by the factor $\cos(\Delta \phi)$. Table 1 shows some example conditions of the two shots in this configuration.

In configuration 2, the index modulation is achieved by setting $\theta_1$ and $\theta_2$ equal to 0 and $\pi$, respectively, and let the intensity $I_1 = mI_0$. The total UV dose is $2I_0$. Equation 3 in the previous page shows the superposed UV fringe amplitude distribution. The net phase shift at $x_i$ is zero, and the final fringe amplitude is determined by the factor $(m-1)$. Table 1 shows some example conditions of the two shots in this configuration.

<table>
<thead>
<tr>
<th>Configuration</th>
<th>$\theta_1 = -\theta_2 = \Delta \phi$</th>
<th>$\theta_1 = 0, \theta_2 = \pi$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_1 = I_2 = I_0$</td>
<td>$I_1 = mI_0, I_1 + I_2 = 2I_0$</td>
<td>$I_1 = mI_0, I_1 + I_2 = 2I_0$</td>
</tr>
</tbody>
</table>

Table 1 The conditions of the two configurations.

The proposed methods have been experimentally tested by exposing photosensitive optical fibers to 244-nm UV radiation, which has a Gaussian fringe profile with its beam size $1/e^2$ about 6.5 mm. The UV dose for achieving maximum index modulation is divided into two shots, and sequentially exposes onto the fiber at the same location. To ensure linear index response, the fibers are performed with UV treatment in advance [4]. The grating index modulation envelopes are measured by side-diffraction method [4] and the transmission dips are measured by the optical spectrum analyzer. Figure 2(a) and 2(b) shows some measured grating index profiles. The index modulation profiles of the gratings are Gaussian-shaped, similar to the shape of UV writing profile. By using the measured Gaussian index envelopes in the transform matrix calculation to fit the transmission spectra, the curve between the normalized index modulation and phase shift $\Delta \phi$ is shown in Fig. 2(c), and the curve between the normalized index modulation and $I_1/I_0$ is shown in Fig. 2(d). The simulation curves are derived from Eq. 2 and Eq. 3. The experimental data in dots fit well with the theoretical predictions. These correlation results for index modulation can then be used in actually fabricating complex FBGs, which is in progress right now and further results will be reported during the conference.

In conclusion, a simple method for attaining pure apodization is proposed for FBG sequential UV writing. The UV dose at every exposed location is divided into two, and the ac index modulation is adjusted by controlling the superposed phase and amplitude of the two imprinted UV fringes. The average refractive index is kept constant due to the constant net UV flux per grating section, and the ac index modulation can be locally and independently changed in both configurations.

3 References
The refractive index sensing of different types of long-period gratings

Jaw-Luen Tang¹, Jian-Neng Wang², and Ping-Chang Jui³

¹Department of Physics, National Chung Cheng University, 168 University road, Chia-Yi 621, Taiwan
TEL: +886-5-272-0586, FAX +886-5-272-0587, Email: phyjlt@.ccu.edu.tw
²Department of Construction Engineering, National Yunlin University of Science and Technology, Yun-Lin 640, Taiwan
TEL: +886-5-534-2601-2723, FAX +886-5-531-2049, Email: wangjn@yuntech.edu.tw
³IR Section, Materials and Electro-Optics Research Division, Chung Shan Institute of Science and Technology, Lung-Tan, Taiwan

Abstract:
We report a comparative study of sensing characteristics (temperature, strain, refractive index, and chemical solution) for three different configurations of long-period gratings (LPGs), which include normal LPGs, phase-shifted LPGs, and LPGs inscribed in tapered fibers.

Introduction
Long-period fiber gratings (LPGs) are widely used for many applications in telecommunication and sensor systems such as band-rejection filters [1], and temperature/strain/refractive index (RI) sensors [2, 3]. The sensing applications are of particular interest partly due to their relatively simple of fabrication, low back-reflection, and particularly high temperature, bend, and load sensitivity. An LPG is a periodic axial RI variation in the core of a singlemode optical fiber that couples light from the fundamental guided mode to some co-propagating cladding modes, resulting in several attenuation bands at discrete wavelengths that are highly sensitive to RI of surroundings. Recently, several highly sensitive LPG-based RI sensors have been demonstrated for chemical and biological applications [4-6]. In this work, we investigate on analyses and sensitivity to surrounding medium’s RI (SRI) of three types of LPG-based devices for RI and chemical sensing, which include a normal LPG, a phase-shifted LPG (PS-LPG), and a LPG written in biconical tapered fiber (BT-LPG).

Experimental
The LPGs were fabricated on singlemode standard communication fibers using a pulsed 193nm ArF excimer laser (laser energy of ~135 mJ/cm² and total exposure time of ~3 minutes) with an amplitude mask technique, as shown in Fig. 1. The fiber was hydrogen loaded at a pressure of 120 bar over a period of two weeks at room temperature. The length of three types of LPG-based sensors was about 2.7 cm and the grating period was set to 380 µm. The phase we used in PS-LPG is π. The technique to fabricate a BT-LPG was similar to that used in Ref. 7. The total length of tapered fiber is 1 cm with an approximate minimum fiber radius of 34 µm. After the laser exposure, all the LPGs were annealed at 150 °C for 24 h to stabilize their optical properties. With suitable fabrication parameters such as laser power, exposure time, and grating period, the resulting resonance wavelengths ranging from 1200 nm to 1600 nm with a greater than 20 dB peak depth were obtained. In this work, a number of these three types of LPG-based sensors associated with various attenuation bands (corresponding to various cladding modes) were measured and analyzed.

Results and Discussion
The characteristics of LPGs such as temperature and strain were first investigated. The thermal responses of the LPG were measured by heating the grating from 30°C to 150°C in incremental steps of 15°C, using a temperature-controllable chamber. When temperature increased the resonance wavelength of LPGs is shifted to the longer or the shorter wavelength, showing that temperature sensitivity (dλ/dT) of LPGs exhibited either positive or negative sensitivity. The found temperature sensitivities for the investigated LPG, PS-LPG, and BT-LPG was 0.06-0.09 nm/°C, 0.04-0.09 nm/°C, and 0.07-0.08 nm/°C, respectively, which are several times larger than those of FBG (~0.01 nm/°C). It can be seen that all three LPG-based sensors possess almost similar responses in magnitude to temperature changes. The spectral responses to the strain of the LPG were performed by mounting the grating on a translation stage that was moved outward to induce a strain in the optical fiber. For all LPGs used in this work the strain sensitivities of various resonance wavelengths were found to be very small (~0.09 ± 0.01 pm/µε) and exhibited nearly insensitive to strain variations.

In order to characterize the LPG as a RI sensor or a chemical concentration sensor, measurements with sucrose and sodium chloride aqueous solutions were performed. The surrounding RI was controlled through the use of sucrose solutions with various concentrations [6]. The ability of three LPG-based sensors to detect changes in the SRI or chemical solution concentrations was then studied. For precise RI measurement, experimental setup and sample solution were maintained at the same temperature (within ~0.1°C fluctuation). Therefore, the results reported here were not influenced by temperature and strain effects. The experiment used to measure refractive index sensitivity was followed by Ref. 7. When the concentration and, hence, the refractive index of a sucrose solution
increased in the range of 1.34-1.41, the transmission spectrum of the LPG sensor exhibited either a linear decrease (LPG and PS-LPG) or a linear increase (BT-LPG) in the insertion loss and a shift in the peak wavelength. Fig. 2 shows wavelength shifts of a typical PS-LPG against sucrose concentration and, hence, the corresponding changes in SRI. A linear regression approach [6] was employed to calculate the RI sensitivity of each type of LPG sensor. Our results show that RI sensitivities of LPG, PS-LPG, and BT-LPF are $\frac{d\lambda}{dn} = -45.3 \text{nm/RI}, -(74.4-87.5) \text{nm/RI}, +271.7 \text{nm/RI}$, respectively, which leads to a limit of detection in index, $1.3 \times 10^{-5}$, $(6.9-8.1) \times 10^{-4}$, $-2 \times 10^{-4}$, respectively. The investigation shows that BT-LPG exhibited the highest sensitivity to SRI among three LPG types of sensors.

The sensitivity of LPG to concentration change in chemical solution was further investigated using sodium chloride (NaCl) solutions. The LPG under test was immersed in a container of deionized water (volume 100 cc) and increments of 5 cc of 4 molar NaCl solution were added. The temperature kept constant to 0.1°C. Fig. 3 shows wavelength shifts of a typical PS-LPG against molar concentration of NaCl and the corresponding changes in SRI, showing that sensitivities of LPG, PS-LPG, and BT-LPF in terms of NaCl molar concentration are $-0.3 \text{nm/Mol}$, $-0.33 - 0.68 \text{nm/Mol}$, 0.57nm/Mol, respectively, leading to a limit of detection in molar concentration, 0.2, 0.08-0.18, 0.1, respectively. That small sensitivity of BT-LPG to NaCl molar concentration remains unknown and the investigation is under way. However, studies presented here demonstrate the feasibility of using the design of LPG-based sensor for RI or chemical sensing with high sensitivity.

Conclusion
We report a comparative study of sensing capabilities (temperature, strain, RI, and chemical solution) based on three different LPG configurations (LPG, PS-LPG, and BT-LPG) using a pulsed UV laser with an amplitude mask technique. Our results demonstrate that LPG fiber sensor can provide a resolution of $10^{-4}$ to $10^{-3}$ for refractive indices in the range of 1.34 to 1.41, suggesting that these devices may be suitable for use with aqueous solutions for chemical sensing.

References
Cross Phase Modulation by Pump Pulses Emitted from High-Power YAG Laser on Fiber Grating Couplers

Fatemeh Abrishamian*, Keiji Narumi, Shinya Sato, Masaaki Imai†
Muroran Institute of Technology, 27-1 Mizumoto-cho, Muroran, Hokkaido, 050-8585 Japan,
Tel†: +81-143-46-5523, Fax: +81-143-46-5501,
Email*: f_abrishamian@ieee.org,
Email†: mimai@mmm.muroran-it.ac.jp

Abstract

The optical-switch due to cross-phase modulation (XPM) that intense high power YAG pump laser induces on the signal light propagating through the Bragg grating coupler is experimentally and theoretically investigated.

1 Introduction

The capacity of optical transmission systems has been recently increased significantly by using high-speed wavelength division multiplexing (WDM). Fiber grating couplers (FGC) is a novel key component in optical add/drop multiplexers. The FGC can be fabricated by writing the Bragg grating in the tapered waist of fiber coupler. Recently, all-optical switch of fiber Bragg gratings and FGCs has been proposed and demonstrated experimentally [1-3]. In this paper, we develop the nonlinear optical fiber switching using the so-called Kerr nonlinearity in fiber grating coupler [4]. The switching is based on the cross-phase modulation (XPM) when the intense high power pump copropagates in the grating region simultaneously with the signal light [5]. The fact will lead to the change in reflection spectra of Bragg grating and the resultant output intensity at the signal wavelength will be decreased when the power of high pump pulses increases.

2 Experimental set of all-optical switching

The experimental setup for all-optical switching is shown in Fig. 1. A narrow linewidth tunable laser, nearly centered in the reflection spectra of the FGC, is used as a low probe (signal) light. This CW probe is coupled at the fused portion of the FGC to high pump pulses of a Nd:YAG laser. The output frequency of YAG laser is used as a reference signal of a lock-in amp. This frequency is equal to the laser output pump pulse repetition. The pump pulse source emits its spectra around 1064nm. The repetition rate of the pump pulse is 1kHz and the pulse width is approximately 70ns. Minimum YAG laser peak power is around 4kW at input end of WDM coupler. The grating used in the FGC is 10mm long and the index profile is Gaussian apodized. The insertion loss of the grating coupler fabricated in our laboratory is 0.89dB. The splitting ratio of the coupler is 1: 32.4 as a through-port power versus a coupled-port power. In our experiment, a drop efficiency of 58% at maximum is obtained at the Bragg wavelength $\lambda_B$ =1538.26nm. The reflected signal light from FGC taken out of an optical circulator is detected with photo-diode (PD). The pump pulse, propagating in the grating of the FGC, induces a nonlinear variation of the refractive index that shifts the whole reflection spectra toward longer wavelength that is red-shifted.
The operation of all-optical switching has been confirmed by means of coherent detection using a lock-in amp. [6].

3 Results
By detecting the change of output from the lock-in amp. in the presence and absence of pump light, the output is plotted as a function of signal wavelength (Fig. 2).

![Fig.2 Experimental results of lock-in amplifier output vs. wavelength for a variety of pump peak powers using YAG pump pulse.](image)

![Fig.3 Experimental results of lock-in amplifier output vs. wavelength for a variety of pump peak powers using EDFA pump pulse.](image)

The lock-in amp. output corresponding to the dropped signal power from port 2 decreases as the pump peak power increases. This is due to the fact that as the pump peak power increases, reflection spectra shift to a longer wavelength according to the formula of red-shifted Bragg wavelength \( \lambda'_{B} = 2n\Delta \) where \( \Lambda \) is the grating period and \( n \) is modulated refractive index. Modulated refractive index is a function of pump peak power, polarization coefficient and effective core area. By comparing the experimental curve (Fig. 2) with theoretical ones for outputs from a lock-in amplifier, we can estimate the red-shift of Bragg wavelength of FGC. This wavelength shift is comparable to the results of Fig. 3, where we used an EDFA pulse laser as the pump source [6]. The pump pulse from an EDFA emits its spectra around 1534.2nm-1561.1nm with a peak power of 3.55kW at maximum. The repetition rate of the pump pulse is 20.6MHz and the pulse width is approximately less than 1ps. Using EDFA pump cause the mixture of pump pulse components into signal light problem. To overcome this problem in this experiment we perform an efficient switching using high power YAG laser because its duration of pump pulse is longer than the grating within a FGC. We estimate the red-shift of Bragg wavelength 0.06nm at maximum pump power of 2.1kW.

4 Conclusions
Using the fabricated FGC in our laboratory, we experimentally demonstrated a relatively low-power, all-optical switching induced by intense pump light at 1064nm in a fiber optic-grating coupler. It was confirmed that the dropped signal power decreases as the pump peak power increases at the wavelength \( \lambda<\lambda_{B} \) and the red-shift of the Bragg wavelength is 0.04-0.06nm/1.8-2.1kW.

References
Effective Automatic Gain Spectrum Adjustment of Bi-directionally Pumped Broadband Raman Amplifiers

Zhi Tong, Jilai Zhang, Dayin Sun and Shuisheng Jian
Institute of Lightwave Technology, Beijing Jiaotong University, 100044, Beijing, P.R.China
email: zhtong@center.njtu.edu.cn

Abstract: An effective feedback algorithm is proposed to dynamically control the pump powers of bi-directionally pumped broadband (>70nm) Raman amplifiers. By introducing a simple saturation factor, wide-range and robust gain spectrum adjustment can be achieved.

1. Introduction
Multi-wavelength pumped Raman amplifiers (RAs) have been widely used in ultra-long haul transmission systems. To meet the practical requirements, the gain spectra of RAs should be controlled in real-time according to various system requirements. Apparently, dynamic gain spectrum control (DGSC) of RAs can be derived cost-effectively by direct adjustment of Raman pump powers. To date, several methods have been reported to realize DGSC of backward pumped RAs, which include pump grouping method [1], linear-matrix method [2], and iterative pseudo-inverse matrix method [3]. However, no control methods for RAs with bi-directional pumping are proposed. In this article, we demonstrate an effective feedback algorithm to achieve automatic DGSC of bi-directionally pumped broadband RAs for the first time. By introducing a simple saturation factor, wide range gain spectrum control can be derived even for a 100nm high-gain RA within 5 iterations, which will take a millisecond or so by using built-in high-speed microprocessors.

2. Theory model and control algorithm
Firstly we consider an optimized steady-state RA with N pump wavelengths and M signal channels. According to Equ.2 in ref. [2], when neglecting the high-order term, the approximate pump power vector (N×1) at Δz distance can be deduced (detailed prove will be presented in conference) as

\[ \Delta P_p(\Delta z) = A(0) \cdot \Delta P_p(0) + sB(0) \cdot \Delta P_p(0), \]  

where constant s is the saturation factor, and pump powers are launched at 0. A and B are both N×N matrices and assumed unchanged within Δz length, which is correct when Δz is small enough. We have

\[ A(0) = \begin{bmatrix} G_{p_1}(0) & 0 & \cdots & 0 \\ 0 & \ddots & \vdots & \vdots \\ \vdots & \ddots & 0 & 0 \\ 0 & 0 & \cdots & G_{p_N}(0) \end{bmatrix}, \]  

\[ G_{p_i}(0) = \exp[\sum g_{pp_i} \bar{P}_p(0,\Delta z) - \alpha_{p_i} \cdot \Delta z], \]  

where \( \bar{P}_p(0,\Delta z) \) is the average pump power form 0 to Δz, subscripts

\[ P_1, \ldots, P_N \]  

denote different components of vector \( P_0 \), and \( g_{pp_i}, \alpha_{p_i} \) are pump-pump Raman coefficient and fiber loss, respectively.

While

\[ B(0) = \begin{bmatrix} 0 & B_{12} & \cdots & B_{1N} \\ B_{21} & 0 & \cdots & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ B_{N,1} & \cdots & \cdots & 0 \end{bmatrix}, \]  

\[ B_{M,N} = \frac{g_{pwp_w} \bar{P}_w(0,\Delta z)}{C_{p_p}(0)} \cdot (G_{p_w}(0) - 1) \]  

and

\[ C_{p_p}(0) = \sum g_{p_p} \bar{P}_p(0,\Delta z) - \alpha_{p_p}. \]  

Then we can derive approximate pump integral vector along fiber span as

\[ \Delta I_p(z) = [\Delta z + [A(0) + sB(0)] \Delta z + \cdots] \cdot \Delta P_p(0) = \mathbf{H} \cdot \Delta P_{pin}, \]  

where \( \Delta P_{pin} \) denotes the change of input pump powers, and \( \mathbf{H} \) is the N×N control matrix. Next, applying Equ.(2) to a bi-directionally pumped RA with Nf forward and Nb backward pump wavelengths we have

\[ \Delta P_{pf, in} = H_f^{-1} \cdot g_{spf}^{-1} \cdot \Delta G_s(z) \]  

and

\[ \Delta P_{pb, in} = H_b^{-1} \cdot g_{spb}^{-1} \cdot \Delta G_s(z), \]  

where subscript f and b denote forward and backward pumping schemes, \( \Delta G_s \) is an M×1 vectors denoting the variations of signal gain profile (in decibel), and \( g_{spf}^{-1} \) is the pseudo-inverse matrix of \( g_{spf} \), which is a M×N matrix representing the pump-signal Raman coefficient. Here we assume the saturation factor is identical for both forward and backward pumps, and the Raman interactions between the forward and backward pumps are neglected. As a result we can adjust forward and backward pump powers independently. By carrying out feedback pump adjustment based on Equ.3 iteratively, wide dynamic range can be achieved with an appropriate saturation factor.

3. Simulations and results
When applied to practical systems, \( g_{sp} \) can be
measured experimentally, and $\Delta G_s$ can be determined through real-time detection of the gain spectra or gains of several probe channels. Moreover, the original pump power distributions $P_p(z)$ of a steady-state RA can be measured directly by using modified OTDR method [4]. It should be highlighted that saturation factor $s$ is the key parameter of our algorithm, which can be achieved by either experiments or simulations. The principle is to find the optimal $s$ value which can lead to the minimum difference between the target and the actually output gain profile after one iteration.

To achieve general conclusions, we consider the non-degenerate pump wavelength setup, which uses unequal forward and backward pump wavelengths, and 100km NZ-DSF is used as gain medium, while the conclusions are also valid for other fiber types. Genetic algorithm is adopted to optimize the pump powers as well as wavelengths of the steady-state RAs. Through intensive computations, we show the typical $s$ values at different conditions in Tab.1. It can be found the broader the gain band is, the less the $s$ value is. And $s$ also decreases slightly as the total signal power and the fiber length increase. As shown in Fig.1, larger than 24dB dynamic range (-16~8dB net gain, which equals about 4~28dB on-off gain) can be achieved within 4 iterations by using $s = 0.4$ when Raman gain bandwidth is less than 85nm. However, when the gain band is up to 100nm, the converge speed of the algorithm will be slower if a lower gain level is required, which is mainly due to decreased pump-pump interactions. This problem can be solved by adding an accelerating factor to the longest wavelength pump. In Fig.2a, 22dB dynamic range can be obtained for a 100nm RA with 3 forward pumps (1401,1418 and 1428nm) and 5 backward pumps (1412,1458,1468,1478 and 1517nm) by using $s = 0.08$, and it can be found the relationship between output gain level and input pump powers is nonlinear as shown in Fig.2b. Moreover, tilted rather than flat gain profiles can also be obtained by using this method as shown in Fig.2c, which tells the robustness of the proposed algorithm.

At last, it will take less than 0.3ms to complete one iteration for detecting stable gain profile. Consequently the total adjustment time will be 1ms or so for most conditions, which is enough for practical applications.

4. References


Table.1 Typical s values at different conditions.

<table>
<thead>
<tr>
<th>Bandwidth</th>
<th>60nm band, 2 forward and 2 backward pumps</th>
<th>80nm band, 2 forward and 4 backward pumps</th>
<th>90nm band, 2 forward and 5 backward pumps</th>
<th>100nm band, 3 forward and 5 backward pumps</th>
</tr>
</thead>
<tbody>
<tr>
<td>S value</td>
<td>0.6</td>
<td>0.4</td>
<td>0.2</td>
<td>&lt;0.1</td>
</tr>
</tbody>
</table>

Fig.1 (a) -16~+8dB gain spectrum adjustment of a 100 channel, 80nm RA with 2 forward and 4 backward pumps, (b) the adjusting process from -16dB to 8dB. Total input signal power is 20mW.

Fig.2 (a) -15~+7dB gain adjustment of a 100nm RA with 3 forward and 5 backward pumps, (b) various gain levels vs. pump powers, and (c) achieved gain profiles with negative and positive slope. Total input signal power is 50mW.
Configuration Comparison of C/L-band EDFA for DWDM Systems

Tsair-Chun Liang, Yi-Lun Liu, and Hsi-Shan Huang

Graduate Institute of Electro-Optical Engineering, National Kaohsiung First University of Science and Technology, Kaohsiung, Taiwan 811, R.O.C.
Tel: 886-7-6011000 ext. 2712, Fax: 886-7-6011096
E-mail: tcliang@ccms.nkfust.edu.tw

Abstract

We theoretically investigate various schemes of C/L-band erbium-doped fiber amplifier (EDFA) for offering the maximum gain and moderate low noise figure in DWDM systems. A total of six EDFA with single forward pumping schemes is examined and compared. The investigation provides the best EDFA configuration selection to design both power and in-line amplifiers.

I. Introduction

EDFAs have emerged as vital components for optical fiber networks, serving a wide range of applications from WDM network repeaters for power amplifiers and in-line amplifiers. The WDM optical fiber amplifier with a flatten amplification region from 1530 to 1560 nm (conventional band, C-band), and the long wavelength region from 1570 to 1610 nm (L-band), has been widely investigated [1]. In order to improve gain characteristics, the EDFAs through a double-pass technique [2], or by partial amplified spontaneous emission (ASE) was feedback into the EDF causing a gain increment in the L-band [3], or by recycling unused ASE into under pumped EDF regions have been suggested and experimentally verified.

Here we theoretically investigate and compare various EDFA configurations without using external gain equalization technique to amplify 32 digital WDM channel signals. These configurations considered include one-stage and two-stage single forward pumping schemes for achieving high-power low-noise-figure characteristics.

II. EDFA Configurations

Fig. 1 shows the traditional configuration EDFA in one-stage single-pass (1S). Fig. 2 depicts the block diagram of various EDFA configurations. There are two optical circulator blocks (OC1 and OC2), and five erbium-doped gain fiber blocks (L1, L2, L3, L4, and L5) in Fig. 2 used to construct various amplifier configurations. For each configuration, the 980/1550-nm or 1480/1580-nm wavelength selective coupler (WSC) is employed to act as the pump/signal optical combiner. In these schemes, the circulator is used to route the amplified signal power output and the fiber mirror was used to reflect pump and signal lights back to the EDF3 with a reflectance ratio is assumed to be 100%. Table 1 list five kinds of DWDM EDFA configurations, their compositions using the blocks, and their configuration symbols.

<table>
<thead>
<tr>
<th>Pumping Schemes</th>
<th>Configuration Symbols</th>
<th>Configuration Compositions</th>
</tr>
</thead>
<tbody>
<tr>
<td>One-stage double-pass</td>
<td>1D</td>
<td>(P_in, WSC, Pump LD, OC2, L3, Mirror, P_out2)</td>
</tr>
<tr>
<td>Type A</td>
<td>(P_in, WSC, Pump LD, L2, OC2, L3, Mirror, P_out2)</td>
<td></td>
</tr>
<tr>
<td>Type B</td>
<td>(P_in, WSC, Pump LD, OC2, L3, Mirror, L5, P_out2)</td>
<td></td>
</tr>
<tr>
<td>Type C</td>
<td>(P_in, OCT, L1, WSC, Pump LD, L2, Mirror, P_out2)</td>
<td></td>
</tr>
<tr>
<td>Type D</td>
<td>(P_in, OCT, WSC, Pump LD, L2, Mirror, L4, P_out2)</td>
<td></td>
</tr>
</tbody>
</table>

Fig.1 One-stage single-pass EDFA configuration

Fig.2 Various configurations of C/L-band EDFA's

The EDFA model used in this work, which is based on the model by Giles and Desurvire. The erbium-doped fiber with the erbium ion density of $1.4e+25/\text{m}^3$ was used. Its core radius is assumed to be 1.8 $\mu$m, and numerical aperture is assumed to be 0.23. The insertion loss of WSC coupler is assumed to be the same with 0.5 dB at both 980 and 1550-nm bands and both 1480 and 1580-nm bands. The insertion loss and
isolation of optical isolator and circulator are assumed to be 0.5 dB · 50 dB, and 0.8 dB · 50 dB, respectively. The output power of pump laser diode is assumed to be 135 mW and 980-nm-pumped for C-band and 1480-nm-pumped for L-band. A 32-channel signal in C-band the 1544.53 〜1556.93 nm wavelength range with a channel spacing of 0.4 nm, and in L-band the 1574.54 〜 1600.60 nm wavelength range with a channel spacing of 0.8 nm are considered. The input power level of each digital channel for each EDFA configuration is set to be –23 dBm.

III. Characteristics Comparison

For investigating the optical gain and noise figure characteristics of each two-stage reflection EDFA scheme, the first step is to find the optimum EDF length of each amplifier configuration for simultaneously achieving the highest channel output power with differential channel gain variation $\Delta G (= G_{\text{max}} - G_{\text{min}})$ of $\leq 0.6$ dB among 32 digital channels and the noise figure of $\leq 5.5$ dB. The range of EDF length set in the simulation is 1 〜 100 m; the increment length of each EDF used in the iteration loop is 0.5 m. When the EDF was pumped by laser diode, we called the first-stage of the amplifier. The second-stage is a signal path amplifier with the ASE from the first-stage amplifier.

Fig. 3 shows the spectral (a) gain and (b) noise figure characteristics of six amplifier configurations for C-band EDFA. In Fig.3, the one-stage double-pass configuration is better than all two-stage for achieving high-power characteristic. We found that the optimum EDF length for one-stage double-pass is 5.5m. And the total EDF length for one-stage single-pass, two-stage type A, type B, type C, and type D are 10.5 m, 7.5 m, 9.5 m, 4.5m, and 6.5 m, respectively.

Fig. 4 shows gain and noise figure characteristics of the comparative L-band EDFA at the output port with the optimum EDF length of the six different type EDFAs. In Fig. 4, the two-stage type A has the lowest noise figure with output gain of about 25.2 dB. From the maximum output power with satisfied noise figure point view, Type A and Type B of two-stage schemes can get the best channel output power of about 1.66 mW per channel with a moderate noise figure of $< 5.5$ dB.

IV. Conclusions

We have theoretically investigated six configurations of C/L-band gain-flattened EDFAs, two one-stages and four two-stages, using single forward pumping for DWDM system applications. Among them, one-stage double-pass scheme and type A and type B are the best amplifier design for C- and L-band DWDM lightwave systems, respectively.

References
Polarization-stabilized single-mode vertical-cavity surface-emitting lasers

Ki Soo Chang, Young Min Song, and Yong Tak Lee
Gwangju Institute of Science and Technology, 1, Oryong-dong, Buk-gu, Gwangju , Korea, 500-712
Tel: +82-62-970-2206, Fax: +82-62-970-3128, Email: ytlee@gist.ac.kr

Abstract
We developed polarization-stabilized single-mode vertical-cavity surface-emitting lasers (VCSELs) with a side mode suppression ratio of over 35dB and orthogonal polarization suppression ratio of over 14dB for the entire drive current range by introducing a mode selective aperture (MSA) and asymmetric current injection (ACI).

1 Introduction
Vertical-cavity surface-emitting lasers (VCSELs) have become a vital optical component in short distance optical data communication applications, such as local area networks (LANs), storage area networks (SANs), and parallel optical interconnects [1]. In recent years, VCSELs extend their application area to various non-communication fields, such as bio-sensors, gas sensors, atomic clocks, laser mouse sensors, and proximity sensors [2]. May of these applications require single-mode and stable linearly polarized output beam. However, the standard oxide-confined VCSEL designs typically lead to lase in multiple transverse modes due to the large effective index step induced by the low refractive index of the Al-oxide layer. Furthermore, the light from VCSELs do not has a well-defined linear polarization state due to the isotropic material gain and symmetric device geometry.

In this presentation, we report a simple and robust technique for obtaining a single-transverse-mode and polarization-stabilized operation of an oxide-confined VCSEL by introducing a combination of a mode selective aperture (MSA) and asymmetric current injection (ACI).

2 Device structure and fabrication
For experimental confirmation of the proposed idea, we fabricated oxide-confined ACI VCSELs with an MSA as shown in Fig. 1 and Fig. 2. Asymmetric current injection (ACI) structure by restricting p- and n-contacts to the opposite side of the mesa provides an asymmetric gain for two orthogonal polarization states. For conduction and heavy hole band transition, the transition probability becomes a maximum when the angle between the momentum vector of electron and the electric field polarization of emitted photon is 90° [3]. Therefore, perpendicular polarization to the current path is preferred in ACI VCSELs. For the suppression of high-order modes, we introduced MSA in the top DBR mirror. The MSA, formed naturally in the top DBR during the selective oxidation step for the current aperture (CA) formation, provides a large optical loss in the outer region of the top mirror. If the MSA diameter is larger than the fundamental mode width and smaller than the high-order mode width, only the lasing threshold for the high-order mode becomes high. Thus, a VCSEL with an MSA can maintain a stable fundamental mode operation.

We measured the near-field profile to investigate the spatial intensity distribution of the transverse modes in the VCSEL. The full width at half maximum (FWHM) of the fundamental mode and lowest high-order mode are 4.3 and 8.4 μm, respectively, for VCSEL with CA of 5μm. In order to experimentally confirm the validity of the proposed concept, VCSELs with MSA of 7μm and CA of 5μm in diameter were fabricated and characterized.

The epitaxial layer is a typical structure for 980nm intracavity-contacted oxide-confined VCSELs. However, the epitaxial layer structure is not optimized yet for high-power single-mode operation because 45nm-thick oxidation layers for current aperture are located in the middle of the node and antinode of standing wave pattern [4]. The CA and MSA were formed simultaneously in the
wet thermal oxidation furnace at a temperature of 400 °C. Standard intracavity-contacted VCSEL processing was used to fabricate ACI VCSELS with MSA without additional process steps due to the simultaneous formation of MSA and CA via single step oxidation by properly controlling the mesa diameters.

3 Results

Fig. 3 shows CW lasing spectra and light output characteristics of fabricated ACI VCSEL with MSA at room temperature. Completely stable single-mode operation was achieved with side mode suppression ratio (SMSR) of over 35dB for entire drive current range. Relatively high threshold current and low slope efficiency can be attributed to the increase of optical loss not only for high-order transverse mode but also for fundamental mode. Single-mode operation for entire current range and small slope efficiency indicate that the maximum single-mode output power can be increased by optimizing the MSA diameter relative to the CA diameter. Fig. 4 shows polarization resolved output power and orthogonal polarization suppression ratio (OPSR) versus current. The ACI VCSEL with an MSA clearly exhibits a dominant polarization perpendicular to the current path as expected. The maximum OPSR is 17dB at 7mA. The difference between the total output power and the sum of the powers in both polarizations is due to the loss by collimation optics and polarizer.

The most obvious advantage of fabricating polarization-stabilized single-mode VCSELS using this method is that standard VCSEL processes can be used without any additional process steps.

This work was supported by MOST through the TND project.

4 References

In this report, 1.3 µm oxide confined InGaAsN/GaAs VCSELs grown by MOVPE were fabricated. The lasing spectra with blue-shift oxide mode were observed. The related DC characteristics of the fabricated VCSELs are also reported.

1. Introduction:
This report will demonstrate the oxide mode of InGaAsN vertical cavity surface emitting laser (VCSEL) in 1.3 µm range. The wavelength blue shift amount could be up to 40 nm, which is larger than previous reported results observed in highly strained InGaAs quantum well VCSELs [1]. This blue shift oxide mode can be attributed to the large detuning between the gain peak and cavity resonance. And the effective optical thickness shrinkage of the oxide layer will also result in a new cavity resonance, which is so called oxide mode [2].

2. Experiment
Samples used in this study were all prepared by metalorganic vapor phase epitaxy (MOVPE), and grown on 2-inch n⁺-GaAs substrates. The VCSEL structure consists of a InGaAsN/GaAs double quantum well (DQW) active region sandwiched by a Si-doped n-distributed Bragg reflector (DBR) mirror and a C-doped p-DBR mirror. The DQW active region consists of two 7 nm-thick InGaAsN well layers and 10.5 nm-thick GaAs barrier layers. The indium composition in the well layers was around 0.35 and N composition was 0.01. Both the n- and p-DBR were composed of interlaced 1/4 λ-thick GaAs and Al₀.₉Ga₀.₁As layers with a 20 nm-thick interface grading to reduce the series resistance. There were 40.5 and 26 AlGaAs/GaAs pairs used as the n- and p-DBR mirrors, respectively. A 20 nm-thick AlAs layer was inserted at the interface of the first p-DBR pair and the active region to define the oxide aperture after selective wet oxidation. The detail processing procedure and parameters were well described in previous work [3].

3. Results and Discussion
Figure 1 shows the continuous-wave (CW) light output-current-voltage (L-I-V) characteristics of the fabricated VSCELs under room temperature (RT).

Fig.1 LIV characteristics of the fabricated devices. The inset in the lower right corner shows the top view microphotograph of the device, and the inset in the upper left corner is the cross-section SEM photograph of the oxide-confined VCSELs, the tapered oxide layer can be easily observed.
It is worth noting that the threshold currents of these devices are about 1mA instead of 5mA. The devices had already lased in oxide-mode wavelength when driving current exceed 1mA, though the output power seemed small, and until driving current higher than 5mA the main mode began to lase and output power increased rapidly. This phenomenon had been manifested in our previous works [1].

Fig.2 Simulated reflectance spectra before (blue curve) and after (red curve) oxidation. The cavity resonance wavelength of two different mode correspond to the observed lasing spectrum.

From the SEM picture, it was found that the thickness of the tapered oxide layer front was around 107 nm. According to the SEM measurement result, we simulated the reflectance spectrum of the VCSELs sample with a 107 nm-thick AlOx layer as a replacement of the unoxidized semiconductor layer. The simulated reflectance spectra before and after oxidation were represented in Fig.2, and the lower part also shown the corresponded lasing spectrum, whose main peak wavelength was 1.285 µm and oxide-mode was 1.247 µm. Such an observation can be attributed to the large detuning between the gain peak and cavity resonance. And the effective optical thickness shrinkage of the oxide layer will result in a new cavity resonance, which is so called oxide mode [2]. The blue shift oxide modes were resulted from these two factors combination. After oxidation, the reflective index of AlAs was changed from 2.91 to 1.55 when it was converted into AlOx. Thus, the effective optical thickness will also decrease. When the driving current was small, carriers tended to distribute along the perimeter of the oxide aperture. Since the refractive index difference between the oxide layer and unoxidized GaAs is larger than that of the Al0.89Ga0.11As/GaAs DBR pair adjacent to the cavity, the reflectivity of the area beneath the oxide layer is thus larger than that inside the aperture. Hence, photons underneath the oxide layer require less threshold gain to reach stimulated emission. As a result, the emission wavelength blue shifted from the designed cavity resonance [2]. Besides that, the gain spectrum was closer to the shorter wavelength initially. As a result, the emission wavelength blue shifted from the designed cavity resonance under lower injection current. Once the drive current increased, both of the cavity resonance and gain spectrum red shifted and the designed cavity resonance began to lase. This phenomenon also influenced the light output power versus current (L-I) characteristics obviously.

4. Conclusion

In this report we demonstrated the fabrication and optical property of InGaAsN oxide confined VCSELs emitting in 1.3 µm region, and describe the oxide mode phenomenon. The simulated reflectance spectra shown well agreement with the measured lasing spectra.

Reference


Etching Characteristics of Low-k Polymer for Wide Modulation Bandwidth VCSELs

Akihiro Matsutani, Naoki Jogan, Fumio Koyama and Kohroh Kobayashi
Tokyo Institute of Technology, Precision and Intelligence Laboratory, R2-21 4259 Nagatsuta, Midori-ku, Yokohama 226-8503, Japan
Email: matsutani.a.aa@m.titech.ac.jp

Abstract

We investigated etching characteristics of low-k polymer for wide modulation bandwidth VCSELs. We used an O2-plasma etching process for patterning a low-k polymer passivation film. The parasitic capacitance of VCSELs could be reduced by using the low-k polymer.

1. Introduction

Vertical-cavity surface-emitting lasers (VCSELs) are expected to be a key device for high-speed local area networks, which enables low power consumption and small packaging. A small signal modulation bandwidth of 21.5GHz [1] and direct pulse modulation at 20Gb/s [2] and 25Gb/s [3] were reported so far. To reduce the parasitic capacitance is an important issue for improving the modulation bandwidth. Polyimide has been often used as a passivation film. However, the dielectric constant k of the polyimide is as large as 3.2-3.5 at 1 MHz. Therefore, it is important to reduce the dielectric constant of the passivation film for decreasing the parasitic bonding pad capacitance.

In this paper, we report O2-plasma etching characteristics of the low-k polymer for wide modulation bandwidth VCSELs.

2. Experiment and Result

We used the AL-polymer (ASAHI GLASS Co., Ltd) as low-k materials for a passivation film. The dielectric constant k of the AL-polymer is 2.4-2.5, which is 30% lower than that of the polyimide. The AL-polymer is not photosensitive. Therefore, a patterning process using resist mask and dry etching is needed for fabrication of devices. We prepared approximately 2 μm thick AL-polymer as samples. These samples were cured in N2 atmosphere after spin coating. The AZ photoresist was used as a mask material. We used a RIE etching system (converted Samco RIE-1N) with a RF frequency of 13.56 MHz. A sample table is cooled by water. We used pure O2 as an etching gas.

Figure 1 shows the etching rate of the AL-polymer as a function of RF power. The AL-polymer was etched at a gas pressure of 2 Pa and an O2 gas flow rate of 50 sccm. The etching rate of the AL-polymer increases with increasing of the RF power. Etching rate ratio between the AL-polymer and the AZ photoresist was maximized at 100W.

Figure 2 shows the etching rate of the AL-polymer as a function of the process pressure. The AL-polymer was etched at an RF power of 100W and an O2 gas flow rate of 50 sccm. The etching rate of the AL-polymer increases with increasing of the process pressure. The etching rate ratio reached at a maximum value of 1.2 at 2Pa.

Figure 3 shows the etching rate of the AL-polymer as a function of the O2 gas flow rate. The AL-polymer was etched at 100 W and 2 Pa.

We fabricated VCSELs for wide modulation bandwidth using the AL-polymer as shown in Fig. 4. The parasitic capacitance of the p-electrode could be reduced to 0.1pF [4]. The capacitance using the AL-polymer was as small as a half of that using polyimide for the same device structure. The reduction of the parasitic capacitance is primarily due to the smaller dielectric constant of the AL-polymer. In addition, a
3dB-modulation bandwidth of 17GHz was obtained in the VCSEL with a small mesa structure [4].

3. Summary
In summary, we investigated etching characteristics of low-k polymer for wide modulation bandwidth VCSELs. We reported O₂ plasma etching characteristics of the AL-polymer as a low-k material. The maximum etching rate ratio between the AL-polymer and the AZ photoresist was approximately 1.2. The AL-polymer is useful for insulating high-speed optical devices including wide modulation bandwidth VCSELs.

References
A Quantitative Postweld Shift Compensation Technique in Butterfly Laser Module Packages

Maw-Tyan Sheen¹, Yi-Cheng Hsu², and Wood-Hi Cheng³

¹Department of Electronic Engineering, Yung-Ta Institute of Technology and Commerce, Ping-Tung, Taiwan,
²Department of Biomechatronics Engineering, National Pingtung University of Science and Technology, Pingtung, Taiwan,
³Institute of Electro-Optical Engineering, National Sun Yat-sen University, Kaohsiung, Taiwan
Tel. (886)75252000 ext. 4453, Fax. (886)75254499, E-mail: whcheng@mail.nsysu.edu.tw

Abstract: A quantitative postweld shift (PWS) measurement and compensation technique employing a high-magnification camera with image capturing system (HMCICS) in butterfly-type laser module packages is investigated. The results show that the direction and magnitude of the fiber alignment shifts induced by the PWS in laser-welded butterfly-type laser module packaging can be quantitatively determined and compensated. The measured coupling powers in laser module packages after welding and compensation were clearly confirmed the measured fiber shifts that determined by the translational and rotational parameters. Therefore, the fiber shifts due to the PWS could be realigned back closer to their original optimum position after applying welding compensation, and hence the coupling powers loss due to the PWS could be regained. In comparison with previous studies of the PWS compensation by a qualitative estimation technique, this HMCICS technique has successfully provided the mechanical adjustment through a quantitative compensation to the effect of the PWS on the fiber alignment shifts in butterfly-type laser module packages. The reliable butterfly-type laser modules with high yield used in lightwave transmission systems can be developed and fabricated.

Introduction: There are many types of laser module package designs. The coaxial and boxed-type are the most common. The TO (transistor outline)-Can laser packages¹-² based on coaxial-type design are often selected when the fabrication cost is important and the performance requirement in lightwave transmission systems is not so high. The butterfly laser packages³-⁴ of the boxed-type designs, by contrast, are often used in higher performance devices that require a higher power output, a higher speed, a higher reliability, a thermoelectric cooler, and multiple components. In the process of packaging a butterfly laser package, a dual-beam laser welding system was used to connect the fiber ferrule assembly to the laser diode³-⁴, whereas a three-beam laser welding system was used to connect the fiber ferrule to the TO-Can laser diode in a TO-Can laser package¹-². However, in a two-beam laser welding system to weld laser modules, due to the fiber ferrule is not solidly constrained in the vertical direction, the welded region solidification shrinkages introduced in the butterfly-type laser packages may be critical. Therefore, how to compensate this shift quantitatively is a very important subject to improve its yielding in butterfly-type laser modules.

Package Construction and Measurement: A typical butterfly-type laser module consisted of a 1.5 or 1.3 μm laser diode, a single-mode fiber, a thermoelectric cooler (TEC), and an Invar housing, as shown in Fig. 1(a). The metal plate, saddle-shaped clip, and fiber ferrule of the Invar housing materials were chosen to minimize thermal stresses and strains. In a butterfly package construction, the fiber ferrule and the submount-to-metallic plate assemblies were soldered joints, while the fiber ferrule-clip (FFC) and the clip-metallic plate (CMP) assemblies were welded joints. There were a total of ten welded spots on the laser-welded FFC and CMP joints, as shown in Fig. 1(a). In general, the laser welding of the CMP was implemented initially, followed by the laser welding of the FFC³-⁵. The laser welding of the CMP was applied sequentially at the center, front, and rear. After the laser welding of CMP, the clip was fixed to the plate.

The fiber alignment shifts due to PWS in laser-welded laser modules is a three-dimensional movements and may be described as translation and rotation displacements between the laser diode and the fiber axes. Since the center of the fiber ferrule in a butterfly-type laser module package is the fiber axis, the approach for measuring the alignment shifts due to the PWS is equivalent to probing the displacement of the center of the fiber ferrule, thus to probe the position and angle changes of such fiber ferrules. A technique employing a high-magnification camera with image capturing system (HMCICS) was used to measure and compensate for the PWS in butterfly type laser module packages. The HMCICS consisted of a 230x high-magnification video camera, a high-performance image capture card, and a computer, as shown in Fig. 1(b). In this proposed mechanical adjustment, the quantitative compensation in the opposite direction of PWS was used. By measuring the PWS parameters before welding, after welding, and after compensation, the PWS measurement and compensation were obtained. Then the measured coupling powers in laser module packages before welding, after welding, and after compensation were obtained and compared to the measured fiber alignment shifts determined by the PWS parameters.

Results: Fig. 2 shows the measured PWS before welding,
after welding and after compensation. This indicates the PWS could be realigned back closer to their original optimum position after applying compensation. Fig. 3 illustrates the coupling power before welding, after welding, and after the PWS compensation for the eight different modules. From Figs. 2 and 3, it clearly shows that the power loss and PWS can be measured and compensated by applying the HMCICS technique.

Fig.4 illustrates the dominant displacement $\Delta y$ of the PWS for different modules. Due to PWS’ effects, the results showed that the measured $\Delta y$ were different before and after the welding, however, the fiber shifts could be realigned back closer to their original optimum position after compensation. From Figs. 2, 3 and 4, the measured coupling powers in the laser module packages before welding, after welding, and after compensation clearly confirmed with the measured fiber alignment shifts determined by the dominant parameter of $\Delta y$ that the fiber shifts induced by the PWS could be realigned back closer to their original optimum position after compensation, and hence the coupling powers loss due to the PWS could be regained.

**Conclusion:** In summary, we have successfully demonstrated quantitative measurement and compensation for the effects of the PWS on the fiber alignment shifts in laser-welded butterfly-type laser module packages by HMCICS. The benefits of using the HMCICS technique to determine the PWS direction and magnitude are the quantitatively measure the fiber alignment shift and the quantitative compensation to the effect of the PWS on the fiber alignment shifts during the laser-welded butterfly-type laser module packages, when compared to the currently available qualitatively estimated techniques to compensate the PWS in laser module packages.

**References**

---

**Fig.1.** Schematic diagram of a butterfly laser package showing (a) laser module and (b) postweld shift measurement

**Fig.2.** The measured PWS before welding, after welding, and after a PWS compensation for different laser modules.

**Fig.3.** Illustrates the coupling power before welding, after welding, and after a PWS compensation for different laser modules.

**Fig.4.** Illustrates the dominant displacement $\Delta y$ in Y-axis of the PWS measurement for different laser modules
Growth and characterisation of InSb Films on GaAs

Substrate grown using molecular beam epitaxy

Lin LI1,2, Guo-jun LIU, Zhanguo LI, Mei LI and Xiao-hua Wang
1 National Key Lab of High Power Semiconductor Lasers, Changchun University of Science and Technology, Changchun, 130022, China
2 Institute of Industrial Science, University of Tokyo, 4-6-1 Komaba, Meguro-ku, Tokyo 153-8505, Japan
E-mail: lilincust@hotmail.com; Fax: 86-431-85384517; Tel: 86-431-85303282

Abstract

InSb thin films were grown using molecular beam epitaxy, characterized by using SEM, TEM and XRD measurements. These results indicate that the InSb buffer layer grown at low temperature plays an important role in the growth of InSb films on GaAs substrates.

Epitaxial InSb thin films on GaAs substrates are studied with low temperature (LT) InSb buffer layers having different thicknesses. The LT InSb buffer layers are critical to improve the quality of the InSb epitaxial films, and reduce the threading dislocation density therein. The LT InSb buffer layers are adapted to improve the quality of InSb thin films under appropriate growth conditions. Using transmission electron microscopy (TEM), we confirmed attenuation of threading dislocations by the buffer layer. Surface roughness was analyzed by Scanning electron microscopy (SEM). Furthermore, strain relaxation and mosaicity were investigated by X-ray diffraction (XRD) measurements. The measurements indicated that the lattice mismatch between InSb and GaAs generated the formation of plane defects in the InSb epilayer near the InSb/GaAs heterointerface. Structural properties of the InSb/GaAs heterostructures were strongly affected by the growth temperature.

These results indicate that the InSb buffer layer grown at low temperature, by eliminating the defects due to the lattice mismatch, plays an important role in the growth of InSb films on GaAs substrates. The films grown were n type and the highest electron mobility obtained was 5.7E+4 cm²v⁻¹s⁻¹ at room temperature.
Fig. 4 TEM image of the InSb/GaAs heterostructure grown on GaAs substrate at 420°C.

Fig. 5 TEM image of the InSb/GaAs heterostructure grown on GaAs substrate at 440°C.
Reconfigurable Photonic Microwave Bandpass Filter Based on Polarization Modulation

Choong Keun Oh, Tae-Young Kim, Chang-Soo Park
Department of Information and Communications, Gwangju Institute of Science and Technology,
1 Oryong-dong, Buk-Gu, Gwangju, 500-712, Republic of Korea,
Tel: +82-62-970-3145, E-mail: csp@gist.ac.kr

Abstract
We propose a reconfigurable photonic microwave bandpass filter based on polarization modulation. The experimental results show that a filter response can be reshaped by changing the polarization state and power of optical sources.

1. Introduction
Photonic microwave filters (PMFs) have attracted great interest because they allow the processing of radio frequency (RF) signals in the optical domain with the advantages of wide bandwidth, immunity to electromagnetic interference, and low loss [1].

These PMFs can be operated at either coherent or incoherent regime. However, the incoherent filters are preferred because they are more stable due to the free from the interference. Only, the incoherent filters are mostly implemented by the sum of optical powers tapped along the optical path and thereby negative coefficients cannot be achieved. For this reason, a few methods to obtain negative coefficients have ever been presented such as differential detection [2], cross gain modulation in a semiconductor optical amplifier [3], slicing of a broadband source [4]. However, these methods are limited in filter characteristics as far as the system configurations related to amplitude and/or sign for the filter coefficients are not changed.

In this paper, we propose a novel reconfigurable photonic microwave bandpass filter with negative coefficients based on polarization modulation. A uniform and a Hanning weighted filter with five taps are demonstrated.

Fig. 1. (a) Polarization modulation due to the switching voltage of \( V_s \) with the LiNbO\(_3\) crystal, (b) Measured non-inverted and inverted data patterns to the input of ‘10111100’ at 5-Gb/s.

2. Description of the proposed filter
Fig. 1-(a) depicts the principle of polarization modulation to implement positive or negative coefficient to the input polarization states. At the zero voltage level of the signal applied to the LiNbO\(_3\) crystal, the input beam with the polarization state of ‘B\(_{00}\)’ appears after the linear polarizer (LP). However, for the switching voltage of \( V_s \), the LP passes the beam only with the polarization state of ‘A\(_{00}\)’ because the input beam is rotated to the counter-clockwise direction by 90° due to the anisotropy of the electro-optic coefficient of the crystal [5] and thereby the input beam with ‘B\(_{00}\)’ is blocked. Fig. 1-(b) shows the output after LP to the applied signal with the data pattern of ‘10111100’ at 5-GHz, measured with a sampling oscilloscope. The non-inverted and inverted data signals observed show the exact replica of each signal with extinction ratio of 10-dB. Thus, they can be used as positive and negative coefficients required for the implementation of transversal filters, respectively. Also, the sign of the coefficients can be controlled by giving
beams (LD$_2$, LD$_4$), resulting in the change in the signs of the corresponding coefficients. And, in Fig. 3-(c) and (d), the amplitudes of the filter were selected from the Hanning window function by adjusting optical power of input beams. From the results, the mainlobe to sidelobe ratio of more than 30-dB could be achieved and well-matched to the theoretical results with the measured free spectral range (= 1/T$_d$) of about 102-MHz. Thus, the proposed method can be used to implement various kinds of filter shapes without modifying the filter structure.

4. Conclusion

We have reported a reconfigurable photonic microwave bandpass filter based on polarization modulation. By adjusting the polarization state and power of optical sources, various filter responses which have different filter coefficients were obtained.

Acknowledgement

This work is partially supported by KOSEF through grant No. R01-2006-000-11088-0.

References

All-optical interconnection for two-dimensional data using photorefractive organic film

Satoshi Honma1, Shinzo Muto1, Hiroshi Kobayashi1, Masatoshi Bunsen2, Atsushi Okamoto3
(1 Interdisciplinary Graduate School of Medicine and Engineering, Yamanashi University, Takeda, 4-3-11, Kofu, Yamanashi, 400-8511, Japan,
2 Faculty of Engineering, Department of Electronics Engineering and Computer Science, Fukuoka University, Japan
3Graduate School of Engineering, Hokkaido University, Kita 13-Nishi8, Kita-ku, Sapporo, 060-8628, Japan
Tel: +81-55-220-8412, Fax: +81-55-220-8412, Email: shonma@yamanashi.ac.jp)

Abstract
An all-optical interconnection for two-dimensional data using photorefractive organic film is proposed. An optical image transmission between two nodes controlled by pure optical control was demonstrated. It achieved over 10% connection efficiency.

1 Introduction
It has been demanded to transmit and process a large amount of information at high speed with rapid progress of studies in fields of information technology. Under the circumstances, we keep up interest in the optical parallel processing for improvement of processing performance remarkably. For example, Fourier transform using optical lens and Joint transform correlation using non-liner optics with over 1000 flame/sec rate are known well. Furthermore holographic data storage technologies have again been in the limelight as a next-generation storage system because of not only large storage capacities but also high transfer rates. Batch read out of two-dimensional data achieved over 1 Gbps data transfer rate. Hybrid systems which combine conventional electrical system and optical computing with parallel processing have been expected to be developed.

We propose a new technology of optical interconnection for two-dimensional data by using organic photorefractive film (PRF) and cross-polarized four-wave mixing (CP-FWM). PRF has some advantages such as large effective area, high efficiency and low cost. In the interconnection module, the optical image inputted to the center unit is automatically delivered to only receiving nodes which are sending a requesting beam to the unit. Using the organic photorefractive film (It is made by Nitto Denko inc.) and He-Ne laser as photorefractive material and optical source, it was demonstrated that optical image is connected to the receiving nodes by pure optical control. We measured the connection efficiency of optical signal relative to the incident intensity ratio of the pump beam to the requesting beam. The connection efficiency, which depends on the incident angle of the beam with respect to the direction of the applied voltage on the PRF, was larger than 10% when the intensity ratio was adjusted to the optimum value.

Fig. 1 Conceptual diagram of interconnection module
Fig. 1 shows the conceptual diagram of the all-optical interconnection module. The optical image is input to a center unit from a sender unit. On the other hand, receiver units turn the requesting beam on. The optical connections between the center unit and the receiver units are built up automatically and the input image is delivered to the desired receiver. The state of optical connection is controllable by turning the requesting beam on or off.

Fig. 2 shows the schematic diagram of the interconnection module by using CP-FWM. A p-polarized signal beam is incident to PRF from the transmitter, and simultaneously the s-polarized pump beam is turned on in the opposite direction. On the other hand, the s-polarized requesting beam is incident to PRF from the receiver. The pumped beam and the requesting beam are plane waves and the signal beam has two-dimensional data.

When s-polarized requesting beam is incident to the PRF from the receiver unit, it interferes with the pump beam and then a refractive index grating is induced via photorefractive effect. The signal beam is diffracted by the grating toward the receiver unit. The diffracted signal beam propagates in opposite direction with the requesting beam, but the optical pass of the two beams can easily be divided using a polarized beam splitter because of mutual orthogonal polarization.

It takes several tens of milliseconds for buildup of the optical connection, but it is possible to transmit dynamic image in light velocity because the index grating is constantly induced by interference of the s-polarized beams in PRF after the connection is once established.
3 Image transmission and connection efficiency

Fig. 3 shows the optical setup of the interconnection. The organic photorefractive film (It is made by Nitto Denko inc.) whose thickness is 50[μm] and He-Ne laser whose wavelength is 633[nm] were used as photorefractive material and optical source. M, BS, PBS, and BE indicate mirrors, beam splitters, polarized beam splitters and beam expanders. SLM is a spatial light modulator PPM 8267 made by Hamamatsu Photonics. Applied voltage to the photorefractive film is 8[kV].

P-polarized signal beam, S-polarized pump beam and requesting beam illuminate on the PRF. The optical image whose size is 1.7[mm] × 1.7[mm] was given to the signal beam by SLM. The signal beam was switched to channel 1 (CH1) or CH2 by using optical shutter Sh2 or SH1 for data transmission from CH1 to CH2 or CH2 to CH1, respectively.

Fig. 4 shows the input optical image given by SLM and the output results on CCD camera 1 and 2. We found that the optical images is transmitted to CCD camera and the quality of output image is better than the result using conventional optical setup[1]. We also found that connection efficiency depends on the direction of the input signal beam with respect to the direction of the applied voltage to PRF. In this experiment, the connection efficiency in case of connection CH2 to CH1 was larger than that in case of connection CH1 to CH2.

Index modulation in photorefractive materials is proposal to intensity contrast ratio of interference fringes. Therefore the connection efficiency also depends on the intensity ratio of the requesting beam to the pump beam. We measured the connection efficiency of the signal beam against the intensity ratio of the requesting beam to the pump beam, where the intensity of the signal beam is adjusted equal to the pump beam. The result is plotted in Fig. 5. The maximum values of the connection efficiency were 15.2% and 11.8% in case of connection CH1 to CH2 and CH2 to CH1 respectively. We have to adjust the intensity ratio of the requesting beam to the pump beam for high connection efficiency. Here, the connection efficiency increases with increasing applied voltage to PRF.

4 Conclusion

We proposed the new optical design of the interconnection for image transmission. It was demonstrated optical image was delivered to the receivers transmitting the requesting beam. It achieved over 10% of the connection efficiency from the transmitter to the receiver.

References
Self-Written Micro Optical Pin for Optical Interconnect

Takuya Tokuhara, Osamu Mikami
School of Information Science and Technology, Tokai University
1117 Kitakaname, Hiratsuka, Kanagawa, 259-1292 JAPAN,
TEL: 0463-50-2016
6adgm014@keyaki.cc.u-tokai.ac.jp

Abstract – An easy fabrication method of micro optical pin using Self-Written Waveguide (SWW) having a 45-deg mirror was proposed and fabricated. An optical pin with a short length of 160 μm was fabricated successfully. Positional tolerance of SWW-optical pin is compared with that of the fiber optical pin theoretically.

INTRODUCTION

An optical coupling scheme using optical pins has attracted much attention to realize board to board and chip to chip optical interconnection. Conventionally, the optical pin is fabricated using a multi-mode fiber (MMF) and the tip is cut by a rotating blade[1,2]. This may induce a difficulty in adjusting the length, controlling the positional alignment and handling. On the other hand, self-written waveguide (SWW) optical pins using UV curable resin by a photo-mask transfer were also studied[3]. However, this has a problem in positional alignment just on the optical devices.

In this paper, we have proposed a new fabrication method of SWW micro optical pin. Much reduction in the implementation cost and much improvement in the productivity can be expected by using this technology.

APPLICATION OF OPTICAL PIN

We have proposed an optical surface mount technology for optoelectronic (OE) printed wiring boards (PWBs) as an optical interconnection technology [1]. An OE-PWB has optical and metallic wiring inside the board, and optical and electrical devices are mounted on the board. Most laser diodes (e.g., vertical cavity surface emitting lasers (VCSELs) and laser diodes (LDs)) and photodetectors (PDs) are mounted face down. Therefore, the optical paths of optical devices (e.g., between a VCSEL and a PD) and optical wiring are crossing at right angles. In this case, 90-deg path redirection devices such as a 45-deg mirror, prism, and 45-deg-ended waveguide are needed for optical signal transmission. One effective redirection device is an optical pin, which is a waveguide having a 45-deg mirror on its end-face.

The optical pin has a mirror on the tip and has a function of beam path conversion by 90 degrees. However, fabrication of optical pins is difficult and it takes a lot of time for processing. Especially, short optical pins with a length of 100-200μm are needed in applying into the board level.

Fig.1  Application of optical pin in OE-PWB

The application of an optical pin is schematically shown in Fig. 1 for optical interconnection inside an OE-PWB. We have already demonstrated this type of optical model and its performance characteristics [2]. The optical pin has a large positional tolerance along x, y, z-axes and rotation. Therefore passive alignment is possible.
EASY FABRICATION METHOD

We have proposed a new method to fabricate a very short SWW optical pin having a 45-deg mirror. The proposed method has following advantages. An optical pin with very short length less than 100 µm and a precise length can be easily obtained. The short SWW optical pin can also be fabricated just on the window of such as VCSEL and PD.

Fig. 2 is the experimental setup. We used dye mixed UV curable resin and a green laser beam. We put a GI-MMF having a 45-deg mirror at the tip vertically to the glass plate. Distance between the glass plate and the GI-MMF was set to be about 150 µm. Next, we filled the dye mixed resin into the space between the GI-MMF and the glass plate. After that, by launching a green laser beam into the GI-MMF, an SWW grew and reached the glass plate. Finally, non-hardened resin was removed off.

A micrograph of fabricated SWW micro optical pin is shown in Fig. 3. A 45-deg mirror was successfully fabricated on the tip and the total length was 160 µm. The results are summarized in Table 1.

Coupling efficiency and positional tolerance of optical pins having a conventional fiber structure and SWW are compared theoretically. In the case of making an optical pin just on the window of VCSEL, the positional alignment point of SWW pin can be reduced and the clad-less structure has a great advantage in coupling between the 45-deg mirror and a fiber. One dimensional array micro pins can be fabricated by using a taped fiber and/or multi-channel waveguide with a 45-deg at the tip.

CONCLUSION

New fabrication method of SWW micro optical pin was proposed and the feasibility was investigated. The positional tolerance of conventional fiber pin and SWW pin was compared theoretically. Effectiveness of the easy fabrication method was confirmed.

REFERENCES

Optical parallel interconnects using organic photorefractive polymer

Kohei Shimayabu1, Kazuhiro Harasaka1, Atsushi Okamoto1, Shuji Rokutanda2
(1Graduate School of Information Science and Technology, Hokkaido Univ, 2Nitto Denko Technical Corporation)
N14-W9, Kita-Ku, Sapporo, 060-0814 Japan, Tel: +81 11 706 6522, Fax: +81 11 706 7836,
E-mail: shimayabu@optnet.ist.hokudai.ac.jp

Abstract
Optical parallel interconnects using photorefractive polymers as optically controllable Bragg gratings are proposed. Spatial resolution of about 50lp/mm which is comparable with other parallel optical devices is achieved by using thinner polymers than 100μm.

1 Introduction
Optical parallel interconnects attract much attention as a method to realize high-speed optical transmission in relatively short length such as board-to-board. Among them, that using photorefractive effect is actively studied because they can reconfigure the connection pattern all optically. We have proposed optical parallel interconnects with photorefractive connection module (PRCM), which is an application of photorefractive four-wave mixing, by using photorefractive polymers[1]. Signal beams which carry two-dimensional data can be selectively extracted from paralleled spatial channels by optically induced Bragg diffraction gratings in the polymer. Since the gratings are optically rewritable, signal beams to be extracted are switched by matrix patterned control beams from paralleled light emitting devices such as VCSEL arrays. Since E/O conversions are not required, this module simplifies the configuration and speeds up extracting signal beams.

We use polymeric materials instead of generally used crystals as the photorefractive material because thin gratings induced in the polymer suppress the distortion of the diffracted signals compared to thick crystals. Furthermore, diffraction efficiency comparable with crystals is obtained due to about 10² times larger index modulation than crystals, while the diffraction efficiency is generally proportional to square of the grating thickness. The large index modulation is caused by the applied electric field via orientational enhancement[2]. The applied field is increased as the thickness becomes small, and the index modulation is more increased. Therefore, thinner photorefractive polymers realize both higher spatial resolution and sufficient diffraction efficiency.

In this paper, we estimate the spatial resolution of the Bragg diffracted beam by the photorefractive polymer. Analytical and experimental results reveal that spatial resolution of 20-50lp/mm which is comparable with the pitch of general VCSEL arrays (~50μm) for optical parallel interconnects is achieved with reducing the deterioration of the diffraction efficiency.

2 Photorefractive connection module
PRCM is an application of photorefractive four-wave mixing for optical parallel interconnects. PRCM is composed of a photorefractive polymer and a polarization beam splitter (PBS) (Fig.1). Here, we selectively extract signal beams which carry two-dimensional data from spatially paralleled 2 x 2 channels. First, pump beam, which has uniform intensity in its cross section, and the signal beams are incident on as they counter-propagate in the polymer. Control beam, which has matrix patterned intensity distribution, is incident on and the Bragg gratings are induced by interference between the control beam and the pump beam. Accordingly, the signal beams in channels to be extracted are branched by the Bragg grating into the diffracted beams which propagate in the direction of the control beam, and the transmitted beams. The diffracted beams are separated from the control beam by the PBS. Extracted channels are switched over by changing the intensity pattern of the control beam. Once the gratings are induced in the polymer, the signal beams are extracted in a moment, and the gratings disappear when the control beam is cut off.

We can construct optical bus line which distributes optical signal beams by sharing the transmitted pump beam and the signal beams into next PRCM (Fig.2).

3 Estimation of spatial resolution
We estimate the spatial resolution of the Bragg diffracted beam by the photorefractive polymer. Angular selectivity of the diffraction efficiency depends strongly on the grating thickness by phase matching condition. From Fourier optics, intensity modulated
parallel signal beams are described as angular spectrum of plane waves and the propagation angle $\Delta \theta$ is expressed as $\sin \Delta \theta = f_s \lambda$, where $f_s$ is the spatial frequency in x-direction, and $\lambda$ is the wavelength. Accordingly, the diffraction efficiency with respect to each spatial frequency component can be estimated by phase matching and Fourier analyses.

First, we show the phase mismatch characteristic of the diffraction efficiency derived from coupled wave theory. The analysis model is shown in Fig.3. We used material parameters of the polymer from Ref.3, and calculated for four different thicknesses (See Table.1). The applied voltage is experimental value so as not to cause dielectric breakdown of each polymer. Fig.4 shows that the angular selectivity of diffraction efficiency becomes wider with the thickness get thinner. It indicates that thinner polymers can diffract the higher spatial frequency components. In contrast, the diffraction efficiency decreases with a decrease in the polymer thickness. However, the efficiency is less decreased than general Bragg gratings (the efficiency decreases in proportion to square to square of the grating thickness). It is because larger electric field is applied to thinner polymer, and it enhanced the index modulation. Therefore, thinner photorefractive polymers improve the spatial resolution with suppressing the decrease in the diffraction efficiency.

Next, we consider the spatial resolution for intensity modulated signal beams as square wave. Fig.5 shows the analysis flow chart. The signal waveforms after Bragg diffraction are calculated with spatial frequency filtering by the diffraction efficiency. We assume that the signal beams are intensity modulated like a 3-Bar filter. Fig.6 shows the experimental setup with bias voltage of 3.0kV, 4.0kV, and 9.5kV are applied, respectively. The spatial resolution is evaluated on the contrast defined as $C_{\text{max}}-C_{\text{min}}/(C_{\text{max}}+C_{\text{min}})$.

The analytical and experimental results (Fig.7) show that the higher contrast is kept at larger line pairs as the thickness gets thinner. Temporal increase in contrast for 100µm around 50lp/mm in Fig.7a is caused by the side lobe of the phase mismatch characteristics (Fig.4). From Fig.7a, it can be seen that the contrast for 100µm decreases rapidly around 20lp/mm, which is similar to Fig.7b. However, the experimentally obtained contrasts for 10µm and 30µm are decreased around 25lp/mm while the calculated values remain high. It is probably due to imaging system which forms an image on the vertical view plane (CMOS) from tilted polymer[4]. Therefore, the spatial resolution for 100µm on Bragg diffraction is about 20lp/mm, and higher resolution over 50lp/mm is expected by using thinner polymers about 10-30µm. To confirm the resolution limit for thinner polymers than 100µm, a higher resolution CMOS than present one (up to 28lp/mm) has to be used. The spatial resolution of 20-50lp/mm is comparable with the pitch of VCSEL arrays. Therefore, parallel signals can be extracted by photorefractive polymers without losing information.

4 Summary

Optical parallel interconnects using photorefractive polymers as Bragg gratings are proposed. Spatial resolution of 20-50lp/mm can be realized with suppressing the decrease in the diffraction efficiency.

5 References

1 K. Harasaka et al., Tech. Dig. OECC2006, No.4P-8, 2006
Measurements of Lattice Constant and Electrooptic Constant in Reverse-Proton-Exchanged LiNbO₃ Waveguides

Shoji Kakio, Mokoto Maeda, Hitoshi Watanabe and Yasuhiko Nakagawa
Interdisciplinary Graduate School of Medicine and Engineering
University of Yamanashi, Takeda-4, Kofu 400–8511, Japan
Email: kakio@yamanashi.ac.jp

Abstract—Reverse-proton-exchanged waveguides in Z-cut LiNbO₃ are found to have an almost perfectly recovered lattice constant and an electrooptic constant $r_{13}$ of 8.6 pm/V, which is equal to the bulk value.

I. INTRODUCTION

Reverse proton exchange (RPE) has been proposed as a method of exchanging lithium ions and protons in the surface of a proton-exchanged waveguide[1][2]. By utilizing the RPE process as a method of fabricating buried waveguides, the highest normalized difference frequency generation (DFG) efficiency was achieved in quasi-phase-matched DFG devices[3]. On the other hand, Rams et. al reported that the nonlinear optical coefficient $d_{33}$ in the RPE waveguide is very similar to that of the substrate for Z-cut LiNbO₃[4]. If the RPE waveguide has electrooptic, photoelastic, and piezoelectric constants very similar to those of the bulk, for instance, an improvement of the driving power in a waveguide-type acoustooptic modulator[5] can be expected.

In this study, the lattice constant of the RPE layer formed in Z-cut LiNbO₃ was evaluated by measuring the X-ray diffraction pattern. Moreover, the electrooptic constant $r_{13}$ in the RPE channel waveguide was measured by the phase-modulation method[6].

II. LATTICE CONSTANT IN RPE LAYER

An RPE planar waveguide with a depth of 2.0 µm was fabricated. An initial PE layer with a depth of 4.8 µm was formed by immersing a Z-cut LiNbO₃ substrate in pure phosphoric acid at 220°C for 26 h. Then, an RPE layer was formed by immersing the sample in an equimolar mixture of LiNO₃-NaNO₃-KNO₃ at 350°C for 10 h 40 min.

Figure 1 shows the X-ray diffraction pattern for the (0 0 12) plane of LiNbO₃ before and after the RPE process. The satellite peak appeared at a diffraction angle of around 83.3° after the PE process, showing that the PE layer undergoes a lattice constant change of 0.33%[6]. After the RPE process, the PE waveguide was buried under the RPE layer and was simultaneously annealed at the temperature of the RPE process. $\Delta c/c$ for the PE layer decreases with the annealing time[7], as is shown in Fig. 1. In spite of a PE layer deeper than the initial PE depth, the peak intensity at the diffraction angle corresponding to the substrate increased after the RPE process. Therefore, the peak with $\Delta c/c$ equal to approximately 0.02% corresponds to the RPE layer. The lattice constant is found to be almost perfectly recovered by the RPE process.

III. ELECTROOPTIC CONSTANT IN RPE WAVEGUIDE

Figure 2 shows the configuration of the phase modulator used for the measurement of the electrooptic constant. First, a PE planar waveguide with a depth of 3.7 µm was formed on the whole surface of the Z-cut LiNbO₃. For the fabrication of RPE channel waveguides, RF-sputtered SiO₂ masks with a thickness of 0.25 µm and a mask width of 3 µm were formed by a lift-off technique. Next, an RPE channel waveguide with a depth of 1.0 µm was formed by RPE for 3 h 15 min. A SiO₂ buffer layer was sputtered on
the surface of the sample to prevent the absorption of the electrical field of the guided wave. Finally, after polishing the end face, two parallel strip electrodes with a gap d of 10 µm and a length l of 14 mm were formed using an aluminum film. These parameters were chosen so that the RPE channel waveguide maintains a single-mode guided wave in the depth direction.

By means of the phase-modulation method[6], the electrooptic constant in the RPE waveguide was measured. The applied RF electric field is parallel to the Z-axis and perpendicular to the optical electric field of the TE-like mode. Therefore, the electrooptic constant sensed by the guided wave in the RPE waveguide is $r_{13}^S$.

A He-Ne laser beam with a wavelength $\lambda$ of 0.633 µm was focused by an object lens onto the fabricated sample on its end face so that the TE-like single-mode guided wave was excited. An RF voltage at a frequency of 100 MHz was applied between the electrodes. The spectrum of the modulated output beam passing through a scanning Fabry-Perot interferometer was detected using a photomultiplier and was observed on an oscilloscope.

Typical examples of the spectra of the output wave are shown in Fig. 3. When no voltage is applied, only the spectrum of the carrier corresponding to the wavelength of the He-Ne laser appears. As a voltage $V_{p-p}$ is applied, the power is shifted from the carrier to the sideband according to the degree of modulation. The measured results of the ratio of the power in the carrier to that in the first sideband for the applied voltage cases are shown as solid circles in Fig. 4. The power ratio is given as the square of the ratio of the Bessel functions of the zeroth and first orders. The solid lines in Fig. 4 show the theoretical result of $(J_1(u)/J_0(u))^2$ obtained for a parametric value of $r_{13}^S$. Here, $u$ is the phase-modulation index and is expressed as

$$u = \frac{\pi n_o r_{13}^S l}{\lambda d} \Gamma V_{p-p}/2,$$

where $n_o$ is the ordinary refractive index and $\Gamma$ is the correction factor for the applied electric field. Here, the computed value for a similar waveguide structure was set as $\Gamma=0.34[6][8]$. $r_{13}^S$ was determined so that the theoretical curve is fitted to the measured values by the least squares method. The determined $r_{13}^S$ was 7.94 pm/V, which is 92% of the bulk value (8.6 pm/V).

The measured value derived above is the effective value containing the characteristics in the annealed PE (APE) waveguide due to the penetration of the guided wave into the cladding layer. It can be assumed that the electrooptic constant in the APE waveguide is 70% of the bulk value[8]. By taking the cross section of the electric field distribution of the guided wave in the core and cladding into account, $r_{13}^S$ in the RPE waveguide is determined to be 8.6 pm/V and is equal to the bulk value.

IV. CONCLUSIONS

The lattice constant and the electrooptic constant $r_{13}^S$ in RPE waveguides in Z-cut LiNbO$_3$ were evaluated. The lattice constant in the c-axis direction was found to be almost perfectly recovered by the RPE process. $r_{13}^S$ in the RPE waveguide was determined by the phase-modulation method to be 8.6 pm/V, which is equal to the bulk value.

REFERENCES

Low Birefringence Silicon-on-insulator Waveguide and Its Optical Interconnection using High Numerical Aperture Fiber

Shih-Hsiang Hsu
Department of Electronic Engineering
National Taiwan University of Science and Technology
No. 43, Sec. 4, Keelung Rd., Taipei, Taiwan
TEL: +886-2-2737-6359
Fax: +886-2-2737-6424
E-Mail: sshu@mail.ntust.edu.tw

Abstract: 5 µm thick silicon-on-insulator waveguide was demonstrated to low birefringence of 4.6x10^{-5}. Instead of complex processing from 3D mode-size-converter, the high numerical aperture fiber was utilized to get competitively coupling loss of 0.4 dB.

Introduction
Silicon-on-insulator (SOI) is a critical platform for integrated optoelectronic circuits since it offers the potential of monolithic integration for photonic and electronic functions on a single substrate [1]. Integrating photonics functions on a silicon platform will be a low cost solution if integrated optoelectronics circuits are feasible. Birefringence is a serious problem faced by many photonics devices, which results in polarization dependent wavelength (PDW) shift and polarization dependent loss (PDL). However the beam propagation method (BPM) with semi-vectorial capability to simulate SOI waveguides shows that the waveguide height of a few microns has lower birefringence than one micron or less height waveguide, which is approximately 10^{-2}, two orders of magnitude larger. Therefore a bigger core size for SOI waveguide is preferred due to higher birefringence effect on a smaller core. On the other hand, thicker SOI waveguides will cause significantly large bend radius. A suitable core size for SOI waveguides should compromise between low birefringence and competitively compact device size.

The mode field diameter (MFD) from thinner SOI waveguides will have significant coupling loss associated with a SMF-28 fiber. A complex mode-size-converter taper was then utilized on this kind of SOI waveguides to make optical modes extended to match fiber modes via three dimensional expansion [2]. A simple SOI waveguide process is necessary for further multiple layers integration and cost reduction. Therefore, a high numerical aperture (HNA) fiber with small MFD was used as the bridge between a standard telecommunication fiber, SMF-28, and higher refractive-index-contrast SOI waveguides [3]. To optimize the coupling loss, the SOI waveguide mode can be chosen as similar as the HNA fiber. In this paper, the birefringence and PDW shift of SOI waveguide with suitable thickness will be discussed and demonstrated based on the waveguide geometric width and etch depth. A low coupling loss between HNA fiber and SOI waveguides will also be presented.

Birefringence Study for SOI Waveguides
The mechanical stress to an optical substrate modifies the optical properties of a material by the dielectric impermeability. Thus, a homogeneous and isotropic material subject to mechanical stress will become optically anisotropic. The phenomenon is known as the stress birefringence or photoelastic effect. The changes in the indices of refraction are due to the effects of stress imparting changes in the dielectric impermeability that alter the size, shape, and orientation of the index ellipsoid.

The correlation between PDW shift and PDL can be illustrated by the array waveguide grating (AWG) demultiplexer. The modal birefringence can be observed as PDW shift on the central wavelengths, which can be given by

$$PDW=\frac{\Lambda}{L/m}$$

Here \(\lambda\) and \(n\) are the wavelength and effective index of the waveguide, respectively; for transverse electric (TE) and transverse magnetic (TM) polarization states; \(\Delta L\) is the difference of two adjacent grating lengths; and \(m\) is the grating order in an AWG. Once the PDW shift is verified, the difference between highest and lowest optical power on the central wavelength is called PDL. The lowest birefringence and PDL are required for telecommunication system design.

A BPM simulation shows that if the waveguide dimensions shrink to around 1 µm or less in high refractive-index-contrast SOI waveguide layer structures [4], it is not only that the birefringence effect is large, but also the PDL becomes significant. The overlap integral simulation suggested that the most efficient coupling with a SMF-28 fiber is 12 µm thick SOI waveguides. The large core thickness is then preferred for decent optical performance. Nevertheless, the large core needs a very big bend radius to maintain the single mode operation and the device layout cannot be maintained as compact as expected. For example, 12 µm thick core of SOI waveguides satisfies the single mode conditions only when 20 mm minimum bend radius is applied. For the points of view with regard to low birefringence, low PDL, compactness, and efficient interconnection, another compromise thickness for SOI waveguides is needed.

The recent progress on fusion splicing can join two different cores of fibers within 0.1 dB coupling loss. The HNA fiber is a good selection to play an intermediate role to connect SOI waveguides and SMF-28 fibers. The MFD of HNA fiber is typically around 5 µm, which corresponding bend radius is around 7 mm for single mode operation. Therefore, the BPM simulation tool, BeamPROP, was implemented onto 5 µm thick SOI waveguide to understand the birefringence performance. In Fig. 1 (a), a reasonable etch depth, 2.65 µm, to get a compact bend radius, 7 mm, was utilized to study the width effect on birefringence variation for 5 µm thick SOI waveguide. The birefringence and PDW will have a minimum value on 3.8 µm width. And then 3.8 µm width is fixed, the etch depth is varied before reaching the multimode region, 2.8 µm etch depth. The lowest birefringence and PDW are 5x10^{-5} and 3 GHz, respect-
respectively, on 2.76 μm etch depth and 3.8 μm width on 5 μm thick SOI waveguide.

![Graph](image)

**Fig. 1** (a) PDW vs. waveguide width for 5 μm thick waveguide and 2.65 μm etch depth (b) PDW shift vs. etch depth for a waveguide with 5 μm thickness and 3.8 μm width (Wavelength is assumed as 1550 nm)

<table>
<thead>
<tr>
<th>5 μm Thick SOI Waveguide with 3.8 μm Width</th>
<th>Etch Depth (μm)</th>
<th>PDW Shift (f_{21}-f_{12}) (GHz)</th>
<th>Birefringence</th>
</tr>
</thead>
<tbody>
<tr>
<td>SOI AWG1 Central Channel</td>
<td>2.45</td>
<td>4.375</td>
<td>8.1E-05</td>
</tr>
<tr>
<td>SOI AWG1 Central Channel</td>
<td>2.65</td>
<td>3.125</td>
<td>5.8E-05</td>
</tr>
<tr>
<td>SOI AWG2 Central Channel</td>
<td>2.45</td>
<td>5</td>
<td>9.2E-05</td>
</tr>
<tr>
<td>SOI AWG2 Central Channel</td>
<td>2.65</td>
<td>2.5</td>
<td>4.6E-05</td>
</tr>
</tbody>
</table>

**Table 1** Birefringence comparison on various etch depths

![Graph](image)

**Fig. 2** Birefringence study from a 17-channel AWG

**Experiment Results and Discussions**

To avoid the multimode region for SOI waveguides and the processing variation, the reasonable etch depth for 5 μm thick SOI waveguide was only applied up to 2.65 μm and its bend radius can be controlled within 7 mm. The birefringence of the central channels from two AWGs, located in the same SOI wafer, are listed in Table 1, which indicates that the PDW shift gets improved when the etch depth is deeper. Due to the warpage of SOI wafer, the experimental data for PDW shift of Table 1 demonstrate 2 GHz higher than theoretic simulation of Fig. 1 (b). Finally a 17-channel AWG was employed to check the birefringence distribution across the whole 4" SOI wafer, shown in Fig. 2, which insertion loss is, excluding Fresnel effect in the power budget, around 5 dB. The residual birefringence induced by the stress via the photoelastic effect shows 6 GHz PDW shift variation on the 4" SOI wafer. A more uniform SOI wafer with less warpage is necessary for better birefringence control. The HNA fiber was then implemented to couple to the horizontal tapered SOI waveguides in the input/output areas of 17-channel AWG with a thickness of 5 μm, width of 5.5 μm, and etch depth of 2.65 μm. Another additional coupling loss of 0.4 dB needs to be considered for each interface. The splicing between HNA and SMF-28 fibers can be optimized to get the coupling loss as low as 0.1 dB per splicing.

The three-dimensional taper is a gradual transition from a large cross-sectional waveguide area to a smaller one. A vertical taper requires a differential etch rate along the length of the taper [4]. Recently a simpler two-step etch and regrowth can be applied together to remove the difficulties for a smooth vertical taper transition and achieve the coupling loss of less than 0.5 dB/facet on the interface [2]. Compared with two approaches for tapering, the two-dimensional tapers are relatively straightforward to fabricate because this lateral taper is essentially just an etching process from the top of the silicon wafer. The suitable HNA fiber is then applied for efficient coupling. Due to the MFD from HNA fiber is a little bit bigger than 5 μm, our new design with 5.5 μm thick core of SOI waveguide can achieve a better coupling loss, within 0.2 dB, with the HNA fiber.

**Summary**

We demonstrated the low birefringence of 4.6×10^{-5} for a SOI waveguide with 5 μm thickness, 3.8 μm width, and 2.65 μm etch depth. The effect on birefringence from the wafer warpage still needs to be considered for better performance control. An AWG demultiplexer is a good indicator to demonstrate the birefringence distribution across the whole wafer. The HNA fiber was successfully utilized to connect 5 μm thick SOI waveguides and SMF-28 fibers, respectively, with 0.4 dB per interface and 0.1 dB per splicing for joint junction loss to achieve excellent coupling efficiencies as well as low birefringence.

The authors thank Professor S.-L. Lee and Professor S.-K. Liaw for their technical support, Chun-Min Huang and Dong-Han, Shen for the processing development.

**References**

4. G. T. Reed and A. P. Knights, “Silicon Photonics – an introduction”, (John Wiley & Sons Ltd., 2004), Chapter 7 and 4
Comparison of Transmission Media for Next Generation Home Network

Shohei Terada1, Yu Kakishima1, Shingo Yamakawa1, Kunio Tojo1, Teruyuki Taniguchi2, Takashi Kawakami2, and Kimio Oguchi1**

1Graduate School of Engineering, Seikei University, 3-3-1, Kichijo-Kitamachi, Musashino 180-8633, Japan
Tel: +81-422-37-3732, Fax: +81-422-37-3871, Email: *dm053210@cc.seikei.ac.jp, **oguchi@st.seikei.ac.jp
2Sekisui Chemical Co., LTD, 2-2, Kamichoshi-cho, Kamitobaba, Minami-ku, Kyoto, 601-8105, Japan

Abstract Transmission media for the next generation home network (NgHN) are compared. Interference from microwave ovens is measured for wired/wireless transmission. Characteristics of optical and metallic cables are also compared. It is concluded that optical cable is the most promising medium for the NgHN.

1. Introduction

The number of FTTH (Fiber-To-The-Home) users is increasing, but their number in Japan is exploding. At the end of 2005, the number was over 4 million and at the end of Sept., 2006, it was over 7 million. The number will exceed 30 million by 2010. The rate of acceptance is significantly higher than for any other broadband medium such as ADSL (Asymmetric Digital Subscriber Line). The major service is the triple player of video, Internet, and VoIP [1]. The new players will be broadband services such as a higher resolution video and on line gaming.

Next generation home network (NgHN) will accommodate high resolution video/TV, a variety of home appliances/terminals, VoIP (Voice over IP), and plenty of sensors [2]. Research is rapidly advancing in several areas [3]. We consider here the broad classes of transmission media i.e. wired/wireless with various bandwidths.

Video applications create burst traffic with steep peaks and idle periods. DVTS (Digital Video Transport System), which will become common in the home, outputs peak traffic of more than 100 Mbit/s while its average is around 30 Mbit/s [4]. In order to maintain the required QoS (Quality of Services), wired, or optical fiber transmission is indispensable for NgHN.

The NgHN configuration that uses optical fibers to the fullest extent has been not adequately discussed because the kinds of terminals/appliances and their QoS requirements have not been settled yet. The authors presented the characteristics of transmission cable in terms of cable length for an NgHN house [5]. This paper first compares the transmission characteristics of both wireless and wired transmission by evaluating the interference from microwave ovens, a common home appliance. Wired media are also compared.

2. Comparison of wireless/wired transmission

The average home has several appliances that emit electromagnetic waves. This may cause electro-magnetic interference (EMI) problems for the transmission medium configuring NgHN. The EMI from a microwave oven, a very common appliance, was measured. Figure 1 shows the experimental set-up. A data quality analyzer (Anritsu, MD1230B) transmitted 54Mbit/s IP traffic with packet size of 1508 Bytes. The traffic was returned to the analyzer via wireless LAN (IEEE802.11g) access point and a wireless bridge. The distance between the commercially available microwave oven (600 W power) and the access point was 7m considering the maximum room size. The wireless bridge was located on concentric circle: Positions a, b, c.

![Fig.1 Experimental set-up](image)
Packet loss was measured while changing the distance between the access point and wireless bridge. Figure 2 shows the measured results. The result clearly shows that packet loss varies from several percent up to 30 percent depending on the location of the wireless devices. The EMI from the same microwave oven on optical or metallic cables was also measured using 1 Gbit/s IP traffic with the same analyzer as shown in Fig. 1. Optical cable used was plastic optical fiber (POF) [6] while the metallic cable was Ethernet category 5e. No packet loss was observed for these cables.

The quality of the video streaming application, which will be a dominant NgHN service, is degraded by packet loss. Therefore, wireless transmission is impractical for NgHN use.

3. Comparison of optical and metallic cables

To compare optical and metallic cable, bandwidth and cable length are important points. Video terminals have the highest demands, ranging from several Mbit/s to several tens of Mbit/s depending on the quality. Future video terminals may have much higher demands. The video application creates burst traffic with steep peaks and idle periods. The DVTS (Digital Video Transport System) outputs peak traffic of more than 100 Mbit/s with average rate of around 30 Mbit/s [4]. The total bandwidth needed will be around a few Giga bit/s. Both optical and metallic cables satisfy this requirement, but metallic cable offers a transmission range of only 100m. Optical fiber is superior in terms of transmission range and bandwidth and by using WDM (Wavelength Division Multiplexing), more effective broadband transmission can be realized [5].

Cable weight and diameter are also important points in implementing the NgHN. UTP (Unshielded Twist Pair) cable weight for Ethernet is 34 g/m, and its diameter is 6 mm. Multi-mode optical fiber cable weighs 10 g/m, and its diameter is 4.6 mm. General duct diameter in an average home is 24 mm. Therefore, more than 20 multi-mode optical fiber cables can be installed in a duct, while only several UTP cables could be. Therefore, optical fiber has much higher scalability than UTP cable to expand the network.

Considering the cable length needed, traffic characteristics and cable characteristics, optical fiber will be preferred for the NgHN. Metallic cables might be used, however, depend on the length, bit rate used, and the number of cables installed in a duct. Other considerations such as connectivity, cost, and ease of installation should be examined to choose the optimum medium.

4. Conclusion

The transmission medium for the next generation home network (NgHN) was discussed. Interference from microwave ovens on wired and wireless transmission was measured. Wireless transmission suffered high bit error rates and so is problematic for the NgHN.

Traffic characteristics for the dominant NgHN application, video, were elucidated. Some cable characteristics were compared. Considering all of the requirements, optical fibers are the most promising medium for the NgHN.

References

Challenges and Requirements for ASON/GMPLS Photonic Networks

(Wataru Imajuku and Yasunori Sameshima)

NTT Network Innovation Laboratories, 1-1 Hikari-no-oka, Yokosuka, Kanagawa, 239-0847, Japan
NICT Tsukuba Research Center, 2-5-5 Azuma Tsukuba-shi, Ibaraki, 305-0031, Japan
Tel: +81-46-859-4315 Fax: +81-46-859-5541, E-mail: imajuku.wataru, sameshima.yasunori)@lab.ntt.co.jp

Abstract: This paper discusses the issues and requirements pertaining to Automatically Switched Optical Networks (ASONs) and Generalized Multi-Protocol Label Switching (GMPLS) control functionality to achieve dynamically reconfigurable photonic networks.

I. INTRODUCTION

There are mainly two directions for the evolution of photonic network architecture. The first one is the convergence of functionalities in photonic transport equipment. In particular, this is exemplified by the integration of packet switching functionality such as an Ethernet frame switching capability into photonic transport nodes such as Optical Cross-Connects (OXC’s) and Reconfigurable Optical Add Drop Multiplexers (ROADMs). The other direction is the continuous effort to expand the dimensions of transparent optical networking. Thanks to the evolution of long-haul and high capacity optical transmission technologies, transparency in the ROADM ring has become a very common network architecture, and now transparent optical networking even in optical mesh networks is a reality. Both of these directions provide opportunities for network operators not only to provide low-cost and reconfigurable network operation, but also to enhance network reliability if elaborate network design is successfully achieved.

The Automatically Switched Optical Network (ASON) [1] and Generalized Multi-Protocol Label Switching (GMPLS) technologies [2] represent a logical framework that support or drive such technical trends by unifying the control of various types of switching capabilities. The concept of an LSP (Label Switched Path) hierarchy in particular ensures flexibility and multiple types of switching capabilities. The concept of an LSP (Label Switched Path) hierarchy in particular ensures flexibility and scalability of control for such converged networks. The objective of this paper is to present an evaluation of the ASON/GMPLS control plane technology, which includes unifying the control of various network elements given in Table I [6] at Tokyo and Isocore in Washington D.C. This trial was a test to ensure the routing scalability of the GMPLS controlled networks through the evaluation of principal technologies required in multi-area routing architecture. Since the IETF specification recommends “not” to advertise link state information to outside areas for the sake of routing scalability, the Area Border (ABR) OXCs in Fig. 1 located in the sub-area border should perform “per-area hop route calculation.” Namely, the ABR-OXCs should calculate and determine the LSP route within the sub-areas to which the ABR-OXCs belong [7].

Regarding the “per-area hop route calculation,” we successfully confirmed the insertion of an Explicit Route Object (ERO) into the Resource ReSerVation (RSVP-TE) messages to assign the route within the sub-routing area at each ABR-OXC. Table II shows a list of the successful LSP creation scenarios and the round trip time for a two-way RSVP-TE signaling message.

Another issue is the optical LSP routing across multi-rate links. The ITU-T G.709 based Optical Transport Network (OTN) architecture provides the capability to transport various rate client signals [8]. We evaluated the capability of the routing protocol and path computation elements to route optical LSPs over GbE/STM-16 hybrid links as shown in Fig. 1. The trial did not yield successful results and this remains as a real issue even in the current OTN [9]. More detailed implementation agreements such as the advertisement of wavelength channel status information and transportable signal rate are required to adapt the GMPLS control plane even to the transparent optical networks.

II. FIELD TESTS OF MULTI-AREA (G)MPLS NETWORK

Many network engineers contributed to the standardization activities in the Internet Engineering Task Force (IETF), and have performed a number of MPLS/GMPLS interoperability trials [4,5]. Thanks to these efforts, the main technical target of the ASON/GMPLS technology has progressed toward further realistic deployment issues such as the ASON/GMPLS control over intra/inter-carrier multiplex routing domains and GMPLS network deployment in existing IP/MPLS networks.

The latest inter-operability test on the GMPLS technologies reports on intra-carrier multi-area MPLS/GMPLS ROADM/OXC hybrid photonic networks evaluated using the network elements given in Table I [6] at Tokyo and Isocore in Washington D.C. This trial was a test to ensure the routing scalability of the GMPLS controlled networks through the evaluation of principal technologies required in multi-area routing architecture. Since the IETF specification recommends “not” to advertise link state information to outside areas for the sake of routing scalability, the Area Border (ABR) OXCs in Fig. 1 located in the sub-area border should perform “per-area hop route calculation.” Namely, the ABR-OXCs should calculate and determine the LSP route within the sub-areas to which the ABR-OXCs belong [7].

Regarding the “per-area hop route calculation,” we successfully confirmed the insertion of an Explicit Route Object (ERO) into the Resource ReSerVation (RSVP-TE) messages to assign the route within the sub-routing area at each ABR-OXC. Table II shows a list of the successful LSP creation scenarios and the round trip time for a two-way RSVP-TE signaling message.

Another issue is the optical LSP routing across multi-rate links. The ITU-T G.709 based Optical Transport Network (OTN) architecture provides the capability to transport various rate client signals [8]. We evaluated the capability of the routing protocol and path computation elements to route optical LSPs over GbE/STM-16 hybrid links as shown in Fig. 1. The trial did not yield successful results and this remains as a real issue even in the current OTN [9]. More detailed implementation agreements such as the advertisement of wavelength channel status information and transportable signal rate are required to adapt the GMPLS control plane even to the transparent optical networks.

Table I. List of Evaluated Network Elements

<table>
<thead>
<tr>
<th>Vendor</th>
<th>Equipment Type</th>
</tr>
</thead>
<tbody>
<tr>
<td>A,D</td>
<td>IP/MPLS routers</td>
</tr>
<tr>
<td>B, C</td>
<td>IP/MPLS/GMPLS routers</td>
</tr>
<tr>
<td>E, F</td>
<td>IP/MPLS/GMPLS testers</td>
</tr>
<tr>
<td>G, H</td>
<td>TDM-XC(STM-16c/OC-48c XC)</td>
</tr>
<tr>
<td>I, J, K, L</td>
<td>OXC</td>
</tr>
<tr>
<td>M, N</td>
<td>ROADM</td>
</tr>
</tbody>
</table>

Table II. List of Successful Scenarios

<table>
<thead>
<tr>
<th>LSP Route</th>
<th>RTT (msec)</th>
</tr>
</thead>
<tbody>
<tr>
<td>B1-L1-K-L2-I-B2</td>
<td>694</td>
</tr>
<tr>
<td>B1-N1-N2-L1-L3-I-B2</td>
<td>5,216</td>
</tr>
<tr>
<td>B1-L1-L2-I-B2</td>
<td>583</td>
</tr>
</tbody>
</table>
III. OPERATIONAL EVALUATION IN JGN II TESTBED

Figure 2 shows the JGN II ASON/GMPLS network testbed comprising two administrative domains connected by 10-Gbit/s SONET/SDH links. Each domain consists of different types of GMPLS controlled optical cross-connects (OXCs) and routers, called Type A and Type B. The Type A and Type B OXCs are based on a three-dimensional micro electro-mechanical system (3D-MEMS) optical switch fabric and planar lightwave circuits (PLCs) controlled using the thermal effect, respectively. The domain comprising a Type-B OXC provides an O-UNI interface. Namely, the Tokyo-B-R and Osaka-B-R are located in different administrative domains from the Type-B OXC.

Considering the basic inter-carrier operational environment, independent LSP control in each domain is desired. Furthermore, the support of not only end-to-end LSP recovery operation, but also a domain-to-domain basis recovery operation is desired to prevent affecting failure from one domain to another. The “logically” hierarchical LSP architecture as shown in Fig. 3 is one solution to enhance the independency of LSP control in each domain while ensuring end-to-end LSP control over multiple administrative domains [9].

Figure 4 shows the statistical performance of sub-network restoration for inter-domain LSP over User Network Interfaces (UNIs) between the Tokyo-B-R and Osaka-R based on the “logically” hierarchical LSP architecture. Here, the fiber accommodating the primary optical path was cut at the egress side of the optical path. The average time to recover from the cut fiber to the STM-64 signal level is 432 msec. This is longer than the 200 msec measured in the optical domain evaluation of the 2-hop backup path scenario. Here, the recovery time of the STM-64 signal is measured using the SDH analyzer by monitoring the Alarm Indication Signal (AIS).

IV. CHALLENGING ISSUES AND REQUIREMENTS

Through the evaluation of the ASON/GMPLS, the “logically” stitched LSP architecture provides a novel solution to achieve independent operation of LSPs across multiple domains. However, the scalable routing architecture and effective routing scheme in the layer converged networks is still an open issue. The use of a Path Computation Element (PCE) [11] with inter-layer traffic engineering capability is one choice in such a network. Another issue is the extension of the GMPLS capability to route optical LSPs in transparent networks. The GMPLS control plane should take over the capability to solve not only the Routing and Wavelength Assignment (RWA) problem but also other constraints such as selectivity of links in ROADMs. The dynamic control mechanism of the chromatic dispersion compensation mechanism [12] and the employment of a fast response optical receiver becomes indispensable, if the restoration functionality is allocated to the transparent optical networking layer.

V. CONCLUSIONS

Interoperability and operational evaluation results showed that the ASON/GMPLS technology can be applied to unify the control of networks across multiple domains. The ASON/GMPLS technology, especially the signaling mechanism is now becoming a mature technology toward commercial service deployment. However, the routing mechanism is still requires further development from the viewpoints of architecture and implementation. Continuous research and inter-operability evaluation as well as standardization activities are indispensable toward achieving layer-converged and transparent photonic networks.

REFERENCES

Abstract: This paper gives the evaluation of optical switching, packet loss, died lock and signaling conflict. Based on the evaluation, the capacity is managed to support for Globus Toolkit Version 4, which provide the virtual infrastructure for grid computing.

Keywords: Grid networking, GT4, died lock, conflict.

1. Introduction

2. Evaluation of switching and packet transmission
Does the dynamic bandwidth (e.g. bandwidth on demand) work well? We simulated the BoD service in the computer. The simulation includes an ingress module, a buffering module and an egress module. The ingress module is designed to generate the burst traffic. The buffering module receives the burst traffic, saves such traffic as packets to a large buffer, and then sends these packets to the egress module. The egress module requests and modifies bandwidth for transmitting the burst traffic in the buffer.

Fig.1(a) illustrates the burst traffic generated by the ingress module. Based on such burst traffic the egress module requests and modifies the bandwidth in order to avoid congestion. One important consideration is the choice of control policies for modifying the bandwidth. In our experiment the control policy is based on the buffer size. Fig.1(b) shows the modification of bandwidth and the variety of egress traffic. As the change of bandwidth can’t fully keep up with the variety of burst traffic, the egress module discards packets if the buffer is full. Fig.1(c) shows the packet loss and Fig. 1(d) shows the variety of buffer size.

We can see that the dynamic bandwidth can’t decrease the rate of packet loss.

Does bad control policy cause more packet loss? We also applied two different control policies to control the bandwidth. The first simulates fixed bandwidth over Ethernet lightpaths. The second simulates the request and modification of bandwidth. Figure 2 shows the results, where no matter what models, the packet loss still exists. The Ethernet circuit with 600Mbps of fixed bandwidth causes less total packet loss, while a bad control policy for the modification of bandwidth causes more packet loss and extra expense.

In our opinion, the dynamic bandwidth can’t enhance the total transmission performance, so do optical burst switching.
(OBS). Therefore, we connect computers with grid network using the functionality of self-controlled traffic scheduling or fixed bandwidth rather than dynamic bandwidth.

3. Evaluation of signaling delay and conflict

Does signaling call cause the additional delay? Do two signalings request different capacity and cause the capacity allocation died lock? Do two signalings request the same capacity in different node and cause the conflict?

We measured the signaling delay in a node with various number of requests. The measurement in Fig. 3 shows that the delay per signaling increases when the ingress has more signaling packets to handle. Connection-oriented communication may cause died lock and conflict in the distributed resource allocation using signaling. The test in Fig. 4 shows that when the number of signaling request increases the blocked number (failed number) increase as well.

4. Grid networking support for GT4

No middleware supports the optical transmission directly using optical networks. Moreover, the distributed computing or grid computing can’t easily change their structure and software. In our solution (in Fig. 5), we give and define various service based on the management of optical capacity, TCP (transmission control protocol) and interfaces in order to support several requirements including the bulk data transmission. The service is bound to GT4 infrastructure in order to support the ordinary structure of grid computing.

5. Conclusion

Grid networking is a new communication infrastructure for computational grids. However, Challenges including died lock and conflict will still existed and slow the step to use optical network as ease-use capacity. Evaluation of transmission performance is crucial. The grid networking support for GT4 will enhance the performance of grid computing and the usage of optical capacity.

References

Fault management by prioritized alarm correlation in multi-layer GMPLS networks

Masanori Miyazawa, Kenichi Ogaki, Tomohiro Otani
KDDI R&D Laboratories Inc., 2-1-15 Ohara, Fujimino-shi, Saitama 356-8502, Japan
Tel: +81-49-278-7559, Fax: +81-49-278-7559, Email: ma-miyazawa@kddilabs.jp

Abstract
Fault management by prioritized alarm correlation was proposed in multi-layer GMPLS networks. The prototype of the GMPLS network management system successfully analyzed the root cause of the failure during the GMPLS recovery operation.

1. Introduction
Generalized multi-protocol label switching (GMPLS) technology enables the unified control of network elements in different layers from the wavelength to the packet and is intensively investigated [1]. In order to accelerate the deployment of such network architecture, a GMPLS integrated network management system (NMS) represents one of the big challenges to develop, because the existing network is generally managed and operated through a set of EMS’s (Element Management System) per network element vendors or an NMS per type of network. We have so far proposed and preliminarily developed a prototype of the MPLS/GMPLS multi-layer network management system [2], which can manage inventories and simple alarms, as well as provide a label switched path (LSP) and linkage MPLS and GMPLS LSPs. However, alarm correlation over multiple layers to specify the cause of failure is not considered. Generally speaking, many failures occur over dense wavelength division multiplexing (DWDM) links, and therefore the cause of failures cannot be specified without considering alarms generated from DWDM equipment.

In this paper, the alarm correlation mechanism has been investigated during the occurrence of network failure and recovery operation. We extended the GMPLS network management system so as to include DWDM equipment, which interworks with photonic cross-connect (PXC) in the level of the GMPLS control plane, and which has prioritized all alarms during a certain period. By using these extensions, alarm correlation over multiple layers could be successfully implemented.

2. Fault management by prioritized alarm correlation
The alarm correlation function is the key to comprehending the cause of failure on a timely basis as well as the influence on services because the current network environment, i.e. multi-vendor or multi-layer networks, increase the difficulty to localize the cause of failure more than before. By correlating alarms, we can identify the cause of failure and filter the unnecessary alarms, even across multiple layers. To realize the alarm correlation, the NMS must implement a mechanism to specify the alarm of the root cause from another alarm generated by the same cause within a specific period of time. Fig.1 shows our proposed flowchart of the NMS fault management. When failure occurs and multiple simple network management protocol (SNMP) trap messages are received, these messages are internally transferred to the filtering process to select the specified messages such as WDM trap, link management protocol (LMP)-MIB trap, IF-MIB trap and MPLS-TE MIB trap. These are defined in the filter configuration file. Subsequently, the selected messages are transferred to the alarm correlation process to identify the cause. To narrow down the received message generated by the same cause, the transferred messages are categorized within a specific period of time as well as with respect to the associated network resources. Finally, these classified messages are prioritized in order of data-link, traffic-engineering (TE)-link and LSPs, based on the priority configuration file as shown in Fig. 2. This alarm correlation process selects the series of highest prioritized alarm messages as the root cause of failure, and transfers them to the event process, which summarizes these messages. Consequently, the alarm correlation function is performed and the root cause is displayed.

![Fig.1 Flowchart of the alarm correlation function](image)

![Fig.2 Detailed priority of generated traps](image)
3. WDM link management

In order to identify the cause of failure lying from the lower layer to the packet layer, the management of TE-links related to DWDM equipment is quite significant. In the GMPLS network, such TE-links are generally managed by the standardized LMP-WDM protocol [3] between DWDM and PXC equipment. To date, the NMS has been developed to collect and manage TE-link information through the OSPF-TE database. However, TE-links between DWDM and PXC, hereafter referred as LMP-WDM TE-links, cannot be obtained from such a database, and another mechanism to manage such TE-links is required. Fig. 3 shows the conventional TE-links between two PXCs, and the LMP-WDM TE-links between the PXC and DWDM. The NMS is developed in order to correlate these different TE-links and Data-Links by looking for a TE-link interface identifier (IF-ID) with the same Data-Link IF-ID. Such information of IF-IDs can be retrieved from the teLinkTable object in TE-LINK-STD-MIB [4], the LmpDataLinkTable object in LMP-MIB [5] and the ifStackTable object in IF-MIB [6], respectively. Consequently, the NMS is able to resolve the relationship between the nodes and TE-links, and therefore all the network topology in the MPLS/GMPLS networks can be comprehended and displayed.

4. Demonstration of the multi-layer NMS

To validate our proposed fault management function in the multi-layer network, we evaluated the developed NMS using an MPLS/GMPLS network testbed. Fig. 4 shows the network topology view of the MPLS/GMPLS network testbed. DWDM equipment could be appropriately managed and displayed in addition to PXCs, GMPLS routers and MPLS routers. Two LSPs were provisioned between two pairs of GMPLS routers. A failure to one of the LSPs via PXC1, PXC2 and PXC3 was caused by removing a fiber from the input of one GMPLS router1. Once the failure occurred, the NMS received multiple SNMP traps from each node, which are IF Down traps sent from GMPLS routers, mplsTunnelDown traps sent from GMPLS routers, TeLinkDegraded sent from PXCs, and mplsTunnelDown sent from MPLS routers, respectively. Thanks to the implemented correlation functionality, the NMS successfully resolved the correlation of MPLS and GMPLS networks and indicated the node having detected the failure, while the event view table appropriately indicated the root cause of failure after the alarm correlation of the MPLS and GMPLS networks, as shown in Fig. 5. Furthermore, the NMS visually displayed the restored LSPs with the updated route information collected by SNMP. Thus, we confirmed the feasibility of the alarm correlation function in the multi-layer network by the developed NMS. As a next step, the evaluation of developed NMS is required under actual operational environment for concept proof.

5. Conclusion

We proposed and developed the GMPLS NMS with the fault management by prioritized alarm correlation for multi-layer MPLS/GMPLS networks, and successfully identified the cause of failure and updated the route information of the restored LSP via the MPLS/GMPLS testbed. This concept is broadly applicable and is expected to improve network operation.

6. References

A Study on Optimized Assignment of Dispersion Compensation Capability for Dynamic Optical Paths

Shoichiro Seno¹, Yoshimasa Baba¹, Takashi Mizuochi¹, Takashi Sugihara¹ Kuniaki Motoshima¹ and Tetsuo Ideguchi²
(¹Information Technology R&D Center, Mitsubishi Electric Corporation, 5-1-1 Ofuna, Kamakura, 247-8501 Japan, ²Faculty of Information Science and Technology, Aichi Prefectural University, Nagakute-cho, Aichi, 480-1198 Japan. Tel: +81-467-41-2881, Fax: +81-467-41-2819, Email: Senoo.Shoichiro@de.MitsubishiElectric.co.jp)

Abstract
For dynamic optical paths over all-optical networks requiring dispersion compensation, this paper proposes a new node architecture where compensation devices are shared as well as an extension to GMPLS to advertise compensation capability information for optimized assignment of compensation devices.

1 Introduction
The authors proposed extensions to GMPLS (Generalized Multi-Protocol Label Switching) to offer dynamic control of chromatic dispersion compensation for dynamically established optical paths, and called it GMPLS-Plus (Photonic layer usability support) [1]. GMPLS-Plus extends the three protocols of GMPLS [2], namely LMP, OSPF-TE [3], and RSVP-TE, to automatically discover link chromatic dispersion, to exchange it between nodes for selection of the best route for optical paths taking into account chromatic dispersion, and to optimize compensation upon path establishment. GMPLS-Plus defines additional objects and procedures in the Control Plane to perform these tasks, while electronic pre-distortion [4] with a wide compensation range is to be used to compensate for dispersion in the Data Plane.

In the future, dynamic establishment and release of optical paths over all-optical networks will become common with the aid of tunable electronic pre-compensator. Then it will be necessary for them to support a wide range of compensation capability. For example, Fig. 1 illustrates dispersion ranges of dynamic optical paths between randomly chosen source/destination pairs over the NSF network topology shown in Fig. 2. In this example, 2/3 of links are assumed to have 3 times higher chromatic dispersion than other links and each link can support up to 20 wavelengths. A dynamic optical path may be established over a minimum length route (MinLength) or a minimum dispersion route (MinDispersion) if the same wavelength is available on each link through the route. Traffic demands for optical paths were 20 erl over the whole network.

Fig. 1 tells us that strong compensation capability is required at each node to fully compensate for dispersion that may be accumulated over an optical path destined to an arbitrary node. Yet such strong capability is necessary for a small percentage of all paths, e.g., about 5% of all paths will require compensation capability stronger than twice the average dispersion when each link supports 4 wavelengths. This percentage will be larger if available wavelengths become fewer or traffic demands grow.

![Fig. 1 Path dispersion range over the NSF network](image)

**Fig. 1** Path dispersion range over the NSF network

(a) NSF network  
(b) lattice network

**Fig. 2** Network Models

Considering that there are many compensation technologies and device cost may differ corresponding to different compensation capability, we arrived at a new optical node architecture where compensation devices, such as electronic pre-compensator, with different capability are shared by dynamic optical paths so that assignment of compensation capability will be optimized. In the following, we propose the new architecture as well as an extension of GMPLS, in which each node’s compensation capability is advertised for optimized assignment of compensation devices upon route selection.

2 New optical node architecture with shared compensation devices
We propose a new optical node architecture shown in Fig. 3, where compensation devices with different capability are shared by dynamic optical paths. In this
3 Advertisement of compensation capability information and its effects on routing performance

To achieve optimized assignment of compensation devices with different capability upon path establishment, up-to-date availability of compensation devices at a path’s ingress node as well as at its egress node shall be obtained by the route selection process. Thus we propose to advertise each node’s compensation capability information by an extension to OSPF-TE for GMPLS.

Compensation capability information may consist of the number of available compensation devices and their compensation ranges. Its advertisement will take some abstracted form of such information because it should not depend on a particular compensation technology.

To evaluate effects of advertisement of compensation capability on routing performance, we performed simulation on the two network models, NSF network and lattice network, shown in Fig. 1. In the simulation, each node is assumed to have the same number of two types of compensation devices, Type_A and Type_B. Type_A devices could compensate for dispersion exceeding the threshold, i.e., 75% over the average path dispersion, while Type_B device could not.

Fig. 4 summarizes simulation results comparing compensation capability-aware (Optimized) routing and unaware (Conventional) routing over the NSF network and the lattice network with varied wavelengths, with respect to failure ratio due to inappropriate assignment of Type_B devices to high dispersion paths. Optimized routing assigns Type_A devices prioritized over Type_B devices to paths exceeding the threshold, whereas Conventional routing assigned compensation devices based upon availability. Fig. 4 shows Optimized routing resulted in fewer failures, which demonstrates effects of advertisement of compensation capability.

4 Conclusion

For optimized assignment of compensation devices such as electronic pre-compensator with different capability at optical nodes, we proposed a new optical node architecture where compensation devices can be shared by dynamic optical paths, as well as an extension to GMPLS to advertise each node’s compensation capability. It was confirmed that compensation device assignment based upon advertised capability information would result in better routing performance by simulation.

Acknowledgement

This work was in part supported by a project of National Institute of Information and Communications Technology (NICT), as part of a program of the Ministry of Public Management, Home Affairs, Posts and Telecommunications of Japan.

References

Title: Verizon Optical Network – Strategic Vision

William C. Uliasz  
Director – Access and Transport Network Architecture  
Verizon Technology Organization (VTO)  
40 Sylvan Road  
Waltham, MA USA 02451  
Phone: 781-466-2404 Fax: 781-466-2319  
email: william.c.uliasz@verizon.com

Abstract:  
Through network introduction of emerging optical technologies Verizon continuously enhances our end-to-end optical network. Spanning across Access, Metro and Long Haul domains, leading edge deployments include: xPON, ROADMs, WSSs, GMPLS, and 40G+ wavelength support.

Introduction:  
Continuous advances in optical components and systems are being disclosed on an almost daily basis, promising to reduce costs, gain efficiencies and support advanced features and services.

Verizon is focused on transforming its network to meet emerging bandwidth requirements and applications. Use of next generation optical components, sub-systems and control plane enabled network elements is key to meeting such requirements. With strong emphasis on standards based interfaces.

Our vision is to build a fully automated end-to-end all optical network capable of supporting advanced services for Residential, Enterprise, Government, and Wholesale applications. Support for IP and TDM traffic must be done efficiently and transparently. To achieve this, advanced optical technologies such as Next Generation (NG) PON, ROADMs, WSSs, 40G/100G line rates, and others must meet the challenge of driving down network costs while being easy to deploy for rapid network scaling.

Presentation:  
Presentation will cover how strategic vision and target architectures are being realized in overall network deployments.

In the ULH domain, supporting Global and National backbone network demands, a photonic optical mesh network is a fundamental necessity. These DWDM based networks must be capable of supporting multiple Tbps of traffic, 100+ wavelengths, optical reaches with out regeneration often exceeding 3,000 Km for national and 12,000 Km for submarine applications, and scaling of client interfaces from 10G to 40G and beyond (e.g. 100G).

Reconfigurable Optical Add Drop Multiplexers (ROADMs), equipped with Wavelength Selective Switches (WSS), tunable optics, multi-rate client ports, and SONET ADM on a “blade” or “wavelength” have allowed for efficient system installation and network scaling. The combined impact of these optical component advancements has reduced build costs by over 50%, when compared to previous methods. Hybrid packet and TDM fabrics will further reduce network costs and allow for better integration and convergence of traditional network boundaries.

SONET mesh enabled with GMPLS Optical Control Plane will support the Metro/Core and ULH domains offering advanced new services and network protection/restoration scenarios. Service activation will take minutes for fractional and full rate GigE and OCn
services to be provisioned, and network faults will remain transparent but offer improved network robustness as routing alternatives will increase.

OTN G.709 promises better management of services across domains, and improves the ability to offer higher rate transparent services to our largest Enterprise, Government, and Wholesale customers.

Deployment of PON technology in the access domain has extended the last leg of the optical network to the mass market residential and small business customers on a massive scale. Evolving from B-PON to G-PON is already underway, and advancements will continue as the components evolve to support NG PON systems such as 10G, WDM, and hybrid PON.

Handling the growing expansion of fiber terminations and connections has created an opportunity for mechanizing the ‘tried and true’ manual optical patch cord and patch panel method of operation. MEMS, Beam Steering and Robotic switches are now available or nearing availability to automate fiber terminations and connections. Several critical factors still need to be addressed for large scale acceptance to occur. Depending upon the application, issues include: cost per connection, scalability of device, connection performance, and stability of connection during power failures. Moving to these devices will allow for: automated provisioning, better performance monitoring, centralized testing, physical layer network grooming, and additional fault/restoration techniques.

Conclusion:
As new services and applications drive the need for ever increasing transport capacity in the optical network, continuous scaling must occur by rapidly integrating new technology that support standards based interfaces. The transport network must also transition from an OSS (operational support system) controlled network to a dynamic control plane enabled network. A self aware network capable of recognizing topology, capacity, and elements to establish service routes, bandwidth allocations, and protection schemes to meet rigorous yet changing customer demands. In doing so, the end-to-end optical network must remain: accessible, predictable, scalable, reliable and survivable.

Adherence to open standards, with cooperation through the system vendor development cycles to reach common interpretation and implementation of industry standards will help improve time to market of new technology and enhance the customer experience.

About the author:

William C. Uliasz received his B.S.E.E. degree from Northeastern University, Boston, MA in 1987. He has held various technical assignments focused on the access and transport network during his 19 years at Verizon. He is responsible for setting the target architecture for Verizon Telecom a Director in the Verizon Technology Organization (VTO).
A Field Trial of On-Demand Optical Grid Lightpath Network Services

Sugang Xu and Hiroaki Harai
National Institute of Information and Communications Technology
4-2-1, Nukui-Kitamachi, Koganei, Tokyo, 184-8795 Japan
TEL: +81-42-327-6927, FAX: +81-42-327-6680; Email: {xsg,harai}@nict.go.jp

Abstract—This paper summarizes our field trial of on-demand optical grid lightpath network service for e-Science applications. The end-users can take advantage of the flexibility on network control provided by our GMPLS based optical grid network infrastructure. To support end-to-end lightpaths, we propose a RSVP-TE extension for bidirectional lightpath with the same wavelength in both directions.

1. Introduction

Scientific research communities need very-long distance communication networks because their observation and computing facilities are distributed in world-wide range. For example, e-VLBI (Very Long Baseline Interferometry) [1] as one effective observation approach has been employed on the most frontiers of the research in Astronomy, Geodesy, and Spacecraft Navigation. During observation 256M-1Gbps (60TB per day) of data is generated at each observation site. Three essential requirements of e-VLBI (long baseline for high angular resolution, large bandwidth for high sensitivity, and real-time data transfer for fast/rapid turnaround) make a WDM based lightpath network the ideal candidate for high-quality data transmission.

GMPLS (Generalized Multi-Protocol Label Switching) suite of protocols allows carriers to automate the provisioning and management of the network, and promises to lower the cost of operation by several orders of magnitude compared with manual operation. More automation is helpful to the scientific communities. We provide GMPLS’s user interfaces that drive GMPLS function for supporting application users.

In this paper, we address our developed WDM based lightpath network called optical grid infrastructure. We also propose a new bidirectional lightpath setup method.

2. Optical Grid Network

Figure 1 depicts the network topology that we installed. This is an optical grid infrastructure in which multiple lightpaths can be setup by a user’s (i.e., end-host PC) request [2]. Two pairs of optical fibers between Koganei and Otemachi (about 50km), Otemachi and Akihabara (about 5km) have been employed. All sites are located in Tokyo. The fibers are JGN2 [3] service. Three PCs were distributed at these three sites to fulfill specific application tasks. Two L2 switches were placed at Koganei and Otemachi, respectively, for switching data outside optical grid infrastructure. End hosts or L2 switches have multiple MCs (media converters) equipped with DWDM wavelength and 1000Base-T SFP modules. These MCs were used as the fixed wavelength receivers. The wavelength is any of 1548.5, 1549.3, 1550.1, 1550.9 nm. Optical switches (i.e., no O/E/O) pass lightwave in wavelength granularity. AWGs were employed for wavelength multiplexing and de-multiplexing only on optical fibers which connect two sites.

The dashed lines in Fig. 1 show the Control-Plane of the network. It was constructed with a VLAN established via JGN2 VLAN service, which was separated from the DWDM circuits. In fact, besides VLAN, we can construct the Control-Plane with any other feasible approaches from private infrastructure to the Internet. Using this Control-Plane, a lightpath network, which is a set of multiple lightpaths, was established among three sites, say, Koganei, Otemachi and Akihabara. We should note that to simplify our system, end host PCs for application use were running as GMPLS edge nodes as well. The wavelength lightpaths were connected to network interface cards (NICs) of these PCs directly to attain the full bandwidth of the lightpaths for performance requirement of the application.

We have used such infrastructure for e-VLBI application. Before the data transmission, when a demand of lightpath network comes, at a proper network node, this network request was decomposed into multiple bidirectional lightpath creation requests starting at different source end/edge nodes. Then these bidirectional lightpaths were established automatically with a path message extension of RSVP-TE [4]. Totally, there are six bidirectional lightpaths were established on four different wavelengths (1548.5, 1549.3, 1550.1, 1550.9 nm). Three lightpaths started from Koganei L2 switch, with each to one end-host PC for correlation calculation purpose, another three were started from Otemachi L2 switch, terminated at each PC. After setup of the lightpaths, datum from Haystack and Kashima observation sites were transferred to the three PCs for parallel correlation computation via Otemachi and Koganei L2 switches.
3. Bidirectional Lightpath on Same Wavelength

As we mainly employed commercial product Ethernet NICs (1000Base-T), which need to work in full-duplex mode, we have to establish bidirectional lightpaths for node pairs who have data to exchange. Moreover, because we hoped to use relatively cost-effective fixed wavelength MCs as the fixed wavelength transceiver arrays, each end-host NIC is fixed to one MC, and DWDM SFP module of each MC is physically connected to one wavelength channel/port of AWG or OXC, it implies that both transmitting and receiving of a MC work on the same wavelength. Thus, we also need that the bidirectional lightpath uses the same wavelength on both directions, say upstream and downstream directions. Although drafts for Upstream Label [4] even Upstream Label Set [5] have been proposed, and they do deal with the general cases which need bidirectional LSPs, there is not any definition for bidirectional lightpath on the same wavelength.

![Diagram of bidirectional lightpath setup]

Fig. 3. Support for bidirectional lightpath on the same wavelength:
(a) Path Message, and (b) Bidirectional Lightpath Flag Object

Considering our system configuration, we need to add a new function into RSVP-TE to support bidirectional lightpath with same wavelength on both directions. Our lightpath setup procedure is briefly described as below:

1) Ingress node adds new lightpath type indication in Path message. It is propagated in the Path message in the same way as a Label Set for downstream;

2) Upon receiving Path message containing both such indication and Label Set, the receiver checks the local LSP database to see if the Label Set TLVs (type-length-value) are acceptable on both directions jointly. If there are acceptable labels, then copy the values of them in to new Label Set TLVs, and forward the Path Message to the downstream node. Otherwise the Path message will be terminated, and a PathErr message with a “Routing problem/Label Set” indication will be generated;

3) Upon receiving Label Set combined with new lightpath type indication, egress node verifies whether the Label Set TLVs are acceptable, if at least one label is available on both directions, then the first available label is selected. A Resv message is generated and propagated to upstream node;

4) When a Resv message is received at an intermediate node, the intermediate node allocates the label to interfaces on both directions and update internal database, then configures the local OXC for lightpaths of both directions.

The descriptions of other procedures are omitted due to space limitation. Regarding the new lightpath type indication in Path message, several options might be considered. For example, we can make use of the reserved bits in Label Set, or Upstream Label/ Upstream Label Set. However, as we consider our future possible RSVP-TE extensions which might also need new indications, without changing other formats, we newly define a multiple-purposes object. It will be used to carry different Flags, for example, the Bidirectional Lightpath Flag Object for our purpose is depicted in Fig.3.

4. Demonstration

To verify the possibility of new on-demand services that support the emerging large-scale distributed computing applications, we have implemented a GMPLS based lightpath control system on a DWDM optical network. Moreover, for special users who might require dynamical underling network reconfiguration according to their data transmission needs, it is necessary to provide them the capability and flexibility of fast lightpath setup/release directly. Thus we created GMPLS’s user interfaces that drive GMPLS function for supporting application users. Figure 4 is a snapshot of a RSVP-TE console showing the users issued unidirectional and bidirectional lightpaths, where the bidirectional lightpath employs the same wavelength on both directions.

![Snapshot of RSVP-TE console]

Fig. 4. A snapshot of RSVP-TE console showing a “Bdlp” field (bidirectional and unidirectional lightpath provisioning).

5. Conclusion

To provide on-demand high-performance networks service, in the field trial, we have successfully established a GMPLS based DWDM optical grid network infrastructure, which supports bidirectional lightpath with the same wavelength in both directions. The end-users can take advantage of the flexibility on network control.

Reference:
[1] e-VLBI at NICT (http://www2.nict.go.jp/w/w114/itsi/e/research.html)
Demonstration of a GMPLS-controlled Transparent Optical Network with Wavelength Continuity Constraint
Hongxiang Guo, Takehiro Tsuritani, Noboru Yoshikane, Tomohiro Otani
KDDI R&D Laboratories Inc., 2-1-15 Ohara, Fujimino-shi, Saitama 356-8502, Japan
Tel: +81-49-278-7864, Fax: +81-49-278-7811, Email: ho-guo@kddilabs.jp

Abstract The applicability of a GMPLS extension supporting global wavelength information was investigated to efficiently establish an end-to-end light path. A GMPLS-controlled multi-ring transparent optical network was successfully demonstrated considering wavelength continuity and a multi-vendor environment.

Introduction
By eliminating expensive O/E/O conversion in intermediate optical nodes, an optical transparent network can simplify architecture and reduce capital expenditure as compared with an existing opaque network. Furthermore, its bit-rate and data-format free features enable the optical transparent network not only to offer higher and flexible transparent bandwidth to the client, but also to be easily upgraded to meet future unpredictable bandwidth demands if required. On the other hand, the introduction of a Generalized Multi-Protocol Label Switching (GMPLS) control plane [1] into the optical transparent network is expected to bring more intelligence and to control the end-to-end light path in a cost-effective manner.

However, in order to provision a light path with acceptable signal integrity in such a transparent network, some physical parameters need to be fully considered. One of them is the wavelength continuity constraint due to the lack of wavelength converters, which may result in a high blocking probability and inefficient utilization of wavelength channels if dealt with improperly. Therefore, it is necessary to advertise wavelength availability information and consider it in the Constrained-Based Shortest Path First (CSPF) calculation. In the current Open Shortest Path First with Traffic Engineering extensions (OSPF-TE) [2] definition, the Link Color attribute can be used to carry the wavelength availability information [3]. Only 32 bits are available at maximum in the Link Color field, which is not enough to cover all wavelengths in the current DWDM channels with more than 80. Although the introduction of a dedicated sub-Type Length Value (TLV) for the top-level TE-link TLV in the OSPF-TE type 10 opaque Link Status Advertisement (LSA) was proposed [4] in order to provide a wavelength mask with a variable length, actually the single wavelength bit-mask only has local meaning and could not precisely identify the specific DWDM wavelength. This will result in the degradation of network scalability or even fail to interwork in a multi-vendor environment unless the same bit-mask is commonly configured at all nodes.

As for Resource Reservation Protocol with Traffic Engineering extensions (RSVP-TE) [5], the current solution only provides a Label Set object to restrict labels equivalent to wavelengths, which could be used on the downstream side. When precise wavelength availability information is not available, however, a loose hop expansion scheme has to be employed, which may require crankback signaling and inevitably degrade the performance. Therefore, wavelength information with global meaning in the whole network is indispensable [6].

This paper presents experimental results for the applicability of a GMPLS extension to define a standardizing wavelength label to advertise wavelength availability information with global meaning. A GMPLS-controlled multi-ring network was successfully demonstrated considering wavelength continuity and a multi-vendor environment.

Implemented GMPLS extensions supporting global wavelength information
In order to support the global wavelength information, a definition of wavelength, which we are jointly proposing to standardize [6], as shown in Fig. 1, was employed but modified to support a 200 GHz spacing. Based on the ITU-T DWDM grid, a central frequency of 193.0 THz was selected and all wavelengths were calculated as:

\[ f (\text{THz}) = 193.0 \text{THz} + n \times CS \text{THz} \]

where ‘CS’ is set to 1, 2, 3, 4 or 5 to indicate the channel spacing of 12.5, 25, 50, 100 or 200 GHz respectively; ‘n’ represents the positive (‘s’=0) or negative (‘s’=1) offset to the central frequency.

Therefore, as shown in Fig. 1 (a), when the smallest wavelength is specified in the sub-TLV for the TE-link, each bit in the wavelength mask field can identify the availability of a specific wavelength. On the other hand, when this standardizing label with global meaning is applied in RSVP-TE as shown in Fig. 1 (b), a Label Switch Path (LSP) with an explicitly specified wavelength can be provisioned.

![Fig.1 GMPLS extension supporting wavelength information](image1)

Experimental results
As shown in Fig. 2, a GMPLS-controlled multi-ring transparent network was constructed with two Reconfigurable Optical Add-Drop Multiplexer (ROADM) nodes and one Wavelength Cross Connect (WXC) node, which were based on Wavelength Selective Switches...
(WSS) with 100 GHz spacing. Fig. 3 shows the architecture of the ROADM and WXC nodes. In order to investigate the applicability of standardizing wavelength labels in a multi-vendor environment, two Photonic Cross Connect (PXC) nodes integrated with DWDM MUX/DEMUX optical filters with 200 GHz spacing were intentionally inserted into the ring. Furthermore, core routers were utilized at the ingress and the egress and equipped with a 10 Gigabit Ethernet (GbE)-based colored interface of Optical Transport Network (OTN) framing and a C-band full tunable laser with a 50 GHz step [3]. For simplicity, only two wavelengths (λ1=1551.72 nm / 193.2 THz, λ2=1550.12 nm / 193.4 THz) were supposed as the available TE link resource along rings.

In the experiment, transparent TE links in the ring were advertised as having a switching type of Lambda Switching Capability (LSC) and an encoding type of Lambda. Considering the current limitation of optical transmission, the bandwidth per wavelength of TE links was set to 40 Gbit/s. Such a transparent link can be selected as a transit link for a LSP whose required bandwidth is less than 40 Gbit/s [7]. Wavelength availability information as shown in Fig. 4 was also observed in OSTF-TE flooding messages. Due to the different channel spacing at PXC1, ROADM/WXCs and Routers, the smallest wavelength, which was set to λc (193.0 THz) at all nodes for simplicity, was encoded as 0x34000000, 0x30000000 and 0x2C000000 respectively. In terms of the wavelength mask, PXC nodes set it to ‘011’, where the 2nd and 3rd nonzero bits indicated the two supposed wavelengths λ1 (0x34010000) and λ2 (0x34020000) were available. On the other hand, the bitmask was set to ‘00101’ corresponding to λ1 (0x30020000) and λ2 (0x30040000) at ROADM/WXCs. The all-one bit-mask was used at routers to indicate all C-band wavelengths were supported by the tunable laser.

After these sets of advertised wavelength information were successfully exchanged and injected into the Traffic Engineering (TE)-Database, an enhanced CSPF calculation with consideration of wavelength continuity constraint could be conducted. When a LSP with 10 GbE bandwidth from R1 to R2 was initiated, the ingress link and egress link with 10 GbE bandwidth were firstly selected, and then a series of transit links with wavelength continuity constraint were selected along the shortest path of ROADM1-WXC-ROADM2, and the wavelength would also be determined. Since both λ1 and λ2 were available, the first-fit wavelength λ1 was selected by default. Then, the laser in R1 was controlled to be tuned to λ1, and simultaneously a PATH message with Label Set of λ1 was sent out. All intermediate nodes parsed the label in the received Label Set object based on the local wavelength configuration, and then forwarded it to the downstream side. After the PATH message arrived at the egress router R2, it acknowledged the PATH creation request by sending back a RESV message. R2 also controlled the laser to tune to λ1 since the GMPLS LSP is bidirectional by default. When λ1 in the TE link between ROADM1 and WXC and λ2 in the TE link between WXC and ROADM2 were intentionally occupied and the aforementioned LSP was re-initiated, the previous shortest path was pruned out due to the wavelength continuity constraint and the transit links were selected along the second shortest path via ROADM1, PXC1, WXC and ROADM2.

Finally, the setup latency of LSPs was evaluated. In the case of no tuning, the latency to establish the second route was about 300 ms on the control plane and 350 ms on the data plane. The difference resulted from the switch setup latency, but it was not so large. On the other hand, in the case of tuning to a new wavelength, the final bidirectional path was established up to 4 seconds, as shown in Fig. 5. For faster path establishment, it is indispensable to speed up tuning the wavelength.

Conclusions
We have investigated and implemented a GMPLS extension in support of global wavelength information so that wavelength availability information with global meaning could be advertised by OSPF-TE, and generalized label could be used to control optical nodes by RSVP-TE. By using this extension, an end-to-end light path with consideration of wavelength continuity constraint was successfully established even in a transparent optical network consisting of all-optical switching nodes with different wavelength spacing. We have confidence that this extension is quite beneficial for further improving the network control and management in future transparent optical networks and is expected to be standardized.

Acknowledgement
The authors thank Drs. M. Usami, M. Suzuki, and S. Akiba of KDDI R&D Labs for their continued encouragement.

References
An Effective Light-path Setup Scheme for Dynamic Traffic in Multi-ring Wavelength-routed networks

Mirang PARK†, Yoshimasa BABA†, Shoichiro SENO† and Naonobu OKAZAKI††

†Information Technology R&D Center, Mitsubishi Electric Corporation, 5-1-1 Ofuna, Kamakura, 247-8501 Japan
††Faculty of Engineering, University of Miyazaki, Miyazaki, 889-2192 Japan
Tel: +81-467-41-2430, Fax: +81-467-41-2419, Email: Park.Mirang@dn.MitsubishiElectric.co.jp

Abstract

A simple and fast RMT (Ring Management Token) wavelength reservation method for dynamic traffic in multi-ring wavelength-routed networks is proposed. The RMT is an agent that can control and manage the used wavelength in the network efficiently.

1. Introduction

With the rapid emergence of many multimedia applications required a high-speed rate of data transfer and a high reliability of the transmission channels, a photonic network based on Wavelength Division Multiplexing (WDM) technology attracts attention now. In WDM networks, a light-path (LP) must be established between a pair of source and destination nodes before the data can be transferred. The LP-based optical WDM networks are generally called wavelength-routed networks [1]. Two LPs can use the same link, if and only if they use different wavelengths. In the absence of wavelength converters, an LP would occupy the same wavelength on all fiber links that it traverses, known as wavelength continuity constraints [2].

In this paper, we discuss an LP setup method of the multi-ring WDM networks, in which the wavelength converters are not needed. We aim to improve the LP setup time, and propose an efficient wavelength reservation procedure (WRP) to satisfy the wavelength continuity constraint of an optical path. Now, the procedure for satisfy with the wavelength continuity between source node and destination node is provided by RSVP-TE (Resource Reservation Protocol with Traffic Engineering) [3]. However, there is a problem that the LP setup time becomes long when failure occurs in the wavelength reservation process. In this paper, we propose a simple and fast wavelength reservation procedure, named RMT (Ring Management Token) reservation method, for dynamic multi-ring wavelength-routed networks. We show that the LP setup delay is shortened and the processing time of the signal traffic messages are reduced.

2. WRP of Multi-Ring WDM Networks

2.1. Network Systems Model

In this paper, the multi-ring network system model is targeted, in which two or more WDM ring networks are connected by the ring connection node. Each ring network is composed of photonic nodes, that is, Optical Add Drop Multiplexer (OADM) or Optical Cross Connect (OXC). We assume that wavelength converters are not available in this system. We consider an LP setup method that satisfies the wavelength continuity in the shortest route between two nodes.

2.2. RSVP-TE Light-path Setup Procedure

As shown in Figure 1, the principal signaling operations of RSVP-TE are described as follows. When a connection request arrives, the source node (S) sends an LP request message (Path) along the selected route to the destination (D). At the source node, all available wavelengths are collected as candidates simultaneously. When an intermediate node (I) receives the "Path", it checks the status of these candidate wavelengths on the outgoing link, and only collects the subset of these candidate wavelengths (W) which are currently available. This new set of candidate wavelengths information is carried by the Path to the next node. If the Path arrives at the destination with a non-empty set of candidate wavelengths, the destination will select one wavelength to be used for the connection and send a reserve message (Resv) back to the source node. The data transmission can begin after the source receives the Resv. However, there is a problem that it takes time to select a wavelength along with it when the number of intermediate nodes increases. In addition, it holds the following problems of blocking for the reserved wavelength when the dynamic traffic concentrates on a certain specific link.

(i) Forward Blocking: This is caused when the common wavelength doesn't exist, in the process of transmitting the "Path" message. The intermediate node sends the "PathErr" to the source node.

(ii) Backward Blocking: This is caused when the wavelength selected by the destination node doesn't exist. The intermediate node sends the "ResvErr" to the
destination node.

3. RMT Light-path Setup Scheme

To solve the above problems, we introduce “Ring Management Token (RMT)” that can control and manage the usage of wavelength in the network efficiently while it goes round the ring network. RMT always goes to one direction of the round in the ring networks in a constant order. It is caught by the source node \(N\) of an LP, and used to reserve wavelength in which the wavelength continuity constraint is satisfied. Our proposed RMT LP setup scheme of the multi-ring WDM network can be classified into the cases when:

(i) one RMT exists in whole multi-ring network, or
(ii) one RMT exists in each ring network.

In case (ii), the wavelength that can be used in a ring and the wavelength that can be used between rings are able to be managed separately. The detailed procedures of case (i) are described as follows.

[Wavelength Reservation Procedure by the RMT]

(1) When a connection request arrives, the source node finds a pre-assigned explicit route to the destination. Then it gets the RMT that goes round own ring, to determine which wavelengths are currently available, and checks on empty wavelengths that is based on wavelength usage information in RMT.

(2) The source node selects and reserves a wavelength \(\lambda_i\), and then "Resv Req" message is transmitted along the route.

(3) Each intermediate node preserves notified wavelength \(\lambda_i\) in the part of wavelength usage information on the resource management table.

(4) When the resource reservation to the destination node completes, "Resv Reply" is sent back to the source to configure the appropriate channel on each link along the path.

(5) The data transmission can begin after the source receives the "Resv Reply".

(6) After the communication is ended, the source gets the RMT again, and deletes wavelength \(\lambda_i\) from the part of keeping wavelength usage information on RMT.

There is no possibility that backward blocking occurs though forward blocking has the possibility of occurring.

4. Evaluations

In this section, we evaluate the LP setup time based on the signal traffic by numerical computations. We assume that only the rate of the backward blocking is considered for the simplification. The notations and values used throughout the estimation are shown in Table 1. The LP setup time by RSVP-TE is as follows.

\[
PT_{\text{RSVP}} = 2nP + 2(n-1)T + 2 \sum_{j=1}^{n-1} (1-\varepsilon) \frac{jT + (j+1)P}{(n-1)P} \cdot (1)
\]

On the other hand, RMT setup time of the proposed method is as equation (2). In this paper, we estimate the LP setup time of ring network by comparing (1) with (2).

\[
PT_{\text{RMT}} = \frac{\sum T_i}{N} + \sum_{i=1}^{n-1} P_i \cdot \frac{\sum T_i}{N} + P_i + 2(n-1)T + 2(n-1)P \cdot (2)
\]

First, \(\varepsilon\) and \(\rho\) could be disregarded in the above expressions, when the arrival rate of the path request in each node is low enough. In this case, both LP setup times depend on the number of nodes. We verified that the setup delay of the proposed method is greatly improved compared with the RSVP method so far as the number of nodes increases. On the other hand, it is thought that the rate of failure goes up, too, as the connection-arrival rate rising. Fig. 2 shows the setup time of PT_{RSVP} and PT_{RMT}. From this, we can see that the time of the proposed method is about one to about 47 to 84 percent of the conventional method.

5. Conclusions

We proposed a simple and fast wavelength reservation protocol based on the RMT for dynamic and large scale wavelength-routed multi-ring networks. Its advantages are shorter connection setup time and processing load reduction of network nodes. This is quite favorable for the burst data transmission and fast restoration applications.

References


The First Optically-Virtual-Concatenated Lambdas Over Multiple Domains in Chicago Metro Area Network Achieved Through Interworking of Network Resource Managers

Yukio Tsukishima¹, Akira Hirano¹, Atsushi Taniguchi¹, Wataru Imajuku¹, Masahiko Jinno¹, Yoshihono Hibino¹, Yoshihiro Takigawa¹, Kazuo Hagimoto¹, Xi Wang², Luc Renambot², Byungil Jeong³, Ratko Jagodic³, Sungwon Nam³, Jason Leigh⁴, Tom DeFanti⁴, Alan Verlo⁵

¹NTT Network Innovation Laboratories, NTT Corporation, 1-1 Hikarinooka 239-0847 Japan
Tel: +81-46-859-8594, Fax: +81-46-859-5541, Email: tsukishima.yukio@lab.ntt.co.jp
²Electronic Visualization Laboratory, University of Illinois at Chicago USA, E-mail: xiwang@uic.edu

Abstract
Optically virtual concatenated parallel lambdas over multiple-domains that are provided using reservation-messaging between network resource managers for a high-end parallel visualization application are experimentally shown for the first time in the Chicago metro area network.

Introduction
High-end visualizing applications [1] are targeting parallelism using tiled displays and cluster computers to ensure unlimited scalability. Inevitably there will be a need for load-balanced parallel lambdas to carry their traffic. Then, these lambdas possibly traverse multiple network domains as needed. On the other hand, such applications are quite sensitive to latency deviations among these lambdas. Therefore, we have proposed Optical Virtual Concatenation (OVC) [2,3] to de-skew these lambdas. There have been no reports of the demonstration of load-balanced & OVC-enabled parallel lambdas over multiple-domains.

A network resource manager (NRM) with the capability to accept reservation requests of lambdas, to compute available resources from a visualization middleware, which is the Scalable Adaptive Graphic Environment (SAGE) [1] developed by the Electronic Visualization Laboratory (EVL). It has the capability to dynamically configure parallel lambdas in collaboration with other network devices.

In this paper, we present the first field trial of the interworking of NRMs and OVC over two photonic network domains against advanced reservation requests for network resources from a visualization middleware, which is the Scalable Adaptive Graphic Environment (SAGE) [1] developed by the Electronic Visualization Laboratory (EVL). Two PXC systems were installed in the Chicago metro area network. We jointly defined and implemented a SOAP-compatible web service interface (WSI) for the NRMs interworking of SAGE and OVC between SAGE-Tx and SAGE-Rx systems to exchange compatible web service interface (WSI) for the NRMs.

Messaging among SAGE, NRMs and OVC
Figure 1 illustrates the experimental configuration. Domain #1 and Domain #2 over I-WIRE are multi-vendor PXC networks. GbE link #1 (GbE #1) traverses through Layer 2-switch #1 (L2SW #1), PXC #1, PXC #2, and Layer 2-switch #2 (L2SW #2). In contrast, GbE link #2 (GbE #2) runs through L2SW #1, PXC #3, PXC #4, and L2SW #2. The two GbE links are bundled via the IEEE 802.3ad Link Aggregation Protocol at L2SW #1 and L2SW #2. Domain #1 is controlled by NRM #1, and Domain #2 is under the control of NRM #2. SAGE [1] comprising SAGE-Tx and SAGE-Rx systems manages visualization and high-definition video streams, enables viewing of images on ultra-high-resolution displays such as the EVL 55-tiled LambdaVision display wall, and functions as an interface between SAGE and the photonic networks. In this experiment, the visualizing processes for rendering frames among SAGE-Tx and SAGE-Rx systems are synchronized, and they require two synchronized parallel GbE lambdas.

Figure 1. Experimental configuration

We developed two different NRM prototypes. NRM #1 developed by EVL consists of a SAGE-proxy, a Photonic Domain Controller (PDC) and a WSI. It receives end-to-end connection requests generated by SAGE, converts them into (multi-)domain lambda reservation requests, and sends these requests to the PDC. The PDC provides and manages lambdas in its local domain (Domain #1). It also collaborates with other NRMs through the WSI for inter-domain lambda provisioning. For local domain requests, the PDC finds available resources from the database, registers them in the reservation table, and configures local PXC systems via TL1 at the scheduled time. For a foreign domain (in this case, Domain #2) requests, it converts the requests into the appropriate format and sends them to the
destination NRM (in this case, NRM #2). On the other hand, NRM #2 developed by NTT consists of the NRM-Engine and a WSI. The NRM-Engine is invoked by a reservation request received from the PDC through the WSI. The NRM-Engine computes an available route according to the reservation request. If it finds an appropriate route, it registers the calculated route in the database and returns the results to the PDC through the WSI. NRM #2 periodically checks the characteristics of each of the allocated GbE links kept in the database, and controls PXCs via TL1 or Telnet when the time comes to activate or deactivate the GbE link. Moreover, NRM #1 and NRM #2 are equipped with the WSI to exchange information regarding network topology and reserved lambdas. Based on WSI, NRM can compute the latency to be added to the OVC capable interfaces.

The reservation message diagram is shown in Fig. 2. Before starting the SAGE streaming application, SAGE sends a reservation request with parameters such as the destination and source site IDs to NRM #1. NRM #1 computes the characteristics of the GbE links that the streaming application needs. Then NRM #1 calculates the appropriate routes and checks the availability. If NRM #1 decides that the GbE link over Domain #2 is needed as well, as in the above example, it sends one more reservation request to NRM #2. NRM #2 executes calculations for the available route and lambda allocation processes in response to the reservation request. At the same time, the NRMs compute the latency for de-skewing based on the exchanged network topology information and set the latency for each OVC capable interface when the start time arrives. When the route satisfying the requirement of the streaming application is found and available, NRM #2 returns an Acknowledgement (Ack) message to NRM #1. NRM #1 and NRM #2 then configure the PXCs and activate the GbE links before the start time arrives. Finally, NRM #1 sends an Ack message to the SAGE application when it confirms that the GbE links have been activated. Only at that time, the SAGE application starts streaming.

When we initiated a TCP-based streaming application on SAGE, SAGE sent reservation requests to NRM #1 with the node-IDs (IP addresses) of the SAGE-Tx and SAGE-Rx systems as parameters. At that time, NRM #1 estimated that the required transmission capacity between SAGE-Tx and SAGE-Rx systems for rendering the streaming application on SAGE would be approximately 1.6 Gbps. Then, NRM #1 determined that two GbE connections, GbE #1 and GbE #2, would be required. NRM #1 allocated GbE #1 and sent a reservation request to NRM #2 to reserve GbE #2. NRM #2 allocated GbE #2 as requested. Following the above allocation process, NRM #1 and NRM #2 configured the PXCs. After NRM #1 confirmed that GbE #1 and GbE #2 were established, NRM #1 sent an Ack message to SAGE. Then, SAGE started the streaming application, and the streaming application was successfully rendered on the six-tiled display. When the latencies of GbE #1 and GbE #2 were zero, the rendering capacity was approximately 1.6 Gbps. During the experiments, the rendering capacity decreased with the increase in the round trip time between the SAGE-Tx and SAGE-Rx systems, and with the slowdown of the synchronous processing between the SAGE-Tx and SAGE-Rx systems caused by the increased round trip time.

To demonstrate the OVC functionality, we intentionally introduced additional latency into GbE #1. The rendering capacity of the SAGE-Rx system decreased with the relative latency between GbE #1 and GbE #2. A part of a frame comprising packets streamed on GbE #2 reached the SAGE-Rx system prior to the other part of the frame comprising packets streamed on GbE #1. To render a frame, the SAGE-Rx system buffered the part of the frame comprising the packets streamed on GbE #1 until the other part of the frame comprising the packets streamed on GbE #2 arrived at the SAGE-Rx system. As a result of the waiting process, the rendering performance of the SAGE-Rx systems decreased to approximately 480 Mbps at the relative latency of 10 milliseconds, which is anticipated when we reserve multiple-lambdas over different domains. In contrast, when the relative latency was adjusted to zero by OVC through the collaboration of NRM #1 and NRM #2, the rendering capacity recovered to 1.2 Gbps, which was an expected value with a 10-ms latency for both GbEs.

**Conclusion**
The SAGE parallel visualization application streamed smoothly on parallel lambdas over multiple domains, configured by OVC and two interworked NRMs. OVC realized by the NRM coordination successfully adjusted the relative latency within a millisecond. The experimental results showed the complete recovery of the SAGE rendering performance deterioration induced by the relative latency deviation.

**References**
1. L. Renambot et al., Proceedings of WACE 2004
2. M. Tomizawa et al., Tech. Digest of OFC2005 OThG6
4. A. Takefusa et al., OGF19, 2007
5. StarLight http://www.startap.net/starlight/
Flexible Optical Access and In-Building Networks

A.M.J. Koonen\textsuperscript{1}, P.J. Urban\textsuperscript{1}, H. Yang\textsuperscript{1}, M. Garcia Larrode\textsuperscript{1}, H. de Waardt\textsuperscript{1}

\textsuperscript{1}COBRA Institute, Eindhoven University of Technology, Den Dolech 2, NL 5612 AZ Eindhoven, The Netherlands,
Tel. +31 40 2474806, Fax +31 40 2455197, E-mail a.m.j.koonen@tue.nl

Abstract
Optical routing in access or in-building networks can improve the flexibility with which services are provided to the end users. It also can increase the system’s operational efficiency by remarkably decreasing the call blocking probability.

1 Introduction
In response to the fast growing need for broadband communication capacity of residential users, Fibre-to-the-Home networks are rolled out at increasing pace in many countries. With respect to a point-to-point access network, the point-to-multipoint PON architecture reduces fibre installation (and repair) efforts, but requires more sophisticated line termination equipment [1, 2]. In TDM-PON schemes, the channel capacity is shared among the users which may lead to call blocking due to congestion problems. WDM-PON architectures offer an individual wavelength channel per user, thus implementing logical point-to-point links on a physical point-to-multipoint topology. In this way, call blocking is avoided, at the expense however of a larger amount of terminal equipment in the local exchange, which is also often not deployed up to its full capacity due to the varying traffic loads from the individual users. In particular when growing to larger numbers of users connected, a PON access network may be operated more efficiently by combining the capacity-sharing virtue of a TDM-PON with the congestion-reduction capabilities of a WDM-PON. The operation efficiency of such a hybrid WDM-TDM PON can in particular be improved by introducing an optical level of flexibility [2]. Using flexible wavelength routing, the TDM capacity can be shared among an adjustable set of users, and thus capacity can be allocated to them according to their actual needs, which improves the utilization ratio of the line terminal equipment in the local exchange. Alternatively, the multiple wavelength channels may each host an independent service provider, and/or an independent set of services. By flexible wavelength routing, new or modified subscriptions to services can be quickly arranged per user.

2 Wavelength-flexible FTTH network

Fig. 1 shows a wavelength-routed FTTH architecture which is investigated in the Dutch Freeband project Broadband Photonics Access [3]. It features wavelength-tunable add-drop nodes in a ring-shaped feeder network, which under remote control from the headend station can drop a wavelength channel to one or multiple homes. These nodes also drop a CW wavelength channel to one or multiple homes, which is modulated in a colourless optical network unit (ONU) in order to send the upstream data. The sharing of these wavelength channels, both for downstream and upstream, is done in TDM; each wavelength channel runs a Gigabit Ethernet PON protocol. Eight wavelength channels are transporting the downstream GbE data, and eight ones the upstream GbE data. Together with a variable 1x2 coupler, the ring-shaped feeder provides protection against a feeder cable break.

Fig. 1 Flexible WDM-TDM Fibre-to-the-Home network

The ONU in each home needs to be wavelength-agnostic, as any of the eight wavelengths may be directed to it. Hence a colourless ONU design is adopted, using a reflective SOA for the upstream data modulation. In order to avoid spurious lasing effects, reflections in the fibre link to the ONU need to be minimized.
3 Flexible radio-over-fibre in-building network

After having reached the doorstep of buildings, the next step for providing broadband capacity to the end users is to introduce optical fibre inside homes, hospitals, office buildings, airport departure lounges, etc. Such in-building network preferably carries both wired (such as GbE) and wireless services (e.g. WLAN). Like in access, also in in-building networks optical routing may improve the flexibility and the efficiency with which the network resources can be used.

Fig. 2 Flexible inter-room wireless communication using switched radio-over-fibre links

The in-building (polymer) optical fibre network concept shown in Fig. 2 is being investigated for providing reconfigurable connectivity between broadband wireless pico-cells confined to rooms. The wireless signals are carried between rooms using radio-over-fibre techniques, without any interference into the other rooms. The connections between rooms are established by an optical crossconnect in the central Home Communication Controller, which switches the analog radio-over-fibre signals transparently. The optical transparency enables the support of multiple wireless standards, such as WiFi, WiMAX, emerging 60 GHz WLANs, etc. Thus by optical routing virtual private wireless LANs can be established between rooms, and reconfigured upon user demand.

4 Impact of optical network reconfiguration

By re-allocating capacity in an access or in-building network in response to changing user demands, the call blocking probability can be remarkably reduced. Fig. 3 shows the system blocking probability versus the relative system load for a WDM-TDM FTTH access network according to Fig. 1 with 8 wavelength channels, each at a capacity of 1.25 Gbit/s, and 256 users who generate Poisson distributed calls with a bandwidth of 63 Mbit/s or 125 Mbit/s. The blocking probability has been calculated using Chernoff’s upper bound approximation. As Fig. 3 clearly shows, the system blocking performance is much better when the wavelength channels are adaptively reconfigured among the users, in response to their actual traffic demands, than when the wavelength channels are assigned to the users in a fixed way. Fig. 3 shows that by deploying reconfigurability the system load can be doubled at a blocking probability of $10^{-3}$ for call bandwidths of 63 Mbit/s, and even more at 125 Mbit/s.

Fig. 3 System blocking probability versus relative system load for static capacity allocation and for reconfigurable capacity allocation

5 Conclusions

Flexible reconfiguration of optical access or in-building networks by means of optical routing enables easy and fast modifications in the service provisioning to the end users. When the reconfiguration can adapt to the actual user traffic demands, the call blocking probability can be remarkably reduced, and thus the efficiency with which the network resources are used can be improved.

Acknowledgement

Partly funding by the Dutch Freeband programme in the BB Photonics Access project is gratefully acknowledged.

References


Effectiveness of mode-selective spatial filtering in mode group diversity multiplexing links

C. P. Tsekrekos and A. M. J. Koonen
COBRA Research Institute, Eindhoven University of Technology
PT 12.25, P.O. Box 513, 5600 MB Eindhoven, The Netherlands, Tel: +31 40 247 5479, E-mail: c.tsekrekos@tue.nl

Abstract—Mode-selective spatial filtering (MSSF) is a new optical method of combating the cross-talk among the channels of a mode group diversity multiplexing link. MSSF can greatly improve the robustness and scalability of the link.

I. Introduction
Light in multimode fibers (MMFs) propagates in a multitude of spatial modes. These modes offer spatial degrees of freedom that can be used in transmission. Although this principle is known, the way of realizing such a transmission scheme is non-trivial. Especially in intensity-modulation direct-detection (IM-DD) links, the utilization of the spatial modes becomes very challenging, since the modes are orthogonal with respect to their field and not their intensity profiles. Multiple-input multiple-output (MIMO) techniques used to increase the spectral efficiency and reliability of wireless communication links can be also applied in IM-DD transmission over MMF [1,2]. In this approach, electrical processing of the received signals is used to recover the transmitted signals, and thus no orthogonality at the optical intensity domain is required.

Mode group diversity multiplexing (MGDM) [2] is an IM-DD MIMO technique that aims at realizing parallel communication channels, transparent to the transmission format. MGDM supports the integration of various broadband services over a common MMF infrastructure [2]. When the link is not limited by dispersion, vector \( s(t) \) of the received electrical signals is related to vector \( s(t) \) of the transmitted electrical signals via an \( M \times N \) real-valued matrix \( H(t) \), i.e. \( s(t) = H(t) s(t) + n(t) \), where \( n(t) \) is a noise vector [3,4]. In silica graded-index (GI) MMFs, a bandwidth of several GHz can be achieved below the dispersion limit [5]. To recover the input signals irrespective of their format, matrix inversion can be used. However, at the same time, this enhances the noise and induces a power penalty in order to maintain the signal-to-noise ratio of the 1×1 case [4]. The more diagonal the matrix \( H(t) \), i.e. the larger the spatial diversity, the smaller the power penalty due to the electronic matrix inversion and the more robust the link. Higher robustness allows the link to be more scalable to the number of channels [4]. We have recently introduced mode-selective spatial filtering (MSSF) as an optical way to combat the cross-talk among MGDM channels and increase the spatial diversity of the link [6]. A 2×2 link was demonstrated [6]. In this paper, we show experimentally the high effectiveness of MSSF in links with a larger number of channels.

II. MSSF principle
A simple way to selectively excite a GI-MMF is to use radially offset Gaussian-like beams at its input face. Each beam excites a different group of modes which yields a distinguishable near-field pattern (NFP) on the output face of the GI-MMF [4]. Given the NFPs, the geometry of the detectors can be chosen to minimize the cross-talk [4]. This approach for an \( N \times N \) link is simple (see references within [4,6] for comparison), however, for \( N \geq 3 \) the link is very sensitive to changes in the elements of \( H(t) \).

MSSF is a new optical technique that improves the condition number of \( H(t) \) [7], and therefore increases the robustness of an MGDM link. Fig. 1 illustrates the MSSF principle. In Fig. 1b, two propagating rays in a GI-MMF are shown. The exit points of the rays on the end face of the GI-MMF fall within the area that would be detected by a single segment \( R_i \) of a multisegment MGDM detector (Fig. 1a). However, the angular divergence (\( \theta \)) of the rays when coming out of the GI-MMF is significantly different. In particular, at the GI-MMF output, ray 1 forms an angle \( \theta_1 \) with the fiber axis smaller than the angle \( \theta_2 \) between the fiber axis and ray 2. Let us assume that a lens is used to project the NFP onto the multisegment MGDM detector. If the numerical aperture (NA) at the object side of the lens (GI-MMF output face), \( N_{\text{AA}} \), corresponds to \( \theta_{\text{AA}} \) such that \( \theta_1 < \theta_{\text{AA}} < \theta_2 \), then only ray 1 is gathered by the lens and subsequently projected onto the detector. In the example of Fig. 1, if \( N_{\text{AA}} \) is sufficiently low, \( R_1 (R_2) \) detects light related to ray 1 (2), while \( R_1 \) does not receive light associated with rays 1 and 2.

In GI-MMFs, the NA has a maximum value on the fiber axis and gradually drops to zero at the core-
cladding interface. This means that MSSF will be effective on the area of the output face of the GI-MMF defined by $NA_{GI-MMF}(r)>NA_{lens}$, where $NA_{GI-MMF}$ is the local NA of the GI-MMF and $r$ the distance from the fiber axis. If $NA_{beam}$ denotes the NA of the radially offset input beams, the following relation will generally hold for $NA_{lens}$: $NA_{beam}<NA_{lens}<NA_{GI-MMF}(0)$. The upper bound expresses the condition for MSSF to take effect, while the lower bound gives a practical rule for GI-MMFs with limited mode mixing. In such GI-MMFs, light launched with a beam of $NA_{beam}$ at radial offset $\rho_0$ will propagate with similar NA around $\rho_0$ and larger NA at points closer to the GI-MMF axis (Fig. 1b). In general, $NA_{lens}$ will decrease as the number of channels increases. However, the lower the $NA_{lens}$, the higher the power penalty due to MSSF (not all the optical power is gathered by the lens). For a certain $N$, the value of $NA_{lens}$ is chosen as a trade-off between the total power penalty and the condition number of $H(\tau)$.

III. Experimental results

In [6], the first demonstration of MSSF was performed with a 2×2 link that exhibited remarkable stability over a 25-h period. Here, we show the effectiveness of MSSF in links with a larger number of channels. Fig. 2 shows our experimental set-up. A 635-nm Fabry-Pérot multi-quantum-well laser diode was used to selectively excite a GI-MMF with core/cladding diameter of 62.5/125 $\mu$m and central NA 0.275. The laser is pigtailed to a single-mode fiber (SMF) with mode field diameter (MFD) 4.2 $\mu$m and NA 0.12. A variable optical attenuator (VOA) with SMF pigtails, of similar MFD and NA with the laser pigtail, was used to control the level of the optical power. A microscope projected the NFP at the GI-MMF output onto a charge-coupled device camera. An image of the NFP was grabbed with video processing software. Two GI-MMFs were tested, of lengths 1 m and 1 km, under excitation with the SMF of the VOA at 0, 10, 15, 21, and 26 $\mu$m radial offset, following the design parameters in [4]. The radial offset of the SMF axis from the GI-MMF axis was set by means of computer-controlled translational stages. Two microscope objectives were used. A 50× one with NA 0.75, capturing all the NFP, and a 5× one with NA 0.10, achieving MSSF, its NA being close to the limit $NA_{beam}=0.12$. The results are shown in Fig. 3. The NFP (50× objective) is confined within a disk which is transformed into a doughnut when the 5× objective is used. It is clear that MSSF can be highly effective and a robust link with five channels could be realized. For a smaller number of channels, MSSF could fully mitigate the cross-talk. Quantitative evaluation of MSSF would require a careful analysis that takes into account both the power penalty due to the electronic processing [4] and the one due to MSSF. The optimal value of $NA_{lens}$ should be also defined for a given $N$. This could be done either with an experimental set-up where the NA can vary over a large range, or with simulations. The low magnification of 5× does not allow for a good estimation of the power penalty due to the electronic matrix inversion, since the obtained images have a low resolution. The results of Fig. 3, though, undoubtedly show the effectiveness and great potential of MSSF.

IV. Conclusion

MSSF is a very simple, easily realizable and highly effective optical technique to increase the robustness and scalability of MGDM links. We have shown that for short wavelengths, a five-channel link is well-supported, while for a smaller number of channels, MSSF could eliminate the need for electronic demultiplexing.

Acknowledgment

Funding from the Freeband Impulse Program of the Ministry of Economic Affairs of the Netherlands is gratefully acknowledged.

References

1-Gbps Link System for Home Network with Plastic Optical Fiber Cable

Takashi Kawakami¹, Hiroki Sueyoshi¹, Teruyuki Taniguchi¹, and Kimio Oguchi²
¹Sekisui Chemical Co., LTD, 2-2 Kamichoshi-cho, Kamitoba, Minami-ku, Kyoto, 601-8105 Japan
Tel: +81-75-662-8522, Fax: +81-75-662-8585, Email: t-kawakami@sekisui.ip
²Graduate School of Engineering, Seikei University, 3-3-1, Kichijoji-Kitamachi, Musashino, 180-8633, Japan

Abstract A 1-Gbps link system based on Plastic Optical Fiber (POF) is developed for the home network. Experiments that examine throughput, bit error rate, and reliability etc., confirm its effectiveness.

1 Introduction
The number of FTTH (Fiber-To-The-Home) users is rapidly increasing, especially in Japan. At the end of September 2006, the number was 7.2 million and will exceed 30 million by 2010 [1]. Through the spread of FTTH, we are going to enjoy high definition (HD) video contents via broadband i.e. VOD (video on demand), IPTV and video download service.

At the same time, our home environments have been changing. Digital broadcasting has already started. Digital audio-visual equipment such as flat-panel TVs and Digital Video Recorders (DVRs) have become common [2]. In the near future, those devices and PCs will be connected to a LAN [3], and we can anticipate enjoying HD video contents anywhere in our house.

Since most homes do not have in-home wiring, it is necessary to promote the infrastructure for the home network. Wireless LANs are not enough to transmit HD video contents reliably throughout the house. Un-shielded Twisted Pair (UTP) cable is so thick that it is highly visible as open wiring, and it readily picks up the electromagnetic noise caused by microwave ovens and electromagnetic cookers in the house. Therefore, in order to maintain the required QoS, the next generation home network will be based on optical fiber [4]. Optical fiber based link system for the home network application has just commenced its development, however, not standardized yet [5].

Plastic Optical Fiber (POF), especially graded index (GI) POF, is the most promising candidate for wiring the home network. This paper describes the evaluation results of a link system that consists of POF and O/E E/O media converter (MC) with G-bit switch. It was recently developed to meet the home network requirements.

2 Requirements
The requirements of the link system for the home network are large-bandwidth, simple and invisible installation, and safe handling. Table 1 summarizes the performance of POF.

3 Specifications
Principal specifications of the link system are as follows: transmission rate is 1Gbps, bit error rate is below 1×10⁻¹⁰ and received light power is under –10dBm assuming POF length=30m, bend radius =15mm,6points and anticipated outlet arrangements.

Running the link system to all rooms is not necessary in the near term; the living room is the first stop and will host many devices such as TVs, DVRs, Games, Set Top Boxes for VOD, and PCs that will be connected to the LAN and the Internet. They may send or receive large volumes of data simultaneously, so we developed an MC with Gbit 5-port switch (1 POF and 4 RJ45) for use in the living room. Figure1 shows a block diagram of the system and Figure 2 shows a photograph of the circuit.
TABLE 2. Transmission rate of physical layer

<table>
<thead>
<tr>
<th>Frame rate of data link layer [bps]</th>
<th>81,273</th>
</tr>
</thead>
<tbody>
<tr>
<td>Calculated transmission rate of data link layer [bps]</td>
<td>986,979,312</td>
</tr>
<tr>
<td>Calculated transmission rate of physical layer [bps]</td>
<td>999,982,992</td>
</tr>
<tr>
<td>Bit error rate</td>
<td>-6.7</td>
</tr>
</tbody>
</table>

3) N: N throughputs, bidirectional transfer, were established between the ports to confirm the routing capacity. Again, throughput of 1 Gbps (100%) was confirmed for all ports and all frame sizes.

4) Transfer capacity of the physical layer, bit error rate, and received light power were measured while replicating wiring setup in the house. Figure 4 shows the change points. The evaluation results, 11-port to 21-port with frame size of 1518, are summarized in Table 2. They show that the link system meets the specifications.

4-2 Evaluation results: O/E E/O MC

We also evaluated the O/E E/O MC itself. The lifetime of the transceiver is the key problem, and it is influenced mainly by temperature. The Mean Time To Failure (MTTF) of a transceiver is 100,000 hours with ambient temperature of 50 degrees C. Measurements showed that the ambient temperature of the transceiver was 52 degrees C when that of the MC was 40 degrees C. According to the life acceleration factor, the lifetime is estimated to be over 9 years for house use.

5 Conclusion

This paper evaluated a 1 Gbps home network system and O/E E/O media converter that we developed on POF. The results showed its effectiveness and confirm that POF is a very promising candidate for the home network. We will verify the practicality of this system by installing it in actual houses.

6 References

Low-power Digital Fiber Optic Sensing Network with Microprocessor-controlled Sensor Terminals

Chanty Rotha, Takeshi Nishimura, Yosuke Tanaka, Tatsutoshi Shioda, and Takashi Kurokawa
Graduate School of Engineering, Tokyo University of Agriculture and Technology
2-24-16 Naka-cho, Koganei-shi, Tokyo 184-8588, Japan
Phone/FAX: +81-42-388-7405, E-mail: tyosuke@cc.tuat.ac.jp

Abstract: We propose a low-power digital fiber optic sensing network where microprocessors control varieties of sensors. Low-power modulators such as liquid crystal ones produce the digital optical signal. This low-power system is driven by laser light.

1. Introduction

Sensing networks have attracted much attention in many fields, beginning from such a small-scale system as home security system to such large-scale systems as electrical power plants, chemical plants, bridges, and road transportation systems. The signal of the sensing networks generally consists of wireless, metallic cable, or optical fiber transmission. However, the conventional systems have several problems. The metallic cable network suffers from electro magnetic interference and consumes high power, which limits the long distance signal transmission. The wireless systems cannot be used in the underground and hospitals. Besides, the wireless systems have a maintenance problem because the power is supplied only from batteries. The analogue fiber sensor where the fiber functions as a sensing head lacks in flexibility. It is difficult for this type to measure several physical or chemical parameters at the same time. Furthermore, it takes high cost when using special fibers such as fiber Bragg gratings. In the digital fiber sensor where the fiber is used as a transmission line, each sensor terminal has a laser diode. This terminal takes high cost and consumes high power.

In this paper, we propose a novel low-power digital fiber optic sensing network. This system uses terminals in which microprocessors control sensors. We can measure a wide variety of physical or chemical parameters by choosing appropriate sensors. The digital optical signal is produced with low power optical modulator such as liquid crystal ones. Since the power consumption of the sensor terminals are very little, the driving power of the whole system is provided through O/E conversion of the laser power [1].

2. Principle

Figure 1 illustrates a schematic diagram of the fiber optic sensing network we propose. The signal light is provided from a laser diode installed in the monitoring side; it is propagated through optical fibers to sensing sides. The sensing side consists of a sensor terminal with several sensors. The sensor terminal is composed of a microprocessor, A/D converters, Photovoltaic (PV) cell, and an optical modulator. This configuration allows us to measure a wide variety of physical or chemical parameters just by replacing the sensors. This is also attractive from the viewpoint of reducing the fabrication cost of the sensor module. The analog signals from the sensors are converted into digital signals by the A/D converters. The microprocessor processes the digital signal and adds header signal for discriminating and avoiding collision of the signal from each sensor terminal. Then the obtained signal is converted into optical signal with an optical modulator. Though the optical modulator does not have to be very fast, it should be driven with low power. This requirement is satisfied by a liquid crystal optical modulator [2],[3] or MEMS modulator. Using these devices, the total power consumption per one sensor terminal becomes approximately 100 μW, which is only 0.5% of the conventional systems.

Fig. 1. Schematic of low-power fiber optic sensing network.

The driving power of the proposed system is provided through O/E conversion of the laser power in the PV cell [4]. Since the power conversion efficiency of a PV cell for laser light is as high as 20 to 40%, hundreds of sensors can be operated with several ten mW of the laser power. Recently, such high-power laser diodes with several hundreds mW to several W is commercially available, which have been mainly fabricated for Raman amplifiers.

This sensing system with fiber optic powering is especially useful for remote sensing in the places where DC electrical power supply is difficult. For example, we can apply this
system for monitoring strain of a large bridge, a chemical plant dealing with explosive material, and electrical power plant radiating strong electro-magnetic noise.

3. Basic experiment of sensing system using fiber optic powering

As a basic study of the proposed sensing system, we constructed a remote temperature sensor. Figure 2 (a) shows the fabricated processing unit of a sensor terminal. A microprocessor with 192 kHz clock frequency controls a digital temperature sensor and a liquid crystal optical modulator (LCOM). The photograph of the LCOM is shown in Fig.2 (b), which has fiber pigtails on both sides of the 5cm long body. The LCOM can produce the digital optical signal with kilobits per second.

![Microprocessor-controlled processing unit and Liquid crystal optical modulator.](image)

Fig. 2. (a) Microprocessor-controlled processing unit and (b) Liquid crystal optical modulator.

![Temperature measured with sensor terminal at the end of 5-km-long fiber.](image)

Fig.3 Temperature measured with sensor terminal at the end of 5-km-long fiber.

The extinction ratio of the LCOM was -26 dB at 25 and 50°C and -20 dB at 0°C for the laser light operating at 1.55 µm. These values were high enough to ensure digital signal transmission in the temperature range 0 to 50°C. Using these components, we measured the temperature at the end of 5km-long single-mode fiber. The wavelength of a laser used in this system was 1.55 µm. As shown in Fig.3, the temperature was measured with an accuracy of ±0.5°C, where the reference temperature was measured with an independent temperature sensor with an accuracy of ±0.1°C. Figure 4 shows the power consumption in this system. One sensor terminal was driven with laser power less than 5mW. Since the transmission speed of the signal is as slow as kbps, we can also use multi-mode fibers for shorter transmission length around 1 km. In this case, it is possible to transmit power higher than several W [1].

4. Summary and conclusion

We have proposed a low-power digital fiber optic sensing network. Each sensing module in the network is controlled by a microprocessor. That makes a flexible and low-cost system where a variety of sensing can be realized just by replacing the sensor devices without changing the whole sensing module. The digital optical signal is generated with low power by using a LC optical modulator that is also controlled by the microprocessor. The total power consumption in the whole system is low enough to be driven with a laser power. As a basic experiment, we have fabricated a temperature sensor system. The power consumption of the whole system was 350 µW. This implies that hundreds of sensors can be driven with a several hundred mW LD. This sensing system is promising and attractive for a wide variety of sensing application beginning from such a small-scale system as home security system to such large-scale systems as electrical power plants, chemical plants, bridges, and road transportation systems.

Acknowledgement

We are grateful to K. Maruyama of Technology Research Association for Advanced Display Materials (TRADIM) for providing the liquid crystal.

References

Reliable Sensor Data Transmission Method for Optical Home Network

Kunio Tojo1, Takahiro Murooka2, Shohei Terada1, Shingo Yamakawa1 and Kimio Oguchi1**
1 Information Networking Lab., Graduate School of Engineering, SEIKEI University
3-3-1 Kichijoji-Kitamachi, Musashino, Tokyo, 180-8633 Japan.
Tel: +81-422-37-3732, Fax: +81-422-37-3871, Email: *kunio@navy.plala.or.jp **oguchi@st.seikei.ac.jp
2 NTT Network Innovation Laboratories

Abstract This paper proposes a delivery control method that can ensure reliable sensor data transmission in the next generation optical home network. The proposed method is verified by simulations. The results show the effectiveness of the proposed method.

1. Introduction
The advance in broadband technology in Japan is remarkable. The number of FTTH (Fiber-To-The-Home) users is rapidly increasing and will exceed 30 million by the end of 2010. Along with the spread of broadband access networks, a Gigabit-class network environment is spreading in the home through the use of optical fiber. Devices previously not connected to any network are now IP-capable. It is expected that the home network (HN) will support various appliances. Since so many applications will be communicating at the same time, the transmission of important data such as medical and security data may be degraded. This paper proposes a delivery control method for reliable sensor data transmission. It ensures packet-loss free transmission of key sensor data.

2. Optical Home Network
Some of the services in the optical HN are shown in Fig.1. The HN will host various services with different requirements. All devices are connected to the home gateway by optical fiber. The HN will use optical fibers that enable data transmission via WDM (Wavelength Division Multiplexing), but the number of wavelengths available is limited.

Two types of services in the HN are considered: One requires high communication quality without any packet loss and the other permits some level of degradation in communication quality. The former includes important data such as medical and security data, which may impact people’s life. The latter includes data transmission between PCs.

3. Traffic Classification based on Packet Sending Interval
It might be not effective to classify all applications in detail and satisfy communication quality simultaneously in HN. The most effective approach is to give adequate priority to the transmission of important data. This paper proposes traffic classification based on packet sending interval with high resolution. Figure 2 shows two traffic classes according to packet sending interval: “continuous traffic” and “discrete traffic.” As shown in Fig.2(a), the feature of continuous traffic is that it has periodicity: the packet transmitting interval is relatively short (nanosecond to microsecond). This traffic is mainly output by audio/video streaming devices [2]. A discrete traffic source has a longer packet interval (milliseconds or more). This traffic is mainly output by various health and security sensors. The HN should prevent degradation of the communication quality of discrete traffic.

4. Effective Bandwidth Control Scheme
To realize the reliable transmission of discrete traffic, we propose a token-based delivery control scheme for the IP layer. In proposed method, only the device that acquires the “transmitting right”, the token, can transmit its data. Token-based control allows traffic release from any equipment to be controlled, while past control methods were active only when traffic passes through specific equipment in the network. Therefore, token-based control is superior in terms of scalability to other bandwidth control approaches.

Fig.1 Services in the optical home network
Fig.2 Traffic class according to packet sending interval

320
5. Verification of effectiveness of proposed method

5.1. Simulation Scenario

To verify the functions and performance of the proposed token control method, we wrote a packet delivery control program that operates on the application level. The proposed method measures the packet loss rate of the discrete traffic and continuous traffic. Token sending interval for continuous traffic sources is taken as a variable. Figure 3 shows the flow of the token control program in the simulation. Hereafter, the prototype programs are shown in italics. *TokenRecv* serves to repeat data transmission. The packet in the *TokenRecv* buffer is transmitted to the receiver side only when the token is received. *TokenSend* has enough performance to be used as the token controller.

Figure 4 shows the simulation network configuration. All network links offered 100 Mbps. DVTS [3] was used as the continuous traffic source. The token is transmitted at constant time intervals with order of microseconds. In order to simulate congestion, a sufficient amount of background traffic was applied.

5.2. Results and discussions

Simulation results are shown in Fig.5. The horizontal dotted line (around 5 \times 10^{-3}) in Fig.5 shows the packet loss rate of discrete traffic when the proposed method is not used. It is understood that using the proposed method decreases the packet loss rate of the discrete traffic for interval times longer than about 100 microseconds. This result shows that the proposed method is effective in guaranteeing the quality of discrete traffic. Figure 6 shows the variation of average continuous traffic rate limited with token control active. It shows that the continuous traffic rate was well controlled. Figure 5 also shows that the packet loss rate of discrete traffic was very small, less than 10^{-3} or so, while that of continuous traffic was small at the value of 175 microseconds, which represents the optimal setting with respect to maximizing the communication quality of both traffic streams. The optimal value increases with the load. This implies that it is necessary to change the token interval dynamically according to the load in the HN to realize effective control.

6. Conclusion

This paper proposed a delivery control method for reliable sensor data transmission which generally consists of important information related to health and security services. Specifically, the method uses token control; we evaluated its validity by simulations. The results confirmed that the method could decrease the packet loss rate of discrete traffic (health etc.). We should examine dynamic control of the optimal token interval to increase HN efficiency.

REFERENCES


Ethernet Optical Interfaces: Today and Tomorrow

Osamu Ishida
NTT

This tutorial will cover the beginning of Ethernet in 1973, today’s fastest growing 10 Gigabit Ethernet, and tomorrow’s standardization activities such as 100 Gigabit Ethernet, Ethernet OAM. The main focus will be placed on Ethernet as optical interfaces; how they were introduced, how they work, and where they are heading. Traditional optical interface technologies such as SDH/SONET and Optical Transport network (OTN) will be also reviews for comparison.

Today Ethernet has become ubiquitous network interface not only in LAN but also in service provider’s network. Since IEEE802.3z task force completed optical Gigabit Ethernet specifications in 1998 (that is a year before 1000BASE-T was specified), Ethernet optical interfaces extend their market beyond LAN into access, metro-, and wide-area networks. Its success came from “patch work” of advanced optical technologies already established in other standards.

Ethernet optical interfaces are fully optimized for asynchronous transport of variable-length frames (i.e. IP packets); they use block codes such as 8B/10B (GbE) and 64B/66B (10GbE), identify packet boundaries with special blocks, and are able to use local clocking by IDLE block(s) insertion/drop. These physical coding technologies will be reviewed and compared with those used in traditional optical interfaces such as SDH/SONET and OTN. Pluggable optical modules, such as GBIC, SFP, XENPAK, and XFP, are also reviewed since those are another unique concept introduced with packet-based interface.

Tomorrow Ethernet optical interfaces will penetrate more and more into service provider’s network. Once the optical technologies had been driven by SDH/SONET, and tomorrow they will be driven by Ethernet. Two of the progress in cutting-edge Ethernet standardization activities, 100 Gigabit Ethernet and Ethernet OAM, will be reviewed.
Biography
Osamu Ishida

Osamu Ishida currently leads the Media Networking Systems Research Group of NTT Network Innovation Laboratories, Yokosuka, Japan.

He joined NTT in 1988 after receiving a MS in Electronics Engineering from the University of Tokyo where his thesis research concerned coherent optical-fiber communication systems. His early research at NTT Transmission Systems Labs included the characterization of integrated lasers and waveguides such as tunable DBR lasers and silica-based arrayed-waveguide-grating (AWG) routers, as well as their applications to digital optical-phase-lock loop (OPLL) subsystems and digitally tunable optical filters. He then made contributions to 10 Gigabit Ethernet (10GbE) Standard developed by IEEE 802.3 at 2002, where he pursued the LAN/WAN convergence at the link signaling sub-layer that produced the standardization of Link Fault Signaling (LFS) in 10GbE. Currently he is pursuing the technologies that could shift the paradigm towards Terabit-LAN, and from 2006, he has been managing the five-year research project “Lambda Access” for this purpose under the financial support of an agency of Japanese government, the National Institute of Information and Communications Technology (NICT).

Osamu Ishida is a member of IEEE and IEICE. He received the Young Engineers Award from IEICE in 1995. He has authored over 30 papers/letters, holds 10 patents and has authored two book chapters. Also he is a co-editor of the popular textbook on 10 Gb/s Ethernet Technologies (in Japanese).
Incoherent Optical Field Detection for High-speed Binary and Multilevel Optical Communication

Nobuhiko Kikuchi
Central Research Laboratory, Hitachi Ltd., 1-280 Higashi-Koigakubo, Kokubunji, Tokyo 185-8601, Japan
Tel: +81-423-23-1111, Fax: +81-423-27-7689, E-mail Address: nobuhiko.kikuchi.ca@hitachi.com

Abstract: We review a novel incoherent optical phase (or optical field) detection scheme using coupled differential receivers. It has the potential of significant sensitivity improvement and/or capability of receiver-side numerical chromatic dispersion compensation.

1. Introduction
The modulation and detection of the phase of high-speed optical signal has been significantly advanced in these days, and will play an important role to realize next-generation binary and multilevel optical fiber transmission systems.

For example, the phase-based signaling, such as binary phase-shift keying (BPSK) and quadrature phase-shift keying (QPSK) is known to have various advantages such as high sensitivity and high tolerance to fiber non-linearity, and eagerly studied as one of the very promising signaling format for long-distance optical fiber transmission at 40 Gbit/s beyond [7][8][9].

2. Operational Principle
Figure 1 shows the basic configuration of the orthogonally coupled differential receiver with digital signal processing as implemented in [11][12].

Incoherent Optical Field Detection for High-speed Binary and Multilevel Optical Communication

Fig. 1. Schematic configuration of incoherent optical binary and multilevel receiver with orthogonally coupled differential detectors and digital signal processing (DSP: digital signal processor, ADC: high-speed A/D converter)

3. Orthogonally coupled differential receiver
One of the advantages of the coupled differential receiver is the high-resolution differential phase detection, which is very effective for multilevel signal reception, such as 8-PSK or higher. For example, four sets of binary differential detector are required for an 8-PSK receiver [13], but it can be replaced by a single coupled differential receiver with numerical rotational operations [14]. Furthermore, signal processing offers various performance improvements: The reception of an 8-PSK component in 32-level APSK signal is demonstrated (QASK+8-PSK) with the BER improvement by 2-D adaptive equalization [15]. The
multi-symbol phase estimation (MSPE) uses averaged phase of several previously received symbols as a phase reference and improve receiver sensitivity. Some optical implementations are proposed so far, such as the analogue implementation to improve the tolerance to non-linear phase noise [16], the multilevel DPSK/QAM receiver with LDPC codes [17], the multi-tip differential receivers [18]. Its digital implementation is demonstrated in a 40-Gbit/s DQPSK transmission experiment [11]. The resultant sensitivity is reported to be only 0.5 dB away from that of the coherent Homodyne receiver.

4. Incoherent optical field detection

One of its ultimate goals is the complete optical field characterization like coherent detection. If T≤Ts/2, we can obtain the differential phase Δθ(t) more than twice in a single symbol. By accumulating Δθ(t) to obtain θ(t) and combining separately detected signal power P(t), the original received signal field E(t)=√P(t)exp(θ(t)) can be reconstructed. By digitally implementing it, a 10-Gbit/s DQPSK transmission over 240-km single mode fiber (SMF) with numerical CD compensation [12] is experimentally demonstrated. Alternate implementations without AM branch [19] and the concept of the incoherent CD compensation with a single differential detector [20] are also proposed. It will enable nearly complete compensation of various linear and non-linear impairments [19]. However, its intrinsic problems, such as the accumulation of phase reconstruction error or the “zero-power hitting” problem resulting in the loss of phase information [12], will limit its performance somewhat less than that of coherent detection technique.

4. Summary

In this paper, we review the concept of joint signal processing with orthogonally coupled differential receivers for the incoherent detection of optical phase, or optical field. Currently, only a few experimental implementations with off-line signal processing have been realized. However, it will be very attractive choice for the next-generation high-performance binary and multilevel optical fiber transmissions.

5. References

Linewidth-Tolerant 8PSK by Pilot-Carrier Added Homodyne

Moriya Nakamura, Yukiyoshi Kamio, and Tetsuya Miyazaki
National Institute of Information and Communications Technology (NICT)
4-2-1, Nukui-Kitamachi, Koganei, Tokyo 184-8795, Japan
Tel: +81-42-327-5571, Fax: +81-42-327-7035, E-mail: m-naka@nict.go.jp

Abstract
Linewidth-tolerant 8PSK self-homodyne modulation/demodulation and simultaneous constellation observation were demonstrated at 10 Gsymbol/s using a DFB-LD with a linewidth of 30 MHz and only one modulator based on a polarization-multiplexed pilot carrier.

1 Introduction
Enhancement of spectral efficiency is one of the most important technical issues in the multi-terabit-class photonic networks that are necessary to efficiently accommodate the endlessly growing demand for internet traffic. Multi-level modulation/demodulation formats, especially differential quadrature phase-shift-keying (DQPSK), have been investigated intensively to resolve this problem. To enhance efficiency further, M-ary phase-shift keying (PSK) including 8-ary PSK (8PSK) has been regarded as one of the promising candidates because of its practically achievable multi-level with more than 10 Gsymbol/s [1]. Some 8PSK experiments have been demonstrated using differential detection [2] and a phase-diversity receiver with bit-rate-compatible analog-to-digital converters [3]. However, these multilevel schemes require a narrow laser linewidth of less than about 100 kHz and use costly external-cavity laser diodes (EC-LDs) or special fiber lasers [2-4].

We previously proposed and demonstrated self-homodyne modulation/demodulation using a polarization-multiplexed pilot carrier without using any critical feedback loop or fast electronic devices. We also experimentally demonstrated the linewidth tolerance of this scheme for PSK and QPSK modulation formats [5]. In this paper, we experimentally demonstrated 8PSK using a distributed-feedback laser diode (DFB-LD) with a linewidth of 30 MHz. Any receiver sensitivity degradation was not observed, compared with a result using a EC-LD. We also showed that the proposed scheme can effectively simplify the modulator/demodulator hardware. In our experiment, the 8PSK signal was modulated using only one LiNbO$_3$-based integrated modulator and was demodulated using inexpensive polarization beam splitters (PBSs) [6].

2 Experimental setup
Figure 1 schematically shows the operating principle of the linewidth-tolerant self-homodyne scheme. A pilot carrier, which provides an absolute optical phase reference, is polarization-multiplexed with an optical signal modulated in a multi-bit-per-symbol format. At the receiver side, the polarization state of the pilot carrier is rotated by 90° to perform self-homodyne detection; phase noise cancellation is achieved because the pilot carrier has identical phase noise to the optical signal.
Fig. 2 and Fig. 3 show the experimental setup for 8PSK modulation/demodulation. Continuous-wave (CW) light from a thermally-controlled DFB-LD with a linewidth of 30 MHz was introduced into the pilot-carrier 8PSK modulator (Fig. 3) where the TM and TE polarization components could be modulated independently. The modulation of the TM component was encoded by applying DATA1 for 0-π (PM1) and DATA2 for 0-π/2 (PM2) using a first pulse pattern generator (PPG1). Here, TE component acted as a pilot carrier, although it was modulated by DATA3 from another PPG (PPG2) for 0-π/4 (PM3), to generate 8-ary polarization-shift keying (8PoSK) signal. Consequently, the modulated signal can be detected by using PBSs and balanced photo-detectors (PDs) [6]. Although our scheme is not conventional 8PSK in a single polarization-plane, the phase difference between the TM and TE polarization components is detected as 8PSK by self-homodyne detection. The data rate was 10 Gsymbol/s, resulting in a 30-Gb/s optical signal. A variable differential group delay (DGD) generator composed of polarization-maintaining fibers (PMFs) was introduced to compensate for the DGD of the modulator [7]. At the receiver side, a simple in-phase (I) and quadrature (Q) receiver pair composed of PBSs and balanced PDs was employed for the self-homodyne detection. The bit-error-rate (BER) was measured by an error detector (ED), while simultaneously observing the constellation using an oscilloscope.

3 Results and discussion

Figure 4 (a) shows the observed constellation diagram of the 30-Gbps self-homodyne detection using the DFB-LD. Although conventional 8PSK optical signal is in an identical polarization plane, the generated optical signal provides 8PSK equivalent constellation. Clear phase-point spacing could be observed in spite of the large phase noise of the DFB-LD. The non-uniformity of the spacing was mainly caused by inter-symbol interference (ISI) due to the imperfect frequency responses of the modulator and electrical amplifiers. However, better modulator response can be expected by modifying the modulator design [1, 8]. For comparison, we also attempted the same experiment using an EC-LD, having a linewidth of about 200 kHz, as the light source. The observed constellation diagram for the EC-LD setup is shown in Fig. 4 (b); no noticeable difference was found. Figure 5 shows the BER characteristics, in which closed circles and open triangles denote the characteristics for the DFB-LD and EC-LD cases, respectively. A BER of less than 10^{-10} was achieved without any error correction for both light sources. Furthermore, we could not observe any receiver sensitivity degradation when using the DFB-LD. The experimental results clearly show the advantage of the linewidth-tolerant characteristics even in the case of applying 0-π/4 phase modulation to pilot carrier.

4 Conclusion

10-Gsymbol/s 8PSK was experimentally demonstrated by using a DFB-LD with a linewidth of 30 MHz. No receiver sensitivity penalty due to the use of the DFB-LD was observed. The 8PSK modulation was achieved using only one modulator. The receivers were composed of simple combinations of PBSs and balanced PDs. Our proposed scheme will offer extremely high phase-noise tolerant homodyne multi-bit-per symbol modulation formats.

5 References
1 N. Kikuchi, et al., ECOC’06, Tu3.2.1 (2006).
7 M. Nakamura, ECOC’06, Mo4.2.5 (2006).
Carrier Synchronization in 43Gbit/s Coherent QPSK Receiver
Lei Li†, Zhenning Tao*, Takeshi Hoshida**, and Jens C. Rasmussen**
*Fujitsu Research & Development Center, Beijing, China
**Fujitsu Laboratories Limited, Kawasaki, Japan

Abstract — Solutions for carrier synchronization in optical coherent receiver, feed-back and feed-forward, are compared through theoretical analysis and simulations with 43Gbit/s QPSK to find their individual advantages/disadvantages in both performances and hardware implementations.

Introduction
The first problem in the coherent receiver realization is the carrier synchronization. Nowadays, there are many proposals to achieve carrier synchronization through digital signal processing (DSP). These proposals are attractive owing to their feasibility with current laser and DSP technologies. And according to how the carrier phase (\( \phi_k \) in Figure 1) is estimated (\( \hat{\phi}_k \)), these proposals can be classified into two categories, either feed-back or feed-forward as drawn in Figure 1(a) and 1(b). Feed-back solution (e.g. “digital PLL” in [1]) has been widely used in wireless field and feed-forward (e.g. “carrier phase estimation” in [2]) is being hotly discussed in optical area.

![Feed-back and Feed-forward block diagrams](image)

In this paper, feed-back and feed-forward are compared to show their capabilities to deal with the problems in carrier synchronization such as phase noise (laser linewidths), ASE noise and frequency offset, as well as their hardware implementations such as computational complexities and requirements for the resolution in analog-to-digital conversion. During the comparison, second-order digital phase-locked-loop (PLL) ([3], fig.5) is selected as the typical implementation of feed-back solution; while “carrier phase estimation” ([2], fig.5) is selected as the typical implementation of feed-forward approach.

Performance Comparison
The performances of feed-back and feed-forward are compared in 43Gbit/s QPSK coherent receivers. In the performance comparison, we only consider the case that the control parameters of both feed-back and feed-forward have been tuned to their optimal values to make the best trade-off between suppressing ASE noises and tracking the phase error changes under the given OSNR and laser line widths. Figure 2 shows the simulated performance of 43Gbit/s QPSK receiver with either feed-forward (FF) or feed-back (FB) solution to achieve carrier synchronization. Figure 2 presents the results from two simulation sets with different laser linewidth, one set for linewidth as 100 KHz and the other for 1 MHz, at both carrier and local laser.

![Figure 2. Feed-forward/Feed-back performance in 43Gbit/s QPSK receiver with laser linewidth as (a) 100kHz and (b) 1MHz.](image)

Figure 2 gives us two insights as follows. At first, for both linewidths as 100kHz and 1MHz, the performance of feed-forward is always slightly better than that of feed-back under all ASE noise levels. Secondly, comparing the results for different linewidths, obviously, the performance difference is increased with the increase of linewidth (typically, the performance difference is less than 1dB). The poorer capability of feed-back can be explained by its slower response to the phase random fluctuation; and such slower response is caused by the inherent one symbol delay in the digital feed-back loop as well as the loop filter. And when the laser linewidth is larger, the phase fluctuation is more significant, which makes more serious impact of such slow response on receiver performance.

From the aspect of system design, the ultimate performance of the two types of coherent receivers should be clarified first. Following the similar analysis in [5] for analog PLL with shot noise, we can get the minimal variance of residual phase error (rad\(^2\)) from feed-back under given OSNR and laser linewidths (\( \delta \)) as equation 1.

\[
\sigma_{PLL, \text{min}}^2 = 6.32 \times 10^{-6} \times \sqrt{3\pi\delta V/OSNR}
\]

(eq.1)

Based on the analysis as described in [4], we can get the minimal variance of residual phase error (rad\(^2\)) from feed-forward as equation 2.

\[
\sigma_{FF, \text{min}}^2 = \sqrt{[(\sigma_g^2 + 4.5\sigma_g^4)\sigma_S^2 - \sigma_g^4]/3},
\]

(eq.2)

where \( \sigma_g^2 = 2\pi \times 2\delta V/B \), \( B \), the symbol rate, \( \sigma_S^2 \) the complex ASE noise variance. For QPSK signal with
normalized symbol power, $\sigma_n^2$ can be written in terms of OSNR as $\sigma_n^2 = 4 \times 10^{-11} \times B_r / \text{OSNR}$ . With equation 1 and 2, we can get minimal phase error standard deviations from both solutions under given OSNR and linewidth; and with the relationship between BER, OSNR and phase error for QPSK ([6]), we can draw clear pictures of what can be expected for 43Gbit/s coherent QPSK receiver with either feed-back or feed-forward for carrier synchronization as the carrier-synchronization-induced Q penalties (dB), the results of which are shown by the contours in Figure 3.

In the above only the impact of phase noise and ASE noise in carrier synchronization was considered. If frequency offset is also taken into consideration, the performance of the feed-forward solution would be seriously affected, since it assumes that the phase error within one averaging block is constant. In [2], the author stated that the frequency-offset-induced phase deviation in one block should be smaller than 5°, which means that the allowable frequency offset is only about 10MHz when block size is 10 (typical optimal block size is 10~30 depending on OSNR and linewidth) . However, for feed-back solution, in [1], the author concluded that, the second-order PLL can capture and remove any frequency offsets at least theoretically. And our simulation results show that the performance of feed-back is no longer affected by frequency offsets once the offsets are quickly (several nano-seconds) compensated if the offset is less than twice of PLL natural bandwidth (typically several hundred MHz) . In short, feed-forward solution performs better in dealing with phase noises and feed-back solution is advantageous in tolerating frequency offsets.

**Implementation Comparison**

Aside from the performance, another important issue we have to consider is the feasibility to implement each of these two solutions with current digital processing technology. In this consideration we have to take three factors into count; the first is the complexity of computations required; the second is whether such computations can be implemented through parallel processing since parallel processing could significantly reduce the requirement for hardware speed; and the third is the requirement for the resolution of the front-end analog-to-digital converter (ADC) that converts the analog signals to digital ones which are then processed by the carrier synchronization block. Concerning the first factor, computational complexity, the “phase detector” are the same ([4], eq.1) in feed-back and feed-forward solutions, while Figure 1 clearly shows that aside from the phase detection part, there are more computations required in the feed-back configuration. Regarding the second point, it should be noted that the parallel implementation would increase the loop delay time which further affects its capability to handle phase noises for the feed-back case. Our simulation results show that if digital PLL is implemented through 8 parallel processing units, PLL-induced Q penalty will be increased by 2dB. For the feed-forward solution, on the other hand, there is no such limitation for parallel processing. On the third factor, if the available ADC resolution is rather low, the digital signal processed by either feed-back or feed-forward approach can not provide the actual information about carrier phase, and thus, the carrier synchronization would not be fully successful. Figure 4 shows the simulated effects of the ADC resolution on both solutions. This figure clearly shows that when the ADC resolution drops below 5 bit, the carrier-synchronization-induced Q penalty dramatically increases for both the feed-back and feed-forward approach.

**Summary**

Two solutions, feed-back and feed-forward, are chosen for comparison because they cover most, if not all, possible ways to achieve carrier synchronization through digital signal processing. Simulation results show that the feed-forward approach is strong in its higher capability to suppress large phase noise and lower hardware requirement, but it is weaker in handling frequency offsets. The feed-back approach is more tolerant to frequency offsets, but its higher computational complexity and its limitation on parallel processing increase the cost for its hardware implementation.

**REFERENCES**

Phase Noise-tolerance of Optical M-QAM Signals in Self-homodyne Detection with Polarization-multiplexed Pilot Carrier

Yukiyoshi Kamio, Moriya Nakamura, and Tetsuya Miyazaki
National Institute of Information and Communications Technology (NICT)
4-2-1, Nukui-Kitamachi, Koganei, Tokyo 184-8795, Japan
E-mail: kamio@nict.go.jp

Abstract
We investigated the phase noise-tolerance of a coherent detection in a high-efficiency optical fiber transmission system using QAM formats by making a computer simulation. We confirmed that the proposed scheme has a high phase noise tolerance.

1 Introduction
Enhancing spectral efficiency is very important for keeping up with the explosive growth in Internet traffic. Multi-level modulation/demodulation formats, such as differential quadrature phase-shift keying (DQPSK), have been investigated in an effort to enhance spectral efficiency [1]. Differential modulation and the delayed-demodulator are very complex on a large number of modulation levels because pre-coding is necessary.

For further enhancement, such as that achievable using M-ary phase shift keying (PSK) and quadrature amplitude modulation (QAM) format, synchronous homodyne detection using an absolute optical phase reference is required.

Optical phase-locked loops (OPLLs) [2] and phase-diversity receivers that use bit-rate-compatible analog-to-digital converters (ADCs) [3] have been proposed for this purpose. Moreover, 128-QAM transmission has been demonstrated for 20-M symbol/s [4]. We previously proposed and demonstrated self-homodyne modulation/demodulation using a pilot carrier with polarization multiplexing [5, 6] as an alternative scheme without using any critical feedback loop or fast electronic devices.

In this study, we investigated the bit error rate (BER) estimation method and phase noise tolerance in QAM formats with the differential group delay (DGD) of our homodyne scheme. The results show the feasibility of QAM formats using our scheme.

2 Configuration of Computer Simulations
2.1 Estimation Method for BER
It is necessary to estimate BER using the received data value and a limited number of symbols in a computer simulation. In the case of binary decision, Q-factor is usually used. However, it cannot be used for multi-level decision. For multi-level decision, another estimation method must be considered.

In our proposed method, first, the Eb/n0 is estimated using the variance between the transmitted modulation signal and the received signal. The Eb/n0 is the received signal’s energy per bit compared to the spectral noise density. Then, the BER is estimated from the theoretical expression [7] of the bit error rate of Gray coded M-QAM using the estimated Eb/n0.

Figure 1 shows the results of this estimation when the numbers of transmitted symbols are 512, 1024, and 2048 in a Monte Carlo simulation and theoretical calculation of 16-QAM. In the Monte Carlo simulation, a simple transmission that added only the Gauss distribution was used. This figure shows that the BER can be sufficiently estimated using the 2048 symbol data.

2.2 Simulation Configuration
Figure 2 shows the configuration of the simulation. During homodyne detection using a pilot carrier with polarization multiplexing [6], the TM polarization component of the input lightwave is modulated while the TE polarization component remains unmodulated in the transmitter; this component is a pilot carrier used to provide an absolute optical phase reference. At the receiver, the polarization state of the pilot carrier is rotated by 90° to perform homodyne detection.

QAM signals are generated as follows: First, an antipodal signal is generated by phase modulation using 1 bit; then a multilevel signal is generated by amplitude modulation using the bit of the remainder. The QAM signal becomes two independent signals by vector combining.

2.3 Parameters of Simulation
The modulation rate was 10-Gsymbol/s with non-return to zero (NRZ) format, and the transmission rate was 80-Gbit/s (8-bit/symbol) in 256-QAM. The received power of this scheme includes power from the multiplexed pilot carrier with a power ratio of 50%. The LPFs are cosine roll-off filters in which \( \alpha = 0.5 \) and the
bandwidth is 0.75*SR in DQPSK and QPSK, while the squared cosine roll-off filters in which \( D = 1.0 \) and the bandwidth is 0.9*SR in QAM, (SR is the symbol transmission rate).

The number of simulated symbols was \( 2^{14} \) symbols. In this simulation, any nonlinear distortion factor such as self phase modulation (SPM) is not considered.

3 Results and Discussion

3.1 Received Characteristics and Effect of Phase Noise

Figure 3 shows the BER performance in DQPSK, QPSK, 16-QAM, and 256-QAM at 10-Gsymbol/s in the case of a spectral line-width of 0 and 20 MHz. The I-Q constellation of 256-QAM shows a received power level of –14 dBm.

In DQPSK, the receiver sensitivity was degraded by 4-dB at a BER of 10\(^{-11}\) due to the effect of phase noise. However, in this detection scheme, the receiver sensitivity was not degraded.

Receiver sensitivity penalties of 16- and 256-QAM compared to QPSK at BER = 10\(^{-9}\) in the proposed scheme were 7.3- and 21.3-dB, respectively. These values are close to the theoretical values 6.9- and 19.1-dB, respectively. It should be noted that our proposed scheme was free from phase-noise-penalty even in the case of 256-QAM. This enables us to use an ordinary and inexpensive light source for ultimate multilevel optical communication.

3.2 Effect of DGD on Detection

The performance of our proposed scheme was affected by DGD [5]. In this simulation, the state of polarization of multilevel signal / pilot-carrier was set to principal state of polarizations (PSP) for transmission medium and ideal polarization control in the receiver side is assumed as an extreme case.

Figure 4 shows the power penalty obtained when BER was 10\(^{-9}\) vs. DGD with spectral line-widths of (\( \Delta \nu \)) 1, 10, and 20-MHz at 10-Gsymbol/s. The power penalty is defined as the increase in received optical power required when there is no phase noise.

4 Conclusion

We investigated the phase noise-tolerance of our coherent detection for high-efficiency optical fiber transmission system using QAM formats by computer simulation. The results show the feasibility of using QAM formats with our detection scheme without the need for an expensive narrow line-width light source.

5 References

1) R. A. Griffin, et al., OFC 02, WX6, (2002)
5) M. Nakamura et al., ECOC, Mo 4.2.5, (2006)
12B1-5

Numerical Study of APSK Format for Long-Haul Transmission and Its Performance Improvement by Zero-nulling Method

Hidenori Taga, Jyun-Yi Wu, Wei-Tong Shih, and Seng-Sheng Shu
Institute of Electro-Optical Engineering, National Sun Yat-Sen University
No.70 Lien Hai Road, Kaohsiung 804 Taiwan R.O.C.
Tel. +886-7-525-2000, Fax +886-7-525-4499, e-mail hidenoritaga@mail.nsysu.edu.tw

Abstract
Transmission performance of APSK format was studied theoretically. The extinction ratio of the ASK caused a trade-off of the performance. Improvement by adopting zero-nulling method was proposed and its effectiveness was confirmed by the simulation.

Introduction
Advanced modulation formats are attracting many attentions because they can enable improved transmission performance [1] and improved spectral efficiency [2]. Amplitude and phase shift keying (APSK) format is one example of such technology, and it has been investigated theoretically and experimentally [3],[4]. Even though, there are not enough studies focusing on a long-haul applicability of the APSK format. In this paper, a theoretical study to clarify the long-haul transmission performance of the APSK format was conducted. The transmission performance of the APSK format was examined as a function of the extinction ratio of the amplitude shift keying (ASK), and the trade-off between the ASK performance and the phase shift keying (PSK) performance was clarified. Then, in order to overcome this trade-off, zero-nulling method was proposed, and its effectiveness was confirmed through the numerical simulation.

Simulation model
This study utilized a following simulation model to evaluate the transmission performance of the APSK signal. Figure 1 shows a schematic diagram of the simulation model. This model is impersonating the undersea system like transmission line [5]. The numerical simulator solved the nonlinear Schrödinger equation using the split-step Fourier method [6].

At the transmitter, there were 16 modulated signals ranged from 1545.5nm to 1554.5nm with 0.6nm channel separation. The modulation bit-rate and the pattern were 10Gbit/s and 2^5-1, respectively. There were 256 bits delay between the pattern for the ASK and that for the PSK. The ASK signal had return to zero (RZ) raised-cosine waveform. There was no wavelength selective function in the multiplexer. At the input of the transmission line, the pattern of all transmitters was synchronized to be the same.

For the optical fiber transmission line, a combination of the non zero dispersion shifted fiber (NZDSF) and the conventional 1.3μm zero dispersion single mode fiber (SMF) was used. The chromatic dispersion, the dispersion slope, the loss and the effective area of the NZDSF were -2ps/km/nm, 0.08ps/km/nm^2, 0.21 dB/km and 50μm^2, respectively, and those of the SMF were +18ps/m/nm, 0.06ps/km/nm^2, 0.18dB/km and 75μm^2, respectively. The nonlinear refractive index of both fibers was 2.6 x 10^-20. The repeater span length was 50km. Undersea system like dispersion management [5] was adopted, and ten fiber spans composed one block. The first to the fifth and the seventh to the tenth spans were the NZDSF, and the sixth span was the SMF. The output power of the repeater was +9dBm, and the noise figure was 4.5dB. The wavelength dependent gain of the repeater was ignored. The cumulative chromatic dispersion for each channel was fully compensated at the receiving end in order to clean up the linear waveform distortion due to the chromatic dispersion. The optical demultiplexer had the Gaussian shape with the bandwidth of 0.2nm. The ASK signal is directly detected by the optical receiver. The electrical bandwidth of the receiver was 7.5GHz, and the electrical filter shape was assumed to be the third order Bessel filter. The PSK signal is detected by the delayed demodulation scheme and the balanced photo detector.

The performance of the received signal was evaluated
by the Q-factor [7]. For the ASK signal, the photo current of the receiver was directly converted to the Q-factor. For the PSK signal, differential optical phase was used to calculate the Q-factor following the method of reference [8].

**Results and discussions**

Figure 2 shows the transmission performance of the ASK and the PSK as a function of the transmission distance and the extinction ratio of the ASK. As seen in the figure, the transmission performance degraded as the transmission distance extended. In addition, the transmission performance of the ASK signal became better when the extinction ratio increased. In the mean time, the transmission performance of the PSK signal became better when the extinction ratio decreased. The results showed a clear trade-off between the ASK performance and the PSK performance. The reason of the PSK performance degradation could be attributed to the degradation of the phase information in the space signal of the ASK format. As the space of the ASK format suffers the effect of the optical amplifier noise more severely, the information of the PSK format on this part was damaged, and the overall performance of the PSK signal was degraded. When the performances of the ASK signal and the PSK signal were compromised, the optimum extinction ratio for over 2000km transmission should be around 5 to 6dB.

![Figure 2](image1.png)

Figure 2  Transmission performance of the APSK signal

Then, a method to improve the performance was considered. As the reason of the degradation of the PSK signal could be attributed to the information on the space of the ASK signal, the performance of the PSK signal should be improved if the PSK information on the space of the ASK signal is to be ignored. Figure 3 shows a proposed method schematically. The PSK information is carried only by the mark signal of the ASK, and the information is nulled while the ASK signal is zero. The effectiveness of this “zero-nulling” method was confirmed through the simulation. Figure 4 shows the results. The extinction ratio of the ASK was 13dB. As seen in the figure, the ASK performance was almost identical though the PSK performance of the zero-nulling method was greatly improved. Therefore, the zero-nulling method was proved to be quite effective to improve the long-distance transmission performance of the APSK format.

![Figure 3](image2.png)

Figure 3  A schematic to explain zero-nulling method

(a) Original APSK, (b) Zero-nulling APSK

![Figure 4](image3.png)

Figure 4  Transmission performance of APSK signal with zero-nulling method

**Conclusions**

The transmission performance of the APSK signal was investigated theoretically. There was a trade-off of the performance between the ASK and the PSK due to the extinction ratio of the ASK. In order to overcome this tradeoff, the zero-nulling method was proposed, and its effectiveness was confirmed through the numerical simulation.

**References**

High capacity wavelength division multiplexed (WDM) transmission systems are known to suffer from impairments arising from fiber nonlinear effects, chromatic dispersion, polarization mode dispersion (PMD), and amplified spontaneous emission noise. To improve the transmission performance of the WDM systems, a wide-variety of techniques have been proposed including the use of advanced modulation formats and new fiber types, techniques for dispersion, dispersion slope, and PMD compensation, management of nonlinear impairments, distributed amplification scheme, signal equalization and forward error correction (FEC). Among these techniques, the use of advanced modulation formats has been demonstrated as an effective solution in managing transmission impairments. In general, different data formats exhibit different waveform and spectrum that lead to different signal quality at the receiver end for a given transmission link. Meanwhile, links with different system parameters (reach, channel spacing, fiber type, amplification scheme, …) may also require different optimal data formats.

The ideal modulation format for long haul, high speed, WDM transmission links is the one that has compact spectrum, low susceptibility to fiber nonlinear effects, large dispersion tolerance and simple and cost-effective configurations for generation and detection. There are a number of advanced formats that meet these criteria to varying degrees. They can mainly be classified into amplitude shift keying (ASK) or on-off keying (OOK), phase shift keying (PSK), frequency shift keying (FSK) and their combinations. OOK encodes data by turning on or off the intensity of light, and it includes non-return-to-zero (NRZ), return-to-zero (RZ) and duobinary formats. There are also a number of variations of the RZ format, including simple RZ, carrier suppressed RZ (CSRZ), chirped RZ (CRZ), vestigial side band RZ (VSB-RZ), and dispersion managed soliton based RZ (DMS-RZ). PSK encodes data by modulating the phase of light, and it includes differential PSK (DPSK), RZ-DPSK, CSRZ-DPSK, and differential quaternary PSK (DQPSK) and its pulse carved forms. FSK encodes data via the modulation of the frequency of light, and includes FSK, continuous phase FSK (CPFSK), and minimum shift keying (MSK). Many studies have been focused on improving the performance of available advanced modulation formats as well as exploring some new configurations to improve transmission performance or reduce the required components for multiple channel signal generation with these formats.

Typical optical data generation schemes include either direct or external modulation of a laser. The former can be used for the generation of NRZ, DPSK, FSK and MSK. External modulators include phase modulator (PM), electro-absorption modulator (EAM) and Mach-Zehnder modulator (MZM). PM is used for phase encoding, EAM can be used for OOK data encoding or pulse carving, while MZM can have all these functions depending the driving conditions. RZ based OOK or PSK (DPSK/DQPSK) typically requires two modulators with one for data encoding and the other for pulse carving. Properly operating a dual drive MZM can achieve various functions and usually leads to reduced number of required modulators.

In this tutorial talk, I will review and analyze various schemes for the generation of abovementioned typical data formats, in terms of configuration, cost and performance.
Yang Jing Wen has over 16 years experience in fiber optical communication systems. He received his PhD degree in electronic engineering from Southeast University, China, in 1996. From 1996 to 1999, he worked at Nanjing University of Posts and Telecommunications, where he engaged in nonlinear optical transmission systems. From 1999 to 2005, he was a research fellow then senior research fellow and key researcher, with Australian Photonics Cooperative Research Centre, Department of Electrical and Electronic Engineering, the University of Melbourne, where he engaged in wavelength division multiplexed transmission, high-speed optical signal generation and photonic subsystems. Since 2005, he has been with Institute for Infocomm Research, A*STAR, Singapore, where he is a research scientist and responsible for high-speed transmission systems, subsystems and WDM passive optical networks. Dr Wen has published one book, a book chapter and over 120 journal articles and conference papers in optical communications. He is a senior member of IEEE.
High-speed lithium niobate modulator and its application to huge capacity transmission system

Tetsuya Kawanishi
National Institute of Information and Communications Technology
4-2-1 Nukui-kitamachi, Koganei, Tokyo 184-8745, Japan
Tel +81 42 327 7490, Fax +81 42 327 7938, e-mail: kawanishi@nict.go.jp

Abstract—We describe integrated lithium niobate (LN) modulators for high-speed optical amplitude, phase and frequency modulation. Over 100 Gb/s transmission can be achieved by differential quadrature-phase-shift-keying (DQPSK).

I. INTRODUCTION

Various types of modulation techniques, for example, differential phase-shift-keying (DPSK) [1], [2], differential quadrature phase-shift-keying (DQPSK) [2], [3], [4], [5], amplitude-and phase-shift-keying (APSK) [6], [7], frequency-shift-keying (FSK) [8], [9], [10], [11], [12], single-sideband (SSB) modulation techniques [13], [14], [15] etc, were investigated to obtain enhanced spectral efficiency or receiver sensitivity in optical transmission systems. Orthogonal modulation techniques with OOK and FSK or OOK and DPSK are also attractive for optical labeling in optical systems [10], [16], [17]. Recently, we reported a high-speed optical versatile modulator designed for over 40 Gbaud/s operation based on frequency-shift-keying (FSK) and single-sideband (SSB) modulators [10]. The optical versatile modulator consisting of two sub Mach-Zehnder (MZ) structures embedded in a main MZ structure can directly control optical in-phase and quadrature components, so that optical quadrature phase-shift-keying (QPSK) signals can be obtained by feeding a pair of binary data streams to the two sub MZ structures [4], [5]. The versatile modulator is also applicable for quadrature amplitude modulation (QAM) [18].

II. VERSATILE MODULATOR

The versatile modulator consists of two sub MZ structures (MZ_A and MZ_B) as shown in Fig. 1 (a). For QPSK modulation, a pair of data signals are applied to the two sub MZ structures, to achieve control of in-phase and quadrature components. The output can be expressed by

\[
R = \frac{A_{\text{LW}} e^{i \pi/4}}{2} e^{-2\pi f_0 t} \left[ \cos[g_1(t)/2] + j \cos[g_2(t)/2] \right]
\]

where \( g_1(t) \) and \( g_2(t) \) denote induced optical phase differences in MZ_A and MZ_B, respectively. The main MZ structure has optical phase difference of \( \pi/2 \) between the two arms. By using four symbols of \( g_1, g_2 \) = (0, 0), (2\pi, 0), (0, 2\pi), (2\pi, 2\pi), we can generate a QPSK signal, where the phases of the symbols are 0, \( \pi/2 \), \( \pi \), 3\( \pi/2 \). The electrode lengths of the main and sub MZ structures of a fabricated versatile modulator for QPSK modulation were respectively 16 mm and 32 mm. In a 1.5\( \mu m \) region, the half-wave voltages (\( V_{\pi} \)) of the main and sub MZ structures were, respectively, 4.9 V and 2.5 V in push-pull operation at low frequency. Optical 3 dB bandwidth of each electrode was larger than 27 GHz. The insertion loss of the modulator was 5.1 dB.

III. HIGH-SPEED DQPSK MODULATION

Fig. 1 (b) shows the experimental setup for DQPSK modulation. Each of the sub MZ structures was biased for minimum dc transmission, where optical phase difference between the two sub MZ structures was adjusted to \( \pi/2 \) by using the electrode C1 or C2. A pair of non-return-to-zero (NRZ) data streams at 40 Gb/s were obtained from a 4 : 1 multiplexer that combines four 10-Gb/s sub channels of 2\( ^7 \) \(-\)1 pseudo-random bit sequences. As shown in Fig. 1, one of the streams was fed to MZ_A for I component modulation, and the other was fed to MZ_B for Q component, where the delay between the two streams was adjusted to 115 bit. The amplitude of I and Q signals was 6.5 V (peak-to-peak), corresponding to 2\( V_{\pi} \) at 40 Gb/s, to generate an 80 Gb/s optical DQPSK signal at the output port of the modulator. The DQPSK signal generated at the modulator was decoded by a one-bit delay interferometer whose constructive and destructive ports were connected to a balanced photodetector. However no precoder was employed for our experiment, and hence there was a deterministic mapping of data from input to output. In order to allow bit-error-ratio (BER) measurements, the error detector was programmed with the expected data sequence. We used a single receiver to decode each 40 Gb/s tributary by adjusting the differential optical phase in the one-bit delay interferometer (\( \Delta \phi \)) at \( \pi/4 \) or \(-\pi/4 \). Fig. 2 shows eye diagrams measured at the electric output of the balanced photodetector, and a back-to-back BER curve of a sub channel extracted from each tributary by a 1:4 demultiplexer. In back-to-back transmission, clear eye openings and error-free operation were observed for the two tributaries whose symbol rate was 40 Gbaud/s.

Our versatile modulator can provide over 100 Gb/s DQPSK signals [19]. 2000-km transmission of 107Gb/s DQPSK was reported by using 50% duty cycle RZ curving technique [20]. The modulator can be applied to dense WDM systems. 12.3-
TB/s transmission capacity and 3.2-b/Hz spectral efficiency were demonstrated by using polarization-multiplexed 85.4-Gb/s RZ-DQPSK at 50-GHz channel spacing in the C band, where 77 WDM channels were transmitted over 240 km of SSMF [21].

REFERENCES
[3] A. H. Gnauck, P. J. Winzer, S. Chandrasekher, and C. Dorrer, “Spectrally Efficient (0.8 b/s/Hz) 1-TB/s (25x42.7 Gb/s) RZ-DQPSK Transmission Over 28 100-km SSMF Spans With 7 Optical Add/Drops,” ECOC 2004
A Novel RZ-DQPSK Transmitter Setup and Comparison Regarding their Tolerance in Spectral Efficient 8 x 40 Gb/s WDM Systems

Muhammad Haris 1, Jianjun Yu 1, 2, and Gee-Kung Chang 1
1) School of Electrical and Computer Engineering, Georgia Institute of Technology, Atlanta, GA 30318, USA
2) NEC Labs America, Princeton, NJ, 08540
Phone: 404-894-1917, Email: haris@gatech.edu

Abstract: This paper presents a novel setup to generate a RZ-DQPSK signal which employs only two modulators as compared to three modulators in conventional parallel or serial setup configurations. The robustness of RZ-DQPSK signal generated by this technique is investigated using 8x40 Gb/s WDM systems with 50GHz separation over 500km SMF.

Introduction:
Recently, different modulation formats have been investigated for high bit-rate systems to overcome chromatic dispersion limits and nonlinear effects in fiber. RZ-DQPSK has drawn much attention because for the same data rate, spectral efficiency is almost doubled. Up till now there are two conventional transmitter setups for RZ-DQPSK signal generation. First technique consists of parallel combination of two Mach-Zehnder modulators (MZM) followed by a RZ pulse carver (Fig. 1a) to generate a RZ-DQPSK signal [1]. Another setup comprises (Fig 1b) of serial combination of MZM and a phase modulator (PM) followed by a RZ pulse carver [2]. Both of these techniques use at least three modulators for the generation of RZ-DQPSK signal. Fig. 1c presents a proposed scheme for the generation of RZ-DQPSK signal using only one dual-arm Mach-Zehnder modulator and RZ pulse carver. Precoded data and clock signal is split into two paths and then added after introducing time delay to generate two RF signals to drive dual-arm modulator for the generation of RZ-DPSK signal. Another 90 degree phase shift is introduced by phase modulator, which is connected in serial to generate RZ-DQPSK signal. RZ-DPSK signal by this technique is first presented in [3], this is the first time RZ-DQPSK signal is presented using this method and comparison is drawn with other schemes by numerical simulation. DQPSK signal carries two bits of information per symbol employing four states of optical phase. Time delay module in proposed scheme is set to one bit period, which in this case is 50ps as 40 Gb/s RZ-DQPSK signal has symbol rate of only 20 GHz. System performance is evaluated by eye opening penalty (EOP). Numerical simulation has been performed with the VPI Transmission maker software. EOP penalty is calculated with respect to back-to-back transmission [4].

System setup and results:
The investigated system setup is shown in Figure 2. Balanced detection is used for phase modulated signals at the receiver. Eight channels are considered with data rate of 40 Gb/s and 50 GHz channel spacing. The multiplexer and demultiplexer both consist of third order band-pass Bessel filter with 3-
dB bandwidth of 50 GHz. Each span has 100km of SSMF which is compensated by dispersion compensation fiber (DCF) by post compensation. The optical power launched into SMF is kept at 0dBm per channel and noise figure (NF) of EDFA is kept at 4 dB. Loss of single mode fiber (SMF-28) and DCF is 0.25 and 0.45 dB/km, respectively.

SSMF and DCF dispersion is 17 and -100ps/nm/km, effective core area is 80 and 40 μm², respectively. An optical pre-amplifier is set to provide +2dBm optical power before optical detection. Figure 3 presents eye diagrams of 40 Gb/s RZ-DQPSK signal generated by proposed technique.

I. Input Power
In Fig. 4 the EOPs are plotted against average fiber input power per channel for all the three transmitter setups. Solid lines in graph represent SMF length of 500km and dotted lines represent SMF length of 100km. EOPs for serial and parallel setups are in agreement with each other. As expected EOPs shoot high as average power is increased and non-linear effects in fiber come into play at greater extent. Our setup exhibits similar performance for lower span length but for larger span length EOP is approximately 0.5 dB higher for proposed scheme.

II. Residual Dispersion
Figure 5 shows residual dispersion plotted against EOP after WDM transmission over 500km SMF with 50GHz separation. DCF length is varied in each span to provide cumulative dispersion for 5 spans. If we choose EOP penalty of 2 dB as transmission threshold, parallel setup exhibits dispersion tolerance range of 145ps/nm, while serial configuration and our proposed setup shows 140ps/nm and 125ps/nm of dispersion tolerance range. Figure 6 shows eye patterns after 10 spans of fiber totaling 1000km of transmission.

Conclusions:
A novel transmitter setup is presented to generate RZ-DQPSK signal, which require only one dual-arm Mach-Zehnder modulator and one phase modulator. This scheme is more economical and cost effective since it reduces the number of modulators to two as compared to three in conventional setup. Simulation results are presented for RZ-DQPSK signal generation and WDM transmission with 0.8b/s/Hz spectral efficiency, which explain that performance of this proposed setup is in close agreement with other schemes. Considering the tolerance to fiber non-linearities and dispersion, no particular scheme outperforms the other two.

References:
1. R. A. Griffin, et al., OFC 2002, Anaheim, USA
Experimental Measurement and Numerical Estimation of Optical Phase Jitter in RZ-DPSK Systems

David Boivin (1), Muhammad Haris (1), Marc Hanna (2), Jianjun Yu (3), Gee-Kung Chang (1), John R. Barry (1)
1 : School of Electrical and Computer Engineering, Georgia Institute of Technology, Atlanta, GA 30318, USA
2 : Laboratoire Charles Fabry, Institut d’Optique, 91403 Orsay Cedex, France
3 : NEC Laboratories America, Princeton, NJ 08540, USA

Abstract We propose a novel experimental method to estimate the optical phase variance of RZ-DPSK systems. Numerical simulations confirm the accuracy of the method making it a useful and simple tool to characterize DPSK transmission impairments.

Introduction

Long-haul DPSK-based transmission performances are primarily corrupted by nonlinear phase jitter. Although it is now known that the Gaussian phase noise assumption fails in predicting balanced receiver performance, its use is still reasonably accurate for single-ended detection [1]. Based on this assumption, a differential phase Q factor has been defined [2] to predict the BER. This method essentially addresses the impossibility of error counting in simulations and proposes a useful BER estimate by evaluating the statistical fluctuations of the received signal. If the standard deviations of the differential phase on 0 and π rails are easily evaluated in simulations, their experimental evaluation is far from being straightforward. To the authors’ knowledge, reference [3] is the only experimental investigation that proposes an evaluation of the soliton phase stability through its standard deviation measurement. Nevertheless, the proposed scheme is too specific to the modulation format used and cannot be adapted to evaluate the performances of standard DPSK-based communications.

Here, we report on experimental investigations of phase jitter in RZ-DPSK systems and propose a novel method which provides an estimate of the optical phase variance. The means and variances of the power detected before and after DPSK demodulation are the only physical parameters needed to determine the optical phase variance. This can be easily measured with the histogram function of a fast oscilloscope. Numerical simulations confirm a posteriori the accuracy of the method which can be considered as a useful and simple tool to characterize a DPSK transmission.

Principles

After propagation in the transmission line, the electric field of one pulse can be written as

\[ E_{\text{in}}(t) = (a(t) + n(t)) \cdot e^{i(\phi(t) + \delta\phi(t))} \]  

where \( a(t) \cdot e^{i\phi(t)} \) is the pulse waveform corrupted by amplitude and phase noise \( n(t) \) and \( \delta\phi(t) \). In general these noise are correlated, however for simplicity, we will assume they are independent Gaussian zero-mean random variables with variances \( \sigma_n^2 \) and \( \sigma_{\delta\phi}^2 \). This simplified model is justified a posteriori by our simulation results. The data bits encoded on the differential phase are retrieved by using a Mach-Zehnder delay interferometer (MZDI) with insertion loss \( X_{\text{loss}} \approx 3.14 \text{ dB} \). When the interference is constructive, the variance of the detected power at peak intensity is given by

\[ 10^{-5} \cdot \sigma_{\text{out}}^2 = f(m,v) + 2m^2 e^{-2\sigma_n^2} + g(m,v) e^{-2\sigma_{\delta\phi}^2} + h(m,v) e^{-2\sigma_n^2} \]  

(2)

where \( m \) and \( v \) are the mean and variance of the detected power before demodulation. The functions \( f \), \( g \) and \( h \) are given by

\[ f(m,v) = \frac{1}{8}(v + m^2), \quad g(m,v) = -\frac{1}{8}(2m^2 - v), \]  

\[ h(m,v) = m \sqrt{\frac{m^2 - v}{2}} - m^2 + \frac{v}{2}. \]

The histogram measurements easily performed with a fast oscilloscope provide the numerical values for the physical parameters \( m, v \) and \( \sigma_{\text{out}}^2 \). The numerical resolution of Eq. (2) finally gives the estimation of the phase variance.

Experimental Results

The experimental setup used to investigate the proposed phase jitter estimation method is shown in Fig.1. The output of a 1550 nm CW laser source feeds a phase modulator driven at 10 Gb/s by a pseudo random bit stream (PRBS 2^7 − 1). The Mach-Zehnder modulator driven by a 10 GHz clock with an amplitude \( V_g \) performs the pulse carving for RZ-DPSK modulation. In order to observe nonlinear effects, we intentionally add ASE noise to the RZ-DPSK signal at the transmitter. The noise generation scheme consists in a variable attenuator followed by an optical amplifier and an optical bandpass filter and allows the adjustment of the OSNR before transmission.

The transmission line consists of a 100 km span of SMF preceded by 1 km span of highly nonlinear fiber (HNLF) with a nonlinear coefficient \( \gamma = 10 \text{ W}^{-1} \text{ km}^{-1} \)
to enhance the SPM effect. Dispersion is compensated by a dispersion compensating fiber (DCF) module with 1820 ps/(nm). The signal is preamplified before being demodulated. One output port of the MZDI is connected to the oscilloscope to perform the histogram measurements while the other one is used for BER measurement.

In order to enhance the Gordon-Mollenauer effect, high input powers were used. The input powers into the HNLF and DCF were 16.5 and 2.45 dBm respectively. The optical bandwidth of the oscilloscope for the histogram measurement was 50–GHz.

![Fig. 1: Experimental setup](image1)

**Fig. 1: Experimental setup**

Fig. 2 shows the eye diagrams observed before and after demodulation for 28 and 33 dB noise loader attenuations. The process used to estimate the phase jitter is represented in the first case. As indicated, all the histograms are measured in a 20 ps timing window at the peak power of the eye diagram before and after demodulation. The received peak power prior demodulation is set to 8 mW so that the detection noise remains negligible compared to the amplitude and phase noise we want to characterize.

![Fig. 2: Eye diagrams observed before and after demodulation for different noise loader attenuations.](image2)

**Fig. 2: Eye diagrams observed before and after demodulation for different noise loader attenuations.**

Fig. 3 shows the phase variance obtained with experimental measurements as a function the noise loader attenuation.

These estimations are compared with numerical simulations where the phase variances are directly computed from the optical phase evolution of a PRBS $2^7-1$ sequence propagating in the setup described in Fig. 1. The estimated variance is computed using the best unbiased estimator with 1480 samples.

![Fig. 3: Phase variance as a function of noise loader attenuation (+ experimental points, ◊ simulations).](image3)

**Fig. 3: Phase variance as a function of noise loader attenuation (+ experimental points, ◊ simulations).**

We note a satisfactory agreement between experimental and simulations results. When the attenuation ranges from 14 to 20 dB, the standard deviation measurements do not change significantly. In fact, as shown in Fig. 3 inset where the measured BER is plotted as a function of both attenuation and Q phase ($Q_{phase} = \pi / 2\sigma_p$), the BER in this range is inferior to $10^{-9}$. The amount of phase noise is then gradually increased and measured for BER ranging from $2.41 \times 10^{-8}$ to $1.13 \times 10^{-3}$.

**Conclusions**

We have proposed and demonstrated a novel experimental method to estimate the optical phase variance of RZ-DPSK systems. The method is based on histogram measurements performed before and after demodulation. The estimation of the optical phase jitter relies on the assumption that the phase and amplitude noise are Gaussian and independent. This simplification allows us to relate the phase variance to easily measurable physical parameters and yields estimations in good agreement with the ones obtained by numerical simulations.

**References**

Demonstration of optical quadrature amplitude modulation by using a high-speed optical DQPSK modulator and precise voltage-control technique

Akito Chiba, Takahide Sakamoto, Tetsuya Kawanishi and Masayuki Izutsu

National Institute of Information and Communications Technology, 4-2-1 Nukui-kitamachi, Koganei-shi, Tokyo 184-8795, Japan
Tel: +81-42-327-7490, FAX: +81-42-327-7938, E-mail: kawanish@nict.go.jp

Abstract

We succeeded in observation of constellation map of a 16-level optical quadrature amplitude modulation with a rate of 2 Gbaud by adopting techniques of precise control of modulation voltages and digital carrier-phase estimation.

Multilevel optical modulation technology such as differential quadrature phase shift keying (DQPSK) and quadrature amplitude modulation (QAM) attracts much attention by favor of high-speed optical vectorial modulators [1, 2]. The most feature of these modulation techniques is that increase of data transportation rates can be possible without increase of modulation rates. Especially, optical QAM which is conventionally used in wireless communication technology is the most powerful tool based on the degree of state per one symbol and narrow modulation-data occupation in frequency domain. Nowadays, some proposals and experiments of the optical QAM [3, 4, 5] has been demonstrated though, optical phase locked loop (PLL) was employed in the experimental setup which limits increase in modulation rate of optical QAM per one symbol. In addition, a pilot carrier was used for optical PLL in the experimental setup [5]. Since the pilot carrier decreases effective spectral efficiency, development of alternative method is necessary for realization of high-modulation rate optical QAM. In this paper, we demonstrate 2.0G-baud 16-level optical QAM by a high-speed vectorial optical modulator. By adopting carrier-phase estimation technique [6] instead of rapid optical PLL, we succeeded in demodulation of optical QAM signal. Each levels in the constellation map obtained can be almost discriminated without consumption of frequency resource.

Figure 1 shows an experimental setup. Components enclosed by dashed (dotted) lines are for modulation (demodulation) of light signal. An external-cavity diode laser (Agilent, 8163A and 81689A) was used as a light source. The optical power was set at 6 dBm and was divided by a polarization-maintaining (PM) 50:50 optical directional coupler. Light emitted from one output port of the coupler was launched into an optical modulator while light emitted from the other port was lead to an input port of an optical hybrid coupler and was used as a local oscillator Ls for homodyne detection. After passing the modulator, the light was amplified by an EDFA (Fitel, ErFA1224-LN20-J) and passed through an optical narrowband filter (Ouyoukouden, TFMB-1560-1, \(\Delta\lambda = 1\) nm).

Optical single side band modulator (Sumitomo Osaka Cement), which is a Mach-Zehnder (MZ) structure (MZ0) nesting two MZ structures (MZ1 and MZ2) in each arm of the MZ0, was used for modulating optical signal. Half wave voltages of MZ0, MZ1, MZ2 are 7.0 V, 3.5 V and 3.3 V, respectively. Bias voltage of MZ0 was set so as to ensure \(\pi/2\) phase difference between each arm, while bias voltages of MZ1 and MZ2 were adjusted to be the minimum in light power transmittance of each MZ. Electric modulation signals corresponding to in-phase (I) component and quadrature-phase (Q) component were amplified (SHF, SHF806E, Bandwidth: 40 kHz – 42 GHz) and superimposed on the bias voltage of MZ1 and MZ2, respectively. The I component of the modu-
lation signal was created from a four-level data sequence made of two individual pseudo random bit sequences (period: 32767) followed by "0" and the Q component was created by 8192-point rotation of the I component data. These signals were generated by an arbitrary waveform generator (Tektronix, AWG7102) with a rate of 10 Gsamples/s, where the modulation speed was 2.0 Gbaud/s. One of difficulty in optical QAM is to generate rapid and precise multilevel electric signals applied to the optical modulator, which originates from a patterning effect of rf amplifiers. As this effect is one of hysteresis effect, we decreased input voltages of rf amplifiers to be nearly 0 V after each amplification of electric data signal. By using this technique, we achieved expansion of eye aperture of electric signal applying to the modulator.

We demodulated the optical data signal by using an optical 90-degree hybrid coupler (Optoplex, HB-C0AFC001) and balanced detectors (u2t, BPVD2020R) and evaluated both I-component and Q-component of the modulation signal. The electric signals obtained by the balanced detection were acquired by a digital oscilloscope (Tektronix, TDS6154C) with a sampling rate of 10 Gsamples/s to construct a constellation map, after amplification by rf amplifiers (SHF, SHF200CP). Due to phase instability of the $L_\alpha$, demodulation signals at each level were widely distributed in the obtained constellation map, which forces us to be difficult to discriminate their initial level. However, we succeeded in discrimination of each level in the constellation map, by using a digital synchronous process to eliminate the acquisition data during the level transition and a phase retrieval algorithm [6].

Figure 2 shows constellation maps of demodulation signal. Figure 2(a) is the map before applying the phase retrieval process while Fig.2(b) is the map after applying the process to the obtained data shown in the Fig.2(a). From Fig.2(b), sixteen clusters of points can be confirmed by adopting the carrier-phase estimation algorithm. Q factor of the clusters in the dashed square (1) and (2) of Fig.2(b) are evaluated as 2.036 and 2.167 respectively, each of which corresponds to 12.2 dB and 12.7 dB in S/N ratio, respectively. It implies a possibility of the optical QAM with Gbaud-order modulation rate.

In conclusion, 2-Gbaud optical QAM is realized by using high-speed optical DQPSK modulator, precise control of electric signal and carrier-phase estimation method. Each level can be discriminated by adopting carrier phase estimation algorithm instead of optical PLL.

This work was partially supported by NEDO and JSPS. The authors wish to thank Dr. Tsuchiya of NICT for his encouragement and fruitful discussion.

References
12B3-5

Tolerance of Fiber-Optic Systems to CD and PMD

Marco Secondini, Enrico Forestieri, and Giancarlo Prati
Scuola Superiore Sant’Anna, Via Moruzzi 1, 56124 Pisa (Italy), tel. +390505492166, fax. +390505492194, email: marco.secondini@sssup.it

Abstract—We theoretically compare the tolerance to chromatic dispersion and polarization mode dispersion of various modulation formats, electronic processing techniques, and optical equalizers. The comparison is given in terms of a target system outage probability.

I. INTRODUCTION

The capacity of fiber-optic systems is strongly limited by chromatic dispersion (CD) and polarization mode dispersion (PMD). So far, standard NRZ-OOK systems have been commonly deployed for optical communications. Now, the upgrade of optical core networks to 40 Gbit/s, the realization of 100G Ethernet, and the deployment of dispersion uncompensated reconfigurable metro networks require more sophisticated techniques to counteract the transmission impairments and increase the spectral efficiency. Different solutions have been proposed to increase the tolerance of fiber-optic systems to CD and PMD: alternative modulation formats [1], [2], electronic processing [3], and optical equalization [4], [5]. In the following, we give a theoretical comparison of these techniques in terms of their robustness to CD and PMD and system outage probability performance.

II. PERFORMANCE COMPARISON

The system performance is given in terms of system outage probability, which is defined as the probability that the bit-error rate (BER) is higher than the reference value of $10^{-12}$, given an optical signal to noise ratio (OSNR) margin of 3 dB with respect to the back-to-back OSNR required for the reference BER. The outage probability, that gives a measure of the statistical impact of PMD, is evaluated by a first-order PMD analysis for uncompensated systems, and by the reduced Bruyère-Kogelnik model proposed in [6] for compensated systems. The BER is evaluated by the Karhunen-Loève series expansion method reported in [7], which accounts exactly for any linear filtering and dispersion effect, and can be extended to include polarization effects [8], to consider D(Q)PSK formats [9], and to account for the presence of an optical equalizer [6] or an MLSD receiver [10]. The amount of residual chromatic dispersion is measured by the CD index

$$\gamma = \frac{\lambda^2 R_b d}{\pi c} \quad (1)$$

where $\lambda$ is the reference wavelength, $R_b$ the bit-rate, $d$ the residual dispersion, and $c$ the speed of light. Fiber PMD is statistically characterized by the normalized mean differential group delay (DGD)

$$\Delta \tau_0 = \overline{\Delta \tau} R_b = \alpha_{\text{PMD}} \sqrt{L R_b} \quad (2)$$

where $\overline{\Delta \tau}$ is the mean DGD of the fiber, $L$ the link length, and $\alpha_{\text{PMD}}$ the PMD coefficient of the fiber. Finally, the OSNR is given in terms of the $E_b/N_0$ ratio, where $E_b$ is the mean energy per bit and $N_0$ the power spectral density of noise in each polarization, such that this ratio represents the average number of photons per bit [7]. The use of these dimensionless parameters means that all the presented results are valid for any bit-rate.

First, we consider different modulation formats: NRZ- and RZ-OOK, NRZ- and RZ-DPSK, NRZ- and RZ-DQPSK, and the CAPS code [2]. All the signals are generated by a Mach-Zender (MZ) modulator with 0.8 $R_b$ driver bandwidth. The RZ formats, with 50% duty-cycle, are obtained by a sinusoidally driven MZ pulse carvers. The CAPS code is approximated by narrow filtering the NRZ-DPSK signal [11]. The optical and electrical filters at the receiver and the back-to-back OSNR are the same reported in [12] (unconstrained optimization). Fig. 1 shows the tolerance curves for all the considered formats. Each curve gives the maximum mean DGD that is tolerated by the system with an outage probability lower than $10^{-6}$ (about two minutes per year) as a function of the CD index. In practice, each curve defines the region in the plane $<\gamma, \Delta \tau_0>$ where the system can correctly operate. We note that the slightly different curves reported in [12] were rather evaluated for a fixed OSNR, resulting in an advantage for the formats, like DPSK, with a better back-to-back performance. The DQPSK formats, that are multilevel formats with a narrower bandwidth and a longer symbol time, are more tolerant both to CD and PMD. The CAPS code, thanks to its very compact spectrum, has the same tolerance to CD as DQPSK, but is much less tolerant to PMD. On the other hand, RZ formats, thanks to their shorter pulses, are more tolerant to PMD than NRZ. When CD dominates, the most convenient format is the CAPS code, which has the same robustness as DQPSK but a lower complexity. When PMD dominates, the RZ-DQPSK is the best format, followed by the NRZ-DQPSK. Finally, when both CD and PMD are significant, DQPSK formats are still the most robust, while all the other formats are almost equivalent in terms of robustness.

Then, taking the NRZ-OOK as a reference format, we consider different kind of electronic processing or optical equalization at the receiver. Fig. 2 shows the tolerance curves for a standard uncompensated receiver, a synchronous MLSD receiver, an oversampled (two samples per bit) MLSD receiver, and for 1- to 4-stage optical equalizers. The curves for the MLSD receivers are taken from [10], with 64 trellis states and neglecting signal quantization. The optical equalizers are fractionally spaced (the time-delay is half the bit-time) and have the functional structure described in [6], with the recombined outputs 2-D configuration, that allows the simultaneous
compensation of CD and PMD. Electronic processing is highly effective in combating CD. In particular, the oversampled MLSD receiver tolerates about eight times more CD than the standard one, and its tolerance could be further improved by increasing the number of trellis states. Oversampling is required to get the sufficient statistic. In fact, the oversampled receiver largely outperforms the synchronous one. Moreover, the oversampled MLSD receiver gives the ultimate performance theoretically achievable by any electronic processing strategy (such as feed-forward or decision feedback equalization). Also PMD tolerance is increased by electronic processing, but the improvement is limited by the complete loss of information about polarization after photodetection. On the other hand, optical equalizers are effective against both CD and PMD. The single-stage equalizer, that is equivalent to a classical first-order PMD compensator, can almost double the PMD tolerance but do not affect the CD tolerance. Increasing the number of stages, the tolerance to CD and PMD increases. Indeed, optical equalization can achieve perfect channel equalization with any amount of CD and PMD, the only limit being the number of stages and the complexity of control.

III. CONCLUSIONS

A theoretical comparison of the CD and PMD tolerance of various modulation formats, electronic processing techniques, and optical equalizers has been given. The tolerance has been measured in terms of a target system outage probability. The most robust modulation format is RZ-DQPSK, while the simpler CAPS code gives the same tolerance to CD as RZ-DQPSK, but is the least tolerant to PMD. A much higher tolerance to CD is obtained by oversampled MLSD receivers, while a higher tolerance to PMD is obtained by multi-stage optical equalizers. Indeed, optical equalizers can theoretically achieve perfect channel equalization with any amount of CD and PMD, the only limit being the complexity of the equalizer.

PRESENTED IN [10]

REFERENCES

Polarization-based Monitoring in WDM Systems

Invited paper

Mats Sköld¹, Bengt-Erik Olsson², Magnus Karlsson¹, Peter A. Andrekson¹
¹ Photonics Laboratory, Department of Microtechnology and Nanoscience, Chalmers University of Technology, SE-412 96 Göteborg, Sweden, Email: mats.skold@chalmers.se
² Chalmers Industrieteknik, SE-412 96 Göteborg, Sweden

Abstract Optical performance monitoring in WDM networks and polarization-based methods are discussed. In particular, the feasibility of a polarization modulation based monitoring scheme is addressed and demonstrated in an 820 km installed transmission link.

Introduction to optical monitoring

The functionality of optical wavelength division multiplexing (WDM) networks is governed by routers and switches, combining optical point-to-points links to obtain different communication paths. So far, these connections are established by high-speed electronic switches/routers in combination with optical-to-electrical-to-optical conversions. This means that all optical paths are well defined and can be designed with a fail-safe margin. Consequently, the possible dynamic events or system break-downs are limited to a few events, e.g. add/drop of wavelength channel, fiber cut, transmitter or optical amplifier break-down. This is also reflected by the optical monitoring topology found in today’s optical networks, which basically is restricted to ensure correct channel wavelengths and power levels in transmitters and receivers. The channel power measurements require tunable optical filters followed by a single photo detector or arrayed-waveguide gratings (AWG) in combination with several photo detectors.

The situation will change dramatically as the optical networks develop towards optically reconfigurable networks to increase capacity. The networks will then comprise dynamic optical cross-connects, optical add-drop multiplexers, wavelength converters, optical regenerators, compensators etc. In these systems any optical path can be set-up at any time, with the consequence of dynamically changed optical properties of the network paths. As a result, it is no longer sufficient to measure wavelength and power to ensure network functionality and quality. Here, more advanced optical monitoring is required, such as monitoring of optical signal-to-noise ratio (OSNR), dispersion, polarization-mode dispersion (PMD), polarization dependent loss (PDL), and Q-factor [1]. Among these, OSNR and PMD are of particular interest; OSNR because it is closely related to bit-error rate, and PMD due to its time-varying nature. To achieve complete control of the network performance and of every component of an optical network, monitoring should ideally be conducted at every available point of the network, i.e. at each transmitter, amplifier, router, etc., see Fig. 1. In practice, the number of monitors becomes limited by the complexity (price) of the monitors (usually related to the monitoring capability.) The monitors are then placed at strategic positions in the network, i.e. around routers and at the end of transmission paths. However, this limits the ability to localize the cause of any system malfunction.

Polarization-based monitoring

Many different techniques to conduct performance monitoring have been proposed over the last years including utilization of the polarization properties of the light. Since the polarization properties are bound to each WDM channel, the polarization properties provides channel-individual information independently of how the channels are routed or added/dropped through the network. Most polarization-based monitoring methods are similar to regular power monitoring in a sense that it requires tunable filters or AWGs. However, the polarization information can offer additional information, e.g. the ability to determine the channel OSNR through polarization-nulling [2, 3] or from the degree of polarization [4]. The polarization-based OSNR measurements rely on the property that the amplified spontaneous emission (ASE) noise of optical amplifiers is unpolarized, in opposite to the coherent optical signal. Hence, the in-band ASE noise can be determined by blocking the signal by means of a polarizer. The polarizer is then realigned to the orthogonal state of polarization to pass the optical signal to through the polarizer. The two power measures provide the in-band OSNR of the WDM channel. The technique relaxes the requirements on the optical filters compared to conventional OSNR measurement methods. The method can be implemented with one AWG in combination with N sets of polarization-based OSNR measurement devices (N equal to the number of WDM channels), or by one tunable optical filter and a single OSNR measurement device.

Fig. 1. A schematic optical network illustrating the large number of relevant monitor points.
However, in presence of PMD [5], PDL [6] or by nonlinear effects [7], the accuracy of the OSNR measure may be deteriorated. Consequently, various methods have been developed to mitigate or compensate their influence, usually by additional optical filtering [8-12]. This can at the same time provide an estimate of the system PMD. To obtain the complete polarization properties of the system the Stokes parameters must be measured spectrally with high resolution [4, 10, 13].

Polarization modulation monitoring
A monitoring system that obtains the Stokes components over the C-band with high resolution is possible, but would be rather costly (complete spectrally resolved polarimeter at each monitor.) By introducing a polarization modulation of the channel state of polarization (SOP) at each channel transmitter before transmission, the polarization modulation can be used throughout the system at an arbitrary number of monitoring nodes (see Fig. 2) to achieve monitoring of channel power (modulus of Stokes vector) as well as polarization properties of the system [14]. The spectral resolution becomes limited to the channel spacing of the system, but on the other hand no optical filter is required and each monitor only requires a polarizing detector. Thus, the monitors are inherently low-cost and can be placed in a large number throughout the network. The polarization modulation causes not any restriction; in fact in presence of PMD a well-designed polarization modulation can improve the performance of forward error correction algorithms [15].

The monitoring system is characterized by a calibration node directly after multiplexing of the WDM channels, shown in Fig. 2. The calibration can run continuously and adapt the system to changes in the polarization modulation. This way the polarization modulation can be implemented without the use of polarization-maintaining equipment. The calibration defines the initial channel Stokes vectors, the virtual SOP, to be constant and aligned directly after the multiplexer. This property can be used to detect PMD in the system, since PMD influences the transfer functions differently for different wavelength channels. Thus the monitored virtual SOPs are misaligned in proportion of the amount of PMD and the channel wavelength separation. Consequently, if the channel wavelengths are known, the birefringence can be determined (or vice versa). Further, channel power monitoring before and after optical amplifiers provides information about the individual channel gain of the amplifiers. A feasibility study of the monitoring system was realized in an 820 km field transmission link consisted of 12 dispersion compensated amplifier spans. Five WDM channels on a 100 GHz ITU grid were monitored, one 40 Gbit/s carrier-suppressed return-to-zero (CS-RZ) modulated channel and four 2.5 Gbit/s non-return to zero (NRZ) channels. The link was monitored at nine different distances via a built-in monitoring tap of the optical amplifier modules. The Stokes vectors of each wavelength channel were obtained at all nodes and subsequently power, virtual SOP (shown in Fig. 3), and first-order PMD were computed. In addition, channel wavelengths were monitored at the transmitter and OSNR was estimated based on the channel powers and assumed amplifier noise figure [16].

Fig. 3. The Stokes parameters of one channel measured at three different transmission distances during 24 hours.

References
8. M. Petersson et al, ECOC 2004, Tu3.6.2
15. B. Wedding et al, OFC 2009, WAA1
16. M. Sköld et al, ECOC 2006, Th2.5.1
Performance Evaluation of the Improved Polarization-Nulling Technique for the OSNR Monitoring in Dynamic Optical Networks

H. Y. Choi¹, J. H. Lee¹, S. B. Jun¹, S. K. Shin² and Y. C. Chung¹
1 : Korea Advanced Institute of Science and Technology, Department of Electrical Engineering and Computer Science
373-1 Guseong-dong, Yuseong-gu, Daejeon 305-701, Korea (E-mail: ychung@ece.kaisit.ac.kr)
2 : TeraLink communications, Inc., 6F Campus Hill, 482-3 Goong-dong, Yuseong-gu, Daejeon 305-335, Korea

Abstract We report on the performance of the OSNR monitoring technique based on the improved polarization-nulling method. The results show that the performance of this improved technique is not sensitive to the PMD, nonlinear birefringence, modulation formats (such as NRZ, RZ, DPSK, RZ-DPSK, duobinary, RZ-AMI, and DQPSK), modulation speeds, and transmission distance.

Introduction
For the proper operation of a dynamic WDM network, it is essential to monitor the optical signal-to-noise ratio (OSNR) of each channel. Previously, the OSNR has been measured by using the linear interpolation technique, in which the amplified spontaneous emission (ASE) noise was measured in between the channels and then interpolated into the signal’s wavelength. However, this technique can be quite erroneous in the modern optical network where WDM signals are added/dropped or cross-connected directly in the optical layer. To solve this problem, the polarization-nulling method has been proposed for monitoring OSNR in a dynamic WDM network [1]-[2]. This technique measures the noise power right at the signal’s wavelength by utilizing the different polarization properties of signals and ASE noises. However, the performance of this technique could suffer from the polarization-mode dispersion (PMD) and nonlinear birefringence [1]-[3]. In particular, it has been reported in details that the accuracy of the polarization-nulling method could be deteriorated if the signal is significantly depolarized by the nonlinear birefringence [3]. Several techniques have been developed to overcome this problem [1]-[2], [4]. These techniques either calibrate out the small amount of signal power leaked into the noise in the orthogonal polarization state (due to PMD or nonlinear birefringence) by using an additional optical filter [2] or measured the noise power at the side of the signal’s spectrum to mitigate the effect of PMD [4]. Recently, we have developed an improved version of the polarization-nulling technique by using a narrow tunable bandpass filter [2]. It has been already shown that the accuracy of this technique is not sensitive to the PMD and nonlinear birefringence [2]. In this paper, we report that this improved version of the polarization-nulling technique can also monitor the OSNR accurately regardless of the use of advanced modulation formats. In addition, we investigate the effects of higher-order PMD and confirm the possibility of using the proposed technique even in an ultra-long haul transmission link by using a re-circulating loop.

Principle of operation
Fig. 1 shows the operating principle of the improved version of the polarization-nulling technique by using a narrow tunable bandpass filter [2]. A PMD compensator (PMDC) is used at the input of the OSNR monitoring module to negate the effect of PMD. The tunable filter has a bandwidth much narrower than the signal. We adjust this filter to the center and the slope of the optical signal, and measure the signal powers \( P_2 \) and \( P_3 \) and the ASE noise powers \( P_2' \) and \( P_3' \) polarized orthogonal to the signal, as shown in Fig. 1(c) and 1(d). However, if the optical signal is slightly depolarized after the transmission due to the nonlinear birefringence, there could be a small portion of optical signal in addition to the polarized ASE noise in \( P_2 \) and \( P_2' \). Thus, if we define this small portion of optical signal as \( \epsilon P_2 \) and \( P_2' \) are expressed as,

\[
P_2 = \frac{1}{2}P_{\text{ASE}} + (P_1 - P_{\text{ASE}})\epsilon \quad P_2' = \frac{1}{2}P_{\text{ASE}} + (P_3 - P_{\text{ASE}})\epsilon
\]

where \( P_{\text{ASE}} \) is the power of the ASE noise located within the bandwidth of the tunable optical filter. Since \( P_1 \) is smaller than \( P_2' \), the depolarized portion of the signal power transferred into \( P_2' \) (due to nonlinear birefringence) is also smaller than the portion transferred into \( P_2 \). As a result, in case the signal is depolarized due to the nonlinear birefringence (i.e., \( \epsilon \neq 0 \)), the OSNR derived from \( P_2' \) becomes more accurate than the OSNR obtained by using \( P_2 \) [4]. However, if \( \epsilon \) were large, this technique could still suffer from large errors since the portion of \( P_2' \) transferred into \( P_2 \) cannot be neglected. To solve this problem, we subtract these portions of the optical signal from the measured noises, and estimate the power of ASE noise and OSNR by using equation (2),

\[
P_{\text{ASE}} = \frac{2(P_1 - P_2')}{P_1 - P_3', \quad \text{OSNR} = \frac{P_2 - P_{\text{ASE}} B_s / B_t}{P_{\text{ASE}} B_s / B_t}}
\]

where \( P_s \) is the total power of the optical signal (measured by scanning the tunable filter over the entire bandwidth of the optical signal), \( B_s \) is the bandwidth of the optical signal, \( B_t \) is the resolution bandwidth of OSNR (i.e., 0.1 nm), and \( B_r \) is the bandwidth of the tunable bandpass filter [2].

![Fig. 1. Operating principle of the proposed technique. (a) OSNR monitoring module, (b) Signal spectrum in the presence of the ASE noise and the measurement positions of the tunable filter, (c) Filtered signals, (d) The ASE noise orthogonally polarized from the signal.](image-url)
eight sections of polarization-maintaining fibers and polarization controllers. This PMDE could emulate the PMD of real fiber including the higher-order PMD [5]. Fig. 2(a) shows the probability density function of the emulated PMD at 1553 nm. The emulated PMD had a Maxwellian probability density function. The average PMD, \( <\Delta\tau> \), was 15.2 ps. To evaluate the effect of higher-order PMD, we sent a 10-Gb/s NRZ signal (OSNR: 20 dB) to this PMDE, and then measured the differential group delay (DGD) and OSNR error by using a commercial polarization analyzer and the proposed polarization-nulling technique, respectively. As expected, Fig. 2(b) shows that no significant error was observed in the measured OSNR. This was due to the PMDC placed in front of the monitoring module and the extremely narrow bandwidth (3 dB bandwidth = 3 GHz) of the tunable filter used in the proposed technique. Thus, the effect of the higher-order PMD on the performance of the proposed OSNR monitoring technique could be neglected.

Fig. 2. Effect of all-order PMD (a) Probability distribution of the emulated first-order PMD (\( <\Delta\tau> = 15.2 \) ps), (b) Measured OSNR error of the proposed technique (10-Gb/s NRZ signal, OSNR = 20 dB).

We also evaluated the effects of using advanced modulation formats on the performance of the proposed technique. In principle, since the proposed technique obtains the noise power to calculate the noise contribution in a part of signal’s wavelength, we expect that the proposed technique can monitor the OSNR regardless of spectral shape and width of the signal. For this purpose, we modulated the signal at 10 Gb/s ~ 40 Gb/s in various formats (such as NRZ, RZ, DPSK, RZ-DPSK, duobinary, RZ-AMI, and DQPSK), and then measured their OSNR’s. The results in Fig. 3 confirm that the proposed technique could accurately monitor these OSNR’s, regardless of their modulation formats and modulation speeds.

Finally, to verify the possibility of using the proposed technique in an ultra long-haul transmission link, we implemented a re-circulation loop made of a 640-km long SMF link shown in Fig. 4. The dispersion of this SMF link was compensated by using a dispersion-compensating fiber (DCF) module. We multiplexed 8 WDM channels (spaced at either 100 GHz or 50 GHz), modulated them with 10-Gb/s NRZ or RZ signal (pattern length = \( 2^{31}-1 \)), and then launched into this loop. The signal powers incident on the SMF and DCF were 0 dBm/channel and -3 dBm/channel, respectively. Fig. 5 shows the measured OSNR of the center channel by using the proposed technique as a function of the transmission distance for each modulation format and channel spacing. The measured OSNR’s agreed well with the values obtained by using an optical spectrum analyzer, regardless of the modulation formats and channel spacing. This figure also shows that the proposed technique could monitor the OSNR accurately, even when the transmission distance was as long as 3200 km.

Fig. 4. Experimental setup to verify the possibility of using the proposed technique in an ultra long-haul transmission link.

Fig. 5. Measured OSNR by using an OSA and the proposed technique (a) NRZ with 100-GHz spacing, (b) NRZ with 50-GHz spacing, (c) RZ with 100-GHz spacing, (d) RZ with 50-GHz spacing.

Summary

We have investigated the performances of the OSNR monitoring technique based on the improved polarization-nulling method. Previously, it has been reported that this technique is not sensitive to the first-order PMD and nonlinear birefringence. In this paper, we experimentally show that the proposed technique is also insensitive to the higher-order PMD, modulation formats (including NRZ, RZ, DPSK, RZ-DPSK, duobinary, RZ-AMI, and DQPSK), and modulation speeds. In addition, we verified the possibility of using the proposed technique in the ultra long-haul transmission link by using a 64-km long re-circulating loop. The proposed technique could monitor the OSNR accurately even when the transmission link was longer than 3200 km.

References

Optical Performance Monitoring in Amplitude Sampling Receivers

F.N. Hauske (1), M. Kuschnerov(1), K. Piyawanno(1), B. Lankl (1), E.-D. Schmidt (2)
1 : University of the Federal Armed Forces, Munich, Institute for Communications Engineering, D-85577 Neubiberg,
phone: +49-89-60043924, fax: +49-89-60043921, fabian.hauske@uniw.de
2 : Siemens Networks GmbH & Co. KG, Hofmannstr. 61, D-81379 Munich, Germany

Abstract We present a state-based OPM estimation, which enables electrical receivers to simultaneously monitor chromatic dispersion and SPM induced non-linearity. The robust estimation is proved on measured NRZ-OOK data showing accurate results.

Introduction
As optical dense wavelength division multiplex (DWDM) networks become more flexible, optical performance monitoring (OPM) experiences increased importance. Key issues are residual chromatic dispersion (CD), high-power induced non-linearities like self-phase modulation (SPM) and the optical signal-to-noise ratio (OSNR). Various methods to monitor the signal are known. However, in general they require cost-intensive external devices evaluating the optical spectrum. They also monitor the optical signal, neglecting filters and electric distortions that are crucial for the signal quality at the decision point. In addition, they tap the optical signal reducing the effective receive power. [1][2]

In [3] we have shown a method to estimate the noise power based on the digital receive sequence. This method evaluates electrical amplitude distributions allocated to certain symbol interference patterns, similar to the metrics built in state based equalizers like the MLSE. In the following, we apply the same method to estimate optical signal distortions like CD and non-linearities induced by high launch power (Plaunch).

Within this work we present a method to simultaneously estimate the parameters of CD and Plaunch of a single channel. By the allocation of states to certain bit patterns we can separate intra-channel distortions like CD and SPM from noisy statistical disturbances. The noise power estimation described in [3] is prerequisite to this method. We verify our results by simulations and measurements.

Elimination of Noise Component
During transmission over the optical fiber, the signal is exposed to deterministic linear distortions like CD and non-linear distortions like SPM. Therefore, every bit pattern will lead to a distinct interference pattern at the receiver, which is superimposed by statistic noisy distortions like amplified spontaneous emission (ASE), induced by optical amplifiers. We assume the optical noise components to be statistically independent and Gaussian. Their individual variances add up to the total variance, which defines the total noise power $4\sigma^2$.

After A/D-conversion we sort the received samples $r_i$ by $2^{L}$ states $z$ with the aid of the digital decisions $d_i$, where every state refers to its according bit pattern. The number of states depends on the number of distinct interference patterns or the channel memory length $L$ respectively.

For given statistics and taking into account filtering effects, we know

$$E(r_i) = S_z^2 \cdot \text{filt} + 4\sigma_r^2$$

By estimating the total signal power $4\sigma_r^2$ (see [3]), we can compute the noiseless representation of the received probe signal $S_z^2$.

Comparison with Reference
We now compare the probe $S_z^2$ with a set of references $S_z^2, \text{ref}$ at known parameters, obtained from simulations. Reference simulations were all single channel and single span for all combinations of residual chromatic dispersion CD=[0,200,400,…,3600]ps/nm and launch power $P_{\text{launch}}=[0.2,…,16]$dBm.

Installed fiber transmission lines usually consist of multiple spans. We define the characteristic power $P_{\text{char}}$, which substitutes accumulated non-linearities due to re-amplification in each span for one equivalent launch power $P_{\text{char}}$ (without further re-amplification) leading to the same degree of non-linear distortions. This makes it easy to simulate according references even for probe signals from multi-span transmission lines.

The reference simulations include the receiver with all filters and an ADC. The more accurate the reference model, the more precise is our parameter estimation. According to the state model applied to the received data, we sort the reference sequence by its bit patterns. As the reference should only describe deterministic distortions, we run the simulations without noise. Thus, a noise component elimination is obsolete.

Finally, we compare the pool of reference simulations with the noiseless representation obtained from the received probe signal. The best matching reference $R_{\text{ref, min}}^\text{target}$, indicated by the lowest error power between reference and received signal, leads us to the according parameters for CD and $P_{\text{char}}$.

$$R_{\text{ref, min}} = \min_{\text{ref}} \left\{ \sum_{z=1}^{2^L} \left( S_z^2, \text{ref} - S_z^2, \text{filt} \right)^2 \right\}, \ N = 2^{L+1}$$

350
Results from Simulation and Experiment
We evaluate the performance of our method by Monte Carlo simulations and measured data. Measurements were carried out for 10Gbit/s NRZ-OOK (PRBS 2^{11}-1) with variations in launch power, chromatic dispersion (adjusted by according transmission line) and OSNR. The parameter \( P_\text{m} \) of the probes refers to the actual launch power. After a 50GHz optical band-pass filtering and a photo diode, a sampling oscilloscope (2 samples/bit, 8 bit ADC, 8GHz bandwidth) saved 2\times10^6 samples of the probe signal. Clock recovery was realized by a re-sampling routine. Monte Carlo simulations were carried out under similar conditions.
For the state model, we sorted the probe signal and the reference by 2^4 states (\( L=4 \)). In a first step, the noise power and the OSNR (0.1nm) were estimated. In the experiment the OSNR was measured by the aid of an OSA. Table 1 compares measured and estimated OSNR. The estimated OSNR shows accurate results independently from deterministic signal distortions, which is in excellent agreement with simulations carried out in [3].

<table>
<thead>
<tr>
<th>Condition</th>
<th>Measured</th>
<th>Estimated</th>
</tr>
</thead>
<tbody>
<tr>
<td>( P_\text{m}=0 )dBm, CD=1250ps/nm</td>
<td>10.0 dB</td>
<td>10.3 dB</td>
</tr>
<tr>
<td>( P_\text{m}=0 )dBm, CD=1250ps/nm</td>
<td>13.0 dB</td>
<td>13.1 dB</td>
</tr>
<tr>
<td>( P_\text{m}=0 )dBm, CD=3050ps/nm</td>
<td>14.3 dB</td>
<td>14.0 dB</td>
</tr>
<tr>
<td>( P_\text{m}=15 )dBm, CD=3050ps/nm</td>
<td>15.0 dB</td>
<td>14.8 dB</td>
</tr>
</tbody>
</table>

Table 1: Comparison of measured and estimated OSNR
After the noise power correction, the probe signal \( S_\text{m,ref} \) sorted by states was compared with all references. Fig. 2 and Fig. 3 show the normalized error power \( R_{\text{err}} \) for two exemplary probes with \( \) \( P_\text{m}=0 \)dBm, CD=1250ps/nm] and \( \) \( P_\text{m}=15 \)dBm, CD=3050ps/nm]. Both figures indicate that the minimum of \( R_{\text{err}} \) from measured data (white arrow on black grid) determines the given parameters of the probe signal. Simulations under same conditions show equally accurate results (black circle on gray faced grid). Taking into account that the probe with CD=3050ps/nm was

multi-span, the estimated \( P_{\text{chew}}=16 \)dB considers re-amplification during transmission and actually tends towards a slight underestimation.

Conclusions
We have shown a method for a simultaneous estimation of deterministic non-linear signal distortions caused by the launch power and deterministic linear distortions like residual chromatic dispersion in amplitude sampling receivers with ADC. The state based estimation employs the sampled receive sequence and the according digital decisions from a decision unit. Furthermore, a reference signal at known parameters is utilized. In case of an appropriate state model with sufficient memory length, the comparison of the reference and the probe leads to an accurate estimation of the mentioned parameters. The estimation is independent from statistic signal distortions and delivers an OSNR estimation as a by-product.
Pre-requisite to the estimation is a parametric statistic description of the receive signal dependent on the noise process. An evaluation with measured data proves the use of an appropriate model and predicts equally good results for all OOK modulation formats.
The results show a satisfactory indication for CD with a clear minimum. Supplementary methods to support the indication of \( P_{\text{chew}} \) with its lower gradient are currently under investigation.
No additional measuring device needs to be applied to the receiver, which makes this method a cost effective and easy to implement alternative to customary OPM.

Acknowledgement
We would like to thank M.S. Alfiad for the experimental lab setup and for numerous measurements.

References
1 Kilper et al., JLT 2004, vol. 22, no.1, p.294
2 Kirstaedter et al., APOC 2004
3 Hauske et al., ECOC 2006, Mo4.4.4
In-Service Monitoring of WDM Passive Optical Network Using Novel Optical Reflectometry Based on Correlation Detection

Y. Takushima and Y. C. Chung
Korea Advanced Institute of Science and Technology
373-1 Guseong-dong, Yuseong-gu, Daejeon 305-701, Korea
Phone: +82-42-869-8508. Fax: +82-42-869-3410. Email: ytaku@ee.kaist.ac.kr

Abstract: We propose a novel optical reflectometer based on the correlation detection, in which the data-modulated optical transmitter is used as a probe light source. For a demonstration, we use this reflectometer for the in-service monitoring of WDM passive optical network.

1 Introduction

The in-service monitoring of the back-reflection is vital for the proper maintenance of a fiber-optic network. There have been many efforts to achieve this goal by utilizing the optical time-domain reflectometer (OTDR) [1]. In these techniques, a supervisory channel is typically used for the OTDR pulses so as not to disturb the channels in service. However, in a wavelength-division multiplexed passive optical network (WDM PON) implemented by using an arrayed-waveguide grating (AWG) at the remote node (RN), the drop fibers cannot be monitored by using these techniques since the OTDR pulse is bound to be blocked at the RN. To solve this problem, several techniques have been proposed, including the one utilizing a tunable OTDR [2][3]. However, it should be noted that all these techniques require the termination of the service for the corresponding WDM channel during the process of monitoring the drop fiber.

In this paper, we propose a novel optical reflectometer based on the correlation detection. Unlike the OTDR, the proposed technique does not need any additional short-pulse sources, but uses the data-modulated transmitter itself. The distribution of the reflectivity is obtained by calculating the cross-correlation between the transmitted and back-reflected signals [4]. Thus, this technique can utilize the existing optical transmitter without modifications, and realize the in-service monitoring at the signal wavelength. In this paper, we describe its operating principle and demonstrate the feasibility of using the proposed technique for the in-service monitoring of WDM PON.

2 Operating principle

Fig. 1 shows the schematic diagram of the proposed technique. We modulated an optical transmitter with a binary signal from the data source at the bit-rate \( B \), and measure the power of the back-reflected light. We denote the input signal power and the distribution of the reflectivity including the round-trip loss by \( P_{in}(t) \) and \( R(z) \), respectively. Thus, the power of the back-reflected light \( P_{ref}(t) \) is expressed as

\[
P_{ref}(t) = P_{in}(t) \otimes R(\nu, t/2),
\]

where \( \otimes \) represents the convolution and \( \nu \) is the group velocity of light in the fiber. Then, the cross-correlation function \( q(\tau) \) between the ac-components of \( P_{in} \) and \( P_{ref} \) (denoted by \( P_{in}' \) and \( P_{ref}' \)) can be described as

\[
q(\tau) = \{ P_{in}'(t) \cdot P_{ref}'(t+\tau) \} \times \phi_1(\tau) \otimes R(\nu, \tau/2),
\]

(1)

where \( \{ \} \) stands for ensemble average and \( \phi_1(\tau) \) is the autocorrelation function of \( P_{in}' \). Since \( \phi_1(\tau) \) is similar to a delta function, we can derive the distribution of reflectivity from the cross-correlation function as \( R(\nu) \times q(2\nu/\nu_c) \).

In order to calculate Eq. (1), we digitize \( P_{in}(t) \) and \( P_{ref}(t) \) by using an analog/digital converter (ADC) at a sampling rate of \( f_s \). Since the sampling rate of a typical ADC is much slower than the signal bit-rate \( B \), we need to limit the signal bandwidth by using low-pass filters (LPFs). Due to this filtering, \( \phi_1(\tau) \) becomes a single-peak function with a finite time width. As a result, the spatial resolution is broadened to be

\[\sim 0.22\nu_c/f_s,\]

where \( f_s \) is the cutoff frequency of the LPF’s.

In practice, it will be necessary to use a finite length of sampled data for the calculation of Eq. (1). Thus, we can replace the ensemble average in Eq. (1) with the average of \( N \)-sampled points. However, this process will generate noise background in \( \phi_1(\tau) \). For example, Fig. 2(a) shows the calculated \( \phi_1(\tau) \) for \( N=4096, f_s=10 \) MHz, and \( f_c=3 \) MHz. To evaluate this background noise, we define the background-noise suppression ratio (BNSR) by the ratio of the peak amplitude to the root-mean-square of the background noise of \( \phi_1(\tau) \). A simple statistical analysis shows that the BNSR can be well approximated by \( N^{-1/2} \), and it becomes 18 dB for \( N=4096 \).

Due to this background noise caused by the imperfect
autocorrelation property, the dynamic range of the proposed method can be severely degraded. Thus, if there exists a discrete reflection point with high reflectivity in the link, all other weak reflection points can be hidden by this background noise. To avoid this problem, we have developed the discrete component elimination algorithm. Since $P_p(t)$ can be easily measured, we can obtain $\phi_0(t)$ directly from $P_p(t)$. Therefore, this background noise can be eliminated by the following steps: (a) calculate the cross-correlation function, (b) detect the discrete reflection points from the cross-correlation function, (c) estimate the noise components by using the calculated $\phi_0(t)$, and (d) subtract them from the original cross-correlation function. As will be shown later, this algorithm enables us to realize the optical reflectometry with sufficient dynamic range.

3 Experiments

Fig. 1 shows the experimental setup. We used a distributed-feedback (DFB) laser operating at 1550.6 nm and directly modulated it with 1.25-Gb/s non-return-to-zero (NRZ) data in order to simulate the system in service. The average power launched into the fiber was 0.8 dBm. To detect the back-reflected light, we used a conventional avalanche photo diode (APD). The data and the back-reflected signals were filtered out by 3-MHz LPF's, and digitized by 9-bit ADC at 10 Ms/s. We then calculated their cross-correlation by using a personal computer. In this experiment, the spatial resolution was determined by the bandwidth of the LPF to be 15 m. The number of sampling points was 4096, and the measurement range was 40.9 km.

We first investigated the influence of the background noise originating from the background of the autocorrelation function. To neglect the influence of the other impairment factors such as the receiver noise, we used a relatively short length fiber of 2.2 km. Fig. 3(a) shows the measured cross-correlation trace. The reflection at the fiber end was clearly observed without averaging. The corresponding BNSR of the trace in Fig. 3(a) was 17.3 dB, which agreed well with the theoretically calculated value of 18 dB. We then applied the discrete component elimination algorithm to improve the BNSR. The result in Fig. 3(b) shows that the proposed algorithm could reduce the background noise level substantially, and improved the BNSR by about 8 dB.

To demonstrate the feasibility of using the proposed technique for the in-service monitoring in WDM PON, we used the setup in Fig. 1(b). The system under test consisted of two AWG’s, 10-km long feeder fiber, 6.2-km long drop fiber, and an optical network unit (ONU). The insertion loss of each AWG was about 4.1 dB. Fig. 4(a) shows the cross-correlation trace measured without averaging. The discrete reflections at the central office (AWG1), the RN (AWG2), and the ONU could be observed clearly even without averaging. In order to improve the S/N ratio further, we measured the cross-correlation trace 400 times and averaged them. Fig. 4(b) shows the result. The background noise level (shown by the dotted line) was lower than the Fresnel reflection at the ONU by as much as 27 dB, which corresponded to the return loss of 41 dB (65 dB including the round-trip loss). In the current standards on PON, the tolerable reflectivity is defined to be in the range of -32 ~ -26 dB [5]. These values are much higher than the measurement limit of this setup. Thus, the proposed technique can be used for the detection of abnormal reflections along the transmission line. However, we noted that the noise level was still too high to measure the distributed Rayleigh scattering (-72 dB/m). To demonstrate the possibility of using the proposed technique for localizing the fiber failure, we intentionally made a break point having a reflectivity of -39.8 dB just before the ONU and measured the cross correlation. The result in Fig. 4(c) confirmed that the proposed technique could indeed localize this break point.

4 Summary

We proposed a novel optical reflectometer based on the correlation detection. This reflectometer utilized the data-modulated transmitter itself, and detected the reflection points by calculating the cross-correlation between the transmitted and back-reflected signals. We demonstrated that this reflectometer could be used for the in-service monitoring of WDM PON.

References

[1] N. Frigo et al., JLT 38(11), 2641-2652, 2004

Fig. 3: (a) Cross-correlation trace when a 2.2-km fiber with open end was measured. (b) Cross-correlation trace when the discrete component elimination algorithm was applied to (a).

Fig. 4: In-service monitoring of the WDM PON system.
In-band OSNR Monitoring System based on Link-by-link Estimation for Reconfigurable Transparent Optical Networks

Jun Haeng Lee, Noboru Yoshikane, Takehiro Tsuritani, and Tomohiro Otani
KDDI R&D Laboratories Inc., 2-1-15 Ohara, Fujimino-Shi, Saitama 356-8502, Japan
TEL: +81-49-278-7857, FAX: +81-49-278-7811, E-mail: jh-lee@kdlilabs.jp

Abstract

A cost-effective optical performance monitoring system, based on link-by-link OSNR estimation, has been proposed for a reconfigurable transparent optical network. In-band OSNR monitoring was successfully demonstrated at 42.7 Gb/s RZ-DQPSK signals, reflecting the transmission performance.

1. Introduction

Optical performance monitoring (OPM) is an essential part of a transparent optical network for the efficient operation and maintenance of the network [1]. In particular, the channel power, wavelength, and optical signal-to-noise ratio (OSNR) of wavelength-division-multiplexed (WDM) signals are the basic parameters to be monitored for providing the quality of service (QoS) of the network. To date, several reliable and cost-effective techniques have been commercialized for the monitoring of channel power and wavelength [2]. However, it still seems difficult to monitor OSNR without a substantial increase in network cost, although several techniques capable of monitoring the in-band OSNR accurately have been proposed [3].

In this paper, we propose a cost-effective method, which can monitor the in-band OSNR in the reconfigurable transparent optical network by simply measuring the total input power to every optical amplifier (OA) and the power of each channel at intermediate optical nodes. We also demonstrated the OPM system, implemented by using the proposed OSNR monitoring technique in a reconfigurable transparent optical network testbed with 4 nodes. The element management system (EMS) indicated the monitored end-to-end OSNR appropriately, reflecting the transmission performance.

2. Principle of the proposed technique

In the proposed technique, the link OSNR of each link (composing a lightpath) is monitored separately instead of measuring the OSNR of an optical signal directly. Subsequently, the end-to-end OSNR is estimated by using the link OSNRs as $\text{OSNR}_{\text{end-to-end}} = 1/\sum_i(1/\text{OSNR}_{\text{OA}_i})$, where $\text{OSNR}_{\text{link}_i}$ is the link OSNR of the $i$-th link. Fig. 1 shows the schematic diagram of the proposed technique for monitoring the link OSNR of a link between two optical nodes. To monitor the link OSNR, we measure the channel power at the input and output of the link and the total power at the input of every OA in the link. By using the channel power information obtained at the input and output of the link, the channel power differences in the wavelength band of the link are calculated. Subsequently, the distribution of channel power at the input of each OA is estimated by using these channel power differences, assuming that all OAs have equivalent gain tilt. This is a reasonable assumption in most cases, where the optical link is composed of the same type of OAs from a single vendor. The link OSNR is estimated as $\text{OSNR}_{\text{link}} = 1/\sum_i(1/\text{OSNR}_{\text{OA}_i})$, where $\text{OSNR}_{\text{OA}_i}$ is the OSNR of an optical signal at the OA output when the input OSNR of the signal is infinite. $\text{OSNR}_{\text{OA}_i}$ is estimated from the measured total input power and the channel power distribution by using the relation as $\text{OSNR}_{\text{OA}_i}(\text{dB}) = P_{\text{in}}(\text{db}) + \alpha$, where $P_{\text{in}}$ is the input channel power and $\alpha$ is a constant related to the characteristics of the OA [4]. Since the proposed technique utilizes the signal powers only for the OSNR estimation, it is not sensitive to the spectral filtering caused by network nodes or other impairments, such as polarization effects or nonlinearities. The basic idea of the OSNR estimation based on the operating conditions of OAs has been reported previously [5]. However, it was required to monitor the channel powers at every OA and allocate polarization modulators at the transmitter side, which could substantially increase the network cost. We solved this issue by monitoring the channel power at the input and output of the link only and measuring the total input power (instead of the channel power) at other OAs.

Fig. 1. Schematic diagram of the proposed OSNR monitoring technique.

3. Experimental setup

Fig. 2 shows the experimental setup of a reconfigurable all-optical network testbed (with 4 nodes) and the proposed OPM system. At Node1, the outputs of 32 DFB lasers operating at 1538.19 nm ~ 1563.05 nm (100-GHz ITU grid) were multiplexed and modulated at 21.3 Gb/s. The return-to-zero differential quadrature phase shift keying (RZ-DQPSK) format (equivalent to 42.7 Gb/s). After tapping and decomposing the modulated signals into 3 wavelength groups by using a wavelength selective switch (WSS1), one group (channel 13 ~ 22) was sent to an add port of Node2 while another group (channel 23 ~ 32) was sent to one of Node3. The other portion of the modulated signals was launched to Link1, which consisted of four 80-km-long standard single mode fiber (SMF) spans. At the input of Erbium-doped fiber amplifier1 (EDFA1), 10% of the optical signal was tapped and sent to an optical channel monitor (OCM) to monitor the channel information (the channel powers and wavelengths of the WDM signals). Since the resolution bandwidth of the OCM (0.062 nm) was much smaller than the signal bandwidth, the channel power was obtained by integrating the optical spectrum over 0.7 nm span. Node2 and Node3 were configured by a dynamic wavelength cross connect (WXC) using two WSSes. Two OCMs were used at the input and output of a WSS pair respectively. The channel information monitored by these OCMs was also used as feedback information for the dynamic equalization of WDM signals after add/drop operation. Ten wavelength channels (channels 13 ~ 22) were dropped at Node2, and the wavelength group from WSS1 was added at this node. At Node3, a further ten channels (channels 23 ~ 32) were...
replaced by another wavelength group from WSS1. Both Link2 and Link3 were composed of two 80-km-long SMF spans. All EDFA, except the boost amplifiers (EDFA1, 6, and 9), were 2-stage amplifiers with a dispersion compensating module (DCM) inside. The amplifiers used for Link1 and Link2 were automatic gain-controlled (AGC) EDFA from the same vendor while those in Link3 were automatic power-controlled (APC) EDFA from another vendor.

In the proposed OPM system, the health of each link was monitored and managed by a node manager (NM) located in each node. For example, NM2 in Node2 was in charge of monitoring Link1. NM2 periodically retrieved the channel information of the Link1’s input from NM1 in Node1 and the operating conditions of each EDFA in Link1 via the simple network management protocol (SNMP) through the management plane. In addition, the channel information of the Link1’s output was obtained by a local OCM in Node2. Subsequently, NM2 calculated the link OSNR of each channel, based on these sets of information on a per second basis. Link2 and Link3 were monitored by NM3 and NM4, respectively. The EMS could retrieve the link OSNRs from every NM via SNMP through the management plane and calculate the end-to-end OSNR for each lightpath.

Acknowledgement
The authors thank Drs. M. Usami, M. Suzuki, and S. Akiba of KDDI R&D Labs for their continued encouragement.

Reference
New fiber designs and fabrications for data transmission and novel fiber devices

P. Nouchi, L. Gasca, P. Sillard, L. A. de Montmorillon, G. Melin, A. Pastouret
Draka Comteq, route de Nozay, 91460 Marcoussis, France
e-mail: pascale.nouchi@draka.com

Abstract

We review some recent results on novel fibers for both transmission and devices, covering bend-insensitive fibers, compensating fibers, highly non-linear fibers and Er-doped fibers.

Introduction

Optical fiber landscape has been constantly evolving. The last few years have witnessed the penetration of fibers deeper into the network with Fiber To The Home (FTTH) applications now booming. In the field of components, more and more applications are foreseen for specialty fibers, that can be telecom, as well as non telecom based.

Research in fiber area is very active to keep pace with this wide variety of demanding requirements. It is the purpose of this paper to review current research works for some key applications. In the field of transmission, advanced fibers have been developed to meet the specific requirements of FTTH; compensating modules have evolved to account for stringent Ultra Long Haul (ULH) system constraints. On the component side, PCFs (Photonic Crystal Fiber) have clearly paved the way to realize improved non-linear optical fibers that can target a large range of applications. At last, we will cover recent advances in the fabrication of Er-doped fibers.

Bend-insensitive fibers for access applications

FTTH is now a reality with transmission needs of 100Mb/s or more. This has spurred the development of fibers exhibiting reduced bending sensitivity compared to today Single Mode Fibers (SMF), to better account for FTTH specific needs. Lower volume at the storage points, as well as increased resistance towards incidental bends originating from improper fiber deployment, and sharp bent for installation in corners are indeed required.

For classical step-index SMF, well-known solution to reduce bending sensitivity consists in decreasing the mode field diameter over cutoff wavelength ratio (MAC value) by increasing cut-off wavelength and/or decreasing mode-field diameter [1-3]. However, bending loss levels remain significantly high when applying incidental kinks with radii in the order of 1 to 10 mm. Moreover, there is not much room to decrease the MAC value if fiber is to be kept compatible with the ITU-T G.652 standard -that guarantees well-defined ranges of values-.

This is the reason why other types of structures have been proposed over the past few years. Trench-assisted solutions [4, 5] consist of a classical step-index core with a cladding that includes a depressed layer (so-called “trench”). It is possible to find an optimized trench design that improves the fundamental mode confinement without reducing its mode field diameter; yielding significant reduced bending sensitivity while keeping G652 compatibility. Fibers compatible with both ITU-T G652.D and new G657 A&B recommendations, have been demonstrated with such type of structure [5]. One main advantage of “trench” solutions is their compatibility with mature process deposition technologies, such as versatile PCVD (Plasma Chemical Vapor Deposition); thus enabling large-scale production. Hole-assisted solutions are interesting alternatives [2, 4]. Trench layer is here replaced with air-holes structure in the cladding. Air-holes physical impact is similar to the trench one but it could bring solutions with even more bend insensitive performances in the future.

System constraints on DCM design

At present, fiber-based dispersion-compensating module (DCM) is the preferred technology to achieve the robust dispersion management required in terrestrial WDM systems.

In the past few years, major progress has been made in DCM developments, in order to account for always evolving ULH system constraints [6-9].

In [8], we established a simple analytical expression of the achievable distance of WDM systems as a function of both insertion loss and non-linearity of DCMs. We showed that the most efficient way to increase the system performance was to design dispersion-compensating fibers (DCF) with high negative dispersions. Reaching values of -250ps/nm-km at 1550nm allows to gain -15% (resp. -5%) in distance for G652 (resp. G655.D) fiber-based systems compared to typical -100ps/nm-km value. The drawback is that decreasing dispersion can alter both PMD and residual dispersion (RD) that also have impacts on system performance [9].

As a general rule, it is more difficult to manufacture DCFs when the dispersion is more negative, and especially with respect to PMD. But nowadays typical values of less than 0.10ps/km can be reached through tight controls of the manufacturing process, even at -250ps/nm-km [6]. Besides, systems tolerate a fixed cumulated DCF PMD (linked to the ~10% bit-time limit that also includes contributions of line fiber and amplifiers) and moving from -100 to -250ps/nm-km allows to gain a factor of \(\sqrt{(250/100)}\sim 1.58\) on the maximum DCF PMD value. As a consequence and to a certain extent, there is no real PMD drawback when one targets more negative dispersion to improve system performance.

Concerning the RD, the situation is more complex because it depends on the dispersion-over-slope ratio (DOS) of the line fiber. Decreasing the dispersion has an impact on the RD of high-DOS DCMs (that is for G652 fiber compensation), but RD stays small anyway, within \(\pm 0.05\)ps/nm-km over the Cband [6,9]. For low-DOS DCMs (that is for G655.D fiber compensation), decreasing the dispersion hardly influences the RD that remains high within \(\pm 0.25\)ps/nm-km over the C-band (see Fig.1). Taking into account this effect, we find that, at 40Gbps bit-rate, reaching -250ps/nm-km for high-DOS DCM reduces the gain in distance from ~15% to ~10%.
compared to -100 ps/nm-km, whereas for low-DOS DCMs the gain stays ~5% [8, 9].

**Highly non-linear fibers**

Highly non-linear fibers (HNLF) have attracted a lot of attention either for telecom (all optical signal processing, parametric or Raman amplification) or non telecom applications (white light sources, fiber Raman lasers).

Most of HNLFs are derived from SMF by modification of both the guiding parameters and the material non-linear parameter. Higher index contrast and smaller core diameter lead to a strong reduction of the effective area. Whereas the corresponding increase of Ge (germanium) content in the core also allows to increase the $n_2$ coefficient. Best fiber, using conventional technology, has Ge concentration of more than 40wt%, effective area in the range of 10 µm², yielding non linear coefficient $\gamma$ of about 20 W⁻¹km⁻¹ and loss of 0.5 dB/km [10]. Nevertheless there are some limits to such modifications and chromatic dispersion properties can only be tailored within a limited range of wavelengths.

To get even higher non-linearities, shifting from silica-based fibers to fibers made from multicomponent glasses, such as Bismuth-based, telluride or chalcogenide glasses, has also been extensively studied. Nevertheless such fibers suffer from much higher loss and practical issues like reliability and splicing to SMF. These can represent some drawbacks, essentially for telecom applications. The potential of Photonic Crystal Fibers (PCF) to enhance nonlinearities has been recognized at an early stage of their development. This is essentially due to the high index contrast between air and silica that allows to further decrease the Effective Area. Values as low as a few µm² have been reported. Doped rod in the core can be used as well to enhance non-linearities [11]. Even if loss is still a limiting issue compared to classical-technology fibers, PCFs can be designed to truly adjust chromatic dispersion properties. For example, zero chromatic dispersion wavelength $\lambda_0$ can be adjusted from visible up to telecom bands [12].

**Erbium-doped fibers**

Erbium Doped Fiber Amplifier (EDFA) is widely acknowledged as a key technology that has enabled the dramatic capacity increase from Gbit/s to Tbit/s, witnessed over the last decade in telecom applications. It has spurred the development of WDM technologies and the extensive use of dispersion management.

EDFAs are based on specific Erbium-Doped Fibers (EDF) that allow optical amplification through WDM window thanks to the incorporation of aluminum in the vicinity of the active Rare-Earth dopant. Further doping with aluminum allows unmatched properties [13]. Most common process to manufacture such fibers involves incorporation of Er and Al in a porous core layer through a preform-soaking step, as these elements are very difficult to incorporate using vapor-based processes. However, today doping technologies based on soluble chloride precursors mixing have achieved their limits in terms of doping homogeneity and capabilities. In this frame, nanotechnology is now offering new potentialities thanks to molecular engineering that allows to define and finely build Er chemical environment in a preliminary nanoparticle (NP) manufacturing stage. It is then possible to introduce these NPs into fiber core using an innovative doping concept [14].

By using NPs that include Al and Er ions, the EDFA gain shape of classical highly aluminum-doped EDFs can be achieved using far less aluminum, and with an improved Er doping homogeneity all along the fiber. Such performances pave the way for even flatter and wider gain EDFA by using a smart selection or Er ions co-dopants.

**References**

Ultra-high-density optical fiber cable and its application as pre-connectorized cable with adjustable excess length

Yusuke Yamada, Kunihiro Toge, and Kazuo Hogari

NTT Access Network Service Systems Laboratories, NTT Corporation, 1-7-1, Hanabatake, Tsukuba, Ibaraki, Japan
Tel: +81 29 868 6123  Fax: +81 29 868 6142  e-mail: yamada_yusuke@ansl.ntt.co.jp

Abstract: This paper proposes small diameter and ultra-high-density 100-fiber cable with low bending loss optical fibers and pre-connectorized cable with adjustable excess length that employs the 100-fiber cable. We confirmed the feasibility of these cables experimentally.

1. Introduction

The growth in the demand for broadband services using optical access networks has led to a rapid increase in the number of FTTH subscribers during the expansion period. This has resulted in a shortage of ducts and spaces for the installation of optical fiber cables, because many metallic cables are already housed in the available ducts and spaces, and many optical fiber cables are required. Therefore, we need high-density optical fiber cables with a smaller cable diameter that can be easily installed in the small spaces between cables in ducts. However, ultra-high-density cable that uses conventional single-mode (SM) fiber suffers from mechanical stress and so does not have stable optical loss characteristics.

On the other hand, various types of low bending loss optical fiber have been proposed with a view to achieving ease of fiber handling and wiring on residential and business premises. These fibers can be bent with a smaller diameter than conventional SM fiber. However, there has been no practical investigation of cables using these fibers except for optical drop and indoor cables.

Therefore, this paper describes ultra-high-density 100-fiber cable that uses 0.25 mm diameter mono-coated optical fiber with a low bending loss. To enable us to select the optical fiber with the optimum low bending loss and thus obtain the maximum fiber density, we manufactured the cable and examined its characteristics. Moreover, we propose its application as pre-connectorized cable with adjustable excess length designed to improve fiber joint workability. This paper also describes our experimental results.

2. Optical fiber cable structure

Figure 1 shows the structure of an ultra-high-density 100-fiber cable. This cable is composed of 0.25 mm diameter mono-coated optical fibers with a low bending loss, strength members, rip cords and a polyethylene (PE) sheath. The cable has five 20-fiber units and each unit is SZ stranded. Colored threads for identification are wound around units of twenty 0.25 mm diameter mono-coated fibers. Individual fibers in a 20-fiber unit can be identified by the color of their outer coating. The advantages of this cable are its ease of manufacture, easy mid-span access, small diameter and light weight.

3. Cable design and characteristics

We manufactured the ultra-high-density 100-fiber cable shown schematically in Fig. 1. The diameter and weight of the cable were 6.3 mm and 0.03 kg/m, respectively, and its fiber density was about 3 fiber/mm². We incorporated fibers with various low bending loss characteristics and conventional SM fiber in the cable for selecting to allow us to determine the optimum fiber.

We measured the bending loss increases of each fiber at a wavelength of 1.55 μm for different bending radii. The minimum allowable bending radius of each fiber was estimated from these results. Here, the minimum allowable bending radius is defined as the radius for which there is a 0.2 dB/10-turn loss increase caused by bending. The loss change induced by temperature cycling depended strongly on the minimum allowable bending. The measured temperature cycling results are shown in Fig. 2. We found that the optical loss caused by the temperature cycling increased rapidly when the minimum allowable bending radius of the fiber was more than about 12 mm.
Moreover, we measured the transmission, mechanical and midspan access characteristics of the cable using fiber with a minimum allowable bending radius of less than about 12 mm. The results are shown in Table 1. We found that loss changes due to transmission and mechanical test were less than the measurement accuracy. As regards mid-span access, the maximum loss change for all operations is small in relation to the allowable loss of 1 dB. Moreover, the mid-span access operation time is almost the same as that for the conventional cable. Therefore, to obtain stable characteristics, it is necessary to use fiber with a minimum allowable bending radius of less than about 12 mm.

### 4. Pre-connectorized cable with adjustable excess length

When mono-coated optical fibers are used, a large number of fibers must be joined, and fiber joint workability is unsatisfactory. Thus, the fiber joint workability must be improved if we are to construct optical fiber access networks effectively. One method of improving the workability involves the use of optical fiber cables that are pre-connectorized at each end in the factory. When these cables are used and access networks are designed, the optical cable length is generally overestimated to allow for facility data errors and other such factors. Thus, spaces are also required for accommodating the excess cable length. Such spaces are usually hard to find in manholes, and the use of optical fiber cables pre-connectorized at each end has been restricted solely to facilities with more available space. The ultra-high-density 100-fiber cables described above have the attractive features of a small diameter and stable characteristics for small radius bending. Thus, when the cable is coiled with a small diameter and the coiled cable part is expanded and contracted, the estimation error in cable length can be covered and optical fiber cable with both ends pre-connectorized and a coiled cable part can be accommodated in a small space.

Figure 3 shows a pre-connectorized 100-fiber cable with adjustable excess length. The connectors at the both ends are composed of pre-assembled mechanically transferable (MT) connectors. The coil diameter is 150 mm in consideration of the maximum cable accommodation space in a manhole and the loss characteristics related to the expansion and contraction of the coiled part. The coiled cable length is 10 m long and the length of the coiled part is 0.15 m.

![Schematic of pre-connectorized optical cable with adjustable excess length](image)

**Table 2** shows the measured optical loss characteristics of the coiled part when using fiber with a minimum allowable bending radius of less than 12 mm, which is the same as that of the cable. Other important factor for pre-connectorized cable is the connection loss. Table 3 shows the example for connection loss of MT connectors when using hole-assisted optical fiber as one of low bending loss optical fibers. The connection loss is almost same as that of conventional optical fiber. Therefore, we confirmed the feasibility of pre-connectorized cable with adjustable excess length.

### 5. Conclusion

This paper proposed small diameter and ultra-high-density 100-fiber cable with a low bending loss optical fiber and pre-connectorized cable with adjustable excess length utilizing this 100-fiber cable. We confirmed the stable optical loss characteristics of these cables. We also confirmed the low connection loss in the preconnectorized cable when we use a hole-assisted optical fiber. We revealed the feasibility of proposed cables for practical use.

**References**


**Table 1** Transmission, mechanical and mid-span access performances

<table>
<thead>
<tr>
<th>Item</th>
<th>Result</th>
</tr>
</thead>
<tbody>
<tr>
<td>Optical loss increase in cabling processes</td>
<td>≤ 0.04 dB/km</td>
</tr>
<tr>
<td>Lateral force (1590 N/100 mm)</td>
<td>≤ 0.03 dB/km</td>
</tr>
<tr>
<td>Bending (R=160 mm)</td>
<td>≤ 0.01 dB/km</td>
</tr>
<tr>
<td>Squeeze (392 N, R=250 mm)</td>
<td>≤ 0.01 dB/km</td>
</tr>
<tr>
<td>Twist (±90 degrees)</td>
<td>≤ 0.01 dB/km</td>
</tr>
<tr>
<td>Midspan access Operation procedure</td>
<td>≤ 0.7 dB/km</td>
</tr>
<tr>
<td>1. Extract rip cord from sheath</td>
<td>(Operation time ≤ 10 min)</td>
</tr>
<tr>
<td>2. Cut and remove sheath</td>
<td></td>
</tr>
<tr>
<td>3. Extract fibers</td>
<td></td>
</tr>
</tbody>
</table>

**Table 2** Optical loss characteristics of coiled part

<table>
<thead>
<tr>
<th>Item</th>
<th>Result</th>
</tr>
</thead>
<tbody>
<tr>
<td>Optical loss increase due to coil</td>
<td>≤0.02 dB/fiber</td>
</tr>
<tr>
<td>Expanding (up to 10 m)</td>
<td>≤0.04 dB/fiber</td>
</tr>
<tr>
<td>Temperature Cycling (-30~+70, 2cycle)</td>
<td>≤0.02 dB/fiber</td>
</tr>
</tbody>
</table>

**Table 3** Connection loss of MT connector

<table>
<thead>
<tr>
<th>Item</th>
<th>Result</th>
</tr>
</thead>
<tbody>
<tr>
<td>SMF-HAF</td>
<td>0.09dB (Av.), 0.35dB (Max.)</td>
</tr>
<tr>
<td>HAF-HAF</td>
<td>0.19dB (Av.), 0.56dB (Max.)</td>
</tr>
</tbody>
</table>

Measured wavelength: 1.55 μm
10 Gb/s WDM transmission at 1064 and 1550 nm over 24 km PCF with negative power penalties

K. Kurokawa, K. Tsujikawa, K. Tajima, K. Nakajima, and I. Sankawa

NTT Access Network Service Systems Laboratories, NTT Corporation
1-7-1, Hatabashke, Tsukuba, Ibaraki 305-0805 Japan
Tel: +81 29 868 6430, Fax: +81 29 868 6440, E-mail: kurokawa@ansl.ntt.co.jp

Abstract

We achieved the first 10 Gb/s WDM transmission at 1064 and 1550 nm over a 24 km photonic crystal fiber (PCF) with negative power penalty by using the pre-chirp technique with a conventional Z-cut LN modulator.

1. Introduction

Photonic crystal fibers (PCFs) are very attractive transmission media since they have unique features unavailable with conventional single-mode fibers, namely they can be endlessly single-mode and are capable of dispersion tailoring [1,2]. The ultra-wide single-mode region of PCF has provided the possibility of building communication systems with a bandwidth of over 160 THz [3]. Progress on reducing the optical loss of PCF has generated several reports on transmission experiments using PCFs [3-5]. On the other hand, the construction of a high-speed network such as a 100G Ethernet [6] is attracting a lot of attention with the rapid growth of data-centric services. A new optical communication band at 1000 nm in a PCF is very attractive for high-speed networks. This is because we can use ytterbium (Yb3+) doped fiber amplifiers (YDFA), which have a broad bandwidth in the 1 - 1.2 μm wavelength range [7]. Recently, we achieved a 10 Gb/s transmission at 1064 nm over a 24 km PCF with a negative power penalty by using the pre-chirp technique [5].

This paper reports the first 10 Gb/s WDM transmissions at 1064 and 1550 nm over a 24 km PCF.

2. Properties of PCF and experimental setup

We fabricated a 24 km PCF that had 60 holes and the structural parameter d/λ was 0.5. Here, d and λ denote the hole diameter and pitch, respectively, and λ was 7.6 μm. The effective core areas at 1064 and 1550 nm were 44 and 50 μm², respectively. The optical losses were 0.94 and 0.49 dB/km at 1064 and 1550 nm, respectively. The zero dispersion wavelength (λD) was 1184 nm, and the dispersions at 1064 and 1550 nm were -20 and 33 ps/(nm.km), respectively, which corresponds to group velocity dispersions (β₂) of 12 and -42 ps²/km, respectively.

Figure 1 shows our experimental setup. The light sources were a DFB fiber laser operating at 1064 nm and a DFB laser diode (LD) operating at 1550 nm. The input light at each wavelength was modulated at 10 Gb/s with a 2³¹-1 PRBS by using a Z-cut LN intensity modulator. Here, we used an NRZ format. The optical signals at both wavelengths were multiplexed with a WDM coupler and guided into a 24 km PCF. The input powers into the 24 km PCF were set at 7.5 and 3.8 dBm at 1064 and 1550 nm, respectively. The transmitted optical signals were demultiplexed with a WDM coupler. The signals at 1064 nm were amplified with another YDFA and detected with a PIN photodiode. The signals at 1550 nm were also detected with a PIN photodiode. We did not use a preamplifier at 1550 nm.

3. Results and discussion

Figure 2(a) shows the bit error rate (BER) characteristics of the transmission at 1064 nm. The received power was measured at the input of the YDFA. The solid lines with filled and open circles show the baseline BER and the BER after the 24 km transmission, respectively. A BER of 10⁻¹¹ was achieved when the received optical power was -24 dBm. We also confirmed an improvement in the BER performance after the transmission, namely a “negative power penalty” of -0.5 dB at a BER of 10⁻⁹. Since the β₂ (=12 ps²/km) was positive at 1064 nm, and the chirp parameter C of the LN modulator was -0.7, the β₂C value became negative. Therefore we can expect to observe pulse narrowing after the transmission as a
result of the pre-chirp effect [8]. In fact, there was pulse narrowing after the transmission and we confirmed that this was the main reason for the negative power penalty of -0.5 dB [5].

Figure 2(b) shows the BER characteristics of the transmission at 1550 nm. Since the $\beta_2$ value (= -42 ps$^2$/km) was negative at 1550 nm, we effectively changed the chirp parameter $C$ of the LN modulator at 1550 nm to +0.7 by shifting its DC bias voltage by $V_b$. In order to achieve the pre-chirp effect, therefore, we measured the BERs with the inverse polarity condition at 1550 nm. A BER of $10^{-11}$ was achieved when the received optical power was -18.8 dBm. We also confirmed a negative power penalty of -0.3 dB at a BER of $10^{-9}$ after the transmission. To obtain a quantitative analysis of the BER improvement at 1550 nm, we measured the pulse waveforms using a sampling optical oscilloscope and a fixed data pattern of $<$10101011...$. Figure 3 shows the measured normalized waveforms. As seen in the figure, pulse narrowing was observed after the 24-km transmission. We calculated the root-mean-square (RMS) widths from Fig. 3 and obtained a pulse broadening factor $f_b(=\sigma/\sigma_0)$ of 0.90. Here $\sigma$ and $\sigma_0$ are the RMS widths of the output and input pulses, respectively. Therefore, the dispersion-induced power penalty $\delta_0$ is roughly estimated to be -0.5 dB from Eq. (1) [9].

$$\delta_0 = 10 \log_{10} f_b = 10 \log_{10} (\sigma/\sigma_0)$$  \hspace{1cm} (1)

In addition, when we returned the chirp parameter $C$ of the LN modulator to -0.7, we observed pulse broadening. Therefore, we consider the observed negative power penalty at 1550 nm to be also mainly the result of the pulse narrowing caused by the pre-chirp effect.

Figure 4 shows the variation in the $f_b$ value with wavelengths calculated by using Eq. (2) [8], the dispersion characteristics of the PCF and the experimental parameters ($T_{0}=50$ ps, $m=1.5$, $z=24$ km), where we assume that the input pulse is a chirped super Gaussian ($m=1.5$) pulse.

$$f_b = \left[ 1 + \frac{\Gamma(1/2)C_0Z}{\Gamma(3/2)m} + \frac{\Gamma(2-1/2)Z}{\Gamma(3/2)m} \left( \frac{\Gamma(1/2)}{\Gamma(3/2)m} \right)^{1/2} \right]$$  \hspace{1cm} (2)

Here $\Gamma$ is the gamma function, and $T_0$ is the half-width at the 1/e-intensity point of the input pulse. We assumed that $C$ is -0.7 for $\lambda < \lambda_0$ and +0.7 for $\lambda > \lambda_0$ in order to employ the pre-chirp technique. The chirp parameter $C$ of the LN modulator is effectively changed to +0.7 by shifting its DC bias voltage by $V_b$. As shown in Fig. 4, $f_b$ is not larger than 1 for all wavelengths between 1060 and 1600 nm. This indicates that we can expect to realize 10 Gb/s transmission over a 24 km PCF with negligible BER degradation in the 1060 to 1600 nm wavelength region by using the pre-chirp technique with a conventional $Z$-cut LN modulator.

4. Conclusion

We achieved 10 Gb/s WDM transmission at 1064 and 1550 nm over 24 km with negative power penalties by employing the pre-chirp technique with a conventional $Z$-cut LN modulator. We also showed theoretically that we can expect to realize 10 Gb/s transmission over a 24 km PCF with negligible BER degradation in the 1060 to 1600 nm wavelength range by using the pre-chirp technique.

References


361
Splice Characteristics of Trench-Assisted Bend-Insensitive Fiber Having Equivalent Dispersion Characteristics to SMF

T. Nunome, T. Yoshida, J. Takahashi, S. Mat suo
Fujikura Ltd.
1440, Mutsuzaki, Sakura, Chiba, 285-8550, Japan
Tel: +81-43-484-2197, Fax: +81-43-481-1210, E-mail: nunome@lab.fujikura.co.jp

Abstract:
Splice characteristics of trench-assisted bend-insensitive fibers with optimized dispersion characteristics are presented. Since the electrical fields of the fibers are very close to a Gaussian distribution, the fibers have high compatibility with the C-SMF.

1. Introduction
Fiber To The Home (FTTH) service has been spreading rapidly in Japan. Conventional single-mode fibers (C-SMF) that are compliant with ITU-T Recommendation G.652 are widely used for FTTH systems since an economical 1.3-µm Fabry-Perot laser diode is applicable in the system. However, C-SMF used in FTTH is required to have improved bending loss [1], [2], [3]. For instance, the C-SMF for indoor cables and drop cables may be bent in a small radius in a closure or at the corner of a wall.

C-SMF has a trade-off between a bending loss and a mode field diameter (MFD). The MFD of commercially available bend-insensitive fibers with a step-index profile is smaller than that of the C-SMF with a 9.2-µm MFD. However, the small MFD is not preferable in the viewpoint of splice loss.

We have proposed a bend-insensitive fiber with a trench-assisted index profile (hereafter “trench fiber”) [4]. The trench fiber has unique characteristics that the bending loss of the fiber with the same MFD as the C-SMF can be smaller than that of the C-SMF.

Recently, bend-insensitive trench fibers that are compliant with G.652 have been presented [5], [6]. We have reported the design optimization on a trench position for a dispersion characteristic equivalent to the C-SMF. In addition to dispersion characteristics, the trench position should be carefully designed in the viewpoint of splice loss because the distortion of an electric field distribution may be caused by the existence of the trench [7].

In this paper, we investigate the splice characteristics of the trench fiber that shows equivalent dispersion characteristics to the C-SMF.

2. Fiber Design and Fabrication
Figure 1 shows a refractive index profile of a trench fiber. We experimentally fabricated three trench fibers with different MFDs to investigate the splice characteristics of the trench fiber. According to the results presented in [6], r2 / r1 were designed from 3.5 to 4.0 to realize equivalent dispersion characteristics to the C-SMF.

Table 1 shows the measured characteristics of the fabricated trench fibers. The multi-mode reference technique was employed for measuring the cutoff wavelengths of the manufactured fibers [8]. All optical properties meet G.652 Recommendation. Even fiber C, which has the largest MFD among the fabricated fibers, shows a bending loss of 0.09 [dB/turn] at 1550 nm for a bending radius of 10 mm. The bending losses of each fiber are about one-tenth of those of C-SMF having the same MFD as the trench fiber. Also, the chromatic dispersions are very close to that of the C-SMF.

![Fig. 1 Refractive index profile of trench fiber.](image)

Table 1 Measured characteristics of fabricated trench fibers.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Fiber A</td>
<td>1310 nm</td>
<td>1314</td>
<td>1311</td>
<td>1316</td>
<td>0.088</td>
</tr>
<tr>
<td>Fiber B</td>
<td>1550 nm</td>
<td>1550</td>
<td>1549</td>
<td>1550</td>
<td>0.085</td>
</tr>
<tr>
<td>Fiber C</td>
<td>1550 nm</td>
<td>1550</td>
<td>1550</td>
<td>1550</td>
<td>0.086</td>
</tr>
</tbody>
</table>

3. Electric field distribution of trench fibers
If single-mode fibers have Gaussian electric fields, the splice loss between two fibers is calculated by Eq. (1) [9].

$$ T = \left( \frac{2 w_1 w_2}{w_1^2 + w_2^2} \right)^2 \cdot \exp \left( -\frac{2 d^2}{w_1^2 + w_2^2} \right) $$

where T is a power transmission coefficient, 2w1, 2w2 are MFDs of the fibers, and d is a lateral offset between the centers of electric fields of the fibers. Since the electric field of C-SMF is very close to a Gaussian distribution, the splice loss can be estimated by Eq. (1).

On contrast, a fiber with a non-Gaussian electric field, a splice loss is calculated by Eq. (2) [10].
with the C-SMF even in terms of splice characteristics. The dispersion-optimized trench fiber shows high compatibility of the C-SMF. These results indicate that the splice loss of the trench fiber is 0.01 dB in average, which is equivalent to that of the C-SMF. The dispersion-optimized trench fibers were less than 0.06 dB in average. The homogeneous splice losses between the fabricated trench fibers and the C-SMF are less than 0.06 dB in average. The homogeneous splice loss of the trench fiber is 0.01 dB in average, which is equivalent to that of the C-SMF. The dispersion-optimized trench fibers have high compatibility with the C-SMF.

Table 2 Calculated splice losses of trench fibers to C-SMF.

<table>
<thead>
<tr>
<th>Fiber</th>
<th>Splice loss [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>By Eq. (1)</td>
<td>fiber A</td>
</tr>
<tr>
<td>0.05</td>
<td>0.03</td>
</tr>
<tr>
<td>By Eq. (2)</td>
<td>0.05</td>
</tr>
</tbody>
</table>

4. Measurement results on splice loss

We demonstrated splice tests on the fabricated trench fibers using a commercially available fusion splicer (Fujikura Ltd. fusion splicer FSM-40F) without splice condition modified from that for C-SMF. The wavelength for measurement was 1550 nm.

Table 3 shows the measured heterogeneous splice losses between the fabricated trench fibers and C-SMF on 30 samples. Figures 2, 3 and 4 show histograms of measured heterogeneous splice losses respectively. Averaged splice losses are 0.01 to 0.06 dB, which are equivalent to the calculated values. Moreover, the measured homogeneous splice loss of the fiber B is 0.01 dB in average on 30 samples, which is equivalent to that of the C-SMF. These results indicate that the dispersion-optimized trench fiber shows high compatibility with the C-SMF even in terms of splice characteristics.

Table 3 Measured splice losses of trench fibers to C-SMF.

<table>
<thead>
<tr>
<th>Fiber</th>
<th>Splice loss [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Average</td>
<td>fiber A</td>
</tr>
<tr>
<td>0.06</td>
<td>0.04</td>
</tr>
<tr>
<td>Maximum</td>
<td>0.08</td>
</tr>
<tr>
<td>Minimum</td>
<td>0.05</td>
</tr>
<tr>
<td>Standard deviation</td>
<td>0.01</td>
</tr>
</tbody>
</table>

5. Conclusion

We have shown the splice loss characteristics of the trench fibers that their trench positions are optimized for equivalent dispersion characteristics to the C-SMF. Excess splice losses due to distortion of the electric fields from a Gaussian distribution have been confirmed to be negligible. The heterogeneous splice losses between the trench fibers and the C-SMF are less than 0.06 dB in average. The homogeneous splice loss of the trench fiber is 0.01 dB in average, which is equivalent to that of the C-SMF. The dispersion-optimized trench fiber shows high compatibility with the C-SMF in the viewpoint of the splice characteristics.

6. References

Novel fibers for dispersion and nonlinearity management

Siddharth Ramachandran

OFS Laboratories, 25 Schoolhouse Road, Somerset, NJ 08873; sidr@ieee.org

Abstract: Higher order modes in specially designed, few-moded fibers, fabricated by conventional fabrication techniques, are attractive for building all-fiber short-pulse delivery media, sources and amplifiers, because they offer significantly greater flexibility in tailoring dispersion and \( A_{\text{eff}} \) compared to SMF.4

Ultra-short optical pulses, with pulse widths ranging from 50-500 femtoseconds, are attractive for a variety of industrial and scientific applications. However, realising ultra-short pulses in fiber-based systems faces two challenges—dispersion that leads to temporal dilation of pulses as they propagate and nonlinearity, which leads to pulse distortions that are typically non-recoverable.

The ratio of the dispersion length \( L_D \) to nonlinear length \( L_{NL} \) governs whether pulses propagating in a fiber primarily undergo dispersive or nonlinear changes. For \( L_D/L_{NL} \ll 1 \), nonlinear effects are not important, and the pulse operates in the dispersive regime; for \( L_D/L_{NL} \gg 1 \), nonlinear phase changes primarily dominate; and for \( L_D/L_{NL} \sim 1 \), both effects play a role. This quantity is directly related to fiber properties by

\[
\frac{L_D}{L_{NL}} \propto \frac{1}{D \cdot A_{\text{eff}}}
\]

where \( A_{\text{eff}} \) is the effective area of the spatial mode of the fiber in which propagation occurs, and \( D \) is its dispersion.1

Conventional single mode fibers (SMF) offer limited flexibility in obtaining different \( L_D/L_{NL} \) values. In the technologically important wavelengths of 700 – 1060 nm, their dispersion parameter is always negative (normal), and their \( A_{\text{eff}} \) are typically small (10-30 \( \mu \text{m}^2 \)). Photonic crystal (PCF) and air-guided photonic bandgap (PBG) fibers have recently generated a lot of interest due to their enhanced dispersion-engineering capabilities.2

In this talk, we will review an alternative class of fibers that offer the possibility of obtaining a range of dispersion and \( L_D/L_{NL} \) values. Instead of propagating light in the fundamental mode of a fiber, we will show that signal excitation in a single, well-defined higher order spatial mode (HOM) of a specially designed fiber can offer a vastly enhanced design space for tailoring dispersion3.

A commonly adopted solution for transmitting or delivering high-power pulses with fibers is to use large-\( A_{\text{eff}} \) fibers, since this decreases \( L_D/L_{NL} \), thereby mitigating nonlinearities.4 However, arbitrarily increasing \( A_{\text{eff}} \) causes two problems: it makes the fiber more bend-sensitive, and it makes them multimoded and hence susceptible to modal noise.5 Note, however, that \( L_D/L_{NL} \) can be decreased and nonlinearities mitigated also by increasing the dispersion of the fiber. This is readily achieved with HOM fibers, in which dispersion can be designed to be an order of magnitude higher than in conventional single mode fibers. Dispersion as high as -900 ps/nm-km at 840 nm have been achieved in HOM fibers, resulting in distortion-free delivery of pulses with energy up to ~ 1 nJ – an order of magnitude higher than in SMF.6 Such HOM fibers would prove very attractive for building flexible 2-photon endoscopes and delivery media, since they offer nonlinearity-free performance without the need for sacrificing bend-resistance and robust propagation (as done with conventional fibers by increasing \( A_{\text{eff}} \)).7

A second attractive attribute of HOM fibers is the ability to achieve anomalous dispersion at \( \lambda < 1300 \) nm – the zero-dispersion wavelength for silica. A majority of short-pulse devices operate in the spectral range of 700-1100 nm, where conventional SMF can yield only normal (negative) dispersion. HOMs, in contrast can yield large anomalous dispersion in this spectral range, with \( A_{\text{eff}} \sim 10x \) larger than PCFs, due to their unique modal evolution properties. This novel class of HOM fibers will have far reaching implications for the design of fiber-based short-pulse devices in the visible and near-IR wavelength ranges, because this fiber provides the low-loss, bend- and nonlinearity- resistant operation of conventional fibers in a wavelength range where conventional fibers cannot achieve anomalous dispersion.8

A ring-type polarization-maintaining \( \lambda/4 \)-shifted distributed feedback fiber laser pumped by a 0.98 \( \mu \)m laser diode

Quynh Anh HO THI, Akihito SUZUKI, and Masataka NAKAZAWA
Research Institute of Electrical Communication, Tohoku University
2-1-1, Katahira, Aoba-ku, Sendai-shi 980-8577, Japan
Tel: +81-22-217-5525, Fax: +81-22-217-5524, Email: quynhanh@riec.tohoku.ac.jp

Abstract: We report a ring-type polarization-maintaining \( \lambda/4 \)-shifted DFB fiber laser with 0.98 \( \mu \)m pumping. Such output characteristics as output power, linewidth, and relative intensity noise have been greatly improved.

Introduction
\( \lambda/4 \)-shifted distributed feedback (DFB) fiber lasers [1], [2] are attractive for dense wavelength-division multiplexing (WDM) systems and sensing applications due to their robust single-mode operation, narrow linewidth, fiber compatibility, and ease of fabrication. A reduction in the threshold and an improvement in the linewidth have been reported by employing \( \lambda/4 \)-shifted DFB fiber lasers with a ring configuration [3]. However this ring-type \( \lambda/4 \)-shifted DFB Er-doped fiber laser suffers from self-pulsation, which is caused by ion pairs in a highly Er-doped fiber acting as saturable absorbers [4].

We have recently reported a ring-type polarization-maintaining \( \lambda/4 \)-shifted DFB fiber laser using 1.48 \( \mu \)m pumping [5], in which the self-pulsation can be suppressed by adopting a ring configuration and co-doping erbium fiber with aluminum [6]. In addition, with the ring type configuration, we were able to achieve a narrow linewidth and low relative intensity noise (RIN) characteristics.

In this paper, we report a large improvement in the output characteristics of the ring-type polarization-maintaining \( \lambda/4 \)-shifted DFB fiber laser by using 0.98 \( \mu \)m pumping. We achieved an output power of 0.6 mW with a pump power of 100 mW, a linewidth as narrow as 3 kHz, and low RIN of less than -130 dB/Hz for frequencies above 0.6 MHz.

PM \( \lambda/4 \)-shifted DFB fiber laser using 0.98 \( \mu \)m pumping

A PM \( \lambda/4 \)-shifted DFB fiber laser was fabricated in a PANDA-type Er-doped germanosilicate fiber. To reduce ion clustering, a large amount of aluminum was co-doped into the fiber. The \( \lambda/4 \)-shifted FBG was inscribed by scanning a UV beam (266 nm). The UV laser consisted of the fourth-harmonic generation of a Q-switched YAG laser. The 140 mm FBG was fabricated by using a phase mask and the scanning beam technique [7]. A permanent \( \lambda/4 \)-shift FBG was established by using a phase mask with a phase shift of \( \pi \) in the middle. From the measured reflection spectrum of the fabricated FBG, the 3 dB bandwidth of the stop-band was 4 GHz. In our calculation this corresponds to a grating strength of \( kL=8 \).

With 0.98 \( \mu \)m pumping, the fabricated PM \( \lambda/4 \)-shifted DFB fiber laser oscillated in a single longitudinal mode. However, the laser exhibited self-pulsation when the pump power was less than 60 mW, and CW operation was achieved when the pump power was increased to more than 60 mW. The linewidth measured by the delayed self-heterodyne technique with an 80 km delay line was 18 kHz. The RIN was less than -130 dB/Hz above 0.6 MHz and the peak RIN value was -90 dB/Hz at a relaxation oscillation frequency of 0.3 MHz with 100 mW pumping.

Ring-type PM \( \lambda/4 \)-shifted DFB fiber laser using 0.98 \( \mu \)m pumping

Figure 1 shows the configuration of a ring-type PM \( \lambda/4 \)-shifted DFB fiber laser using a 0.98 \( \mu \)m pump LD (laser diode). Pump light from the 0.98 \( \mu \)m LD was coupled to a \( \lambda/4 \)-shifted EDF-FBG through a PM WDM coupler. We suppressed the self-pulsation by co-doping the EDF with a 0.4 wt% concentration of Er and a 12 wt% concentration of Al. With a 20/80 PM coupler, 80 % of the power was led to the laser output and 20 % was returned to the ring cavity to obtain a high output power. In this ring configuration, a linearly polarized light aligned with a slow axis propagated unidirectionally. The laser output from the PM coupler was coupled to a 2 nm optical band pass filter to eliminate the pump light.
Because of the ring configuration, the signal passes repeatedly through the narrow passband filter at the lasing wavelength that is formed by the \( \lambda/4 \)-shifted grating. The laser signal is stabilized as a result of the noise reduction realized by this ring cavity configuration.

**Output characteristics**

The oscillation wavelength of this laser was 1542.5 nm and the optical signal-to-noise ratio was 70 dB. Figure 2 shows the output power as a function of pump power with 0.98 and 1.48 \( \mu \)m pumping. The laser threshold was 10 mW in both cases. Heterodyne beat measurement with a single-mode semiconductor laser confirmed the CW operation of the laser. It is clearly seen that the output power was significantly increased by using 0.98 \( \mu \)m pumping. For example, for a pump power of 100 mW, the output power was 0.6 mW with 0.98 \( \mu \)m pumping, which was 2.5 times larger than that with 1.48 \( \mu \)m pumping.

The optical linewidth was measured using the delayed self-heterodyne technique with an 80 km delay line. The beat spectrum is shown in Fig. 3. With a Lorentzian fitting, the linewidth is estimated to be 3 kHz for pumping at 0.98 \( \mu \)m and 2 kHz for pumping at 1.48 \( \mu \)m. The small difference is due to the limited measurement resolution. With an 80 km fiber delay, the measurement resolution was limited to 1.3 kHz. In addition, as shown in Fig. 3, the tail of the spectrum is narrower with 0.98 \( \mu \)m pumping than with 1.48 \( \mu \)m pumping, and the relaxation oscillation peak at 0.25 MHz is suppressed.

Figure 4 compares RIN measurements with 0.98 and 1.48 \( \mu \)m pumping. The peak RIN value was -110 dB/Hz at a relaxation oscillation frequency of 0.25 MHz with 0.98 \( \mu \)m pumping and -108 dB/Hz at a relaxation oscillation frequency of 0.31 MHz with 1.48 \( \mu \)m pumping. For pumping at 0.98 \( \mu \)m, the measured RIN was less than -130 dB/Hz for frequencies above 0.6 MHz. For all frequencies, the RIN was below that obtained with 1.48 \( \mu \)m pumping. These improvements are attributed to the high population inversion and large absorption cross-section realized with 0.98 \( \mu \)m pumping.

**Conclusion**

By employing a previously proposed ring cavity configuration and using 0.98 \( \mu \)m pumping, we greatly improved the output power characteristics of a PM \( \lambda/4 \)-shifted DFB fiber laser, where we took advantage of the noise reduction that resulted from 0.98 \( \mu \)m pumping. As a result, the output power was increased 2.5 times, and the linewidth and RIN characteristics were also improved.

**References**

5. A. Suzuki, et al., ECOC2006, Th2.3.2.
Frequency Locking of a Single-Frequency Fiber Laser with Dual-Wavelength External Frequency-Stabilized Light Source

Kaoru MORI†, Motoharu MATSUURA,† and Naoto KISHI†

†Department of Information and Communication Engineering, University of Electro-Communications
1–5–1 Chofugaoka, Chofu-shi, Tokyo, 182–8585 Japan
E-mail: †{mkaoru, matsuura, kishi}@ice.uec.ac.jp

Abstract—We investigated the frequency control characteristics of a narrow line-width single frequency fiber laser locked to one of the laser frequencies of two external frequency-stabilized lights with each different wavelength.

Introduction

In future dense wavelength division multiplexing (DWDM) systems, narrow line-width light sources with high frequency stability are indispensable for DWDM backup sources with ITU-T grid. Fiber lasers enable us to realize narrow line-width operation due to the long cavity length. So far, various fiber lasers locked to the ITU-T grid have already been reported using periodic filters [1-3]. However, it is very difficult to control precisely a oscillation frequency, because the frequency fluctuates easily due to environmental perturbation compared to semiconductor laser diodes. To control such frequency, we proposed previously a single-frequency fiber laser using a saturable absorber controlled with a frequency-stabilized external light source [4, 5]. The oscillation frequency was successfully tuned to the frequency of the external light, and the high frequency stability was achieved.

In this paper, we apply this frequency locking technique for the external light source with dual-wavelength spectrum. The oscillation frequency of the fiber laser can be locked to one of the external light source by adjusting the bandpass filter in the laser cavity. The obtained frequency stability is as high as the stability of the external light source.

Experimental setup

Figure 1 shows the setup of a single-frequency fiber laser [4,5]. Traveling-wave laser cavity is employed. This cavity involves un-pumped erbium-doped fiber (EDF) as a saturable absorber. The lasing wavelength is roughly determined by a band-pass filter (BPF) of 0.6 nm band-width. The gain and absorption peak wavelengths of the employed EDF amplifier (EDFA) and un-pumped EDF are around 1560 nm. The laser cavity length is 18.2 nm. The measured line-width of the fiber laser is approximately 4 kHz. At the wavelength of 1560 nm, the slope efficiency and the oscillation threshold without the external light injection are 18.56 % and 11.23 mW, respectively. To optimize the interference between the oscillation and external lights in the saturable absorber, the polarization controller (PC) is employed to adjust to the state of the polarization of the oscillation light.

When the wavelength difference between oscillation and external lights is large, the oscillation frequency is not locked to the external light. On the other hand, the frequency can be locked as the oscillation wavelength is close to the wavelength of the external light. Such frequency control is achieved by cooperatively induced spatial hole-burning (SHB) of the absorption formed by the lasing and the external lights in the saturable absorber [4, 5]. Previously, we demonstrated the frequency control of the single-frequency fiber laser using an external light. In this work, we employed a dual-wavelength external light source whose wavelength spacing corresponds to the frequency grid of ITU-T. By tuning the BPF in the cavity, the oscillation light can be frequency-locked to one of the external lights. We investigate the frequency stability of the fiber laser for various injected powers and wavelength spacings of the two external lights.

Experiment and results

To investigate the performance of frequency locking of the proposed method, we measured the frequency stability with changing the wavelength spacing between the two external lights. The frequency
stability is quantitatively estimated by employing Allan variance [6, 7]. For example, Allan variance of the fiber laser at free-running operation is in order of several hundreds MHz², whereas the value of the employed external light source is about 10 MHz².

If the BPF is tuned to one of the external light wavelength, the frequency of the fiber laser is locked to the external light. Figure 2 shows Allan variance of the fiber laser controlled by the dual-wavelength external light as the function of the injected powers of the external lights. In the case of 0.8 nm wavelength spacing, the wavelength of the injected external lights are 1546.95 nm and 1547.75 nm, respectively. In Fig. 2 (a), the oscillation wavelength of the fiber laser is locked to the longer wavelength external light. When the injected power of the shorter wavelength external light (un-locking light) is changed, no remarkable change of the frequency stability is observed. On the other hand, it can be clearly seen that the power of the locking light strongly affects the stability. Moreover, we obtain similarly high stability for wavelength spacing above 0.2 nm. On the other hand, for the wavelength spacing of 0.1 nm, good frequency stability is not obtained regardless the injected powers of the locking and un-locking lights. We believe that this behavior is due to broad bandwidth of the employed BPF (0.6 nm) in the cavity. Further improvement of the stability will be possible using BPF with a narrower bandwidth.

Figure 3 shows Allan variance of the fiber laser when the oscillation wavelength is tuned in the range of 1546.94 nm to 1547.26 nm by the BPF. The wavelengths of the external lights are 1547.00 nm and 1547.20 nm, respectively. These injected powers are set to 5.5 dBm. As shown in Fig. 3, the lasing frequency is successfully locked to either of the external light with high frequency stability when the oscillation wavelength of the fiber laser is close to the corresponding external light wavelength.

**Conclusion**

We proposed and demonstrated a frequency locking with dual-wavelength external light injection and investigated the frequency control characteristics. The oscillation frequency was successfully locked to the frequency of the external light with high frequency stability for the external wavelength spacing above 0.2 nm. If external light sources with the ITU-T grid are employed, the proposed single-frequency fiber laser will be applicable to a narrow line-width operation corresponding to the ITU-T grid. Such light source will be useful for future DWDM transmission systems.

**References**

Active Birefringent Optical Loop Filter using an SOA as a Phase Shifter
Noriaki Onodera, Takafumi Mansyo, Kenichiro Tsuji, Terpe Yamaguchi, Jungmin Kim and Masatoshi Saruwatari.
Department of Communications Engineering, National Defense Academy
1-10-20 Hashirimizu, Yokosuka 239-8686, Japan.
Phone: +81-46-841-3810(ext.3371), Fax: +81-46-844-5911, E-mail: noriaki@nda.ac.jp

Abstract
Polarization-independent active birefringent optical fiber loop filters using a semiconductor optical amplifier (SOA) as a phase shifter are studied experimentally. By driving the SOA with DC/sinusoidal current, the SOA is operating as a phase shifter thus the active control of the periodic transmission/reflection properties of the loop filter are realized.

Introduction
Optical fiber loop filters with a birefringent fiber and a polarization controller (PC) were reported as periodic wavelength filters for applications in wavelength division multiplexing (WDM) or multi-wavelength fiber lasers [1,2]. Recently, active operation of the loop filters introducing a pair of phase modulators inside the loop has been reported [3]. However, the filter requires more complicated structure and the characteristic differences between the two phase modulators degrade the performance of the filter. In this paper, we report the polarization-independent, active birefringent fiber loop filters using a 1.3μm SOA as a phase shifter inside the loop. To our knowledge, this is the first experimental report of an active birefringent loop filter using an SOA inside the loop.

Experimental setup
Figure 1 shows the basic configuration of an active birefringent fiber loop filter with an SOA as a phase shifter. The loop filter consists of two polarization controllers (PC-A and PC-B), a 1.3μm SOA, polarization-maintaining (PM) fiber, and a PM 3dB fiber coupler. As light sources, an external cavity laser (ECL) and ASE light from an EDFA are combined together using a 7dB fiber coupler, and then introducing to the loop filter through a circulator which separates the reflected signal from the filter. The PC-A is set to produce a polarization rotation of 90 degrees and the PC-B is set to adjust the direction of the fast/slow axes of the PM fiber and the SOA since the SOA has ordinary single-mode pigtail fibers. The basic operating principle is the same as written in [3].

In order to evaluate the polarization mode dispersion (PMD) of the SOA, we perform the polarization mode dispersion measurement of the SOA using a cross-connect method. Since the PMD of the SOA is estimated to be very small, it is hard to measure the value directly. So, we take the following steps. For simplicity, we assume the slow axis of the SOA is along with x-axis.

Step 1: By controlling both the PC-A and PC-B, we set the half-wave (HW) plate to completely flatten the transmission spectrum with minimum output power. This situation corresponds to the polarization rotation angle (PRA) at AC=0 degree and CB=0 degree.

Step 2: By rotating the HW plate of the PC-A, we create the periodic transmission spectrum with maximum extinction ratio. This corresponds to PRA at PC-A=90 degrees and PC-B=0 degree.

Step 3: Then rotating the HW plate of the PC-B, we create the flat spectrum again. This corresponds to PRA at PC-A=90 degrees and PC-B=90 degrees.

Step 4: Then rotating the HW plate of the PC-A to create the periodic transmission spectrum with maximum extinction ratio again. This corresponds to PRA at PC-A=0 degree and PC-B=90 degrees.

In case of step 2, the total PMD of the loop becomes \( \Delta \text{PM} + \Delta \text{SOA} \), and in case of step 4, the total PMD becomes \( \Delta \text{PM} - \Delta \text{SOA} \), where \( \Delta \text{PM} \) and \( \Delta \text{SOA} \) correspond to PMDs of the PM fiber and the SOA used in the loop, respectively. Figure 2 shows the measured results of transmission spectra corresponding to the “Step 2” and “Step 4” at 30mA driving current of SOA. In this measurement, a PM fiber with 1.43ps PMD is used. By calculating the total PMD for both spectra, PMD for the SOA is obtained as 0.036ps.

Figure 3 shows the measured results of the PMDs of SOA depending on driving current. It is clear that by increasing the driving current, the PMD of SOA is...
decreasing from approximately 37fs to 20fs. This means that by changing the driving current of the SOA, the PMD, i.e., the phase of the signal which passes through the SOA can be controlled by the current. Based on the above results, we perform the control of the transmission spectrum using DC driving current.

Figure 4 shows the measured transmission spectra of the loop filter using the ASE and ECL light. In the experiment, PC-A is set to 0 and PC-B is set to 90 degrees, thus the injection current increases the phase shift inside the loop filter. In this figure, the driving current of SOA is set to 0mA, 50mA, 100mA and 150mA. By increasing the driving current, the peak of the periodic transmission spectrum (P_0) shifts to longer wavelength, and the transmission peak at 150mA (P_{150}) becomes at the reflection peak at 0mA. This means that the ECL signal which is set to the transmission peak at 0mA is switched from the transmission port to the reflection port in fig. 1. Also by increasing the driving current, the periodic spectrum intensity of the ASE is decreasing. The reason is that the loss of 1.5μm wavelength is increasing by increasing the drive current of the SOA.

Figure 5 shows the modulated outputs at the transmission and reflection ports using a 10-kHz sinusoidal wave. The sinusoidal wave is connected to the external modulation port (Bandwidth: 300 kHz) of the drive current source. Both outputs are modulated at 10-kHz frequency but the waveforms are inverted because of the switching of the output port. As shown in the figure, the modulation amplitude at the reflection port is almost 50% of that of the transmission port. This is due to the loss of the 1.5μm wavelength at 150mA driving current as indicated in figure 4. Figure 5 indicates that the SOA-introduced birefringent loop filter can actively switch the input signal between the input and output ports by the driving current of the SOA.

Conclusions

By using an SOA as a phase shifter, the active birefringent loop filter is realized. The periodic transmission/reflection characteristics can be shifted by the injection current to the SOA, and the signal switching between the transmission and reflection ports can be realized. By improving the drive electronics, the proposed filter can be operated in the GHz frequency region which is suitable for high-speed optical networks.

References

High birefringence ring filter with a reflector: application in single-frequency fiber lasers

Guoyong Sun, Dae Seung Moon, Dusun Hwang and Youngjoo Chung
Department of Information and Communications, Gwangju Institute of Science and Technology (GIST)
Oryong-dong, Buk-gu, Gwangju 500-712, Korea
Tel: +82-62-970-2214, Fax: +82-62-970-3137, Email: yehung@gist.ac.kr

Abstract
A high birefringence fiber ring filter with an inline reflector is newly proposed. Narrowband reflection peaks with large effective free spectral range are achieved by the intrinsic vernier effect between the traveling orthogonally polarized lights along the fast and slow axes in the high birefringence fiber. The special properties facilitate its application in single frequency fiber lasers.

1 Introduction
Single frequency fiber lasers [1-2] are of considerable interest for various applications in medical surgery, communication systems, and sensors. There are a number of methods to achieve stable single frequency operation in fiber lasers. For example, the single frequency erbium-doped fiber lasers have been obtained in both linear and ring configurations. However, it is necessary to shorten the length of the gain medium to increase the free spectral range (FSR) in linear cavities, which limits the attainable output power [3]. On the other hand, complex rings [4] are usually used as frequency selectors for lasing single frequency in ring-cavity fiber lasers. However, the complex rings are sensitive to environment fluctuations although they increase the FSR to alleviate the cavity-length problem through the vernier effect. Recently, a feedback Mach-Zehnder resonator with a reflector was proposed to increase the FSR [5], but it may not be sufficiently stable just like the common Mach-Zehnder comb filter for the optical field circulating in two different resonant cavities. A common problem with most of the coupled-cavity approaches is the stringent environmental stability required.

In this paper, characteristics of a high birefringence fiber (HBF) ring structure with an inline reflector are described. Distinct from the common fiber ring resonator, the single mode fiber is replaced with a piece of HBF. By employing the special setup, we can resolve the problems encountered by the ring resonators used as narrowband filters in single frequency fiber lasers as mentioned above. The proposed setup provides improved performance while possessing a simple structure, which dramatically facilitates its application in single frequency fiber lasers.

2 Principle of operation
Figure 1 shows the setup of our proposed HBF ring resonator. The simple configuration consists of one fiber coupler (OC) with intensity coupling ratio $k$, an inline reflector with intensity reflectivity $r$ and a piece of HBF with length $L (= L_1 + L_2)$. The refractive indexes of the fast and slow axes in the HBF are $n_f$ and $n_s$, respectively.

![Fig. 1 The schematic setup of our proposed HBF ring resonator.](image)

When the optical field with amplitude $E_i$ is input from the left port, it will be decomposed along the slow and fast axes in the HBF according to the formula (1). Through calculation, we obtain the amplitude ratios:

$$ E_i = A_1 E_i^s + A_2 E_i^f $$

(1)

$$ \frac{E_i^s}{E_i^f} = \frac{\frac{-ik \sqrt{Re^{2\pi i/n_f}\lambda L}}}{\sqrt{(1-k)e^{2\pi i/n_f}\lambda L}} = -\sqrt{2(1-k)(1-R) + e^{2\pi i/n_f}\lambda L} $$

(2)

$$ \frac{E_i^f}{E_i^s} = \frac{\frac{-ik \sqrt{Re^{2\pi i/n_s}\lambda L}}}{\sqrt{(1-k)e^{2\pi i/n_s}\lambda L}} = -\sqrt{2(1-k)(1-R) + e^{2\pi i/n_s}\lambda L} $$

(3)

It can be seen from the Eqs. (2) and (3) that the two terms are both periodic in $1/\lambda$ with slightly different periods. Only when the circulating fields along the fast
and slow axes in the HBF ring are both resonating, the reflectivity is maximal. It follows that the effective FSR is enhanced by the vernier effect between the fast and slow axes. Namely, the FSR is \( \frac{n_f}{n_r - n_f} \) times that of the circulating light along the fast axis, \( i.e.: \)

\[
FSR = \frac{c}{(n_r - n_f)}L = \frac{n_f}{n_r - n_f} FSR_f = \frac{n_r}{n_r - n_f} FSR_r
\]

3 Calculation and discussion

In our calculation, all losses are ignored since they have little impact on the performance. The input signal is assumed to be adjusted by a polarization controller such that \( |A_1|^2 = |A_2|^2 = 0.5 \).

Figure 2 shows the reflection spectra with \( k=1\% \) and the corresponding critical \( R \) value [5]. The values of other parameters involved are: \( L=0.8m, \ n_r=1.48, \ n_f=1.46 \). Therefore, the ratio of FSR, to FSR, is 73/74. As a result, the effective FSR is enhanced by 73 times, which can be seen from Fig. 2(c). The inset shows one reflective peak with full width at half maximum as narrow as about 280 kHz. The maximal reflectivity occurs at frequencies where the circulating light along the fast and slow axes are both resonating. At frequencies where only one of the fast and slow components resonates, the reflectivity drops by 3 dB. There is no power reflected from the input port at frequencies that neither circulating light is resonating. The effective FSR is thus dramatically increased by the vernier effect without shortening the fiber length. The side reflectivity suppression ratio is 3 dB, which is enough to obtain the lasing field only at the reflection peak when used in single frequency fiber lasers. The smaller the difference between the fast- and slow-axis indexes, the more the effective FSR is enhanced.

4 Conclusion

A HBF ring resonator with an inline reflector is newly proposed. Owing to the difference between the refractive indexes of the fast and slow axes in the HBF, the effective FSR of the HBF ring can be dramatically increased by the vernier effect. Since there is no need for compound resonators for generating the vernier effect, the proposed structure is simple but still provides with enhanced performance. All special properties improve the feasibility of making single frequency fiber lasers by employing the HBF ring resonator as a filter.

Acknowledgements

This work was partially supported by KOSEF through Grant No. R01-2006-000-11088-0 from the Basic Program Project and by the Second Phase of the Brain Korea 21 Project.

5 References

Glass Material Technology for Optical Amplification Media: Past, Present and Future

Yasutake Ohishi

Research Center for Advanced Photon Technology
Toyota Technological Institute
2-12-1 Hisakata, Tempaku, Nagoya 468-8511, Japan
E-mail: ohishi@toyota-ti.ac.jp

In order to reduce the costs for the transmission, it was necessary to minimize expensive elements like transmitters, receivers and regenerators in the optical network. At the same time the maximum data rate and transmission distance had to be increased drastically in the early days of optical telecommunications. This goal could be reached with the invention of the erbium-doped fiber amplifier (EDFA). Since this device allowed the amplification of the signal in the optical domain, transformation into the electrical domain was no longer necessary. Hence a receiver, transmitter, and regenerator were spared with every optical amplifier.

The bit rate in the communication systems in the 1980s was 155 Mbit/s, ten years later it increased to 2.5 Gbit/s. Today 10Gbit/s is widely used and 40 Gbit/s is being used. Data rates up to over 10Tbit/s have been performed in the laboratories. Since the middle of the 1990s, optical telecommunication devices have been transmitted more than one wavelength over a fiber span to better exploit the usable bandwidth of the optical fibers. The technique, dense wavelength division multiplexing (DWDM), transmits different channels at the same time in one fiber with different carrier wavelength. The optical amplifiers have played an important role in this innovation. The studies on the usable bandwidth expansion for rare-earth-doped fiber amplifiers have been performed energetically in the 1990s and new fiber amplifiers have been developed. The research of rare-earth-doped fiber amplifiers has led to the development of high power and ultra-short pulse fiber lasers.

We should remind of nonlinear effect in optical fibers, such as four wave mixing (FWM), Raman and Brillouin scattering. They are able to amplify optical signals in areas that can never be reached by rare-earth-doped fiber amplifiers. They are opening new aspects for optical signal processing as well as optical amplifications. In this talk, I will review glass material technology for optical gain media and give a future prospect of novel materials for future system applications.
Biography

Yasutake Ohishi

Yasutake Ohishi received the B. S. and M. S. degrees in physics from Tohoku University, Sendai, Japan in 1978 and 1980, respectively. In 1980, he joined the Electrical Communication Laboratory, Nippon Telegraph and Telephone Public Corporation (now NTT), where he was engaged in research on fluoride fiber technology. In 1988, he was awarded a Ph.D. by Tokyo Institute of Technology, Tokyo. From 1989 to 1990 he was a visiting scholar at Rutgers University. After returning to NTT in 1990, he was engaged in research and development of rare-earth-doped fiber amplifiers. He invented praseodymium-doped fiber amplifiers and the tellurite-based erbium-doped fiber amplifiers. In 2002, he joined Toyota Technological Institute, Nagoya. He is a professor of Future Industry-oriented Basic Science and Materials and the Director of Research Center for Advanced Photon Technology.

Dr. Ohishi is a member of the Optical Society of America, IEEE, the Institute of Electronics, Information, and Communication Engineers of Japan, the Japan Society of Applied Physics, and the Ceramic Society of Japan.
11 dB gain in Tm-doped Silica Glass Fiber

Pramod R. Watekar, Seongmin Ju and Won-Taek Han
Department of Information and Communications
Gwangju Institute of Science and Technology
1 Oryong-Dong, Buk-Gu, Gwangju, 500-712, South Korea.
Tel. +82-62-970-2215, Fax. +82-62-970-2204, e-mail: wthan@gist.ac.kr

Abstract
We report for the first time, highest ever reported gain for Tm-doped silica glass fiber. We measured the gain of about 11 dB at 1470 nm upon pumping with 1064 nm Ar-ion laser.

1 Introduction
Recently, increasing demand for optical bandwidth has led researchers to investigate the bands other than popular EDFA band (1530 nm - 1600 nm). Various reports have been published for 800 nm and 1470 nm (S-band) optical amplifiers [1-4] using Tm as the doping element in optical fiber core. However, because of high phonon energy associated with the Tm-doped silica glass fiber, most of the reported amplifiers so far are with ZBLAN fiber [1] and fluoride fiber [2], which are incompatible with the existing silica glass fiber network. Our group has recently demonstrated a small gain in Tm-doped silica glass fiber [3, 4].

In this paper we report, for the first time to the best of knowledge of the authors, highest ever measured gain at 1470 nm in Tm-doped optical fiber upon 1064 nm pumping.

2 Tm Energy Levels
The energy level diagram of Tm-ions in silica glass fiber is shown in Fig. 1. When Tm-doped optical fiber is pumped with a 1064 nm laser source, the ground state ions in \(^3\text{H_6}\) energy level are excited to the energy level \(^3\text{H_5}\) and further from level \(^3\text{F_4}\) to \(^3\text{F_2}\) (or \(^3\text{F_3}\)) due to strong excited state absorption (ESA) at \(^3\text{F_4}\) level for 1064 nm pump. The \(^3\text{H_4}\) level-initiated emissions can be at 2300 nm (\(^3\text{H_4}\rightarrow^3\text{H_5}\)), 1470 nm (\(^3\text{H_4}\rightarrow^3\text{F_4}\)) and 800 nm (\(^3\text{H_4}\rightarrow^3\text{H_6}\)), of which 800 nm emission is the most efficient [1-4].

![Energy Level Diagram](image)

Fig. 1. The energy level diagram of Tm-ions in silica glass optical fiber.

3 Experiment
The alumino-germano-silicate optical fiber preform was fabricated using the MCVD process and it was doped with the Tm\(^{3+}\) ions using the modified solution doping technique developed earlier by our group [5]. The optical fiber was drawn from this preform with the outer diameter of 125 \(\mu\text{m}\). The numerical aperture was 0.2 and the core diameter was 3.5 \(\mu\text{m}\). The estimated concentrations of Tm ions was about \(7 \times 10^{22}\) ions/m\(^3\). The measured attenuation was 1.8 dB/m at 1064 nm and 0.07 dB/m at 1470 nm. The gain...
characterization of Tm-doped silica glass fiber was carried out by coupling pump power at 1064 nm (Ar-ion laser, 2W) and signal at 1470 nm (ASE source, -72 dBm at 1470 nm) into Tm-doped silica glass fiber using a WDM coupler. The resulting emission spectrum was measured using the optical spectrum analyzer.

4 Results and Discussion

When the Tm-doped optical fiber was pumped with the 1064 nm Ar-ion laser, emission centered at 1462 nm was observed as shown in Fig. 2. The 3 dB bandwidth of emission was around 97 nm.

Fig. 2. Typical emission spectrum of Tm-doped silica glass fiber upon pumping with 1064 nm Ar-ion laser.

The gain characteristics of Tm-doped silica glass fiber (10 m) upon 1064 nm pumping is shown in Fig. 3. The threshold pump power to start amplification is about 500 mW. The maximum gain at 2 W of pump power was 11.3 dB. The quantum conversion efficiency of Tm-doped silica glass fiber amplifier was found to be very low, in the present case it was about 0.016%.

Fig. 3. Gain variation of Tm-doped silica glass fiber with respect to pump power at 1064 nm. (Fiber length = 10 m).

5 Acknowledgements

This work was supported in part by: GIST Technology Initiative (GTI), South Korea; Brain Korea-21 Information Technology Project, Ministry of Education and Human Resources Development, South Korea and, National Core Research Center (NCRC) for Hybrid Materials Solution of Pusan National University, South Korea.

6 References

Energy transfer analysis for Tb$^{3+}$-Yb$^{3+}$ codoped fluorophosphates glasses

T. Yamashita$^{1,2}$, S. Horiguchi$^{1}$, Y. Arai$^{1}$, T. Suzuki$^{1}$, and Y. Ohishi$^{1}$
$^{1}$Toyota Technological Institute, $^{2}$Toyota Central R&D Labs., Inc.

Abstract Efficiency for upconversion pumping of Tb$^{3+}$ at the 0.98 µm band was improved in Tb$^{3+}$-Yb$^{3+}$ co-doped system by adopting a low-phonon glass as a host. Tb$^{3+}$-Yb$^{3+}$-codoped fluorophosphates glass is a promising candidate for the 0.54 µm band laser medium pumped at 0.98 µm.

1. Introduction Recently, there is a growing interest for visible fiber lasers and amplifiers because of their potential applications in the fields of optical data storage, biochemical spectroscopy, and visible light communication network. Recently, we propose the use of a Tb$^{3+}$-doped fluoride fiber as a practical gain medium for fiber lasers and amplifiers at 0.54 µm based on a spectroscopic investigation and the first demonstration, to our knowledge, of signal amplification at the 0.54 µm band[1]. Furthermore, it has been studied that co-doping of Yb$^{3+}$ in Tb$^{3+}$-doped glass leads to appearance of visible luminescence, which is explained by the Yb$^{3+}$-Tb$^{3+}$ cooperative upconversion process[2]. In this process, as illustrated in Fig.1, two excited Yb$^{3+}$ ions simultaneously transfer their energy to a Tb$^{3+}$, so that Tb$^{3+}$ can be excited to the $^5D_4$ level. Because Yb$^{3+}$ ions have the strong absorption cross section in the region of 0.98 µm, we can take advantage of high-performance 0.98 µm laser diode to excite Tb$^{3+}$ ions. Thus, if this energy transfer process can be used as practical pumping scheme for Tb$^{3+}$, it would be useful for the reduction in size of laser devices. But the study of the energy transfer efficiency among Yb$^{3+}$ and Tb$^{3+}$ has been superficial, so far. In this study, we investigated energy transfer efficiency of Tb$^{3+}$-Yb$^{3+}$-codoped system, focusing especially on phonon-assisted transfer (PAT) effects on the energy transfer from Tb$^{3+}$ to Yb$^{3+}$. As a result, it was found that the energy transfer from Tb$^{3+}$ to Yb$^{3+}$ which degrades a net pumping efficiency of Tb$^{3+}$-Yb$^{3+}$ codoped system can be greatly reduced by adopting a low-phonon glass as a host.

2. Experiment In this study, we use a fluorophosphate glass which is a composition of 84NaPO$_3$-16BaF$_2$ mol% as the host material for the Tb$^{3+}$ and Yb$^{3+}$ ions. The doping concentrations of the rare-earth ions were set to between 1–20 mol.% for both Tb$^{3+}$ and Yb$^{3+}$. The emission spectra of Tb$^{3+}$ and Yb$^{3+}$ were measured with a monochromator, using a photoelectron multiplier tube as a detector. The emission measurements for Tb$^{3+}$: $^5D_4 ightarrow ^7F_5$ transition were carried out with an Argon-ion laser with a wavelength of 488 nm as an excitation source. A 974 nm LD was used for the measurement of Yb$^{3+}$: $^7F_{5/2} ightarrow ^7F_{7/2}$ transition. Emission lifetimes were obtained from the first e-folding time of emission decay curves. A ceramics-heater was used to hold the sample temperature constant in the range of 300 ~500 K for temperature dependence measurements.

Fig. 1 Simplified energy diagram of Tb$^{3+}$-Yb$^{3+}$ codoped system.

3. Results and discussion Figure 2 shows the upconversion emission spectrum of Tb$^{3+}$-Yb$^{3+}$-codoped fluorophosphates glass with Tb$^{3+}$ $2.9 \times 10^{21}$ ions/cm$^3$ ($N_{Tb}$) and Yb$^{3+}$ $4.7 \times 10^{21}$ ions/cm$^3$ ($N_{Yb}$). The green emission corresponding to the $^5D_4 ightarrow ^7F_5$ transition of Tb$^{3+}$ could readily be seen by the naked eye under excitation power densities lower than 40 mW/mm$^2$ at 974 nm.

The efficiency of energy transfer from Yb$^{3+}$ to Tb$^{3+}$ (forward transfer) can be calculated simply from[3]:

$$\eta_{FT} = 1 - \frac{\tau_{Yb-Tb}}{\tau_{Yb-rad}},$$

(1)

where $\tau_{Yb-Tb}$ is the lifetime of Yb$^{3+}$ in the presence of Tb$^{3+}$ and $\tau_{Yb-rad}$ is the intrinsic lifetime of Yb$^{3+}$. Figure 3 shows the energy transfer efficiency dependence on the Yb$^{3+}$ concentration. The $\eta_{FT}$ was increased with increasing the
concentration of Yb$^{3+}$, although it tended to saturate with increasing Yb$^{3+}$ concentration. The energy transfer efficiency as high as 26±7 % was attained, and it was found to have much higher than the values observed in crystals[2].

On the other hand, the energy transfer rate from Tb$^{3+}$ to Yb$^{3+}$ (backward transfer):

$$ W_{BT}(T)=\frac{1}{\tau_{Tb}^{Yb}-1/\tau_{Tb}^{rad}}, $$

where $\tau_{Tb}^{Yb}$ is the lifetime of Tb$^{3+}$ in the presence of Yb$^{3+}$ and $\tau_{Tb}^{rad}$ is the intrinsic lifetime of Yb$^{3+}$. According to T. Miyakawa[4], if the PAT can be ascribed to the cause of backward transfer, the rate of PAT is given by:

$$ W_{PAT}(T)=W_{PAT}(0)\left[\exp(h\omega/kT)\exp(h\omega/kT)-1\right]p, $$

where $\Delta E=h\omega$ represents the energy gap difference between the (Tb$^{3+}$: $\varepsilon_D^4\varepsilon_F^7$) and (Yb$^{3+}$: $\varepsilon_F^7\varepsilon_F^{5/2}$), and $p=\Delta E/h\omega$ is the number of phonons emitted in the process. Thus, the contribution of PAT process to backward transfer can be clarified in terms of temperature dependence on the lifetime of the $\varepsilon_D^4$ level of Tb$^{3+}$. The temperature dependence of $W_{PAT}$ obtained from Eq.(2) is shown in Fig. 4. Measured $W_{PAT}$ was increased with increasing temperature, and agreed well with the fitting curve by Eq.(3) using the phonon energy of 530 cm$^{-1}$ for the energy gap difference of 4122 cm$^{-1}$, as represented in Fig. 4. A value of 1.8 s$^{-1}$ was found for $W_{PAT}$ at room temperature. This value is considerably lower compared to about 470 s$^{-1}$ of Tb$^{3+}$:Yb$^{3+}$-codoped borosilicate glasses which have higher phonon energy of 1100–1400 cm$^{-1}$. These results show that the energy transfer from Tb$^{3+}$ to Yb$^{3+}$ which degrades a net pumping efficiency of Tb$^{3+}$-Yb$^{3+}$ co-doped system can be greatly reduced by adopting a low-phonon glass as a host.

4. Conclusion The energy transfer from Tb$^{3+}$ to Yb$^{3+}$ which degrades a net pumping efficiency of Tb$^{3+}$-Yb$^{3+}$ co-doped system can be greatly reduced by adopting a low-phonon glass as a host. The energy transfer efficiency from Yb$^{3+}$ to Tb$^{3+}$ as high as 26±7 % was attained in a fluorophosphate glass. Tb$^{3+}$:Yb$^{3+}$-codoped fluorophosphate glass was a promising candidate for the 0.54 $\mu$m band laser medium pumped at 0.98 $\mu$m.

Acknowledgement This work was supported in part by MEXT, the Private University High-Tech Research Center program (2002-2006).

References
Novel Bandwidth Control Mechanism in Fiber Based Tunable Filters

K. Oh¹, Y. Jung², W. Shin³

¹Institute of Physics and Applied Physics, Yonsei University, Seoul, 120-749, Republic of Korea,
Tel: 82-2-2123-7657, Fax: 82-2-365-7657, Email: koh@yonsei.ac.kr

²Department of Information and Communications, ³APRI, GIST 1 Oryong-dong, Buk-gu, Gwangju, 500-712, Republic of Korea

Abstract
Recently developed mechanisms to control the bandwidth of all-fiber filters, are introduced, 1) rotational control in helicoidal LPG pair 2) bending loss edge shifts in a HOF-MSF composite. Their tuning optical characteristics will be explained.

1 Introduction
Periodic grating structures in optical fibers could be equipped with flexible wavelength tuning capability using fundamental natures of mode coupling 1) acousto-optic undulation 1) 2) mechanical strain, and 3) thermo-optic effects. In filter applications, it would be highly desirable to provide tunability in two aspects; spectral position of resonance, and bandwidth of the filter window.

Tunable band rejection filter has been recently achieved by thermo-optical tuning and corrugated long-period fiber grating over silica single mode fiber (SMF), as well as classical SMF acousto-optic tunable filter (AOTF). Thus far, most of interests and efforts have been concentrated in the spectral position tunability and possibility to tune the bandwidth has not been fully investigated.

In this review, two novel schemes are introduces, where bandwidth tuning was successfully demonstrated. Firstly a helicoidal long period fiber grating (LPFG) pair was inscribed by CO₂ laser and its bandwidth was made tunable by applying torsional stress. Secondly, the opposite cut-off shift behaviors of hollow optical fiber (HOF) and micro structure fiber (MSF) has been utilized to result in bandwidth tuning by macro-bending.

2 Helicoidal LPG pair band rejection filter
The proposed fiber filter is schematically illustrated in Fig. 1. The device consists of two helicoidal LPG with opposite helicities, clockwise, and counter clockwise direction.

![Fig. 1. Configuration of the proposed interferometer.](image)

LPG with a helical index modulation was obtained by releasing the residual stress of a pristine fiber while it rotating under a continuous single side CO₂ laser beam exposure. In the helical LPFGs novel peak shift was observed so that co-directional and contra-directional torsions resulted in spectral shift to shorter and longer wavelength, respectively.

![Fig. 2. Resonance peak shifts in helicoidal LPFGs](image)

Transmission [dB]
By the opposite shift of resonance peaks as shown in Fig. 2, the helicoidal pair of opposite helicity can provide bandwidth tuning ability as in Fig. 3. Note that the bandwidth could be almost linearly increased from 0 to 2160° rotational stress over the device.

3. Tunable bandpass filter using HOF-MSF bending

One of notable features of MSF is the existence of short-wavelength bend-loss edges (BLEs). In conventional MSFs, the long-wavelength BLEs are usually located at mid-IR irrespective of bending radii and only the short-wavelength BLEs have been experimentally accounted. This peculiar short-wavelength BLE limits the operation bandwidth of the single mode guidance of MSFs but enables us to implement a tunable high pass filter as shown in Fig. 4.

In contrast HOF has a further extended distribution of optical field into the cladding region, which provides HOF with a bend-sensitive long-wavelength BLE in the longer wavelength. The fundamental cutoff wavelength can be easily shifted by increasing the central air hole size in HOF and air-hole periodicity for MSF. Therefore, the HOF show a bend-sensitive long-wavelength BLE, which can be utilized for implementing a low pass filter as depicted in Fig. 4(b). The low pass filtering function of the HOF is a clear contrast to the high pass one of the MSF. Furthermore, by cascading MSF and HOF and individually controlling the bending curvatures, we could easily realize a tunable bandpass filter as shown in Fig. 5. The bandwidth could be widely tuned from 1000 to 300nm with a high extinction ratio over than 20dB.

We expect that the proposed leading mechanism to control the bandwidth of all-fiber filter would find applications in the realization of fiber laser, fiber sensors, and optical tomographic systems.

This work was supported in part by the KOSEF (Program Nos. R01-2006-000-11277-0, R15-2004-024-00000-0), and the Science and Technological Cooperation program between and from MOST.

4. References

Abstract — The embedding of an optical microfiber coil resonator in Teflon is demonstrated. Resonances in excess of 9dB and Q-factors greater than 6000 have been observed. The device is compact, robust and portable.

Index Terms—Microfibers, optical nanowire, optical fiber coupling, optical resonators, ring resonators

In recent years there has been a great deal of interest on microresonators based on sub-wavelength microfibers, mainly because of the development of new fiber fabrication methods that allow for low-loss evanescent wave guiding along microfibers [1,2]. Microfiber coil resonators (MCR) are powerful microoptical devices with a wide range of potential applications such as optical filter, slow light, sensing, and microlasers. Recently two kinds of MCR were reported in literature: self-touching loop resonator [3,4] and knot resonator [1,5]. Knot resonators have a long coupling region but they present severe drawbacks including the complexity of fabrication, the need for an additional coupler at the output or input of the resonator, the high loss and the presence of only one standard telecom fiber pigtails. The self-touching loop resonator is fabricated by bending a microfiber on itself and keeping two sections of a microfiber together by taking advantage of surface attraction forces (Wan der Waals and electrostatic). This approach is simple, can be fabricated from a single fiber, has two telecom fiber pigtails at the extremities and presents a lower loss that the knot resonator. The main challenge relates to the use of very thin microfibers to achieve strong self-coupling; common diameters of microfibers used for these devices are sub-micrometric. A major drawback of the self-touching loop resonator is its geometrical stability; coupling is strongly affected by the microcoil geometry and a small change in shape implies a great deal of change in transmission properties. Moreover, it has been shown [6] that sub-micrometric wires experience ageing when exposed to air for some days. The embedding of a MCR in a low refractive index medium can provide both protection from fast aging and geometrical and optical stability. In this paper the embedding of a MCR wrapped on a low index rod is experimentally demonstrated.

The microfiber was initially wrapped on a rod consisting of two layers: the inner core is a conventional silica optical fiber which provides rigidity to the structure while the outer layer consists of a coating polymer (Efiron UV373), manufactured by Luvantix, South Korea) which has a low refractive index (n~1.37 at wavelength λ~1.55µm) and limits the leakage of power from the microfiber. This structure is stable in air and preserves all the benefits of the self-touching loop resonator. Subsequently the whole arrangement was coated with Teflon fluoropolymer resin. This provides an extremely low-refractive-index embedding material (n~1.3 at λ~1.55µm). The coating method is very simple and the device is strong and portable. Additionally, coated MCR wrapped on a rod make the fabrication of 3D microcoils resonators [7] and high sensitivity biosensors possible [8]. The microfiber used in the experiment was fabricated using the set-up presented in reference 9 and a microheater (NTT-AT, Japan). The microfiber radius and the length of the uniform waist region were ~1.5 µm and 3.5 mm respectively. The rod had a total diameter D~700mm, silica core with radius ~200µm and was additionally coated with Teflon AF to provide a uniform refractive index surrounding to the microfiber. The microfiber had its pigtails connected to an Erbium-doped fiber amplifier (EDFA) and an optical spectrum analyzer (OSA) to check in real time the resonator properties during fabrication and embedding. The embedded structure was manufactured as follows: at first, with the aid of a microscope the microfiber was wrapped on the rod while one of its ends was fixed on a 3D stage; then the other microfiber end was fixed to another 3D stage and both microfiber ends were tuned to find the optimum resonator spectrum. This methodology was similar to that theoretically predicted for the design optimization of 3D microcoil resonators [10-11]. Fig. 1a shows the three turns MCR: because the microfiber has relatively large radius, bending losses are negligible and the whole structure is stable in air. The transmission spectrum in the wavelength interval 1525-1535nm is shown in figure 2a: the overall resonances extinction ratio is about 7 dB.

The resonator embedding was carried out with the use of the 601S1-100-6 solution of Teflon @AF (DuPont, United...
States). The structure was covered with Teflon solution and the solvent was let to evaporate. The coupling process is extremely challenging because the solution reduces the effect of surface forces and if the microfiber in not tightly wrapped around the rod the resonating conditions are lost. Moreover, because of the rapid solvent evaporation, Teflon particles in solution move rapidly and a collision with the microfiber can displace the microfiber and change the MCR transmission properties. Particular care has to be taken to avoid any particle contamination into the solution because it will eventually get in contact with the microfiber and significantly alter its overall loss. The picture of the MCR after its embedding in Teflon is shown in figure 1b.

where \( \lambda \) is the wavelength, \( n_{ef} \) is the effective index of the mode propagating in the microfiber and \( L \) is the loop length. This model predicts FSR\( \approx 0.8 \text{nm} \) for both the microfiber in air and embedded in Teflon, in good agreement with the experimental results. It is important to note that the spectrum shape of the MCR in figure 2 is different with respect to that of a simple self-coupling loop resonator [7], because in this latter case the coupling is limited to a single point while MCR is a 3D microcoil resonator. Simplified, figure 2 can be taken as the combination of two simple resonators, one of which is dominating the spectrum.

The Q-factor of a resonator depends on the coupling coefficient, coupling length and loss. The embedded MCR has a Q-factor greater than 6000. The Q factor of this resonator is not extremely high because the wrapping has been performed by hand and the microfibers do not have a good positioning. By using precisely controllable rotation and translation stages all-coupling resonators with ultrahigh-Q factors should be obtainable.

In conclusion, the embedding in Teflon of microfiber coil resonators has been demonstrated. A resonance extinction ratio greater than 9dB, a free spectral range of 0.8nm and a Q-factor in excess of 6000 have been observed in the embedded resonator. The embedded resonator provides a solution to the stability and reliability problems observed in microfiber resonating structures.

The authors gratefully acknowledge initiating discussions with Prof. D.J. Richardson and J.S. Wilkinson, and financial support by the EPSRC, under the standard research grant EP/C00504X/1.

REFERENCES

A New Scheme of Hyperbolic-end Microlens Using Fusion Technology

Huei-Min Yang, Wan-Yi Lin, Chung-Ting Chen, Hsing-Fu Liu, Huan-Min Lin and Hung Yen Lin
Department of Communication Engineering, I-SHOU University, Kaohsiung, Taiwan
No1, Section 1, Hsueh-Cheng Rd., Ta-Hsu Hsiang, Kaohsiung, Taiwan, 840, R.O.C
Tel: 886-76577711 ext 6767, Fax: 886-76577205, Email: hmyang@isu.edu.tw

Abstract
We propose a new scheme of the tapered hyperbolic-end fiber (THEF) by etching the fiber end in a hydrofluoride (HF) solution with a thin layer of oil floating on top of the HF. The tapered hyperbolic-end fiber microlens results in a more than 2 dB improvement in coupling efficiency when compared to the currently hemispherical microlenses.

1. Introduction
The coupling efficiency is improved by using a microlens on the end of the fiber to match both the spot sizes and the wavefront of the laser beam with those of the fiber [1-3]. The hemispherical microlenses with taper asymmetry and larger Fresnel’s reflection loss have demonstrated imperfect coupling with a typical coupling efficiency of 50% [4, 5].

Recently, hyperbolic microlenses fabricated directly on the end of the fiber by CO$_2$ laser micromachining have demonstrated up to 90% coupling efficiency [6]. The THEFs have demonstrated up to 86% coupling efficiency for a laser with an aspect ratio of 1:1.5. This study makes it possible to fabricate the hyperbolic microlenses using etching and fusion techniques that result in a more than 2 dB improvement in coupling efficiency when compared to the currently available hemispherical microlenses.

2. Manufacturing Process
The fibers used in this study were Corning step-index single-mode fibers. The glass composition was SiO$_2$-GeO$_2$-P$_2$O$_5$ for the core and pure SiO$_2$ for the cladding with low impurity content. The fiber core diameter, the beam spot size, and the refractive index difference were 8 µm, 5 µm, and 0.3%, respectively, at a wavelength of 1.55 µm. In this study, the THEF was designed with a longer taper length and a smaller taper angle with a small radius of curvature. The THEFs were fabricated by etching the fiber end in a 55% HF solution placed in a teflon beaker, with a thin layer of oil floating on top. Different density of oil floating on the HF solution were investigated. The oil density was defined by weight per volume (g/cm$^3$). Figure 1 shows the taper angle as a function of oil density. This result shows that the taper angle is dependent on the oil density. The oil floating on the HF with oil solution with lower density and much etching effect caused by HF deposition due to vaporization exhibits a smaller taper angle. After the fiber was etched to the desired tapered shape, a hyperbolic microlens for the THEF was formed by heating the fiber tip in a fusion splicer.

Fig. 1. The taper angle as a function of oil density.

3. Measurements and Results
I. A Comparison of Coupling Efficiency Between Hyperbolic-end and Hemispheric-end Microlenses
The feature of the hyperpolic-end microlens have a smaller taper angle, i.e. $2\theta_{hp1} < 2\theta_{hs}$, and $2\theta_{hp2}$ larger than $2\theta_{hp1}$ is necessary. In Fig. 2, both microlenses have a radius of curvature of 9.5 µm.

Fig. 2. Both microlenses have a radius of curvature of 9.5 µm.

Figure 3 shows the coupling efficiency as a function of the lens’s radius of curvature both the hemispherical-end microlens and hyperpolic-end microlens. A maximum coupling efficiency of 86.8% of hyperbolic-end microlens was obtained between a fiber lens with a radius of 9.5 µm and a laser diode with aspect ratio of 1:1.5. But the hemispheric-end microlens was 64.5% only with a radius of 12.5 µm.
II. An Empirical Model for Taper Length Depending on Oil Density

The HF with oil solution can be divided three layers: the bottom layer of uncontaminated HF solution, the interface layer of HF with oil solution, and the most top layer of oil, as shown in Fig. 4(a). Figure 4(b) is the etching process of tapered fiber. The local oil density, \( \gamma(z) \), is assumed to vary in the z-direction. The fiber in layer I can be etched to a long round stick, while the fiber in layer II can be etched to form a longer taper length and smaller taper angle.

\[
\gamma(z) = \begin{cases} 
\text{constant} & e_i < e \\
[p + q \gamma(e)](e - e_0) & e_0 < e < e_i \\
0 & e < e_0 
\end{cases} \quad (1)
\]

where the \( p \) and \( q \) are constants. Based on Fig. 6, the taper angle, \( \theta_t \), is given by

\[
\theta_t = 0.5 \tan^{-1}\left(\frac{d_f}{2e}\right) \quad (2)
\]

where the \( d_f \) is the fiber diameter of 125 \( \mu \)m. From Eqs. 1 and 2 yields

\[
\theta_i = 0.5 \tan^{-1}\left(\frac{d_f}{2e_i}\right) \quad (3)
\]

with the constants \( p \) and \( q \) are \(-0.113\) and \(-0.169\), respectively. The constants \( p \) and \( q \) are from the experimental values of the taper angle related to the oil density.

4. Conclusion

The high-coupling performance of microlenses with a hyperbolic profile was due to the improved mode matching between laser and fiber when compared to currently available hemispherical microlenses, which showed a maximum coupling efficiency of 65% [4,5]. Hyperbolically shaped microlenses fabricated through the optimization of etching and fusion processes have demonstrated up to 86% coupling efficiency for a laser with an aspect ratio of 1:1.5.

5. Acknowledgments

This study was partly supported by the National Science Council, Taiwan under Contracts NSC 95-2221-E-214-067.

References
Abstract

Light emission at the surface plasmon resonance frequencies from gold nanoparticles embedded in the germano-silicate glass fiber upon Argon laser pumping was observed. The quantum efficiencies at 833nm and 1536nm were estimated to be $5.75 \times 10^{-8}$ and $2.01 \times 10^{-9}$, respectively.

Keywords: Photoluminescence; Surface plasmon resonance; Gold nanoparticles

1. Introduction

Optically excited metal surface show no or very little luminescence. For instance, smooth gold films have photoluminescence (PL) efficiency of $\sim 10^{-10}$ following excitation of electron transitions from the 5d and 6sp bands [1]. In other words, only one photon is emitted each $10^{10}$ electron-hole pairs excited. One likely reason for this low PL efficiency is known that nonradiative energy relaxation processes of photoexcited carries in metals, such as Coulomb carrier scattering, are much faster than radiative electron-hole recombination, thus quenching the photoluminescence [2].

A unique exception from the rule of low PL yielding in metals is noble metal nanoparticles (NPs). For instance, PL efficiencies on the order of $10^{-4}$ have been observed in gold nanorods [3]. This is a giant enhancement with respect to $10^{-10}$ efficiency of smooth gold films. In present study, while remaining relative weak, broadband visible to narrowband infrared PL from gold NPs embedded in the germano-silicate fiber was detected and proven to be measurable at room temperature with the normal optical measurement systems. The optical absorption and PL properties including the quantum efficiencies (QEs) were investigated.

2. Experiments

The fiber incorporated with gold NPs was fabricated in house using the MCVD and solution doping processes, which is detailedly described in Ref. 4 reported by our group.

Due to the boiling temperature as high as $2856^\circ C$ [5] Au atoms and their clusters can survive from the MCVD process with the temperature up to $2350^\circ C$. To confirm the formation and existence of the gold NPs in the core of fiber, the cut-back method to determine the composition of the dopants existing in the core of the made fiber.

![Schematic diagram of the experiment setup.](image)

To obtain the PL, the fiber was pumped with a 488nm Argon-ion laser at room temperature. As shown in Fig. 1, the 488nm laser beam from Argon-ion laser was reflected by two mirrors and then coupled into one arm of a 3dB coupler by a collimator. At the same time, one power controller was used to measure the feed-back power from the other arm of the given coupler to determine the actual input power into the tested germano-silicate fiber incorporated with gold NPs. In order to get accurate luminescence signal, high sensitive optical spectrum analyzer (OSA) was used as the detector.

3. Results and discussion

![The absorption spectrum of the germano-silicate fiber incorporated with gold nanoparticles.](image)
Figure 2 shows the absorption spectrum of the germano-silicate glass fiber incorporated with gold NPs. The absorption peak appeared at 498.4 nm (2.48 eV), which was due to the SPR of the gold NPs embedded in the fiber core and is the indirect evidence of the existence of gold NPs.

The broadband visible PL in the vicinity of 833 nm (1.48 eV) to the narrowband infrared PL centering 1536 nm (0.81 eV) was found to appear when the fiber was pumped with 488 nm (2.53 eV) Argon-ion laser as shown in Fig. 3. By using the accurately measured input launched power into the fiber and the calculated PL intensities at 833 nm (1.48 eV) and 1536 nm (0.81 eV), we confirmed that the averaged QEs were 5.75×10⁻⁸ and 2.01×10⁻⁹, respectively. It is important to note that the visible PL intensity at 833 nm was two orders in magnitude larger than that of bulk gold metal [1] and the observation of the infrared PL at 1536 nm was the first to report in the field of fiber optics.

Excitation of the gold NPs incorporated germano-silicate glass fiber by the Argon-ion laser at 488 nm (2.53 eV) may lead to the excitation of d-band electrons into the sp-conduction band (interband transition) and a radiative recombination is followed by an initial electronic relaxation bringing about the visible luminescence. The luminescence between 629 nm (1.97 eV) and 1200 nm (1.03 eV) corresponds to the recombination of the excited electron from higher excited states in the sp-band with the hole in the lower lying d-band (interband transition). The lower energy luminescence band in the vicinity of 1536 nm (0.81 eV) can then be assigned to be the relaxed radiative recombination between the highest occupied (molecular) orbital and the lowest unoccupied orbital (HOMO-LUMO gap) of 1.3 eV within the sp-conduction band (intraband transition). The schematic energy transition diagram is shown in Fig. 4, which shows the origin of the two PL bands. The high energy band is proposed to be due to radiative interband recombination between the sp and d-bands while the low energy band is thought to originate from radiative intraband transitions within the 6sp-band cross the HOMO-LUMO gap. Note that intraband recombination has to involve prior nonradiative recombination of the hole in the 5d-band created after excitation with an (unexcited) electron in the 6sp-band.

4. Summary

We successfully made germano-silicate glass fiber incorporated with gold NPs. Weak but distinct broadband visible to narrowband infrared PL excited by 488 nm Argon-ion laser was found to appear at room temperature. The QEs of the PL fixed at 833 nm and 1536 nm are confirmed to be 5.75×10⁻⁸ and 2.01×10⁻⁹, respectively. Interband transition between the sp and d-band and intraband transition across the HOMO-LUMO electronic gap inside the gold NPs are responsible for the two bands PL mentioned above.

Acknowledgements

This research was partially supported by Korea Science and Engineering Foundation (KOSEF) through grant No.R01-2004-000-10846-0, by the GIST Technology Initiative (GTI) through grant No.K01496, by the National Core Research Center (NCRC) for Hybrid Materials Solution of Pusan National University, and by BK-21 Information Technology Project, Ministry of Education and Human Resources Development, Republic of Korea. The authors would like to thank Dr. Il Ho Ahn and Prof. G. Hugh Song from Quantum Integrated Photonics Lab of GIST for their active help in providing Argon-ion laser as the pumping source.

References:
Hetero III-V-N alloy/Si technologies toward monolithic OEIC chips including high-density light emitting devices

H. Yonezu, Y. Furukawa and A. Wakahara
Toyohashi University of Technology
1-1 Hibarigaoka, Tempaku-cho, Toyohashi, Aichi, 441-8580, Japan
E-mail: yonezu@ee.e.tut.ac.jp

Structural defect-free Si and InGaPN/GaPN double heterostructure layers were grown on a Si substrate. Si MOSFETs and LEDs, which are elemental devices for monolithic OEICs, were implemented into a single chip.

1. Introduction

It is essential to grow III-V compound semiconductors on Si substrates for realizing monolithic optoelectronic integrated circuits (OEIC) in which light emitting devices are implemented into Si large-scale integrated circuits (LSI). However, following two problems have to be overcome for realizing the OEIC.

(1) Generation of a large number of structural defects in the growth of III-V compound semiconductors on Si
(2) Mismatch of fabrication processes between Si LSI and III-V compound light emitting devices.

These problems have been overcome [1, 2]. As a result, elemental devices of Si MOSFETs and InGaPN/GaPN DH LEDs for monolithic OEICs were fabricated in Si/GaPN-based epitaxial layers grown on a Si substrate [3].

2. Suppression of structural-defect generation in the growth of Si/(In)GaPN layers on Si substrate

The epitaxial growth of III-V compound semiconductors on Si substrates contains the specific following problems: The differences in (1) the number of valence electrons, (2) lattice parameters and (3) thermal expansion coefficients. These problems cause the generation of structural defects such as dislocations and stacking faults.

Problem (1) causes the generation of anti-phase domains (APDs), stacking faults and threading dislocations at the initial growth stage. This problem was overcome by growing a thin GaP initial layer on a Si (100) substrate misoriented by 4 degree towards a [011] direction at 450 °C by migration-enhanced epitaxy (MEE). Problem (2) generates misfit and threading dislocations in lattice relaxation process. This problem was overcome by lattice-matched growth using GaPN and InGaPN on the GaP initial layer. It should be noted that a lattice parameter decreases with the increase in N compositions x in the III-V-N alloys, while a bandgap (E_g) decreases. Problem (3) could introduce edge dislocations from a grown surface during cooling process after a relatively high temperature growth. This problem was overcome by growing a Si capping layer since tensile strain is negligibly small. Thus, a structural defect-free Si/GaPN layers were realized in principle on a Si substrate, as shown in Fig. 1 [4].

3. Monolithic implementation of MOSFET and LED for OEICs

Elemental devices for the monolithic OEIC are typically a MOSFET and an LED. In principle, the output of a MOSFET circuit is obtained as the light output of the LED. We have tried to fabricate the elemental devices of the MOSFETs in the topmost Si layer and the LEDs in the InGaPN/GaPN DH layer.

The MOSFETs and the LEDs should be fabricated in a conventional processing flow for MOSFETs.
The fabrication processes are shown in Fig. 2 [2]. Thermal process conditions were matched between MOSFETs and LEDs. Wet oxidation was applied to the growth of a gate oxide.

![Fundamental structure](image)

**Fig. 2 Fabrication processing flow of elemental devices for a monolithic OEIC [2]**

The photograph of a fabricated test chip is shown in Fig. 3, which includes pMOSFETs and LEDs with various sizes [3]. Four LEDs with an active area of 100 μm × 100 μm emitted red light at 10 mA. Drain current I<sub>d</sub> vs. drain voltage V<sub>d</sub> curves of the pMOSFET with W<sub>L</sub> of 30 μm and the spectra of the LED with an active area of 20 μm × 45 μm are inserted in Fig. 3. The optical output of the LED was switched by applying a pulsed voltage to the gate of the pMOSFET. OEICs in which LSIs contain a large number of LEDs could be realized when the threshold voltage of MOSFETs is controlled and the small-size LEDs are fabricated. For high performances, point defects should be reduced in (In)GaPN.

**4. Summary**

Structural defect-free (In)GaPN and layers were grown on a Si substrate. MOSFETs and LEDs, which are elemental devices for OEICs, were monolithically merged in a single chip with a Si layer and an InGaPN/GaN DH layer grown on a Si substrate. The growth and fabrication process technologies are effective for the realization of monolithic OEICs.

**Acknowledgments**

This work was supported by the Ministry of Education, Science, Sports and Culture Scientific Research in Priority Areas, Specially Promoted Research, the 21st Century COE Program “Intelligent Human Sensing”, the Venture Business Laboratory and the Research Center for Future Technology of Toyohashi University of Technology.

**References**

GaInAsP/InP Membrane DFB Lasers Directly Boded on SOI Substrate with Rib-waveguide Structure

Takeo MARUYAMA,1,2 Tadashi OKUMURA,1 Masaki KANEMARU,1 Shinichi SAKAMOTO,1 Shigeo TAMURA,1 and Shigehisa ARAI,1,2

1 Quantum Nanoelectronics Research Center, Tokyo Institute of Technology
2-12-1-S9-5 O-okayama, Meguro-ku, Tokyo 152-8552, Japan
2 CREST, Japan Science and Technology Agency
Kawaguchi Center Building, 4-1-8 Honcho, Kawaguchi, Saitama 332-0012, Japan
Phone: +81-3-5734-2555  Fax: +81-3-5734-2907  E-mail: maruyama@pe.titech.ac.jp

Abstract
Membrane GaInAsP/InP distributed feedback lasers were realized on an SOI substrate integrated with rib-waveguide structure and continuous-wave operation under optical pumping was obtained up to 85°C. The characteristics temperature was 65 K and the thermal resistance was estimated to be 11 K/mW.

I. Introduction
It has been predicted that the continued growth of silicon LSI technology will soon hit the interconnect bottleneck such as the ohmic heating and the RC delay due to the metal lines. One of methods to solve this problem is a replacement of the metal line by an optical line.

Since a silicon on insulator (SOI) substrate has a high index contrast between the silicon core (n=3.45) and the oxide buffer layer (n=1.45) and it makes ultracompact optical circuits at the low-loss fiber communication wavelengths, integrations of functional photonic devices such as lasers and optical amplifiers on SOI platforms are very attractive. Room-temperature continuous-wave (RT-CW) operations of semiconductor lasers fabricated on a silicon substrate by heteroepitaxial growth [1] and direct wafer bonding [2] have been reported, however, it is difficult to couple light outputs to a silicon waveguide because of their thick cladding layers. Membrane-based lasers are very attractive for low threshold operation as well as integrations to an SOI waveguide [3]. Recently, we have reported RT-CW operations of optically pumped GaInAsP/InP membrane buried-heterostructure (BH) distributed feedback (DFB) lasers on an SOI substrate using direct wafer bonding [4],[5]. An injection type semiconductor laser fabricated on an SOI substrate was achieved by BCB/SiO$_2$ bonding [6] and plasma assisted bonding [7].

In this paper, we would like to report relatively high temperature CW operation of GaInAsP membrane DFB lasers on an SOI substrate rib-waveguide structure with narrow stripe width.

II. Device Structure and Fabrication
The device structure of GaInAsP/InP membrane lasers on an SOI waveguide is shown in Fig. 1. A five-quantum-well, consisting of a Ga$_{0.22}$In$_{0.78}$As$_{0.81}$P$_{0.19}$ (6 nm thick) sandwiched by 2-step optical confinement layers of Ga$_{0.22}$In$_{0.78}$As$_{0.48}$P$_{0.52}$ (n$_{g}$=1.2 μm, 10 nm thick) / Ga$_{0.14}$In$_{0.86}$As$_{0.31}$P$_{0.69}$ (n$_{g}$=1.2 μm, 55 nm thick) were grown on a (100) n-InP substrate as an initial wafer by an OMVPE technique. The DFB structure was fabricated on the initial wafer by using an electron-beam lithography, CH$_4$/H$_2$ reactive ion etching (RIE) and OMVPE regrowth [4]. The SOI and the GaInAsP/InP substrates were dipped in a H$_2$SO$_4$:H$_2$O$_2$:H$_2$O solution to make these surfaces hydrophilic, which is important for self-adhesion at room temperature. These substrates were loaded into a furnace and heated at 250°C for 3 hours [5]. After bonding, the membrane structure was formed by selective wet chemical etching of the InP substrate side. Then the stripe geometry was formed by photolithography and RIE dry etching as shown in Fig. 1. Typical (wirelike) active region width and the period of the first-order grating were around 110 nm and 250nm, respectively. The cavity length and the stripe width were 140 μm and 1.5 μm, respectively. The waveguide length and the stripe width in the waveguide region were 500 μm and 1.5 μm, respectively.

![Fig. 1 Schematic structure of GaInAsP/InP membrane DFB laser and InP/SOI waveguide.](image-url)
III. Lasing Characteristics

The device was optically pumped from the top with a 980 nm wavelength CW laser diode through a single-mode optical fiber connected to a micro-PL setup. The pumping light was focused through a cylindrical lens to a spot size of 4 µm x 176 µm. The output power was coupled into a multi-mode fiber connected to an optical spectrum analyzer. This laser was oscillated up to 85 °C. The characteristic temperature $T_0$ of 64K was obtained. Although both sides of this mesa stripe was surrounded with air which is very low thermal conductive material compared with semiconductor, the characteristic temperature was higher than the value of the laser on SOI [7].

Figure 2 shows the temperature dependence of lasing wavelength. The incremental temperature coefficient of the lasing wavelength was $8.9\times10^{-2}$ nm/K, which is approximately 20% smaller than that of conventional 1.55 µm wavelength lasers because an optical field penetrates into air cladding regions [8] and exists in silicon layer whose temperature coefficient of the refractive index is almost a half of that of InP.

Figure 3 shows the pump power dependence of lasing wavelength. The lasing wavelength increased linearly with the pump power above the threshold because the carrier density clamps at the threshold and the incremental coefficient against pump power $d\lambda/dP_{\text{pump}}$ was $2.2\times10^{-2}$ nm/mW. By assuming the absorption coefficient of GaInAsP OCL to be 10,000 cm$^{-1}$ and all absorbed power was changed into heat, the thermal resistance $R_h$ can be estimated from the temperature coefficient of the lasing wavelength to be 9.3 K/mW, which is slightly small value (11 K/mW) of the previously reported air-bridge type membrane BH-DFB laser due to better heat pass through SOI layer [9].

IV. Conclusion

CW operation of GaInAsP membrane DFB lasers integrated with SOI rib-waveguide was demonstrated up to 85 °C. The characteristic temperature of 64K was achieved with the stripe width of 1.5 µm, the cavity length of 140 µm and the waveguide length of 500 µm. This device structure can be applied to realize photonic integrated circuits (PICs) consisting of lasers, modulators, amplifiers, detectors and SOI passive components.

Acknowledgment

We would like to thank Professors Emeritus Y. Suematsu and K. Iga for continuous encouragement, and Professors K. Furuya, M. Asada, F. Koyama, K. Kobayashi, Y. Miyamoto, M. Watanabe, T. Miyamoto, H. Ueno, N. Nishiyama of the Tokyo Institute of Technology for fruitful discussions. This research was partially supported by a Grant-in-Aid for Scientific Research (#17206010, #17760275) from the Ministry of Education, Culture, Sports, Science and Technology (MEXT), Japan.

References

Fabrication of Silicon-on-Insulator arrayed waveguide grating and monolithic power monitor array

P. S. Chan¹, D. X. Dai², A. W. Poon³ and H. K. Tsang¹

1. Department of Electronic Engineering, The Chinese University of Hong Kong, Shatin, Hong Kong, China. 2. Centre for Optical and Electromagnetic Research, Zhejiang University, Hangzhou 310058, China. 3. Dept. of Electronic & Computer Engineering, Hong Kong Univ. of Science & Technology, Hong Kong.

Abstract: The design, fabrication and initial testing of a pair of silicon-on-Insulator arrayed waveguide gratings (SOI AWG) integrated with a monolithic power detector array for power monitoring is described.

The use of reconfigurable optical add-drop multiplexers (ROADM) is attractive for cost effective networks [1]. However changes in the number of channels will affect amplifier performance and produce dynamic gain tilt and gain transient. In this paper, we describe a potentially low-cost compact channel optical power monitor formed by the use of a pair of silicon-on-insulator (SOI) arrayed waveguide grating (AWG) and an integrated photodetector array. The photocurrent is measured by silicon photodetectors made sensitive to the below bandgap energy light via the use of helium implantation. The proposed device thus avoids the need for space consuming waveguide taps and hybrid integrated III-V photodetectors [2].

Fig. 1 shows the design simulation of a typical channel in a Gaussian AWG multiplexer with 40 channels. The target channel spacing was 100GHz (0.8nm).

For the monolithically integrated power monitor, two AWGs were arranged in the folded geometry as shown in Fig. 1. The input channels are first demultiplexed by the first AWG. The demultiplexed signals are individually monitored by individual waveguides which form an array of helium implanted photodetectors. Finally the signals are recombined by the second AWG and coupled to the transmission link. The whole device comprising two 40-channel AWGs and 40 waveguide photodetectors occupy an area of less than 2x2 cm² in size.

Fig. 2 Bidirectional AWG with power monitoring He implanted region.

The waveguide photodetectors were formed by rib waveguides of 2.4um width and 2.3um etch depth. The waveguides were fabricated using deep reactive ion etch (DRIE) at room temperature using a continuous mixture of C₄F₈ and SF₆ gases. To minimize electrical crosstalk between photodetectors, electrical isolation trenches were etched on either side of each waveguide photodetector as shown in the schematic cross section diagram in Fig. 3.

Fig. 1 Spectral response of a typical channel in the AWG
The array of waveguide photodetectors were produced by fabricating lateral p-n diodes with boron and phosphorus implantation respectively on either side of each waveguide. The dose of boron and phosphorus dopants were $5 \times 10^{15}$ cm$^{-2}$ and the implanted diodes underwent rapid thermal annealing at 950$^\circ$C for 1 minute. Metal contacts (Ti/W/Au) were deposited onto the doped regions. A picture of the fabricated device after liftoff of the TiW/Au contacts is shown at the bottom of Fig. 3. The current-voltage characteristic of a single diode in the array is shown in Fig. 4.

Since the photon energy at wavelengths in the 1.55$\mu$m wavelength band are insufficient to excite electrons across the indirect bandgap of silicon, it is necessary to enhance the absorption of below bandgap energy photons by helium implantation into the array of waveguides [2]. The target dose of helium ion implantation was $5 \times 10^{13}$ cm$^{-2}$ and the implanted samples were annealed. Absorption via traps in the bandgap introduced by the ion implantation, together with the thermal excitation of electrons from these defect states into the conduction band energy allowed the generation of photocurrent from the helium implanted waveguides for power monitoring. Although responsivities as high as 64 mA per watt of absorbed optical power and low excess losses in the waveguide photodetectors were previously achieved with single waveguide photodetectors [2], the response was limited in the case of the present monolithically integrated device because of the higher dark currents in the reverse biased diodes. The high dark current limited the reverse bias currents to only a few volts (Fig 4 inset). We speculate that the higher dark currents may be caused by currents flowing via surface states [3].

Work on further optimizing the diodes to improve the diode performance is ongoing. We believe the device will have potential for low-cost in-line channel power monitoring and may find important applications in broadband reconfigurable networks which need individual channel power monitoring in order to allow effective compensation of EDFA gain transients or gain tilt after optical add-drop.

**Acknowledgement:** This work was fully funded by RGC Earmarked Grant 415905.

**References**


CW Operation of Optical Amplification in Organic Dye-Doped Polymeric Channel Waveguide

K. Yamashita1*, K. Hase1, H. Yanagi2, and K. Oe1

1Department of Electronics, Graduate School of Science and Technology, Kyoto Institute of Technology, Matsugasaki Goshokaidocho, Sakyo-ku, Kyoto 606-8585, Japan
2Graduate School of Materials Science, Nara Institute of Science and Technology, 8916-5 Takeyama-cho, Ikoma, Nara 630-0192, Japan
*Phone: +81-75-724-7400, Fax: +81-75-724-7400, e-mail: yamasita@kit.ac.jp

Abstract: We have succeeded in cw operation of optical amplification in an organic dye-doped polymeric channel waveguide. The 19-dB optical gain was achieved for externally input light under 120-mW optical pumping from the waveguide end-face.

1. Introduction

Organic luminescent materials are very attractive for active media in organic lasers and waveguide-type optical amplifiers because of their large fluorescence efficiencies [1]. While there are a lot of reports about amplified spontaneous emission (ASE) and the lasing characteristics in the organic luminescent materials, the optical pumping has been carried out by nanosecond-pulsed lights. In any reports, the pulsed pump beam with tachistoscopically huge power has been focused on the sample surface by a cylindrical lens providing a rectangular excitation stripe, so that the sufficient waveguiding length for optical gain is ensured [2]. While this pumping method is available for evaluation of the optical gain coefficient, it is unrealistic to apply this method to the cw operation. One of clever solution for achievement of the cw operation, which is indispensable for practical applications, is optical pumping from the waveguide edge, as employed in rare earth-doped fiber amplifiers. For this pumping method, a channel waveguide structure is required for confinement of the pumping beam.

In this study, we have fabricated infrared organic dye-doped polymeric channel waveguide and firstly succeeded in observations of cw-stimulated emission and optical amplification by using a ~ 120-mW light source for optical pumping. This remarkable result was accomplished by the optical pumping from the channel waveguide end-face. The experimental scheme shown here is quite reasonable and practical for any organic luminescent materials.

2. Fabrication

The channel waveguide fabricated in this study consists of poly(methyl methacrylate) (PMMA) as host matrix and 4-[4-[4-(dimethylamino)phenyl]-1,3-butadienyl]-1-ethylquinolinium perchlorate (LDS798) as organic active medium [3]. LDS798 is a commercially available near infrared dye, and has maximum absorption at 566 nm and emission band of 650 – 830 nm. A solution of tetrahydrofuran and propylene glycol monomethyl ether acetate, in which powdery PMMA and LDS798 were solved, was spin-coated on a silica substrate. Here, rectangular grooves with 100-μm width and 10-μm depth were fabricated on the substrate by reactive ion etching method. After evaporation of the solvent, the channel waveguide of a reversed ridge-type structure buried in a silica substrate was obtained as shown in Fig. 1. The waveguide sample was cut to 2-mm length with a dicing machine. The dye concentration to the host PMMA was 0.1 wt%.

3. Performance

The fabricated channel waveguide sample was optically pumped from an end-face by using a cw-light source of frequency doubled diode-pumped solid-state laser (532 nm) with output power of 120 mW. The pumping beam was focused by an objective lens, and coupled to one side of the waveguide end-face. We

![Fig. 1. A microphotograph of LDS798/PMMA channel waveguide end-face. The width and height of the waveguide are 100 and 12 mm, respectively.](image-url)
found strong pumping light transmitted to the end-face in another side. At the same time, strong red emission, which is spontaneous emission of LDS798, was observed only at the channel waveguide. These observations indicate that all over the waveguiding cavity is excited and substantially activated by the pumping light.

Figure 2 shows emission spectra monitored at the output end-face of the LDS798/PMMA channel waveguide at various cw pumping power. To avoid degradation of the emission property by thermal deactivation, photo-oxidation, and/or optical aging of the dye, the pumping light was chopped at 100 Hz, so that the optical duty became 3%. With increasing the pumping power, spectrally narrowed emission appeared at ~ 750 nm and rapidly grew up. This spectral evolution shows self-amplification of spontaneous emission in the channel waveguide, i.e., ASE. Following the ASE observation, we have carried out optical amplification measurement for externally input light in the LDS798/PMMA channel waveguide. A semiconductor laser diode with emission wavelength of 770 nm was employed as the light source for the input light. The input light with the power of ~ 1 µW was aligned on almost the same optical axis as the pumping light and coupled to the waveguide at the same time. After the fine optical coupling was confirmed, emission from the output end-face was butt-coupled to an optical fiber with the core diameter of 9 µm. The collected emission was introduced to a CCD spectrometer after filtering the pumping beam. The optical amplification of the input light was found at the pumping power larger than 2.4 mW. Figure 3 shows substantial optical gain as a function of the duty of optical pumping. The pumping power was 120 mW. At the pumping duty of 3%, the optical gain of 19.3 dB was achieved. Furthermore, the gain value was > ~ 5 dB even at the pumping with 41%-duty, which is very nearly to the completely cw condition, while the gain degradation caused by optical aging of the dye was observed.

4. Conclusion

The cw-stimulated emission and optical amplification in polymeric waveguide doped with infrared organic luminescent dye were firstly demonstrated. A channel waveguide structure enabled us to effectively pump the active layer. Since this experimental scheme is quite reasonable and practical for any organic luminescent materials, we can say that the organic luminescent materials have great potentiality for lasers and waveguide-type optical amplifiers.

5. References


Ultrafast Optical Gain Switching for Wavelength Conversion from Silica to Polymer Optical Fibre Communication Wavelengths

Ruidong Xia,* Cora Cheung and Donal D. C. Bradley*
Ultrafast Photonics Collaboration, Experimental Solid State Physics Group, Blackett Laboratory, Imperial College London, Prince Consort Road, SW7 2BZ London UK
Arvydas Rusekas, Dimali Amarasinghe and Ifor D. W. Samuel*
Organic Semiconductor Centre & Ultrafast Photonics Collaboration, School of Physics and Astronomy, University of St Andrews, KY16 9SS St Andrews UK

Abstract Optical gain switching is demonstrated for red emission distributed feedback polymer lasers operating close to the 650 nm polymer optical fibre low loss window, namely at 675 and 692 nm. With control pulses (5kHz, 4ps) at telecommunication wavelengths, 1.28 and 1.32 µm the gain switch operates as a wavelength conversion element.

Introduction
Polymers have been increasingly used to fabricate passive structures within tele- and data-communication systems with robustness, ease of processing, integration and low cost among the key drivers [1, 2]. It has also been recognized that such plastic fibre and waveguide structures might benefit from the incorporation of gain, ultrafast optical switching and other active functionalities, with organic semiconductor materials and devices considered particularly promising due to their potential for ready integration [3].

We report here on optical gain switching in polyfluorene-based red light emitting distributed feedback (DFB) lasers using 1.3 µm optical control pulses to directly modulate their output. Given the polymer optical fibre (POF) defined 650 nm centred target wavelength window [4], we selected red emission copolymer Red F [5] as the gain medium and fabricated polymer lasers with lasing wavelength at 675 and 692 nm. The wavelengths of control pulse (5kHz, 4ps) were specially chosen in the telecommunication range, 1.28 and 1.32 µm. We demonstrate the ability to completely shut down the visible laser output in the presence of the control pulses, providing a wavelength conversion scheme to transfer data from silica to polymer optical fibre communication windows. By tuning the temporal overlap of the laser pump and control pulses it is also possible to more subtly control the laser dynamics, for example to introduce a variable time delay as needed for data re-timing.

Experimental
The Red F polymer used in our study was provided by the Sumitomo Chemical Co., Ltd and was used as supplied. The DFB lasers were prepared by spin-coating toluene solutions of Red F (20 mg/ml) onto one-dimensional grating patterned silica substrates. Two grating patterns were used with periods (Λ), fill-factors and depths: (a) Λ = 420 nm, 50% and 100 nm and (b) Λ = 430 nm, 20 % and 50 nm. After coating 240 nm Red F films these structures showed laser action at 675 nm and 692 nm, respectively.

Pump and control pulses were generated by an amplified 5 kHz Ti:sapphire laser pumped optical parametric amplifier. The pulses were stretched to ~ 4 ps (FWHM) in a highly dispersive TF-10 glass block in order to better match the time resolution of the Hamamatsu synchroscan streak camera used as detector (~ 2 ps). The wavelength of the control pulse was tuned from 1.28 to 1.32 µm, which resulted in a slight change of the pump wavelength from 492 to 498 nm. The optical pump and control beams were spatially overlapped on the sample in a spot of 100 µm diameter but subjected to a variable time delay.

Result and Discussion
Figure 2 shows streak camera images in which the output from the frequency doubled 1.28 µm control beam (left side of each image) and the 692 nm Red F DFB laser (right side of each image) are simultaneously detected. The vertical axis represents the time scale 0-100 ps from top to bottom and the horizontal axis represents a wavelength scale 595 – 725 nm from left to right. The Red F laser was pumped with 36 nJ pulses (~1.5 times the lasing threshold energy). The control pulses were adjusted to 2 µJ, an energy that, after frequency doubling, gave a comparable intensity signal on the streak camera to that of the Red F laser. Figure 2(a) shows the situation in the absence of any control pulses and is thus simply the laser beam profile. Turning the control pulses on causes the intensity of the polymer laser output to decrease significantly as shown in Figure 2(b) and when they are suitably synchronised with the DFB laser pump pulses the 692 nm laser emission is completely quenched (Figure 2(c)). The same behaviour was found for the 675 nm Red F DFB.
laser and for 1.32 µm control pulses, demonstrating the reproducibility of the phenomenon across a range of DFB laser output and control pulse wavelengths.

A more subtle control of the DFB laser emission is also possible, allowing adjustment of the output pulse intensity, temporal profile, and time delay relative to the pump pulse. The latter delay offers the potential for data re-timing, another important function in communication systems. Figure 2b shows just such a temporal shift in the polymer laser output due to control pulses that are synchronized to arrive at the sample after the onset of lasing. The 675 nm DFB laser was pumped at 498 nm with 80 nJ pulses (i.e. four times threshold) and its output was modulated with a 200 nJ control pulse at 1.32 µm, delayed by 3 ps (peak-to-peak) relative to the unperturbed polymer laser output. The polymer laser output pulse was delayed by some 5 ps as an effect of the control pulse but there was very little change in temporal width. In this case, the spectral content was also little altered but there was an approximately 35% drop in pulse intensity. These data are representative of a very large parameter space in regard of relative pump and control pulse intensities and time delays and are, as such, only indicative of the response that can be induced.

Conclusion
We have demonstrated optical gain switching in polyfluorene-based red light emitting distributed feedback (DFB) lasers using 1.3 µm optical control pulses (4 ps, 5 kHz) to directly modulate their output. These results suggest the potential to use ultrafast all-optical gain switching for wavelength conversion between the silica-fibre telecom and plastic-fibre datacom windows in such polymer devices.

Acknowledgement
We thank the Sumitomo Chemical Co., Ltd for providing the polymer used in our experiments and the UK Centre for Integrated Photonics for fabricating the grating structures that we have used in our study. We further thank the UK Engineering and Physical Sciences Research Council (Ultrafast Photonics Collaboration, GR/R55078) and the Commission of the European Community (FP6-IST-026365 POLYCOM) for financial support. We also thank Ian White and Richard Penty for useful discussions.

Reference
Thin film DFB lasers utilizing dye-doped marine-biopolymer DNA-lipid complex films and etch-less gratings

Hideaki Takano, Saki Narisawa, Sunao Takenaka, Junichi Yoshida, Naoya Nakai, Makoto Fukuda and Naoya Ogata
Chitose Institute of Science and Technology
758-65, Bibi, Chitose-shi, Hokkaido, 066-8655, Japan
Tel. & facsimile: +81-123-27-6047  E-mail: yoshi@yoshisv0.spub.chitose.ac.jp

ABSTRACT
DFB laser structure utilizing dye-doped-DNA-lipid films have been studied for potential application to thin-film single longitudinal mode lasers. Although much improvement should be necessary for laser characteristic, we observed single longitudinal mode operation by using etch-less gratings fabricated on a PMMA substrate.

1. INTRODUCTION
Recent research results on DNA-lipid complexes have shown various attractive features such as strong fluorescence, light amplification, selective filtering etc., by the doping of organic dyes into DNA films. We have already reported basic optical characteristics, such as refractive indices, absorbance and fluorescence intensity, and photochromic properties, of organic-dye-doped DNA-lipid complex films, which have been derived from marine biopolymers. On the other hand, waveguide type devices with reduced cost.

2. SAMPLE PREPARATION
Dye-doped-DNA-CTMA films were prepared by the same method described in our previous work except for the solvent for DNA-CTMA complexes. Because of the substrate material is PMMA in our experiment, we used ethanol instead of ethanol/chloroform mixed solution. Regarding the laser-dye, we employed 4-[4-(dimethylamino) styryl]-1-dococylpyridinium bromide (DMASDPB), one of the hemicyannine dye derivatives known as nonlinear optical materials. By using the same material system, spectral narrowing due to ASE has been report above about 0.5-1 mJ/cm² of the excitation energy density. In order to introduce the optical feedback mechanism, we utilized the etch-less grating fabricated on the PMMA substrate. The etch-less grating is the grating structure fabricated by a replica formation process from the original photoresist grating pattern onto a polymer substrate such as PMMA. Details of the etch-less grating fabrication process has been reported elsewhere.

The laser-dye DNA-CTMA was spin-coated or poured on the etch-less grating, and dried for about 3-4 hours. Then, emission spectra excited by the SHG light of pulsed Nd:YAG laser were measured by using HR-2000 multi-channel spectrum analyzer.

3. FLUORESCENCE and ASE SPECTRA
Typical fluorescence and ASE spectra of the DMASDPB-doped DNA-CTMA film are shown in Fig. 1. Note that these spectra were measured for the film coated on a silica glass. The ASE dominated at the region above about 1 mJ/cm² of the excitation energy. This is consistent with the previously reported value by Y. Kawabe et al. Figure 2 compares the fluorescence intensity of DMASDPB-doped DNA-CTMA and PMMA. This figure revealed that efficient fluorescence is attained by using DNA as the host material rather than using well-known organic material such as PMMA.

Fig. 1 Typical fluorescence and ASE characteristics of DMASDPB-doped DNA-CTMA film
4. DFB LASER CHARACTERISTICS

Figure 3(a) shows typical I-O characteristics of the DFB laser structure. Lasing spectra are also shown in Fig. 3(b). As shown in this figure, single longitudinal mode operation is maintained whole range of excitation. Mode components of the lasing light are shown in Fig. 4. Although the single mode operation is obtained, the observed lasing threshold of about 4 mJ/cm$^2$ is slightly higher than the ASE threshold described in §3. This will be due to imperfect guiding structure of the cavity. Overlap of the absorption band of the doped-dye with the emission spectra may cause some degradation of the efficiency. Therefore, to reduce the threshold, it is necessary to introduce proper waveguide structures such as ridge or channel, and to optimize the DFB grating parameters. Furthermore, examination of high-efficient laser-dyes may be also important\(^4\) for achieving much lower lasing threshold.

5. CONCLUSIONS

We have studied thin film DFB laser structure utilizing dye-doped-DNA-lipid films, and observed single longitudinal mode operation by using etch-less gratings fabricated on a PMMA substrate. The lasing threshold of DMASDPB(one of the hemicyanine dye derivatives)-doped DNA-CTMA films was about 4 mJ/cm$^2$. Although the value was almost consistent with the ASE threshold of about 1 mJ/cm$^2$, it is relatively high compared with other system such as LDS798-doped polymeric waveguide structure\(^4\). Introduction of proper waveguide structures and selection of the laser dye which have minimum overlap of emission spectra with absorption spectra should be necessary for further reduction of the threshold energy.

ACKNOWLEDGMENTS

The authors deeply thank Prof. Yutaka Kawabe, Ms. Amane Nakamura and Mr. Kanji Yamaoka for their valuable discussion. This work was partly supported by the Ministry of Education, Culture, Sports, Science and Technology, Grant-in-Aid for Scientific Research, and also Special Coordination Funds for Promoting Science and Technology.

REFERENCES

7. N. Nkai and M. Fukuda, JJAP, 45, 998-1004, 2006

399
Influence of carrier traveling length on transient properties of fast-response organic light sources

Takeshi Fukuda a, Tomoko Okada b, Bin Wei b, Musubu Ichikawa b and Yoshio Taniguchi b
a) Optics and Electronics Laboratory, Fujikura Ltd., 1440 Mutsuzaki, Sakura, Chiba 285-8550, Japan
b) Department of Functional Polymer Science, Shinshu University, 3-5-1, Tokida, Ueda, Nagano 386-8567, Japan
Tel: +81-43-484-3347, Fax: +81-43-481-1210, Email: fukuda@lab.fujikura.co.jp

Abstract
Organic light sources are required for optical interconnect applications because of its flexibility and low fabrication cost. We have investigated transient properties of organic light sources with different thicknesses of organic layers. The cutoff frequency of 12 MHz has been achieved.

1 Introduction
Organic light-emitting diodes (OLEDs) and organic photo-diodes (OPDs) have several advantages over inorganic devices, such as flexibility and low fabrication cost. So, organic devices have been attracted for optical interconnect applications, of which a schematic configuration is shown in Fig. 1 [1, 2].

![Schematic configuration of an example of an optical interconnects with an OLED and an OPD.](image)

Several factors to affect a response time of an OLED have been already investigated, particularly capacitance [3], a fluorescence lifetime (FL) of a light emitting-material [4], and carrier mobility of electron/hole transport materials [5]. To date, the –3 dB cutoff frequency of more than 10 MHz has been achieved [4]. The reported cutoff frequency was less sufficient to apply these OLEDs for optical interconnects. It has been recognized that several hundreds of MHz or over GHz transmission speed is desirable to correspond to expand an application field. Furthermore, other factors to limit the transmission speed are important to examine possibilities of organic light sources.

In this paper, we have investigated that the relationship between the transient property of OLEDs and the carrier traveling length, which is determined as thicknesses of organic layers.

2 Experimental
We fabricated OLEDs using a vacuum evaporation machine. The OLEDs consists of X nm of 4,4′-bis[N-(1-naphthyl)-N-phenyl-amino]-biphenyl as a hole transport layer (HTL), 20 nm of 0.5 mol% 1,4-bis[2-[4-[N,N-di(p-tolyl) amino]phenyl]vinyl]benzene doped 4,4′-bis(9-carbazolyl) biphenyl as a light-emitting layer, 10 nm of bathocuproine as a hole block layer, Y nm of tris(8-hydroxyquinoline)aluminum as an electron transport layer (ETL), 0.4 nm of LiF, and MgAg (9:1 mol ratio) as a metal cathode, upon an indium tin oxide coated glass substrate. Active sizes of all the devices were 1.0 mm² by using a metal shadow mask. Here, we fabricated three devices with different thicknesses in the HTL and the ETL, of which thicknesses are shown in Table 1.

<table>
<thead>
<tr>
<th>Device</th>
<th>X (nm)</th>
<th>Y (nm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Device A</td>
<td>50</td>
<td>30</td>
</tr>
<tr>
<td>Device B</td>
<td>40</td>
<td>25</td>
</tr>
<tr>
<td>Device C</td>
<td>30</td>
<td>20</td>
</tr>
</tbody>
</table>

A transmission speed was estimated from a frequency dependence of an output power when a sine wave voltage was applied using the programmable FM/AM standard...
signal generator (SG-7200, KENWOOD). The output power was recorded using an avalanche photo diode. The amplitude of the sine wave voltage and the bias voltage were 5 V.

We also measured rise and decay times when a pulse voltage with the duration of 1 μs was applied. Rise and decay times were defined as times required for optical responses to change from 10 % to 90 % of the maximum output power and the inverse, respectively.

3 Results and discussion

Figure 2 shows the frequency dependence of the output power of OLEDs with different thicknesses of organic layers. The transmission speed increases with the decrease in thicknesses of organic layers, and the fastest \(-3\) dB cutoff frequency of 12 MHz has been achieved for the thinnest device with thicknesses of \(X = 30\) nm (α-NPD) and \(Y = 20\) nm (Alq3).

![Fig.2 The frequency dependence of the output power of OLEDs with different thickness of organic layers.](image1)

Figures 3 (a) and (b) show rise and decay times as a function of a pulse voltage at a bias voltage of 5 V. Both rise and decay times decreases with increasing the pulse voltage. Carrier mobility of organic materials tends to increase with increasing an applied voltage. Therefore, transient properties are considered to improve with increasing a pulse voltage.

Although the decrease in thicknesses of organic layers causes the increase of the device capacitance, we observed that both rise and decay times shortened with the decrease in thicknesses of the HTL and the ETL. This is mainly contributed to the effect of electric-dependent carrier mobility. However, at high pulse voltage of 10 V, response times of all the devices reach almost the same, indicating that carrier mobility does not limit the transmission speed by applying a high pulse voltage.

![Fig.3 The influence of a pulse voltage on (a) rise and (b) decay times at the bias voltage of 5 V.](image2)

4 Conclusion

We have investigated the relationship between thicknesses of organic layers and transient properties of OLEDs. A transmission speed can increase with the decrease in thicknesses of organic layers, and the maximum cutoff frequency of 12 MHz has been achieved.

5 References

1.3-μm-Wavelength Quantum-Dot Lasers for Temperature-Stable High-Speed Direct Modulation

T. Yamamoto\(^1\), M. Ishida\(^4\), N. Hatori\(^4\), K. Watanabe\(^4\), N. Kumagai\(^4\), K. Otsubo\(^1\), H. Sudo\(^1\), Y. Nakata\(^4\), M. Sugawara\(^4\), and Y. Arakawa\(^4\,5\)
Fujitsu Laboratories Ltd.\(^1\), Fujitsu Ltd.\(^2\), OITDA\(^3\)
10-1 Morinosato-Wakamiya, Atsugi 243-0197, Japan
Tel: +81-46-250-8195, Fax: +81-46-248-5192, e-mail:yamamoto.tsuyoshi@jp.fujitsu.com
Collaborative Institute for Nano Quantum Information Electronics\(^4\), Research Center for Advanced Science and Technology\(^5\), The University of Tokyo
4-6-1, Meguro-ku, Komaba, Tokyo 153-8505, Japan
QD Laser Inc.\(^6\), 1-14-17 Kudan-kita, Chiyoda-ku, Tokyo 102-0073, Japan

Abstract: Low-driving-current temperature-stable 10-Gb/s modulation under fixed driving condition was realized in 1.3-μm-wavelength quantum-dot lasers. For further improvement of modulation characteristics, increased modal gain was also confirmed with antimony-mediated high-density quantum dots.

1. Introduction

Recently, quantum-dot (QD) lasers\(^1\) have been paid much attention because some of their theoretically predicted unique characteristics have been realized owing to the progress in crystal growth technology. Using InAs QDs on GaAs substrates, we can make 1.3-μm-wavelength lasers for optical fiber communication. The drastically improved temperature stability by introducing p-doping to the QD active layers\(^1\,7\) is one major advantage of QD lasers compared with conventional quantum-well lasers in the 1.3-μm wavelength range. We have made first demonstration of temperature-stable 10-Gb/s operation using p-doped InAs QD lasers\(^4\) and improved the characteristics so far\(^5\,7\).

In this paper, we describe our recent activities on 1.3-μm-wavelength QD lasers. We have realized low-driving-current temperature-stable 10-Gb/s operation. We also observed the increase in modal gain using antimony-mediated high-density QDs.

2. Low-driving-current 10-Gb/s operation

In QD lasers, strong damping originated from the carrier capture process to the QDs mainly limits the modulation bandwidth\(^8\). When the threshold gain of the laser is near the maximum modal gain of the ground state of the QDs, carriers occupy most of the ground states under lasing condition. This makes effective carrier capture time longer. Therefore, to realize high-speed modulation, sufficiently high maximum modal gain compared with threshold gain of the laser is required. Based on rate-equation analysis, we have clarified that 10-layer-stacked structure and optimized threshold gain, which is equal to the sum of the internal loss and the mirror loss, of around 15 cm\(^{-1}\) is necessary for 10-Gb/s modulation in the case of the dot density with 3 \times 10^{10} cm\(^{-2}\)\(^9\). To obtain low driving current, it is essential to choose short cavity length and to set the facet coating for satisfying optimum threshold gain condition.

According to our theoretical analysis, we fabricated 200-μm long lasers with 10-layer-stacked QDs as the active layers. P-type doping was applied to the QD active layers to suppress temperature dependence. To obtain optimum threshold gain for 10-Gb/s modulation, both facets were high-reflection coated with the reflectivity of 81%\(^7\). The threshold current

Fig. 1 Temperature dependence of dynamic extinction ratio under 10-Gb/s modulation with fixed driving condition. Insets are eye patterns at 25, 50, 70, and 90°C.
was as small as 2.2 mA at 25°C and the characteristic temperature was 300K between 20 and 50°C. We also obtained output power of more than 5 mW up to 90°C under 40 mA injection current. Fig. 1 shows temperature dependence of dynamic extinction ratio under 10-Gb/s modulation of this laser with fixed driving condition. The bias current was 23.4 mA and the modulation current was 25.5 mA. We obtained dynamic extinction ratio of more than 5dB between 20 and 90°C. As shown in the insets of Fig. 1, we obtained clear eye openings although slight degradation was observed at higher temperatures. Using this QD laser and a receiver with an electric dispersion compensator, error-free transmission over 300-m-long multimode fiber between 25 and 80°C was also achieved under fixed driving condition [10].

3. Increased modal gain using high-density QDs

To improve the modulation characteristics further, it is necessary to increase the modal gain of QD active layers [9]. One effective way is to increase the dot density so that various efforts such as graded composition strain reducing layer [11] and antimony (Sb) mediation in the growth procedure [12] have been made. We have investigated Sb irradiation on the GaAs surface just before the InAs QD growth using MBE system and succeeded in growing high-density QDs with 6 x 10^{10} cm^{-2} [13]. This is twice as high as the density without Sb mediation. Using the 5-layer-stacked Sb-mediated high-density QDs, we fabricated broad area lasers [14]. Fig. 2 shows the relationship between current density and modal gain evaluated from measured threshold current densities of the lasers with different cavity structures at room temperature. We clearly observed increase in the modal gain by utilizing Sb-mediated high-density QDs. This result indicates that Sb mediation is a promising way to improve the modulation characteristics in InAs QD lasers.

4. Summary

We have realized low-driving-current temperature-stable 10-Gb/s operation of 1.3-μm-wavelength QD lasers between 20 and 90°C. We have also shown increase in modal gain using Sb-mediated high-density QD active layers. The temperature stable operation of QD lasers is promising for future 1.3-μm-wavelength uncooled light sources and further improvement will be expected.

Acknowledgement

This work is supported by the ‘IT Program’, the Special Coordination Funds for Promoting Science and Technology, MEXT of Japan and ‘Photonic Network Project’ which OITDA contracted with NEDO.

References

[13] K. Watanabe et al., to be submitted.
High-performance GaInNAs lasers for optical communications

Takeshi Kitatani, Jun-ichi Kasai, Kouji Nakahara, Koichiro Adachi, and Masahiro Aoki
Central Research Laboratory, Hitachi, Ltd.
1-280 Higashi-koigakubo, Kokubunji-shi, Tokyo 185-8601, Japan
Tel: +81-423-23-1111, Fax: +81-423-27-7786, e-mail: takeshi.kitatani.ue@hitachi.com

Abstract
We demonstrated high-performance GaInNAs/GaAs triple-quantum-well (TQW) edge-emitting lasers. Al-free MBE growth significantly improved the crystal quality of the QWs, leading to a record-low threshold current of 4.4 mA at 25°C and 40-Gbit/s operation at 5°C.

1. Introduction
GaInNAs-based quantum-well (QW) systems are promising for the fabrication of temperature-insensitive long-wavelength laser diodes (LDs) due to their strong electron confinement [1]. So far, we have demonstrated 10-Gbit/s operation and 5,000 hours of reliable operation in GaInNAs/GaAs single-QW (SQW) LDs [2]. However, the threshold current of the GaInNAs-LD was still higher than that of the GaInAs-LD because of residual crystal defects.

Takeuchi et al. have reported that Al contamination promotes the 3D growth of GaInNAs grown by MOCVD and thus degrades its crystal quality [3]. We also observed the contamination of Al in MBE-grown GaInNAs [4]. Therefore, the suppression of incorporation of Al in GaInNAs is thought to be one of the key issues for further improvement of the crystal quality of GaInNAs.

In this paper, we report the successful growth of high performance GaInNAs/GaAs triple-QW (TQW) edge-emitting lasers by using an Al-free MBE method, the demonstration of a record-low threshold current (Ith) at room temperature, and 40-Gbit/s operation at 5°C [5, 6].

2. Experimental
GaInNAs/GaAs QW structures were grown on (001) n-GaAs substrates by MBE using N radicals as the N source [1]. The N radicals were generated by an RF plasma cell, and P2 and As2 fluxes were supplied from PH3 and solid As using cracker cells. Elemental Ga and In were used for the group III growth species. The n- and p-type doping were performed with Si and Be. The In and N mole fraction of the GaInNAs layers were 0.32 and 0.01, respectively. Post-growth thermal annealing of the GaInNAs/GaAs QWs was performed outside the MBE chamber.

3. Results and Discussion
3.1 Photoluminescence of highly strained GaInNAs/GaAs TQW
A higher number of QWs is indispensable to increase differential gain, which results in improved modulation bandwidth of LDs. We, thus, grew GaInNAs/GaAs (7nm/20nm) TQW structures. To investigate the influence of the number of QW layers on the crystal quality, photoluminescence (PL) was measured at room temperature by using a 1064-nm line of a YAG laser, which excites carriers only in QW layers.

The measured PL spectra are shown in Fig. 1. The TQW had a strong PL intensity proportional to the number of QW layers, indicating that a high-quality QW structure was obtained even under accumulation of three highly strained GaInNAs layers.

![Fig. 1 PL spectra of GaInNAs/GaAs QW](image)

3.2 Laser performance
The GaInNAs/GaAs-TQW (5.5nm/20nm) laser with a ridge-waveguide (RWG) structure is shown in Fig. 2. The TQW active layers were embedded into GaAs waveguide layers. The n- and p-type cladding layers consisted of a GaInP...
lattice matched to GaAs, instead of Al(Ga)As.

The typical light-output power versus current characteristics of the RWG laser under continuous-wave (CW) operation at 25°C is shown in Fig. 3. The cavity length was 200 µm. We obtained a threshold current \( I_{th} \) of 4.4 mA. The lasing wavelength was 1.287 µm. To our knowledge, this \( I_{th} \) is the lowest value ever reported for GaInNAs-based TQW lasers in the 1.3-µm range. This low \( I_{th} \) is due to good crystal quality in the GaInNAs TQW active layers.

The 40-Gbit/s eye pattern at 5°C is shown in Fig. 4. The bias current and the modulation current were 67 and 53 mA, respectively. As an indication of the high gain coefficient of the GaInNAs/GaAs active region, this TQW device provided an eye opening. Further improvement of the crystal quality will develop the high-speed performance even at high temperature.

4. Conclusions

We demonstrated high-performance GaInNAs/GaAs TQW edge-emitting lasers, which exhibit a record-low threshold current of 4.4 mA at 25°C and a 40-Gbit/s direct modulation at 5°C. These results are due to a reverse-mesa-type RWG laser structure incorporated with high-quality GaInNAs/GaAs TQW active layers grown by Al-free MBE growth. Therefore, we expect that GaInNAs lasers are good candidates for future high-speed cost-effective lasers.

Acknowledgement

A part of this work was supported by the National Institute of Information and Communications Technology (NICT).

References

RECENT PROGRESS OF 40 Gbit/s DIRECTLY MODULATED LASERS

U. Troppenz, J. Kreissl, W. Rehbein, C. Bornholdt

Fraunhofer Institute for Telecommunications, Heinrich-Hertz-Institut, Einsteinufer 37, 10587 Berlin, Germany, Phone: +49-30-31002238, Fax: +49-30-31002558, E-mail: troppenz@hhi.fraunhofer.de

Abstract
High Extinction ratio of 6 dB is demonstrated in 40 Gbit/s directly modulated lasers based on InGaAsP/InP. The increase of modulation bandwidth up to 32 GHz is enabled by integrated optical feedback.

Introduction
For very-short-reach (VSR) optical links and metro application the direct modulation of lasers provides low cost and compact transmitters. Different approaches for high speed lasers have been investigated in the past. With single section lasers (1.55 µm) having 4 dB extinction ratio a four channel 40 Gbit/s transmission over 40 km standard single mode fibre (SSMF) including dispersion compensation was reported [1]. In [2] an optimized composition of the heterostructure and an increased number of QWs were used to improve the modulation dynamic. The resulting overall bandwidth of 19 GHz was limited by parasitic role off. The highest extinction ratio of 5 dB reported for single section edge emitters so far was achieved with Al-containing quaternary active material (1.3 µm) and a very short active cavity [3]. Other approaches for high speed lasers are based on injection locking (e.g. in VCSELs [4]) and multi-section laser concepts [5, 6].

Recently, we could demonstrate the 40 Gbit/s direct current modulation of an InGaAsP passive feedback DFB laser (PFL) with emission wavelength of 1.55 µm [7]. The high speed operation capability is based on a feedback enhanced modulation bandwidth [8]. The present work verifies experimentally the excellent performance of the PFL transmitter under 40 Gbit/s large signal modulation and evaluates the transmission characteristic over a short optical SSMF link.

Device structure and small signal analysis
We have realized PFL-structures consisting of a DFB laser and an integrated passive feedback (IFB) section (Fig. 1).

The compact two section laser device with a total length not exceeding 600 µm is based on a ridge waveguide design. Antireflection (AR) and high reflection (HR, > 90%) coatings have been applied to the DFB and the IFB facets respectively. The active region consists of In1-xGaAs1-xPy strained layer multi-quantum wells (MQWs) embedded between asymmetric quaternary waveguides. Both, index and complex coupled DFB gratings have been used.

The device characterization has been performed on chip level. A Ground Signal Ground (GSG) microwave probe head was used for biasing and modulation of the DFB laser section, an additional dc bias was applied to the IFB section. For the small signal amplitude modulation analysis the DFB facet output was coupled to a lensed fibre from where it was passed to a 40 GHz photodetector combined with an electrical spectrum analyzer.

Fig. 2: 32 GHz small signal intensity modulation bandwidth of PFL structure (DFB: 50 mA).

The investigated PFL devices exhibit a stable single mode emission with a side mode suppression ratio of more than 30 dB. When recording the optical modulation response under different IFB biasing a strong change of the modulation bandwidth is observed. Fig. 2 illustrates the feedback effect in more detail. Without any impact by the feedback section (absorbing IFB section, dashed line) the situation is found similar to that of a single section DFB laser: The modulation bandwidth is limited by the carrier photon...
(CP) resonance frequencies which are typically in the range of 8 to 12 GHz. A high modulation bandwidth of about 32 GHz which exceeds the CP frequency limit by a factor of 3 is measured for optimum feedback conditions. In this case the modulation properties are improved by inducing a photon-photon (PP) resonance close to a desired frequency of 30 to 40 GHz [5]. It has to be noted that the high bandwidth can be accomplished already at moderate DFB current levels. In the following paragraph the high bandwidth operation regime of PFL lasers is tested under large signal modulation.

Large signal analysis

Eye patterns as well as Bit Error Rate (BER) measurements are carried out with $2^7$-1 pseudo random bit sequence (PRBS) data streams for Non-Return-to Zero (NRZ) 40 Gbit/s signals (Fig. 3).

Fig. 3: 40 Gbit/s NRZ eye pattern of directly modulated laser (DFB: 60 mA).

In the chip level measurements the traces result from electrical mismatch between of the 50 Ω high frequency setup and the low impedance laser load. An accurate impedance matching in a packaged device overcomes this problem and will improve the signal-to-noise ratio (SNR) significantly. At DFB currents of 60 mA clear open eyes have been recorded with extinction ratios (ER) of 4 to 6.6 dB, depending on the modulation power. The respective eye SNRs range from 5 (ER of 4 dB) to 4 (ER of 6.6 dB). Results of BER measurements are shown in Fig. 4. The BER for back-to-back configuration follows a straight line where an error free operation is observed down to 1E-12 with no indication of an error floor. In our measurements the quality of eye diagrams suffers from signal degradation by traces. The BER increases after transmission over different lengths of SSMF links with a typical dispersion of 17 ps/nm/km (no dispersion compensation) at the wavelengths of our lasers. Starting from a back to back ER of 6 dB and eye SNR 4.3 the eye pattern shows no detectable decrease of quality after 0.5 km, a slightly degraded but clearly opened eye after 1 km (penalty of 4 – 5 dB) and stronger degraded but still opened eyes after 2 km.

Fig. 4: Results of BER measurements for back-to-back configuration and for transmission over short SSMF links.

Summary

Large signal 40 Gbit/s direct current modulation with 6 dB extinction ratio is demonstrated for a 1.55 µm InGaAsP based DFB laser. The underlying effect of enhanced optical modulation bandwidth is caused by an integrated feedback section. It allows stable error free eye patterns which are still opened after transmission over up to 2 km SSMF links without dispersion compensation. The investigated PFL structure will provide a promising moderate sized laser for modulator free and low cost transmitter solutions in 40 Gbit/s VSR optical links.

Acknowledgment

The project was co-financed by the EFRE-programme of the European Union under contract 10125597 of the IBB.

References

1. K. Sato et al., ECOC 2004, Stockholm, Sweden, We1.5.7.
2. M. N. Akram et al., ECOC 2006, Cannes, France, Mo3.4.5.
5. O. Kjebon, et al. IPRM 2003, Santa Barbara, USA, FA1.2
6. J. P. Reithmaier et al. IPRM 2005, Glasgow, UK, 05CH37633C.
7. U. Troppenz et al., ECOC 2006, Cannes, France, Th4.5.5.
Abstract:

Vertical-cavity surface-emitting lasers with buried tunnel junction at 1.55 µm wavelength with novel high-speed design are presented. The devices show superior modulation bandwidths above 10 GHz. Wide open eye diagrams enable error-free data-transmission at 10-Gb/s.

1 Introduction

Recently, VCSELs for 1.3 µm and longer wavelengths (LW-VCSELs) have become commercially available, cover the wavelength-range from 1.3 to 2.3 µm, show cw-operation beyond 100 °C, provide high output powers and modulation bandwidths above 10 GHz [1, 2]. This enables several important applications, particularly in the fields of optical data transmission. One of the most powerful approaches for the InP-based devices is the BTJ-VCSEL, introduced in 2000.

In this paper, we present our latest results on long-wavelength BTJ-VCSELs in the AlGaInAs-InP material system optimized for high-speed applications.

2 Structure

The VCSELs are based on the InP-material system using a buried tunnel-junction (BTJ). This technique enables the replacement of p-conducting by n-conducting semiconductor material with lower optical losses, resistance and ohmic heating. This concept has yielded high-performance long-wavelength VCSELs for the entire 1.3 to 2.3 µm wavelength range [3].

3 Results

Especially VCSELs at the telecommunications-wavelengths between 1.3 and 1.6 µm require broadband modulation performance to achieve high data rates. The large electro-thermal tunability can be used to stabilize the wavelength in uncooled operation. The threshold current is typically below 1 mA and the single-mode output-power is above 2 mW at high quantum efficiencies around 36 %. Depending on current aperture and bias current, the differential series resistance is between 25 Ω and 75 Ω. Therefore, no matchbox is needed to drive the VCSEL with a standard 50 Ω RF-system.

Fig. 2 shows the small-signal modulation performance of such a device. A 3 dB-bandwidth of 10 to 12 GHz is demonstrated in a wide bias range. This assures open eyes at 10 Gb/s. The measurement was done using a HP 8720D – 20 GHz – VNA (Vector Network Analyzer) and a calibrated HP 11982A photo-detector. The response of the detector was subtracted from the presented data. The chip with its coplanar connectivity was directly connected by a calibrated cascade microprobe. In order to judge intrinsic and parasitic response, we fitted the measured S_{12}-data to the inset equation, a three-pole filter function.

The first part of this equation can be directly derived by the small-signal analysis of the rate-equations above threshold yielding a two-parameter modulation transfer function characterized by the relaxation oscillation frequency $\omega_r$ and the damping-factor $\gamma$. This damping-factor is proportional to the resonance-frequency with

\[ H(f) \propto \frac{f^2}{\omega_r^2 - f^2 + j \omega_r f \gamma} \]

Fig. 2: Small signal frequency response of a high-speed VCSEL. The solid lines are fits to the inset equation.

Keeping the BTJ-design, we recently reworked our device structure according to Fig. 1 with several benefits. With chip sizes around 30 µm this design shows reduced RC-products for higher modulation bandwidths exceeding 10 GHz [1]. In addition, these devices are flip-chip bondable and feature coplanar connectivity. This features are essential for datacom VCSELs at 1.3 or 1.55 µm.
offset $y_0$ and the $K$-factor $K$, yielding a maximum intrinsic bandwidth

$$f_{\text{max}} = \sqrt{\frac{2\pi}{K}}$$

(1)
due to over-damping [4]. The second part of the inset equation in Fig. 1 models the electrical chip-parasitics by a parasitic pole. The advantage of this approach compared to other techniques is that it can account for parasitics that vary over the biasing conditions. For edge-emitting lasers, the oscillation frequency, and the 3 dB-bandwidth is expected to rise with the square-root of the bias-current above threshold, proportional to the modulation current efficiency.

$$f_\text{bw} \propto \sqrt{I - I_\text{th}}$$

(3)
is derived from

$$f_\text{bw}^2 \propto P_\text{avg}$$

(4)
with $P_\text{avg}$ as laser output-power under the assumption that $P_\text{avg}$ rises linearly with bias-current above threshold which, however, is not the case for the VCSEL with its thermal roll-over in the L-I-characteristics.

Fitting all curves in Fig. 2, a $K$-factor of 0.42 ns can be derived as presented in Fig. 3 (b). According to equation (1) this yields an intrinsic limitation by over-damping of 21 GHz.

The transmission performance of the VCSELs was evaluated regarding their feasibility as directly-modulated transmitters in optical communication systems. The laser was biased at 5.2 mA and modulated with non return to zero (NRZ) data of 2^{23}-1 pseudo-random bit sequence (PRBS) pattern length at 10 Gbit/s. Bit-error-rate (BER) measurements in Fig. 4 (a), obtained in back-to-back (BTB) and in transmission experiments using 22 km dispersion shifted fibre (DSF) and 10 km non-zero DSF, demonstrate error-free performance. The corresponding widely open eye diagrams at BER $= 10^{-9}$ are shown in Fig. 3(b).

As compared to the BTB case, the transmission penalty for 22 km DSF fibre is negligible at 0.4 dB due to zero-dispersion around the 1550 nm wavelength with minimum attenuation. Both the bandwidth limitation of the used bias-T and the photodetector cause degradation of the eye diagrams which have to be taken into account. Consequently, the eye-diagram of the electrical driver without laser and PD already showed considerable rise- and fall-times. Our transmission results highlight the potential of the VCSELs as uncooled transmitters in passive optical networks that eliminate the need for optical amplification or costly high-bandwidth DFB laser transmitters.

4 Conclusions

With their high modulation bandwidth above 11 GHz, LW-VCSELs enable a variety of novel applications in the fields of optical data communications.

5 Acknowledgements

This work has been partly founded by the German Research Council (DFG) and the National Natural Science Foundation of China. The authors gratefully acknowledge the fruitful cooperation with M. Ortsiefer (VERTILAS GmbH), and Ning Hua Zhu (National Research Centre for Optoelectronic Technology).

6 References

3 Böhm, G., et al.: ‘Growth of InAs-containing quantum wells for InP-based VCSELs emitting at 2.3 µm’, *J. Crystal Growth*, 2007, accepted for publication
Tapered Hollow Waveguide Multiplexer for Multi-wavelength VCSEL array
Naoto Kitabayashi, Akihiro Imamura, Akihiro Matsutani and Fumio Koyama
Microsystem Research Center, P&I Lab., Tokyo Institute of Technology
4259 Nagatsuta, Midori-ku, Yokohama-shi, Kanagawa 226-8503, Japan
TEL/FAX: +81-45-924-5114, e-mail: kitabayashi.n.aa@m.titech.ac.jp

Abstract: A tapered hollow waveguide multiplexer is proposed to combine the output of a GaInAs/GaAs multi-wavelength VCSEL array. We demonstrated multiplexing of 4-channel output of a VCSEL array formed on a patterned substrate for coupling into a multi-mode fiber.

I. INTRODUCTION
Multi-wavelength vertical cavity surface emitting laser (VCSEL) array is a good candidate for wavelength division multiplexing (WDM) short reach systems to increase the data capacity of local area networks (LAN), optical interconnects and so on. We realized a multi-wavelength VCSEL array exhibiting a wavelength span of 190nm using MOCVD on patterned substrates [1]. In order to realize WDM transceivers based on multi-wavelength VCSEL arrays, however, a low-cost and compact multiplexer of each VCSEL output is necessary. We have developed hollow waveguide optical devices exhibiting various unique features of temperature insensitivity and large tunability and so on [2]. In this paper, we propose and demonstrate a tapered hollow waveguide multiplexer for coupling multi-wavelength VCSEL array output into a multi-mode fiber.

II. DEVICE STRUCTURE
Our proposed tapered hollow waveguide multiplexer is shown in Fig. 1. The structure consists of two high-reflectivity mirror. The core is air and a VCSEL array is integrated on the bottom mirror in the figure. Each output from VCSEL is radiated vertically from the bottom mirror side. An upper mirror and bottom mirror are placed with a taper angle A. After multiple reflections, the reflection angle of the output is simply changed as

$$\theta_{\text{reflection}} = n \times A,$$

where $\theta_{\text{reflection}}$ is reflection angle, n is the number of reflections. Therefore the propagation direction of the output from each VCSEL is converted to a horizontal direction. The output from the tapered hollow waveguide multiplexer can be coupled to a multi-mode fiber and simple and low-cost multiplexer can be realized without losing the advantages of small foot print of VCSEL arrays.

We fabricated the proposed multiplexer using multi-wavelength 7-ch VCSEL array on the patterned substrate [3]. VCSEL array was fabricated using MOCVD on a patterned substrate, consisting of 22pair-Al$_{0.83}$Ga$_{0.17}$As/GaAs p-DBR for the output mirror, 35.5pair-Al$_{0.45}$Ga$_{0.55}$As/GaAs n-DBR for the bottom mirror, an active region of 8nm-thick Ga$_{0.65}$In$_{0.35}$As/GaAs 3 quantum wells (QWs) with 30nm-thick GaAs barriers. In order to use a VCSEL array substrate as a bottom mirror for a slab hollow waveguide, Au was deposited except a 10 μm square output window. An upper mirror was also prepared by depositing a Au on a GaAs substrate and the length is 2mm. A four channel multi-mode VCSEL array was used for an experiment. Figure 2 shows the I-L characteristics of the array. Au was deposited on the surface of the device and thus all the VCSEL was driven simultaneously at 8.6mA per device.

III. RESULT
The air core thickness at an output side and tapered angle was precisely controlled by a PZT actuator and a micro rotation stage. Figure 3 shows the near field pattern of a tapered hollow waveguide multiplexer output when the output air core is 50μm and the taper angle is 1.4 degree. With this condition, the combined output from the VCSEL array was directly coupled with a 50μm-core multi-mode fiber. Figure 4 shows the spectra of multiplexed four channel output coupled into a fiber. The insertion loss was 35dB. This large insertion loss is primarily due to poor reflectivity of the bottom mirror and lateral diffraction in the slab hollow waveguide. We estimated the average reflectivity of the used mirror surface by ray trace and found that it would be deteriorated from a theoretical Au reflectivity of 98% to 80% as shown in Fig. 5. We believe this is due to rough polyimide surface between each VCSEL and mesa steps. By introducing the trench structure for each VCSEL mesa and a highly reflective DBR, of which theoretical reflectivity is over 99.9% for a hollow waveguide, the insertion loss will be reduced below 1 dB.

Figure 6 shows the calculated output angle of ray-trace propagation depending on the VCSEL position in the multiplexer. The output air-core size is assumed to be 50μm for coupling with a multi-mode fiber. The opposite side of the upper mirror is connected to the bottom mirror and the mirror length L is changed. The position is defined as the distance
of each VCSEL channel from the output end of the hollow waveguide multiplexer. For a 2500μm-long multiplexer, the output from VCSELs placed in 1000μm-long area of the tapered multiplexer can be coupled in a multi-mode fiber so that the output from the multiplexer is within the NA of a 50μm-core multi-mode fiber.

IV. CONCLUSION

We proposed a compact tapered hollow waveguide multi-mode multiplexer to combine the output from a multi-wavelength VCSEL array and demonstrated multiplexing of four channel VCSEL output into a multi-mode fiber. The present large insertion loss of the experiment will be dramatically improved by increasing the mirror reflectivity of the hollow waveguide. The proposed hollow waveguide multiplexer is helpful for realizing compact, simple and low-cost WDM transceiver for short reach applications.

ACKNOWLEDGEMENT: This work was partly supported by NICT.

REFERENCES

Transverse-mode and Polarization Characteristics of 1.55µm Micromachined VCSELs

K. Hasebe¹, W. Janto¹, N. Nishiyama²*, C. Caneau², T. Sakaguchi¹, A. Matsutani¹, F. Koyama¹ and C. E. Zah²

¹Microsystem Research Center, Tokyo Institute of Technology
4259-R2-22 Nagatsuta, Midori-ku, Yokohama, 226-8503, Japan
TEL: +81-45-924-5077, FAX: +81-45-924-5977
²Corning Incorporated, SP-PR-02-3, Corning, NY 14831, USA
*Presently at Tokyo Institute of Technology
Email: hasebe.k.aa@m.titech.ac.jp

Abstract: We present the transverse-mode and polarization characteristics of 1.55-µm micromachined VCSELs. The addition of a micromachined structure designed for athermal operations results in single transverse-mode and stable-polarization operation with an orthogonal polarization suppression ratio of over 20 dB.

1. Introduction

Long wavelength VCSELs are attractive for use in various applications such as local area networks and optical sensors because of potentially low manufacturing cost and low power consumption. We have fabricated micromachined VCSELs with very low rate of wavelength variation with temperature [1]. To utilize single mode VCSELs in photonic networks, both single mode operation and polarization stability are desired. However, the conventional VCSELs need a small diameter of current-injection area to achieve single mode operation, for instance, a tunnel junction less than 5 µm in diameter for 1.55-µm InP VCSELs [2]. To control the polarization of output light, there are several ways such as non-cylindrical resonators [3], externally applied stress [4], and so on.

In this paper, we demonstrate 1.55-µm single mode micromachined VCSELs with low threshold current and stable polarization. The integration of a micromachined structure enables high suppression of high-order transverse modes and orthogonal polarization state.

2. Device structure

Figure 1 shows the schematic structure of the micromachined VCSELs. The base structure of the devices was grown in Corning Incorporated, and is similar to that of a conventional InP-based VCSEL with tunnel junction [1]. The cantilever consists of 5.5-pairs GaInAsP/InP DBR and an InP thermal stress control layer which were grown on a GaInAs sacrificial spacer layer above the active region. A 3-pairs Si/SiO₂ dielectric mirror was deposited on the head of the cantilever to increase the reflectivity. Figure 2 shows the SEM view of (a) DBR mirror on the head of freely suspended cantilever and (b) a fabricated micromachined VCSEL. The mirror diameter and cantilever length are 16 µm and 95 µm, respectively.

3. Characterization

To evaluate the characteristics of the micromachined VCSELs, we compared them with conventional structure VCSELs fabricated using the same wafer. The conventional VCSELs were fabricated by etching both cantilever and spacer layer and then depositing a 4-pairs Si/SiO₂ mirror. Figure 3 shows the I-L characteristics of conventional and micromachined VCSELs for 5-µm and 7-µm tunnel junctions. The threshold currents of micromachined VCSELs are lower than those of conventional VCSELs because of the higher reflectivity of the top mirror. The low threshold (300µA) for the 5-µm device shows the low excess loss of the micromachined structure. The low output power would be improved by optimizing the top mirror reflectivity. Lasing spectra of both VCSELs at 4-mA current are shown in Fig. 4 (a) and (b). The micromachined VCSEL with 5-µm tunnel junction exhibits an SMSR of ~50 dB while the conventional VCSEL shows higher-order modes. The air cavity of the micromachined structure is expected to provide an excess diffraction loss for higher-order modes.

To investigate the polarization characteristics for both VCSELs, we measured their I-L characteristics through a polarizer. We adjusted the orientation of the polarizer with respect to the crystal axis of the substrate. Figures 6 (a) and (b) show polarization-dependent I-L characteristics of conventional and micromachined VCSELs with 5-µm diameter tunnel junctions. The direction of a polarization state is along the [011] crystal direction as shown in the inset. The micromachined VCSEL exhibits a single transverse mode with an orthogonal polarization suppression ratio (OPSR) of over 20 dB while that of the conventional VCSEL is only ~4 dB at the same current of 4 mA. The physical understanding of the polarization stability of micromachined VCSELs is open for discussions.

4. Conclusion

We demonstrated polarization controlled single mode 1.55-µm VCSELs with a micromachined structure. The output light was polarized along a fixed direction with an orthogonal polarization suppression ratio of over 20 dB. We expect athermal VCSELs with single mode and stable-polarization operation.

References


Fig. 1 A schematic structure of micromachined athermal InP VCSEL.

Fig. 2 SEM views of a fabricated micromachined InP VCSEL.

Fig. 3 I-L characteristics of conventional and micromachined VCSELs with 5-µm and 7-µm tunnel junctions.

Fig. 4 Lasing spectra of conventional and MEMS VCSELs with tunnel junction of 5 µm at 4-mA current.

Fig. 5 Lasing spectra of conventional and MEMS VCSELs with tunnel junction of 7 µm at 4-mA current.

Fig. 6 Polarization characteristics of (a) conventional and (b) MEMS VCSELs.
10 Gbps/ch 1×10 VCSEL array at 850-nm wavelength

Kazutaka Takeda, Takashi Kondo, Masateru Yamamoto, Ryoji Ishii, and Nobuaki Ueki
Optical System Business Development, Fuji Xerox Co., Ltd.
2274 Hongo, Ebina, Kanagawa 243-0494, Japan
TEL: +81-46-238-3112, FAX: +81-46-238-4647, E-mail: takeda.kazutaka@fujixerox.co.jp

Abstract: 10 Gbps/ch 1×10 850-nm VCSEL array was fabricated for 100 Gbps optical link module. It was found that static characteristics were uniform. In addition, good modulation characteristics of 10 Gbps for each device were realized over extensive environmental temperatures.

Internet traffic is increasing dramatically now and innovative technologic developments are demanded for long distance network, metropolitan area network, access network, and local area network. Especially, in short distance network, economic and scalable high-capacity optical link techniques are demanded. For this purpose, it is important to reduce the device size and its power consumption. As a light source to solve this issue, vertical-cavity surface-emitting laser (VCSEL) is attractive.

We developed 780-nm VCSEL for laser printer, and we have commercialized the world first VCSEL based high performance color laser printer [1]. By applying accumulated techniques in this development, we try to realize the 100 Gbps optical link module with 10 Gbps/ch 1×10 850-nm VCSEL array. In this paper, 10 Gbps/ch 1×10 VCSEL array aimed to improve the uniformity of the optical characteristics and to clarify technological issues for the 1×10 VCSEL array is reported.

Figure 1 shows the top view of the 1×10 VCSEL array. The channel spacing was 250 μm and the p-type contact pad size was 60 μm.

Fig. 1 Top view of 1×10 VCSEL array.

Figure 2 shows the cross section structure of fabricated selective oxide-confined VCSEL. We introduced the in-situ AlAs oxidation probing technique called OPTALO (Optical Probing Technique for AlAs Lateral Oxidation) [2]. The n-type and p-type contacts were both fabricated on the surface. The n-type contact was fabricated by etching to the n-buffer layer. In addition, the lasing mode was controlled by optimizing the inner diameter of the p-type ring contact (Dm) with respect to the oxidation aperture size (Dox) [3].

Figure 3 shows the LIV characteristics of the 1×10 VCSEL array at room temperature. Threshold current was approximately 0.5 mA for each chip, and the maximum light output exceeded 3 mW. Uniform static characteristics were realized.

Fig. 2 Schematic cross section structure of selective oxide-confined VCSEL.

Fig. 3 LIV characteristics of 1×10 VCSEL.
Figure 4 shows the small-signal responses of the 1×10 VCSEL array at 5 mA under room temperature. This result indicates that the high-speed modulation characteristics of more than 10 Gbps will be realized for each channel because \( f_{\text{3dB}} \) exceeded 10 GHz. We have measured the 10.3125 Gbps modulation characteristics with the 10 Gbps Ethernet mask. The good characteristics of the mask margin of more than 20% were realized for all of 10 channels at 25°C.

![Figure 4 Small-signal response of 1×10 VCSEL array.](image)

Figure 5 shows 10 Gbps eye diagrams at –20°C, 25°C, and 85°C. Satisfactory mask margin of more than 20% were obtained under all three temperatures. These results indicate that the high-speed modulation device without temperature control device could be fabricated.

![Figure 5 10 Gbps eye diagram at –20°C, 25°C, and 85°C.](image)

In summary, we fabricated the 10 Gbps/ch 1×10 850-nm VCSEL array. Good characteristics were realized and we believe that this technology has potential for 100 Gbps VCSEL array module application. In the future, we will investigate the reliability of the 1×10 VCSEL array which is important when used as a VCSEL link module.

**Acknowledgment**

The authors would like to acknowledge the members of optical system business development in Fuji Xerox. This work is supported by National Institute of Information and Communications Technology.

**References**


Modeling of Slow Light Modulator with Tilt Coupling Scheme

Keisuke Kuroki, Go Hirano and Fumio Koyama
Microsystem Research Center, P&I Laboratory, Tokyo Institute of Technology,
4529 Nagatsuta, Midori-ku, Yokohama 226-8503, Japan
Tel: +81-45-924-5077, Fax: +81-45-924-5961, Email: Kuroki.k.aa@m.titech.ac.jp

Abstract
We present the modeling of slow light electroabsorption modulators consisting of Bragg waveguides. The result shows the modulator length can be reduced below 20 μm. We predict a low coupling loss by using a tilt coupling scheme for slow light.

1. Introduction
High-speed short reach applications (-100Gbps) have been attracting much interest. A vertical cavity surface emitting laser (VCSEL) is a key device for short reach optical links. However, the modulation speed of VCSELs is primarily limited by the relaxation oscillation frequency. The integration with an external modulator is one of solutions to go beyond the limit. However, a resonant cavity modulator, which has been only reported for integration with VCSELs, provides us narrow optical bandwidth. On the other hand, slow light has been attracting much interest for reducing the size of various optical devices such as optical amplifiers, optical switches and nonlinear optical devices [1]. We have proposed and demonstrated slow light modulator [2, 3] composed of Bragg waveguides [4]. In this paper, we present the modeling of a slow light GaInAs/GaAs electroabsorption modulator with a tilt coupling scheme.

2. Structure of Slow Light Modulator
Figure 1 shows the slow light modulator with Bragg waveguide, which is integrated with a VCSEL. The output light from VCSEL is reflected by a miro-mirror. The reflected light is coupled into the modulator by using a tilt-coupling scheme [3].
Here we carried out a simple model by replacing multilayer Bragg mirrors with a perfect metal mirror [2]. The calculated slow down factor, which is defined as the ratio of the group velocity of slow light versus that in conventional semiconductor waveguides, is shown in Fig.2. The cut-off wavelength is assumed to be 990 nm. The slow-down factor is over 10 in the wavelength range of 980-990 nm. Thus, the electro-absorption effect is enhanced by a factor of more than 10 in this wavelength range and we are able to reduce the length.

3. Modeling Result
The structure is similar to that of VCSELs with a one-wavelength-cavity. Absorption coefficients of 300cm⁻¹ and 1200 cm⁻¹ are assumed for InGaAs QW absorption modulators with zero-bias and 1V-bias voltages [5]. The device length is as small as 20 μm, which is 10 times smaller than that of conventional electro-absorption modulators. Figure 3 shows the calculated insertion loss and extinction ratio as a function of the wavelength. Even for a 20μm long ultra-compact modulator, we expect an extinction ratio of 7 dB over 972 nm, which will be large enough for short-reach optical links. While the large dispersion of the slow light effect and an increased insertion loss would limit the optical bandwidth of the slow light modulator, we expect optical bandwidth of greater than 5 nm.
An issue is how to couple with slow light in a Bragg waveguide. We proposed a simple and practical method of a tilt-coupling scheme [3]. We assumed that an input beam is 28 degrees tilted from the vertical axis. We carried out the full-vectorial numerical simulation using a film-mode-matching method (FIMMWAVE, Photon Design Co.). Figure 4 shows the calculation model and the calculated intensity distribution. The coupling loss is less than 1.8 dB for TE mode with a 4 μm-spot-size Gaussian beam input. The incident angle is optimized so that the output intensity is maximized as shown in Fig. 5. Figure 6 shows the calculated insert loss versus wavelength. It is noted that the insertion loss includes the absorption loss enhanced with slow light effect. The coupling loss excluding the absorption loss is 1.8 dB, indicating a possibility of low coupling loss for slow light excitation.

4. Conclusion
We presented the modeling of a slow light GaInAs QW electroabsorption modulator consisting of a Bragg reflector waveguide. The tilt-coupling scheme gives us low coupling loss and enables us to realize a miniature modulator integrated VCSEL.

References
Fig. 1 Schematic structure of slow light modulator integrated with VCSEL.

Fig. 2 Calculated slow down factor versus wavelength.

Fig. 3 Calculated insertion loss and extinction ratio of Bragg waveguide modulator.

Fig. 4 Calculated mode and calculated field distribution of slow light modulator with tilt-coupling scheme.

Fig. 5 Calculated intensity of input and output intensity for different incident degrees.

Fig. 6 Calculated insertion loss versus wavelength.
PLZT photonic functional devices for future network systems

Hiroyuki Tsuda
Department of Electronics and Electrical Engineering, KEIO University
3-14-1 Hiyoshi Kohoku-ku, Yokohama, Japan
Tel: +81-45-566-1627, Fax: +81-45-566-1529, e-mail: tsuda@elec.keio.ac.jp

Abstract: PLZT photonic functional devices including a high-speed switch, an arrayed-waveguide grating, and a tunable wavelength filter are described. Optical slot switching networks using PLZT switches are also successfully demonstrated.
Detection of multiple hydrocarbon gases by broadband difference frequency generation using apodized $\chi^{(2)}$ grating

T. Umeki, M. Asobe, Y. Nishida, O. Tadanaga, K. Magari, T. Yanagawa and H. Suzuki
NTT Photonics Laboratories, NTT Corporation
3-1 Morinosato Wakamiya Atsugi, Kanagawa, 243-0198 Japan
Tel: +81 46 240 3218, fax: +81 46 240 4300, e-mail: umeki@acl.ntt.co.jp

Abstract
Absorption lines of CH$_4$ and C$_2$H$_6$ are detected simultaneously with a 3 $\mu$m broadband difference frequency generated in a direct-bonded quasi-phase-matched LiNbO$_3$ waveguide using an apodized $\chi^{(2)}$ grating.

1. Introduction
A mid-infrared light source has attracted much attention because it is useful for sensing environmental gases. A 3 $\mu$m band light source is particularly important for sensing hydrocarbons because there are many strong absorption lines in this wavelength range [1] [2]. However, no practical laser source has been developed that can emit a continuous wave at room temperature. On the other hand, quasi-phase-matched (QPM) difference-frequency generation (DFG) is an attractive technique for generating mid-infrared light because we can utilize well-established telecom laser diodes as pump and signal sources. Recently, we demonstrated a 3 $\mu$m band laser light generation in a periodically poled LiNbO$_3$ (LN) ridge waveguide (WG) that was made by using a direct bonding technique [3]. A direct bonded WG is especially suitable for generating mid-infrared light because it avoids any undesired absorption by hydroxyl or other groups. We achieved a high conversion efficiency of 40%/W by making the best use of the advantages of direct bonding [4]. However, the narrow bandwidth of the current waveguide, yielded by quasi-phase matching, limits its application as a multiple-gas sensor. So there is a strong need for a broadband QPM device if we are to realize a widely tunable mid-infrared light source.

In this paper, we demonstrate multiple hydrocarbon gas detection with a 3 $\mu$m broadband difference frequency generated using an apodized $\chi^{(2)}$ grating.

2. Widely tunable 3-$\mu$m-band DFG
In a uniform QPM-grating device, the bandwidths can be increased by reducing the interaction length. This is because the bandwidth scales inversely with length. However, since the conversion efficiency scales quadratically with length, this approach results in a marked decrease in the conversion efficiency. Another commonly used broadband device technology is the chirped $\chi^{(2)}$ grating [5]. It is possible to achieve a linear tradeoff between conversion efficiency and bandwidth by using a chirped grating. However, attempts to obtain a high conversion efficiency with a moderate bandwidth have resulted in large ripples in the tuning curve due to interference arising from various phase-matching conditions of the grating. To suppress these ripples, we achieved apodization in a QPM wavelength converter by changing the duty ratio in a $\chi^{(2)}$ grating [6].

Using the apodization and direct-bonding techniques, we demonstrated a 3 $\mu$m band DFG with a large bandwidth and a high conversion efficiency. We used a 3 inch $x$ cut non-doped LN wafer and a 3 inch $z$ cut LiTaO$_3$ wafer for the waveguide layer and substrate, respectively. First, we formed QPM gratings on the LN wafer. We chirped the poling period linearly from 28.489 to 28.311 $\mu$m, and gradually reduced the duty ratio at both ends of the device to achieve apodization. After fabricating the QPM grating, we directly bonded the two wafers together. We then reduced the thickness of the waveguide layer to 11 $\mu$m by lapping and polishing. Finally, we fabricated 17 $\mu$m wide ridge waveguides using a dicing saw. The finished device was 38 mm long.

Figure 1 shows the measured tuning curve as a function of signal wavelength using a 1.064 $\mu$m pump wavelength. The corresponding idler wavelength is shown on the upper horizontal axis. The theoretical tuning curves for apodized and linear-chirped gratings are also plotted for comparison. The experimental and calculated tuning curves agree well. By using apodization, the ripple in the phase-matching curve was effectively suppressed to $\pm$0.5 dB. A bandwidth as wide as 60 nm in the 3 $\mu$m band and a DFG efficiency of 2 %/W were obtained.

Fig. 1. Measured and calculated DFG tuning curves
3. Detection of multiple hydrocarbon gases

Figure 2 shows the experimental gas detection setup. We used a 1.05 µm laser diode as a pump source. We also used a 1.55 µm band external-cavity laser diode (ECLD) and an erbium-doped fiber amplifier (EDFA) as a signal source. The pump and signal beams were combined with a fiber coupler injected into the QPM-LN WG. Pump, signal, and idler beams radiate from the QPM-LN WG output facet, and the input beams are separated from the DFG idler output beam with a dichromatic mirror. The input light was measured with a powermeter. The pump and signal powers at the output facet of the QPM-LN WG were measured and found to be approximately 600 µW and 100 mW, respectively. The DFG output beam passed through a Ge filter and was divided by a beam splitter into two parts, namely a gas-absorption detection beam and a reference beam. Both beams passed through gas cells or reference cells and were detected by a PbSe photodiode detector. Each of the idler outputs was independently measured with the lock-in amplifier. The QPM-LN WG temperature was set at 25 °C. All the gas cells that we designed and used were made from the same fused silica cylinder and had anhydrous silica windows, which were tilted to prevent reflection. There were two hydrocarbon gas cells and two reference cells. One cell was filled with CH₄ at 9 Torr and the other cell contained C₂H₆ at 5 Torr and buffer gas at 495 Torr. The CH₄ and C₂H₆ cells had path lengths of 20 and 10 cm, respectively.

![Fig. 2. Experimental setup for measuring hydrocarbon gas absorption lines](image)

Figure 3(a) shows the absorption spectrum we obtained when CH₄ and C₂H₆ were observed simultaneously. The spectrum was obtained from a single scan of the signal with a wavelength resolution of 0.01 nm/step. We obtained the absorption spectrum by dividing the transmission spectrum through the gas cells by the transmission spectrum through the reference cells, as shown in Fig. 2. Figure 3(b) shows other experimental data. Each gas cell and the reference cell were used in each gas-absorption measurement. The absorption spectrum agrees with that in the HITRAN database. We successfully demonstrated the detection of multiple gases using the broadband DFG technique.

![Fig. 3. Measured absorption spectrum.](image)

(a) Simultaneously observed spectra for CH₄ and C₂H₆
(b) Superimposed individual spectra for CH₄ and C₂H₆

4. Conclusion

We demonstrated the simultaneous detection of the absorption lines of CH₄ and C₂H₆ with a 3 µm broadband difference frequency generated in a direct-bonded QPM LiNbO₃ waveguide. We obtained a broadband absorption spectrum over 100 nm in the 3 µm region by using the apodization of a χ(2) grating. This technique is useful for application to a multiple-gas sensor and broadband spectroscopy experiments.

References

Novel wavelength filter in a periodically poled Ti:LiNbO$_3$ channel waveguide

(Advanced Photonics Research Institute, Gwangju Institute of Science and Technology, 1 Oryong-dong, Buk-gu, Gwangju 500-712, Republic of Korea, Tel: +82-62-970-3380, Fax: +82-62-970-3419, Email : laks@gist.ac.kr)

Abstract
We have demonstrated the Solc-type wavelength filtering and tuning in a periodically poled Ti:LiNbO$_3$ (Ti:PPLN) channel waveguide which has a domain period of 16.6 µm.

1 Introduction
The quasi-phase-matched (QPM) technique allows us to access a large nonlinear coefficient ($d_{33}$) of LiNbO$_3$. During past ten years, various application, such as efficient wavelength conversion [1,2], optical pulse compression [3], all-optical logic gate [4] and all-optical signal processing [5,6] have been demonstrated in a periodically poled LiNbO$_3$ (PPLN). Recently, several papers on Solc-type wavelength filter based on bulk PPLN device were reported [7,8]. However, up to now, all reported papers on Solc filter have been restricted to use bulk PPLN. In this contribution we report, for what we believe is the first time, the Solc-type wavelength filtering and the wavelength tuning based on a waveguide device.

2 Characteristics of Ti:PPLN waveguide
A channel waveguide of with a width of 7 µm was fabricated by diffusing 980-nm-thick Ti stripes upon the -Z face of a 80-mm-long, 13.2-mm-wide, and 0.5-mm-thick Z-cut LiNbO$_3$ substrate along its X axis. Afterwards, a microdomain structure with a periodicity of 16.6 µm was generated by using an electrical field poling technique with liquid electrodes and annealed to remove the stress which was come from during electrical field poling [2]. We confirmed the qualities of the QPM structure and waveguide in a Ti:PPLN device by the measure of a second-harmonic (SH) curve and the propagation loss of waveguide, respectively. Figure 1 shows the SH curve and the photograph of Ti:PPLN sample after selective etching by a HF:HNO$_3$. The 3 dB bandwidth of SH curve and SH conversion efficiency were measured to be 0.2 nm and 300 %/W, and such a narrow bandwidth and high conversion efficiency indicate that a good QPM structure (duty cycle ~ 50:50) was fabricated through the whole length of the waveguide (80 mm).

Table 1. The characteristics of Ti:PPLN waveguide.

<table>
<thead>
<tr>
<th>Characteristics</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length of PPLN (mm)</td>
<td>80</td>
</tr>
<tr>
<td>Periodicity of microdomain (µm)</td>
<td>16.6</td>
</tr>
<tr>
<td>SHG efficiency (%/W)</td>
<td>300</td>
</tr>
<tr>
<td>Propagation loss @ 1530 nm (dB/cm)</td>
<td>0.12 ™</td>
</tr>
<tr>
<td>Mode size@ 1530 nm (µm)</td>
<td>4.3X2.3™</td>
</tr>
</tbody>
</table>

The propagation losses of TM- and TE-polarization beam at 1530 nm were measured to be -0.11 dB/cm and -0.03 dB/cm, respectively. At the same time, the near-field mode profiles of the PPLN waveguide at both
polarization were measured to be 4.3 µm X 2.3 µm (TM) and 6.3 µm X 4.2 µm (TE). From the waveguide characterization, we confirmed that the Ti-indiffused waveguide can guide TM and TE polarization beam simultaneously with single mode profile and less than a 0.08 dB/cm polarization dependent loss (PDL). The detailed characteristics of Ti:PPLN waveguide were listed in Table 1.

3 Experimental results

The experimental setup to perform the Solc-type filtering based on the Ti:PPLN channel waveguide is shown in Fig. 2. A optical signal from a wavelength swept fiber laser based on a semiconductor optical amplifier (SOA) and a Fabry-Perot tunable filter was collimated and end-fire coupled into Ti:PPLN waveguide which is placed between two crossed polarizers by X10 objective lens. The polarization direction of input beam was adjusted parallel to the Y-direction and the filtering signal was observed by optical spectrum analyzer (OSA). The average power and sweeping frequency of the wavelength swept fiber laser were 10 mW and 15 kHz, respectively. The 3 dB bandwidth of the laser was about 70 nm. Each microdomain of Ti:PPLN works as half-wave plates in a folded Solc filter. In a folded Solc filter, the optical axes of half-wave plates are alternately aligned at angles of +θ and -θ with respect to the plane of input beam's polarization. As the same manner, there is a angle between the optical axes of the positive and negative domains in the Ti:PPLN.

The measured transmission spectra of the 16.6 µm period Ti:PPLN waveguide type Solc filter is shown in Fig. 3. The solid lines indicate experimental data. The left line and the right line were measured at 19 ºC and 13 ºC, respectively. The measured FWHM of the filter was about 0.2 nm, which is almost same as the expected value in theoretical calculation (0.21 nm). As increase of the operation temperature of the Ti:PPLN, the center wavelength of the filter was shifted short wavelength. The wavelength tuning rate by temperature change was about -0.683 nm/ºC.

4 Conclusion

We have demonstrated the Solc-type wavelength filtering and tuning in Ti:PPLN channel waveguide which has a domain period of 16.6 µm. In practical point of view, a waveguide-type PPLN Solc filter is more useful device than a bulk type Solc filter. In our experiments, the FWHM of wavelength filter was a enough narrow (~ 0.2 nm) for using in a tunable filter for all-optical wavelength routing. We believe that the Solc filter based on Ti:PPLN waveguide will be very useful components for future birefringence tunable wavelength filter in optical communication.

5 References

40 Gbit/s Switchable OR/XOR Logic Gates Using a PPLN Waveguide

Jian Wang, Junqiang Sun, Qizhen Sun

Wuhan National Laboratory for Optoelectronics, School of Optoelectronic Science and Engineering, Huazhong University of Science and Technology, Wuhan 430074, Hubei, P. R. China
Tel: +86 27 6284 3946; Fax: +86 27 8779 2225; E-mail: wjhustoe@163.com

Abstract: We report a simple realization of switchable OR/XOR logic gates at 40Gbit/s using sum-frequency generation in a periodically poled lithium niobate waveguide. By changing the input signal powers, OR and XOR logic gates are obtained.

1. Introduction

All-optical logic gates are essential elements to implement various all-optical signal processing in future high-speed optical networks. Recently, several schemes have been explored to enable all-optical logic functions such as NOT, AND, NAND, XOR, NOR, OR, etc. For example, all-optical XOR, NOR, OR logic gates using semiconductor optical amplifiers (SOAs) [1], AND/NAND gates using semiconductor microresonators [2], and NOT gate using a periodically poled lithium niobate (PPLN) waveguide [3, 4] have been reported.

PPLN-based all-optical logic gates are very attractive for several distinct advantages of high nonlinear coefficient, ultra-fast response, and no spontaneous emission noise. However, previous researches on PPLN-based all-optical signal processing mostly focused on all-optical wavelength conversion [5,6], switching [7], and add/drop [8]. Up to now, PPLN-based all-optical OR and XOR logic gates have not been demonstrated yet. In this paper, we propose and demonstrate, for the first time to our knowledge, all-optical switchable OR/XOR logic gates at 40 Gbit/s based on sum-frequency generation (SFG) in a PPLN waveguide. The operation performance including eye diagrams and Q-factor are simulated.

2. Operation principle

![Schematic diagram of the PPLN-based all-optical switchable OR/XOR logic gates.](image)

Figure 1 illustrates the schematic diagram of the proposed switchable OR/XOR logic gates based on SFG in a PPLN waveguide. Two input signals ($\lambda_{SA}$, $\lambda_{SB}$) with independent data streams A and B are launched into the PPLN waveguide and participate in the SFG nonlinear interaction. Note that the SFG process occurs only when both input A and B data bits are ‘1’. During the SFG process, one photon from signal $\lambda_{SA}$ and another photon from signal $\lambda_{SB}$ are annihilated to create one photon at the wavelength of a sum-frequency wave. Thus both of the two signals are depleted when input A and B data bits have the same value of ‘1’. It should be noted that the degree of signal depletion depends on the input signal power level. The higher the powers, the more deeply two signals are consumed [4]. Therefore, by appropriately choosing the input signal powers, the combination of two signals at the output of the PPLN waveguide may generate two specific logic gates corresponding to OR and XOR, respectively. When input signals have a low power level and half of which is consumed in the SFG process with input A and B data bits of ‘1’, the OR logic gate is obtained by combining the two output signals from PPLN. For input signals with a high power level, both of the two signals with input A and B data bits of ‘1’ are almost completely depleted by SFG, which results in a data bit ‘0’ of the combination of two output signals from PPLN, yielding the XOR logic gate. As a result, switchable OR/XOR logic gates can be implemented at different input signal powers with the proposed scheme. The truth table for OR/XOR logic gates is clearly shown in Table 1.

3. Results and discussions

The switchable OR/XOR logic gates are simulated based on the coupled-mode equations describing the SFG process [4, 6]. For simplicity, the waveguide propagation loss is reasonably ignored, while the group-velocity mismatch (GVM) and group-velocity dispersion (GVD) are taken into consideration. Two synchronized input signals ($\lambda_{SA}$, $\lambda_{SB}$) are assumed to be independent 2^{-1, 40Gbit/s pseudorandom bit sequence (PRBS) return-to-zero (RZ) data streams with hyperbolic-secant pulse type and 5-ps pulse width. The peak powers of two signals ($P_{A0}, P_{B0}$) coupled into the PPLN waveguide satisfy the relationship of $P_{A0} = P_{B0} \cdot \lambda_{SA} / \lambda_{SB}$ [4]. The microdomain period of a 40 mm long PPLN

<table>
<thead>
<tr>
<th>A</th>
<th>B</th>
<th>OR</th>
<th>XOR</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>0</td>
<td>1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

The microdomain period of a 40 mm long PPLN
waveguide is assumed to be 18.8 \textmu m to meet the SFG quasi-phase matching (QPM) condition for \( \lambda_{S4} = 1550.0 \) nm, \( \lambda_{SB} = 1538.0 \) nm and the corresponding sum-frequency wavelength at 772.0 nm.

Fig. 2. Temporal waveforms for switchable OR/XOR gates. (a)(b) Input A, B. (c) Input A+B. (d)(g) Output A from PPLN. (o)(h) Output B from PPLN. (f) Output A+B for OR. (i) Output A+B for XOR. The peak power of input signal A is 110mW for (d)-(f) and 1W for (g)-(i).

Figure 2 presents simulation results for switchable OR/XOR logic gates at 40Gbit/s. Fig. 2(a)(b) show sample 20-bit sequences of two independent input signals A (\( \lambda_{S4} \)) and B (\( \lambda_{SB} \)), respectively. Direct combination of input A and B results in 3 power levels as shown in Fig. 2(c) which can not be regarded as OR operation. However, by setting the peak power of input signal A at 110mW, at the output of the PPLN waveguide, both two signals with data bits of ‘1’ are about half consumed during the SFG process as shown in Fig. 2(d)(e), so the OR logic gate can be obtained as shown Fig. 2(f) by combining two equalized output signals with their maximum power of the same value. Note that, by increasing input signal A peak power to 1W, it is found that two signals with data bits of ‘1’ are almost thoroughly depleted by SFG as shown in Fig. 2(g)(h), and the combination of which generates the XOR logic gate as shown in Fig. 2(i). Thus switchable OR/XOR logic gates can be performed by changing the input signal powers. Fig. 3 further depicts the eye diagrams for input A/B, OR, and XOR, respectively. The Q-factor is calculated at approximately 22.5 dB for OR and 34.3 dB for XOR. Fig. 4 shows the dependence of Q-factor on the peak power of input signal A for OR/XOR logic gates. Remarkably, there exists an optimal power value at about 90mW, approaching the largest Q-factor of 37.6 dB for OR. By contraries, the Q-factor for XOR keeps increasing with the increase of input signal power. Therefore, the input signal A peak power should be adjusted close to 90mW for OR logic gate, while the higher the peak power of input signal A, the better the performance of XOR logic gate.

4. Conclusion

A simple scheme of 40Gbit/s switchable OR/XOR logic gates based on SFG in a PPLN waveguide is proposed and demonstrated. The OR and XOR logic gates can be obtained by properly changing the input signal powers.

References

12E1-5

Evaluation of phase change by a HfO$_2$-waveguide

Mach-Zehnder interferometer with a ferro-electric liquid crystal cladding

Hiroki Sato, Katsumi Nakatsuhara and Takakiyo Nakagami

Department of Electrical and Electronic Engineering, Kanagawa Institute of Technology

1030, Shimo-ogino, Atsugi-shi, Kanagawa, 243-0292, Japan, Phone & Fax: +81-46-291-3276

garf.hs@ele.kanagawa-it.ac.jp

Abstract: We fabricated an optical modulator using a HfO$_2$-waveguide Mach-Zehnder interferometer with a ferro-electric liquid crystal cladding, and evaluated its quantity of phase change. We demonstrated a possibility of reducing the device size by using HfO$_2$-waveguide.

1. Introduction

Optical switching devices are needed for Wavelength Division Multiplexing (WDM) networks. We have been studying optical waveguide devices having a ferro-electric liquid crystal cladding (FLC) for optical switching. The FLC cladding provides small device size because of its higher refractive index change than that of conventional electro-optical crystals. In addition, FLC materials have relatively high speed response of sub-microseconds and bistable characteristics that eliminate a state-sustaining voltage required for the device. We have reported an optical switch using a FLC-clad Si-waveguide, and demonstrated its wavelength switching operation at 1550nm wavelength[1-2]. By using a HfO$_2$ instead of Si for the waveguide substrate, we can expect the larger phase change in the waveguide due to the larger influence from the FLC birefringence.

As a preliminary study, we constituted a Mach-Zehnder interferometer modulator with a HfO$_2$ waveguide having a ferro-electric liquid crystal (FLC) cladding at a wavelength of 1550nm and evaluated the phase shift of the propagating light due to the refractive index change in FLC.

In the present paper, we will report the evaluation of phase change characteristics of the FLC-clad HfO$_2$-waveguide by using the developed MZI.

2. Principle of device operation

The devise is based on the interferometric nature that output light power is modulated according to the phase difference between the light waves in the two arms of MZI[3]. There is an output light from the device when they are in-phase waves (2π rad) with equal amplitude, while no output light when opposite phase waves (π rad). The HfO$_2$ core layer is covered with the FLC cladding layer only in the phase change region as shown in figure 1.

In other regions, the HfO$_2$ core layer is covered with a SiO$_2$ cladding layer as shown in figure 2, which is introduced to prevent the effective index change due to FLC layer.

The alignment layer for aligning FLC molecules is coated on the upper indium tin oxide (ITO) electrode. For phase change operation, the voltage is applied between upper ITO electrodes and Si substrate layer.

The molecule orientation of FLC inclines by the tilt angle ±θ with respect to the initial rubbing direction according to the polarity of the applied voltage (Fig.3).

The effective index of the waveguide in the phase change region depends on the FLC molecule orientation. Thus, the output light intensity can be altered corresponding to the applied voltage polarity.

We calculated distribution of electric field intensity of the propagating light wave by effective index method. By assuming that the refractive index of Si was 3.48 and that of HfO$_2$ was 1.98[4]. Figure 4 shows the calculated length of the waveguide in the phase change region to achieve the phase change of π rad.

When the thickness of core layer is 500nm, the required length of the HfO$_2$-waveguide in the phase change region is 95.7μm, while, that of Si-waveguide was 1336.2μm.

It is clearly shown that the phase change region using HfO$_2$-waveguide is greatly shorter than that using Si-waveguide. This is because the refractive index of HfO$_2$ is lower than that of Si and the evanescent field in the FLC cladding of HfO$_2$-waveguide is larger than that of Si.
3. Device structure

The optical waveguides of Mach-Zehnder interferometer are rib waveguide using HfO$_2$ guide layer (thickness 500nm) of the substrate which was loaded on a 1000nm-thick SiO$_2$ cladding layer of the Si wafer. We have reported the HfO$_2$-waveguide fabricated at 1550nm wavelength elsewhere[5].

The waveguide width was 2.0μm. The rib height was 108nm. The relative refractive-index difference was 4%. The length of the waveguide in the phase change region was 4mm equivalent to 41.8π phase shift.

The structure of the arm waveguide in the interferometer was symmetric. The curvature radius of the waveguides in Y-couplers at input and output was 5mm. The overall device size was 18.4mm.

4. Experimental results

We used a 1550nm-wavelength semiconductor laser as a light source. We applied alternating pulse voltages of ±10V with the repetition rate of 10Hz.

A TE-polarized light was input to the MZI through a polarization controller and a polarization-maintaining fiber. A TE-polarized light was input to the MZI through a polarization controller and a polarization-maintaining fiber. A TE-polarized light was input to the MZI through a polarization controller and a polarization-maintaining fiber.

The waveguide width was 2.0μm. The rib height was 108nm. The relative refractive-index difference was 4%. The length of the waveguide in the phase change region was 4mm equivalent to 41.8π phase shift.

The structure of the arm waveguide in the interferometer was symmetric. The curvature radius of the waveguides in Y-couplers at input and output was 5mm. The overall device size was 18.4mm.

We proposed and fabricated a novel HfO$_2$-waveguide MZI switch having a FLC cladding. The fabricated device demonstrated the feasibility of a compact optical switch using HfO$_2$-waveguide. We will make further efforts toward reducing the device size and developing optical functional devices using FLC-clad waveguide.

6. Acknowledgements

This study partially supported by Grant-in-Aid of Scientific Research No.18360189 and SCOPE of MIC Japan (No.042103012)

7. References


Micromachined tunable 1.55µm vertical cavity, multiple air-gap filters and lasers: fabrication, characterization, scaling potential and applications

H. Hillmer, T. Kusserow, N. Dharmarasu, S. Irmer, J. Daleiden, A. Hasse and M. Bartels
University of Kassel, Institute of Nanostructure Technologies and Analytics (INA) and Center for Interdisciplinary Nanostructure Science and Technology (CINSaT), Heinrich-Plett Str. 40, D-34132 Kassel, Germany, Tel. +49-561-8044485, Fax +49-561-8044488, e-mail: hillmer@ina.uni-kassel.de

B. Vengatesan and T. Hayakawa
NanoTech Laboratory, Canon Electric Co. Ltd, 2888-1 Rikka, Kumabari, Nagakute-cho, Aichi-ken, 480-1101 Japan.

Abstract: Widely tunable micro-opto-electro-mechanical-system (MOEMS) based filters are presented, consisting of an air-gap cavity embedded between two highly reflective multiple solid/air-gap Distributed Bragg Reflectors (DBRs) to be applied in fibre-optic communication, sensorics or analytics.

Micromachined tunable Fabry–Perot filters and VCSELs [1-7] are interesting components for communication and sensing. In our structures wavelength tuning is performed by electrostatic actuation of at least one of the vertically stacked membranes to vary cavity length [2,4,5,6]. Approximately, a multiple of half of the transmitted wavelength matches to the cavity (distance to be varied between the two DBR mirrors). In the case of semiconductor DBRs, electrostatic actuation can be achieved by n-doping one DBR mirror and p-doping the other. Varying the reverse bias voltage of this pn-junction enables to control the central spacing and, thus, the filter dip wavelength.

The technological fabrication is based on three main steps: (i) definition of the multiple layer structure by epitaxy or other deposition methods, (ii) vertical dry etching to define the lateral structure (vertical patterning of the mesa) and (iii) removing the sacrificial layers by selective underetching to generate the air-gaps.

Figure 1 displays an SEM image of a 1.55 µm filter structure with four suspensions. In the 3 period top DBR and 3.5 period bottom DBR, the InP membranes have an optical thickness of 3/4. The GaInAs(P) layers are partially etched to form the 3/4 air-gaps. Due to the very high refractive index contrast between InP (n=3.17) and air (n=1), a wide stop band of 550nm spectral width and a high reflectivity of 99.98% are achieved. The optical quality of the two surfaces of each membrane is defined by the quality of the epitaxial heterointerfaces. The surface micromachining fabrication process requires no micro-mounting since the entire structure is fabricated in a batch process. By varying the applied voltage from 0 to 28 V we experimentally obtained a continuous tuning of the filter wavelength of 221 nm (14% of the absolute wavelength) [6] as shown in Fig. 2. The controlled reverse bias tunes the cavity length of the filter which allows a single wavelength to be transparent while blocking the others within the stopband. The resonant wavelength as a function of actuation voltage of the filter is shown in Fig.2. The inset shows the corresponding reflectance spectra. To the best of our knowledge this is the widest experimental tuning range reported for InP/multiple air-gap DBR based vertical-cavity filters with fast tuning. This wide tuning range with relatively high actuation voltage is due to the strain-stiffening effect of the InP membrane that extends the actuation distance [8]. The line width of the spectrum is narrow (~1 nm) at lower actuation voltage, however, it increases at higher actuation voltage due to the large deformation of the membranes. Another device has provided 142 nm tuning using an ultra-low actuation voltage of 3.2 V. In other devices a line width of 0.1nm has been obtained. These micromachined tunable filters can also be implemented in other semiconductor material systems like Si or even dielectric materials. Thus, the optical communication systems can be based on a large variety of different wavelengths, ranging from the visible range (computer architecture, mobile phones, LAN, sensorics) and across the near infrared (LAN and long-haul telecommunication) to the mid infrared (sensorics, transoceanic fluoride fibre based communication).
Membrane/air-gap structures (Fig. 1) are attractive due to their inherently large refractive index contrast. However, the mechanical stability would not be very high if implemented in a macro size: strong accelerations cause noticeable fatigue of material due to the harmful influence of inertia and gravity forces. Fortunately, this does not hold if implemented in micro or nano size. Since all physical quantities differently depend on space coordinates, the relative significance of fundamental physical forces considerably varies with structure size. This is studied by scaling each direction in space by the same factor $1/a$. Figure 3 schematically depicts fundamental forces as a function of the scaling factor. If, in addition, material constants become dependent on dimensions and quantization effects in electron, phonon and photon systems are involved, there will be no strict potential law ($\sim a^{-n}$) for the electrostatic and magnetic forces. For the electrostatic (magnetic) force, the exponent may vary in the range between 0 and 2 (2 and 4), depending on the specific situation. For the schematic plot shown in Fig. 3, the potential relations $a^{-1.5}$ and $a^{4}$ have been chosen as an example for the electrostatic and magnetic forces, respectively. In addition, the absolute values of the forces depend on geometries and materials. In macro world gravity and inertia forces dominate and cause material fatigue. In micro and nano world these forces disappear strongly and electrostatic forces can be used efficiently for actuating the MOEMS, which explains our extremely low voltages (3.2V from above) and our large tuning ranges.

Financial support by DFG and BMBF and stimulating discussions with F. Römer, M. Strasser, and C. Prott are gratefully acknowledged.

Figure 2. Tuning range and corresponding reflectance spectra (inset) of the InP/air-gap filter under electrostatic actuation.
Silicon Photonic Wire Filter
Using Asymmetric Sidewall Long-Period Gratings
in a Two-Mode Waveguide

Young-Bo Cho, Jin-So Shinn, Byung-Ki Yang, Joo-Hyung Lee, Jun-Bo Yoon and Sang-Yung Shin
Dept. of EECS, Korea Advanced Institute of Science and Technology
373-1, Guseong-dong, Yuseong-gu, Daejeon, Republic of Korea
(Phone) 82-42-869-5420, (Fax) 82-42-869-8020 e-mail: whdud@eeinfo.kaist.ac.kr

Abstract
We demonstrate a silicon photonic wire filter using asymmetric sidewall long-period gratings. The period and depth of fabricated grating are 4.3 μm and 5 nm, respectively. The length of long-period grating is 260 μm. The measured maximum attenuation at the center wavelength is about 13 dB. The bandwidth of the transmission dip is 4.5 nm.

1. Introduction
The high refractive index contrast optical waveguide has attracted much attention in recent years. A promising material for high refractive index contrast is silicon because it has a good optical transmission characteristic at infrared optical fiber communication wavelengths. Moreover, it has compatibility with silicon complementary metal-oxide-semiconductor (CMOS) devices [1].

On the other hand, long-period fiber gratings (LPFGs) have been studied and used for various devices in optical communication and optical sensing [2]. They have several advantages such as easy fabrication and low insertion loss. However, one major limitation of LPFG devices is that they need a relatively long device length to achieve desired characteristics. To partially solve this problem, long-period waveguide grating (LPWG) devices have been reported [3,4]. Their device length may be further reduced by employing high index contrast silicon waveguides.

In this paper, we propose and fabricate a silicon photonic wire LPWG device and investigate its characteristics for a filter. It may be applied to realize a polarization splitter.

2. Device structure and operation principle
The proposed LPWG device consists of single mode waveguide sections, taper sections and a two-mode section with asymmetric sidewall LPGs. The structure is schematically shown in Fig 1. Taper sections connect the input and output single-mode waveguides to the two-mode waveguide. The two-mode waveguide supports both $E_{11}$ mode (TE$_{0}$-like mode) and $E_{21}$ mode (TE$_{1}$-like mode). The operation of this device is based on the coupling between these two modes by using LPWGs. Because of the horizontal asymmetry of $E_{21}$ mode, the LPWG should be asymmetric [5].

![Fig. 1 Structure of the filter](image)

If the $E_{11}$ mode is excited in the single-mode waveguide, it evolves to the $E_{11}$ mode of two-mode waveguide through the input taper section. At the wavelength satisfying the phase matching condition, the $E_{11}$ mode is coupled to the $E_{21}$ mode by the asymmetric LPGs on the sidewall of the two-mode waveguide. Since the output taper removes the $E_{21}$ mode, we obtain notch filter characteristics at the output single mode section.
The width and height of the single mode waveguide are 450 nm and 230 nm, respectively. To guarantee the two-mode operation, the width of multimode waveguide is chosen to be 1 \( \mu \)m. The numerically calculated electric field profile of each mode is shown in Fig. 2. Based on the coupled mode theory and numerical simulations, we choose the period and depth of LPWGs as 4.3 \( \mu \)m and 5 nm, respectively. The total length of the LPWGs is 260 \( \mu \)m.

Fig. 2  Electric field profiles of two-mode waveguide

3. Fabrication procedures and experimental results

The LPWG device is fabricated by using conventional processes of semiconductor devices. Fig. 3 shows the steps. First, both LPGs and waveguides are simultaneously formed by using e-beam direct writing and inductively coupled plasma reactive ion etching. As we adopt a sidewall LPG structure, there is no need to carry out another e-beam direct writing and dry etching processes. Furthermore, a precise alignment process between LPGs and waveguides is not necessary. Then, to characterize the device, we fabricated mode size converters[6] at input and output ends of the LPWG filter.

Fig. 3  Fabrication procedure

The measured transmission spectra of the fabricated LPWGs are shown in Fig. 4. If the \( E_{11}^{x} \) mode is exited, we can find that the LPWGs have maximum attenuation level of 13 dB at 1609 nm of the center wavelength. The 3-dB bandwidth of notch filter is about 4.5 nm. On the other hand, if the \( E_{11}^{y} \) mode (TM-like mode) is exited, the LPWGs do not reveal any resonance dip. From this result, the filter using sidewall LPGs may be used as polarization splitter at the center wavelength.

4. Conclusion

We have demonstrated a silicon photonic wire filter using asymmetric sidewall LPGs. Its fabrication processes are simplified by adopting a sidewall LPG structure. The total length of the LPWGs is 260 \( \mu \)m. The device has maximum attenuation level of 13 dB and 3-dB bandwidth of 4.5 nm at 1609 nm wavelength. The LPWGs can be used as filter and polarization splitter.

Acknowledgment

This work was supported by the Korea Science and Engineering Foundation(KOSEF) grant funded by the Korea government(MOST) (No. R01-2005-000-11280-0)

References

1 Ling Liao et al., IEEE J. Quantum Electron., vol. 41, no. 2, pp. 250, 2005
5 Jose M. Castor et al., Opt. Express, vol. 13, no. 11, pp.4180, 2005
Cross Absorption Modulation Enhancement in Silicon Waveguides

Y. Liu and H. K. Tsang

Department of Electronic Engineering, The Chinese University of Hong Kong, Shatin, N.T., Hong Kong, China.
(Tel: +852-26098252, Fax: +852-26035558, Email: hktsang@ee.cuhk.edu.hk)

Abstract
We demonstrate that the modulation depth of cross absorption modulation can be enhanced in silicon waveguides by using helium ion implantation.

Introduction
Recently there has been much interest on silicon optical nonlinear effects based on 3rd order nonlinearities, which have material coefficients two orders of magnitude larger than that of silica [1]. We previously reported that helium ion implantation reduced the nonlinear optical loss to allow a modest continuous-wave (CW) Raman gain to be obtained without the need for an external electrical field [2] to remove the two-photon absorption (TPA) generated free carriers. In this paper, we experimentally measured the cross absorption modulation (XAM) from TPA. Since the mechanism for ultrafast XAM is an ultrafast effect that depends only on the intensity of the pump beam, ion implantation may affect the depth of XAM by reducing the effective carrier lifetime and thus increase the nonlinear effective length and reducing the accumulation of a background density of free carriers particularly for high repetition rate pump pulses or CW pumping. We compare the XAM of non-implanted and ion implanted waveguides and show that the ion-implanted waveguides had a measured modulation depth 7% deeper than the non-implanted waveguides.

Theory and Experiments
When a high power pump pulse signal and CW probe propagate in the silicon waveguide, the absorption of one pump photon excites an electron from the valance band to a virtual state in the bandgap. XAM occurs when the electron in this virtual state absorbs a probe photon and a phonon to excite the electron across the indirect bandgap. Thus one photon from the CW probe and one photon from the pump are absorbed by the photon-assisted non-degenerated TPA process [3]. The CW probe is thus inversely modulated by the high power pump pulses.

We used femtosecond fiber laser as the pump pulse signal and one CW as probe signal as shown in Fig. 1. The femtosecond fiber laser produced 200fs short pulses at a repetition rate of 20MHz, which was used as short pulses for XAM measurements. An optical tunable filter with a 1.1nm 3dB bandwidth was used before the erbium doped fiber amplifier (EDFA) to limit the spectrum shape of the optical pulses. The amplified optical pulses were combined with the probe signal by a 3dB coupler then coupled into the silicon waveguides. Another filter was used to separate the pump and probe signal. The filtered probe signal was detected by an oscilloscope.

Results and Discussions
The silicon waveguides applied in the XAM measurement have a calculated effective area of 6.2μm² and a length of 17mm. The waveguide width, rib height and slab height are 3.8, 4.3, and 2.8μm respectively. The waveguide’s geometry is important for picosecond optical pulses because the dense population of excess free electrons and holes generated by TPA can initially be rapidly depleted by diffusion of...
carriers away from the optical mode. The effective carrier lifetimes can thus be much shorter than the material recombination lifetime.

The pump pulse was measured by the oscilloscope as shown in Fig. 2(a). The pump pulse appeared to be broadened to ~20ps because of bandwidth of the 32GHz bandwidth photodiode in the oscilloscope. The FWHM of the pump pulse was measured to be 3.8ps by an autocorrelator, as shown in Fig. 2(b). The pump pulses induced XAM in CW signal after propagating in the silicon waveguides.

The coupled average input pump power of the optical trace in Fig.3 is 10.8dBm. Fig. 4 shows the modulation depths of the silicon waveguides. The modulation depths increase with the pump power but the XAM depths do not increase linearly with pump power because the accumulation of free carriers and FCA will introduce loss that will tend to reduce the apparent level of 100% transmission. The reduction of the background FCA in the implanted waveguides allows its modulation depth to be increased by 7% at the highest power level used in the experiment (average pump power of 12mW).

**Conclusion**

We showed experimentally that ion implantation can enhance XAM depth and recovery time in silicon waveguides. The results show the potential of ion implantation in carrier lifetime engineering to reduce the nonlinear loss and thus enhance 3rd order nonlinear effects for high speed all-optical signal processing in silicon.

**Acknowledgements:** This work was supported by a RGC Earmarked grant no: 415905.

**References**

Fabrication of Si wire optical waveguides by clad formation by selective oxidation of Si

Koichi Iiyama, Satoshi Asai, Masahiro Wakashima
Division of Electrical Engineering and Computer Science
Graduate School of Natural Science and Technology, Kanazawa University
Kakuma-machi, Kanazawa, Ishikawa, 920-1192 Japan
TEL: +81-76-234-4887, Fax: +81-76-234-4870
Email: iiyama@ee.t.kanazawa-u.ac.jp

Abstract Si wire waveguides are fabricated by forming clad by selective oxidation of Si. Multimode waveguides are fabricated and the propagation loss is decreased as compared to the conventional fabrication process.

Introduction
Extremely high-Δ optical waveguides are widely studied in order to realize highly integrated lightwave circuits owing to their strong optical confinement in the core. A Si wire optical waveguide on a silicon on insulator (SOI) substrate gives large Δ between the Si core and the SiO₂ or air clad, and are actively studied [1-3]. The core width of the Si wire waveguide is 0.3 ~ 0.4 μm and is usually formed by RIE of the Si. The propagation loss of the Si wire waveguide is seriously affected by roughness of the sidewall of the Si core, and then lithography and RIE processes must be carefully optimized to obtain smooth sidewall.

Here we report a Si wire optical waveguide fabricated by forming clad by selective oxidation of Si. An anti-oxidation film with delineated waveguide pattern is formed on a SOI wafer, and then the SOI wafer is oxidized. As a result, the core width is slightly narrowed as compared with the width of the patterned SiN film. Although the optimum width of the SiN film to achieve single-mode propagation is not yet clear, it may be about 1 μm by taking account of lateral oxidation. In our experiment, the width of the SiN film is about 3 μm at this moment, and the fabricated waveguide may be multimode. We are now fabricating waveguides with 1 μm-wide SiN film.

Figure 1 shows the photograph of the waveguide facets for (a) no oxidation, (b) after 4 hours oxidation, and (c) after 7 hours oxidation. The white area is a Si layer. In (a), a horizontal white bar in the middle of the photograph is the Si layer.

Preparation
A commercially available SOI wafer with a 1 μm-thick buried oxide layer and a 340 nm-thick top Si layer was used as a substrate. A 40 nm-thick Si-nitride (SiN) film was coated on the substrate by LP-CVD, which is used as an anti-oxidation film. The SiN film was first patterned with a straight waveguide photomask by photolithography, and then the wafer was oxidized in wet atmosphere at 1100 °C for 7 hours. Since oxidation is an isotropic process, the Si under the SiN film is slightly oxidized from the edge of the patterned SiN film. As a result, the core width is slightly narrowed as compared with the width of the patterned SiN film.
With oxidation, the white bar is narrowed because of selective oxidation outside the patterned SiN film, which cannot be seen in the photograph, and after 7 hours oxidation, the narrow white bar is isolated from the black area, which indicate the oxidized Si and air. The isolated white bar works as a core of the Si wire optical waveguide.

Figure 2 shows a SEM image of the Si wire optical waveguide after 7 hours oxidation. The oxidized area is slightly swollen because of incorporation of oxygen into the Si layer. From a study using spectroscopic ellipsometry for SOI wafers without patterning the SiN film, the oxygen is slightly penetrated into the Si layer. Upper 60 nm in the Si layer is oxidized and then the thickness of the Si layer is reduced to 280 nm. The penetration of oxygen into the Si layer can be perfectly blocked by using a thicker SiN film (100 nm and more).

Measurements

A light from a 1.55 μm laser diode is coupled to the Si wire optical waveguide through tapered single-mode optical fibers. The propagation loss is measured by cutback method. Figure 3 shows the measured input-output transmittance against the waveguide length. The closed triangles are the results for Si waveguide by perfectly etching the unwanted Si by CF$_4$ plasma etching, and the closed circles are the results for Si waveguides fabricated by 7 hours selective oxidation. The core width is 3 μm for both the waveguide, and both the waveguide is multimoded as a result. The coupling and the propagation losses are tabulated in Table 1. The propagation loss is reduced by the proposed fabrication method. The coupling loss is, however, increased. This is due to the narrowed core width and the reduced thickness of the Si by 7 hours oxidation. If we used a thicker SiN film, the increase of the coupling loss is avoided.

Figure 3: Measured transmittance of the Si wire waveguide.

### Table 1: Estimated coupling loss and propagation loss.

<table>
<thead>
<tr>
<th>Fabrication method</th>
<th>CF$_4$ plasma etching</th>
<th>7 hours oxidation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coupling loss</td>
<td>27.0 dB</td>
<td>29.8 dB</td>
</tr>
<tr>
<td>Propagation loss</td>
<td>10.3 dB/cm</td>
<td>6.3 dB/cm</td>
</tr>
</tbody>
</table>

Conclusions

We fabricated Si wire waveguides by clad formation by selectively oxidizing the Si. Although the waveguides are multimode, the propagation loss is decreased as compared to the Si wire waveguide fabricated by plasma etching.

Acknowledgements

We would like to thank Mr. Y. Kurata for his support on etching process. This research was supported by the Ministry of Education, Science, Sports and Culture, Grant-in-Aid for Scientific Research (C), 17560032, 2005-2006.

References

Apodization of photorefractive transmission grating for wavelength filter

Hirofumi Yoshida, Atsushi Okamoto, Kazuhiro Harasaka
(Graduate School of Information Science and Technology, Hokkaido Univ) N14-W9, Kita-Ku, Sapporo, 060-0814 Japan, Tel: +81 11 706 6522, Fax: +81 11 706 7836, E-mail: spa@optnet.ist.hokudai.ac.jp

Abstract
We propose apodization of transmission grating for the wavelength filter controlled by writing beam intensity in photorefractive crystal. The interchannel crosstalk is suppressed by about 4dB compared to that without apodization.

1. Introduction
Wavelength filters based on volume holographic gratings have been expected to be a quality multichannel demultiplexer for WDM systems [1]. Holographic gratings are induced in the photorefractive and photopolymer medium by illuminating writing beams. Numbers of holographic gratings can be multiplexed in the same medium by changing the incident angle of writing beams. The desired optical signal beam is diffracted by the grating from wavelength multiplexed signal beams, and extracted in different angles. This wavelength filter has some advantages such as narrow band filtering characteristic and compact/simple structure. However, as the channel spacing is narrower, the increase of the interchannel crosstalk becomes a serious problem. The apodization of holographic gratings is effective for decreasing the interchannel crosstalk. In the case of photopolymers, apodization of holographic gratings using Gaussian beams have been proposed [2]. On the other hand, in the case of photorefractive crystals, holographic gratings are apodized with planar coherent writing beams (writing beam1 and 2) due to energy transfer between writing beams through photorefractive two wave mixing [3, 4].

In this paper, we propose apodization of transmission grating controlled by writing beam intensity in photorefractive crystal. In the previous study, apodization effect by using absorption of lossy crystal was shown in reflection geometry [4]. However, the absorption loss decreases total energy efficiency. On the other hand, by using transmission geometry, the grating can be apodized without utilizing material absorption property by setting appropriate beam intensity ratio between writing beams. Our analytical results show optimum beam intensity ratio to minimize the interchannel crosstalk of holographic wavelength filter. We also show that the crosstalk is suppressed by about 4dB compared to that without apodization.

2. Principle of the holographic wavelength filter
In this section, we describe the principle of the holographic wavelength filter which can extract a particular wavelength from multiplexed signal beams. Holographic diffraction gratings are induced in the photorefractive crystal with two writing beams. After the gratings are written, wavelength multiplexed signal beams are incident on the crystal. Only the signal beam with the desired wavelength which satisfies phase matching condition is diffracted, and the others pass through the crystal. The phase matching condition is shown as

\[
\frac{2m_w \sin \theta_{w1}}{\lambda_{w1}} + \frac{2m_d \sin \theta_{dif}}{\lambda_{dif}} = \frac{2m_w \sin \theta_{w2}}{\lambda_{w2}} + \frac{2m_d \sin \theta_{dif}}{\lambda_{dif}}
\]

(1)

where subscripts ‘w1’, ‘w2’, ‘sig’, and ‘dif’ denote writing beam1, writing beam2, signal beam, and diffracted beam, respectively. \(\theta_i\) is the incident angle, \(\lambda_i\) is the wavelength, and \(n_i\) is the refraction index for each beam. \(\lambda_{w1}\) and \(\lambda_{w2}\) are the same wavelength sensitive to photorefractive crystal (for example 514nm), and \(\lambda_{sig}\) is the wavelength in C-band (for example 1550nm). \(\lambda_{dif}\) is equal to \(\lambda_{sig}\). \(\Delta \lambda\) is the band width and \(\Delta k\) is phase mismatching value which has only z-component \(\Delta k_z\) (\(\Delta k_x=\Delta k_y=0\)). According to Eq. (1), the extracted wavelength \(\lambda_{dif}\) and diffracted angle \(\theta_{dif}\) are controlled by \(\theta_{w1}\).

3. Analysis
In this section, we derive optimum beam intensity ratio to minimize the interchannel crosstalk of holographic wavelength filter by apodization gratings. Furthermore, we show that the crosstalk is suppressed compared to that without apodization.
In a photorefractive crystal, energy transfer arises between writing beams through two wave mixing [3]. Owing to the energy transfer, the grating strength has nonuniform distribution in the photorefractive crystal. In transmission geometry, an apodized grating is induced by setting the appropriate intensity ratio of writing beams.

According to coupled wave theory, the complex amplitude of the diffracted beam in lossless crystal is shown as

$$ A_{df}(z) = \int_0^L A_{w1} A_{w2} e^{-i\frac{\pi}{2}} dz $$

(2)

where $A_j$ is complex amplitude of each beam and $I_0$ is the total incident intensity. $A_{w1}$ and $A_{w2}$ in Eq. (2) are shown as [3]

$$ A_{w1}(z) = \sqrt{I_{w1}(0)} \left( 1 + m + me^{-\frac{z}{L}} \right) $$

$$ A_{w2}(z) = \sqrt{I_{w2}(0)} \left( 1 + m \right) $$

(3)

where $\gamma$ is the coupling coefficient of the crystal, and $m$ is the beam intensity ratio $I_{w1}(0)/I_{w2}(0)$. By applying undepleted-signal beam approximation, the diffraction efficiency $\eta$ is derived by Fourier Transform of $A_{w1} A_{w2} I_0$ (which denotes the grating strength) according to Eq. (2), and shown as

$$ \eta = \frac{A_{df}(L)}{A_{w1}(0)} \left( 1 + m - e^{-\frac{L}{2}} \right) $$

(4)

where $L$ is the interaction length.

We calculate the wavelength characteristic of the diffraction efficiency for several intensity ratios by using Fast-Fourier-Transform (FFT). The result is shown in Fig.3. It can be seen that the shape of the sidelobe varies according to the intensity ratio of writing beams. We analyze the intensity ratio to minimize the crosstalk. The interchannel crosstalk between interchannels is defined as the dimension ratio of mainlobe to the primary and secondly sidelobes. Fig.4 shows the crosstalk with respect to the intensity ratio for four coupling strengths. We can see that the optimum intensity ratio depends on $\gamma L$, and the crosstalk decreases as $\gamma L$ increases. Fig.5 shows the grating strength distribution when the crosstalk is minimum in Fig.4. According to Fig.5, the crosstalk is minimized when the peak of grating strength is at the center of the crystal. In addition, we can see that the grating is apodized as $\gamma L$ becomes larger.

$$ A_{df}(z) = \int_0^L A_{w1} A_{w2} e^{-i\frac{\pi}{2}} dz $$

$$ A_{w1}(z) = \sqrt{I_{w1}(0)} \left( 1 + m + me^{-\frac{z}{L}} \right) $$

$$ A_{w2}(z) = \sqrt{I_{w2}(0)} \left( 1 + m \right) $$

$$ \eta = \frac{A_{df}(L)}{A_{w1}(0)} \left( 1 + m - e^{-\frac{L}{2}} \right) $$

$$ \gamma = \text{coupling coefficient of the crystal} $$

$$ m = \frac{I_{w1}(0)}{I_{w2}(0)} $$

$$ L = \text{interaction length} $$

$$ \eta = \frac{A_{df}(L)}{A_{w1}(0)} \left( 1 + m - e^{-\frac{L}{2}} \right) $$

$$ \text{Normalized diffraction efficiency} $$

$$ \text{Crosstalk vs. beam intensity ratio} $$

4. Summary

We have proposed apodization of transmission grating controlled by writing beam intensity in a photorefractive crystal. We derived optimal beam intensity ratio to minimize the interchannel crosstalk of holographic wavelength filter. Also, we showed that the crosstalk is suppressed by about 4dB.

5. References

Recent Progress on Waveguide Device Design and Its Applications

Katsunari Okamoto

Department of Electrical and Computer Engineering
University of California, Davis, California 95618, U.S.A.
E-mail: katsu@okamoto-lab.com

Integrated-optic waveguide devices become more and more complicated to realize high functionality. Channel numbers of arrayed-waveguide gratings (AWGs) have been dramatically increased up to 400ch in single wafer. In the multi-chip configuration, 4200ch has been achieved with 5GHz channel spacing. Optical functional devices are important to solve electrical bottleneck issues.

Though silica-based waveguides are simple circuit elements, various functional devices are fabricated by utilizing spatial multi-beam or temporal multi-stage interference effects such as AWGs and lattice-form programmable filters.

Various kinds of optical-layer signal processing devices have been developed; they are reconfigurable optical add/drop multiplexers (ROADM), wavelength selective switches (WSS), dispersion compensators, PMD equalizers, dynamic gain equalizers, optical label recognition circuits, temporal pulse waveform shapers, and etc.

Ultra-compact and CMOS compatible silicon waveguides are important for the integration of an optical component and an electronic circuit aiming at higher level of functionalities. Hybrid integration technologies with LiNbO3, InP, MEMS, and polymer will further enable us to realize much more functional and high-speed devices.

This talk reviews the recent progress and future prospects of waveguide device design and its application to trunk communication systems, FTTH access networks and photonic signal processing applications.
Biography

Katsunari Okamoto

Katsunari Okamoto was born in Hiroshima, Japan, on October 19, 1949. He received the B.S., M.S., and Ph.D. degrees in electronics engineering from Tokyo University, Tokyo, Japan, in 1972, 1974, and 1977, respectively.

He joined Ibaraki Electrical Communication Laboratory, Nippon Telegraph and Telephone Corporation (NTT), Ibaraki, Japan, in 1977, and was engaged in the research on transmission characteristics of multimode, dispersion-flattened single-mode, single-polarization (PANDA) fibers, and fiber-optic components. He proposed for the first time the dispersion-flattened fiber (DFF) and succeeded in fabrication of DFF that had chromatic dispersion less than +/-1 ps/km/nm over a wide spectral range.

From September 1982 to September 1983, he was invited as a guest researcher to Optical Fiber Group, Southampton University, England, where he was engaged in the research on birefringent optical fibers.

From October 1987 to October 1988, he stayed at RCAST (Research Center for Advanced Science & Technology) of University of Tokyo as Associate Professor with Dr. E. A. J. Marcatili from AT&T Bell Laboratories.

Since 1990, he worked on the analysis and the synthesis of guided-wave devices, the computer-aided-design (CAD) and fabrication of silica-based planar lightwave circuits (PLCs) at Ibaraki R&D Center, NTT Photonics Laboratories. He has developed a 256x256 star coupler, various kinds of AWGs ranging from 8ch-300nm spacing AWGs to 128ch-25GHz AWGs, flat spectral response AWGs and integrated-optic reconfigurable add/drop multiplexers (ROADM). 200 GHz to 50 GHz spacing AWGs are now widely used in the commercial WDM systems.

From July 2006, he serves as Professor of Electrical and Computer Engineering at the University of California at Davis (UC Davis). His research at UC Davis includes passive and active photonics devices for high-performance all optical networks.

He has published more than 240 papers in technical journals and international conferences. He authored and co-authored 8 books including “Fundamentals of Optical Waveguides (Elsevier)”. Dr. Okamoto is a member of the Institute of Electrical and Electronics Engineers (Fellow), Optical Society of America and the Institute of Electronics Information and Communication Engineers of Japan.
Multi-wavelength channel selective switch by cascading TO-tunable quadruple series-coupled microrings

Yuta Goebuchi, Masahiko Hisada, Tomoyuki Kato*, Yasuo Kokubun
Yokohama National University, Graduate School of Engineering, *presently with Tokyo. Inst. Tech.
79-5 Tokiwadai, Hodogayaku, Yokohama, JAPAN 240-8501
Tel: +81-45-339-4237, Fax: +81-45-338-1157, Email: ykokubun@ynu.ac.jp

Abstract
We demonstrated a multi-wavelength channel selective switch using individual Thermo-Optic tuning of quadruple series-coupled microring resonators. The extinction ratio and the switching crosstalk were extremely improved to 51.8dB and -23.9dB, respectively, from those of double series-coupled device (27.2dB and -8.4dB).

1 Introduction
Vertically coupled microring resonator filter [1] is suitable for the element of ROADM (Reconfigurable Optical Add/Drop Multiplexer), because it can be densely integrated via stacked configuration [1,2] and cross grid topology [1]. In addition, we fabricated a hitless wavelength selective switch (WSS) [3] without blocking other wavelength channels during the tuning. This switch is based on the individual tuning of resonant wavelength of double series-coupled microring resonator, and can serve as the building block of the wavelength selective switch matrix.

In this study, we proposed and demonstrated a multi-wavelength channel selective switch using the cascaded arrangement of hitless WSS based on the quadruple series-coupled microring resonator as shown Fig. 1, which can realize much more box-like spectrum response than the double series-coupled microring resonator.

2 Principle
In a series-coupled microring resonator, when resonant wavelengths of individual microrings are not matched to each other, all wavelength channels are transmitted to the through port and no spectrum peak appears in the drop port (OFF-state). After shifting all resonant wavelengths and adjusting these resonant wavelengths, a spectrum peak in the drop port response appears at another wavelength channel. Therefore, by controlling individual resonant wavelengths of a series coupled microring resonator using TO effect, a hitless wavelength selective switch can be realized.

3 Experiment
First, we fabricated a multi-wavelength channel selective switch by cascading three WSSs using quadruple series-coupled microring resonators. The basic structure including core and cladding materials and the fabrication process are the same as those described in our previous report [3,4].

By applying electric current to each Cr thin film heater above individual ring resonators separately, we measured multi-wavelength switching characteristics. When the electric current was changed, the measured drop port responses varied as shown in Figs. 2-4.

In the initial stage when no electric current was applied, the resonant wavelengths of individual microrings were slightly different due to the fabrication
The ON-states of $\lambda_1$, $\lambda_2$, and $\lambda_3$ were realized by supplying small amount of electric power to each ring as shown in Fig. 2. These three peaks correspond to these three switch elements. The crosstalk for the drop port was -51.5dB and the shape factor [5] was 0.654, which were much better than those of double series-coupled microrings (-35.3dB and 0.40). The FWHM bandwidth was 0.149nm at $\lambda=1549.8$nm ($\lambda_3$).

Figure 3 shows the measured two wavelength selective switching characteristics. The OFF-state of $\lambda_3$ was realized by supplying electric power to two of resonators in SW#3.

Figure 4 shows the measured zero wavelength selective switching characteristics. All OFF-states of $\lambda_1$ to $\lambda_3$ were realized by supplying electric power to two of resonators in each switch element.

Comparing the peak transmittance of Fig. 2 (three wavelengths selected) with that of Fig. 3 (two wavelength selected), the extinction ratio and the switching crosstalk at $\lambda=1550.64$nm ($\lambda_3$) were 51.8dB and -23.9dB, respectively. However, unexpected wavelength shift occurred for $\lambda_1$ ($\Delta\lambda_1=0.58$nm) and $\lambda_2$ ($\Delta\lambda_2=0.30$nm), resulting from thermal interference. Comparing Fig. 2 with Fig. 4 (zero wavelength selected), the extinction ratio and the switching crosstalk at $\lambda=1549.29$nm ($\lambda_1$) were 51.0dB and -26.4dB, respectively. These characteristics were also much better than those of double series-coupled microring resonators (27.2dB and -8.4dB).

4 Conclusion

Since this hitless WSS has a feature of scalable integration, multi-wavelength and multi-port channel selective switch matrix with more than three wavelengths will be possible.

References


Fabrication and characterization of tunable chromatic dispersion compensator based on hollow waveguide

Tomofumi Takeishi, Takahiro Sakaguchi and Fumio Koyama
Microsystem Research Center, P&I Laboratory, Tokyo Institute of Technology, 4529 Nagatsuta, Midori-ku, Yokohama 226-8503, Japan
Tel: +81-45-924-5077, Fax: +81-45-924-5961, Email: takeishi.t.aa@m.titech.ac.jp

Abstract
We present the design and fabrication of tunable dispersion compensators consisting of a tapered grating hollow waveguide. The length is increased for increasing the amount of tunable dispersion ranges. A dispersion of -119ps/nm is obtained for 10mm long devices with 0.1nm bandwidth.

1. Introduction
Tunable dispersion compensators such as thin-film multi-cavity etalons [1] and fiber Bragg gratings [2] are becoming important for high speed transmission of 40Gbps or beyond. We have proposed a tunable dispersion compensator based on hollow optical waveguide with a variable air core [3]. However, the tunable dispersion range is limited by the length of hollow waveguides.

In this paper, we present the design and fabrication of long hollow waveguide grating devices so that the dispersion range is increased.

2. Structure of hollow wave guide grating
Figure 1 shows the schematic structure of fabricated tunable chromatic dispersion compensators based on a tapered hollow waveguide. Two multilayer mirrors are used to confine light in the air core of a tapered hollow waveguide. The mirrors composed of 6pair Si/SiO₂, is designed to be a quarter wavelength stuck for oblique incident angles. A first-order circular diffraction grating is formed on the mirror surface as shown in Fig. 1, which gives us reflection and focusing at an input port.

The Bragg wavelength of the circular grating is dependent on the spatial effective refractive index with the tapered core thickness. Longer wavelengths of input light result in a shorter group delay than shorter wavelengths in the Fig. 1. Variable tapered angles make it possible to realize tunable chromatic dispersion.

First, we calculated the effective refractive index (nₑₑ) of a fundamental mode in a hollow waveguide by using a full-vectorial Maxwell’s solver (FIMMWAVE, provided by Photon Design Company), which is based on a film-mode-matching method [4] and Transfer Matrix method. Figure 2 shows the calculated maximum dispersion versus optical bandwidth for different grating lengths of 1, 3, 5 and 10mm.

The result indicates the tradeoff between the maximum dispersion and optical bandwidth, and the dispersion/bandwidth product is determined by the grating length. For a 3mm long device, the tunable dispersion range we obtained is -3.0ps/nm to 2.5ps/nm.
3. **Experiment**

Figure 3 shows the fabricated 10mm long hollow waveguide device with 6 pair Si/SiO$_2$ mirrors. The width height and pitch of the SiO$_2$ grating formed on the mirror are 390nm, 500nm and 780nm, respectively. A movable mirror is placed as shown in Figure 1, and the air core thicknesses of the input and output are precisely controlled by using a rotating stage and PZT actuator.

We used the experimental setup shown in Fig. 4.

Figure 5 indicates the measured reflectivity and group delay for different air core thicknesses. The obtained dispersions are -39ps/nm, -29ps/nm and -119ps/nm with 3dB optical bandwidths of 0.6nm, 0.8nm and 0.1nm, respectively.

The dispersion/bandwidth product is in the range of 12-23psec. This number is lower than the value (67psec) we expected in Figure 2 for the 10mm long grating. This would be due to excess losses caused by the low open angle (10 degree) of the fabricated circular grating shown in Fig.3.

4. **Conclusion**

We present the design and the fabrication of a tunable dispersion compensator based on a hollow waveguide with a variable air core. The numerical calculation shows the increased grating length of the compensator is effective for increasing the maximum dispersion of the compensator. At 10mm grating length, we expect the maximum dispersion/bandwidth product of 67psec.

For the fabricated 10mm long device, we obtained the maximum dispersion of -119psec/nm for a bandwidth of 0.1nm and the product of 12-23 psec. Further increase of grating lengths without excess psec enables us to increase the tunable dispersion range, which would be useful for tunable dispersion compensation in high speed photonic networks.

**References**


12E4-1 (Invited)
Compact ROADM devices based on PLC technology
Mikitaka Itoh
NTT Photonics Laboratories, NTT Corporation,
3-1 Morinosato, Wakamiya, Atsugi-shi, Kanagawa, 243-0198, Japan
itohm@aecl.ntt.co.jp

Abstract
This paper reviews recent progress on compact ROADM devices based on planar lightwave circuit (PLC) technology. These compact devices with high levels of optical performance are very attractive for constructing metro-WDM systems at reduced cost.

1. Introduction
Recently the demand has been growing for increased transmission capacity on the metro infrastructure, and thus a large number of carriers and multiple systems operators (MSOs) are deploying WDM systems with fixed and reconfigurable optical add/drop multiplexing (ROADM) functions in their optical networks. Such metro WDM systems should be realized compactly and inexpensively to reduce capital expenditure (CAPEX) owing to the limited number of regional network subscribers.

Since the ROADM function is composed of many optical functions such as WDM filtering, switching, variably attenuating and power monitoring, the most attractive way to make ROADM systems more compact is to integrate these functions into a single package [1-6]. Single-package multi-function modules also enable WDM system suppliers to reduce the cost of complex board design and the reliability testing of individual optical modules as well as the system assembly cost including fiber splicing and electrical soldering. This paper reviews recent progress on such compact ROADM devices based on planar lightwave circuit (PLC) technology. PLC technology enables us to realize a variety of compact ROADM devices whatever their channel number and functions.

2. Comparison of PLC integration techniques
Table 1 summarizes the characteristics of the PLC integration technique for multi-function PLC devices. The conventional integration technique involves connecting different optical elements by using optical fibers. With this technique, fiber splicing and routing increase not only the module size but also the cost of the fiber elements and their assembly. The second integration technique involves the direct bonding of different PLC chips [2-4]. Since most of the fiber splicing and routing can be eliminated, this technique makes the devices very compact especially when a large number of optical functions are integrated. Note that, since this connection is based on the same technique as fiber-PLC bonding, the reliability is sufficient for actual applications such as ROADM devices. The final goal of integration is to realize all the optical functions on a single PLC chip. In this case, except for input/output fiber connections, optical connection with optical fibers is totally unnecessary, which means that optical connection loss can completely be eliminated and optical connection reliability is reinforced.

On the other hand, since all the optical elements are fabricated simultaneously on a single wafer, a high quality optical waveguide fabrication process is needed to maintain a good PLC chip yield. Then, in practice, a suitable integration technique is adopted for ROADM devices after taking optical architecture, channel number and detailed specifications into consideration.

3. Wavelength selective switch (WSS) based on multi-chip direct bonding technique
The WSS is a key device for a multi-degree hub node, which connects plural ROADM ring networks with optical transparency. Currently, WSS modules employ micro-electro mechanical system (MEMS) technology [7]. However, the MEMS-based WSS package becomes thicker because of the use of bulk optics. Therefore, we developed a PLC-type 1 x 4 WSS module with a thinner package by utilizing planar optics [3]. This module can be applied to a 4-degree hub node.

Figure 1 shows the configuration of the 1 x 4 WSS module with 100 GHz spacing and 32 channels. To realize this function, we needed 5 arrayed waveguide gratings (AWGs), a 32 channel 4 x 1 switches switch array with

<table>
<thead>
<tr>
<th>Simplicity of optical connection</th>
<th>Discrete assembly</th>
<th>Multi-chip direct bonding integration</th>
<th>Single chip integration</th>
</tr>
</thead>
<tbody>
<tr>
<td>Smaller packaging</td>
<td>Fair</td>
<td>Good</td>
<td>Excellent</td>
</tr>
<tr>
<td>Reliability improvement</td>
<td>–</td>
<td>Good</td>
<td>Excellent</td>
</tr>
<tr>
<td>Yield of optical connection</td>
<td>–</td>
<td>Good</td>
<td>Excellent</td>
</tr>
<tr>
<td>Yield of PLC chip</td>
<td>Excellent</td>
<td>Excellent</td>
<td>Fair</td>
</tr>
</tbody>
</table>

Table 1 Comparison of PLC integration techniques

In one module case

Fig. 1 Schematic diagram of PLC-based 1 x 4 WSS module.

444
4. Small V-AWG device based on single chip integration technique

Several single chip ROADM devices have been demonstrated, and they all provide the excellent compactness of single chip integration, which eliminates the need for fiber connecting different optical functions [5, 6, 9, 10]. On the other hand, the chip sizes of all the devices are still large because they use conventional 0.8% Δ waveguides [5, 6, 9] and 1.5% Δ waveguides [10]. Then we demonstrated a small V-AWG device by using a 2.5% Δ waveguide. Since this waveguide can strongly confine the transmitted light in its core region, we can increase waveguide density and reduce the chip size compared with conventional ROADM devices [5, 6, 8, 9].

Figure 3 shows the circuit configuration, waveguide layout and a photograph of a fabricated 4-inch wafer of a 100 GHz 16 channel V-AWG with 2.5% Δ waveguides. The use of the 2.5% Δ waveguide reduced the chip size to 20 x 17 mm, and as a result, 12 circuits can be fabricated on a 4-inch wafer. Figure 4 shows the transmission spectra with different attenuation levels. The insertion loss including single-mode fiber coupling at 0 dB attenuation is only 5 dB in spite of use of the 2.5% Δ waveguide. This low loss transmission is realized by using a tolerance-relaxed narrow laterally tapered spot-size converter (SSC), which can be practically and easily adapted to 2.5% Δ PLCs [12].

5. Summary

We demonstrated that both the multi-chip bonding and single chip integration techniques are very attractive and practical ways to construct various PLC-based ROADM devices with a compact size at reduced cost. These PLC devices are indispensable for WDM systems for metropolitan areas.

References
Polarization independent microring resonator filter using internal stress and temperature control

Naoki Kobayashi    Nobuhiro Zaizen    Yasuo Kokubun
Yokohama National Univ., Graduate School of Eng., Dept. of Electrical and Comp. Eng.,
79-5 Tokiwadai, Hodogayaku, Yokohama, JAPAN 240-8501
E-mail: ykokubun@ynu.ac.jp

Abstract   The polarization dependence of resonant wavelength of vertically coupled microring resonator filter was eliminated using the photo-elastic effect induced by the internal stress and the temperature control.

1. Introduction
To realize a polarization independent waveguide filter, two methods were proposed. One is a method using half-wave plate inserted in the device, which was adopted for the arrayed waveguide grating filter[1]. However, this method is difficult to be applied to ultra compact photonic devices such as microring resonators with several tens micron radius, because the thickness of $\lambda/2$ plate is ten micron. Another method is the control of aspect ratio of waveguide cross section. However, this method is also difficult to be applied to high index contrast waveguide devices, such as microring resonators, due to their small tolerance.

In this report, we propose and demonstrate the third method using the control of temperature, which is applicable to high index contrast waveguide devices. In addition, we can compensate for the residual polarization dependence of resonant wavelength due to the fabrication error by temperature control. Therefore, polarization independent ultra compact device can be realized easily by this method.

2. Temperature dependence control of resonant wavelength by internal stress control
We have reported an athermal waveguide filter[2] using polymer as cladding material. Last year, we experimentally realized an athermal vertically coupled microring resonator as shown in Fig. 1 by the control of internal stress as a nobel athermal waveguide filter without using polymers[3]. This athermalization method is based on the compensation of temperature coefficient of refractive index by their photo-elastic effect induced by the difference of thermal expansion coefficient between the waveguide core and the substrate.

The photo-elastic constant of most silica-based glass materials is negative. In addition, the thermal expansion coefficient of $\text{Ta}_2\text{O}_5$-$\text{SiO}_2$ compound glass, which was used as the core material in our device[3], ranges from $0.55\times10^{-6}$ [1/K] to $0.80\times10^{-6}$ [1/K] depending on the $\text{Ta}_2\text{O}_5$-$\text{SiO}_2$ content, which is smaller than that of Si substrate ($2.63\times10^{-5}$ [1/K]). Therefore, if the core layer is formed on a Si substrate, the stress of core layer increases with the increase of temperature due to the difference of thermal expansion coefficient between the core and the substrate.

When the internal stress is negligibly small, the photo-elastic effect can be ignored and the temperature coefficient of resonant wavelength $\frac{d\lambda_0}{dT}$ can be calculated in terms of the temperature coefficient of refractive index of waveguide materials and the thermal expansion coefficient of substrate[5].

On the other hand, when the internal stress is too strong to ignore, the refractive index of core layer suffers the influence of both the thermo-optic effect and the photo-elastic effect. However, the controllability of the temperature dependence by the internal stress control was not made clear. Therefore, we numerically analyzed the photo-elastic effect under a strong internal stress using a finite-element solver (ANSYS by ANSYS, Inc.).

3. Numerical investigation of photo-elastic effect
According to our preliminary experiment, a $\text{Ta}_2\text{O}_5$-$\text{SiO}_2$ straddle mounted beam structure with 100µm length fabricated by RF sputtering deposition was largely deflected by an internal stress, and the deflection was about 3.6µm. The internal stress depends on the sputtering condition. We numerically analyzed the thermo-optic effect of silica-based core layer enhanced by the strong internal stress in the following way.
First, the 100μm long beam structure was assumed to suffer a strong internal stress which deflected the beam by 3.6μm (Fig. 2 (a)). When the beam is kept to contact the substrate, a strong stress of the order of 10^9 [Pa] of magnitude is induced to the beam at this time.

Next, both the bottom edges are assumed to be stretched by 1nm which corresponds to the thermal expansion of substrate by the temperature increase of 10°C (Fig. 2 (b)). In this occasion, the internal stress is decreased by the order of 10^8 [Pa] of magnitude due to the thermal expansion. The change of the internal stress of a few 10^8 [Pa] can result in the change of refractive index by the order of 10^-5 of magnitude per degree⁴, which almost coincides with the experimental result³. The thermo-optic coefficients of TE and TM modes are calculated to be 1.269x10^-5 [1/K] and 1.053x10^-5 [1/K], respectively. It was found from this analysis that the photo-elastic effect can be as large as the thermo-optic effect under a strong internal stress, so as to compensate for the temperature dependence of refractive index.

We also analyzed numerically the photo-elastic effect for the case of no intrinsic internal stress. As a result, when both the bottom edges of the beam were expanded in the same way, the change of internal stress was much smaller by 1/100 than the previous case of strong intrinsic internal stress. Therefore, the change of refractive index is negligibly small without the strong intrinsic internal stress.

4. Elimination of polarization dependence

The temperature coefficient of resonant wavelength of the filter with internal stress has a large polarization dependence³ due to the stress birefringence, compared with conventional microring resonator filters. Therefore, the resonant wavelengths for both polarizations can be equalized at a certain temperature. Fig. 3 shows the temperature dependence of resonant wavelengths of a device with the core width and thickness of 1.4 μm and 0.8 μm, respectively. It is seen that the resonant wavelength for TE and TM polarizations can be equalized at around 70°C. By the precise temperature control, the resonant wavelengths were equalized at 69°C as shown Fig. 4, and a polarization independent filter was successfully realized.

When the spectrum peak of TE polarization is athermalized as realized in Ref. [3], the polarization independent filter can be realized by the temperature control without the wavelength shift of TE peak.

447
Super-high-∆ silica-based flat-passband filter using AWG and cascaded Mach-Zehnder interferometers

Koichi Maru¹²³, Tetsuya Mizumoto¹, Hisato Uetsuka²³
¹Graduate School of Science and Eng., Tokyo Institute of Technology, 2-12-1 Ookayama, Meguro-ku, Tokyo, Japan
²Photo-Electronics Research Center, Hitachi Cable, Ltd., 5-1-1 Hitaka-cho, Hitachi-shi, Ibaraki-ken, 319-1414, Japan
Tel: +81 294 25 3833, Fax: +81 294 43 7487, E-mail: maru.koichi@hitachi-cable.co.jp
³Optoelectronic Industry and Technology Development Association, 1-20-10 Sekiguchi, Bunkno-ku, Tokyo, Japan

Abstract

A super-high-∆ silica-based multi/demultiplexer consisting of a multi-input AWG combined with cascaded MZIs is demonstrated. Very flat passband response and low loss penalty due to the passband flattening were obtained.

1 Introduction

In metropolitan and access-area networks, multi/demultiplexers should have a flat spectral response to allow the concatenation of many filters. To flatten the passband of arrayed waveguide gratings (AWGs) with low intrinsic loss, a technique using a combination of two synchronized routers [1-3] is a promising approach. Low-loss and flat-passband characteristics can be achieved with a Mach-Zehnder interferometer (MZI) or a three-arm interferometer for the input of an AWG [1,2]. Also, back-to-back AWGs [3] can be used to obtain wider spectrum although slab-to-array transition loss is twice that of a single AWG. Aiming at achieving both low-loss and flat response, we proposed a flat-passband filter that consists of a multi-input AWG combined with a cascaded MZI structure, and numerically analyzed its optical performance with a theoretical model of a multi-input AWG [4,5]. In this paper, we demonstrate a super-high-∆ multi/demultiplexer with the proposed structure utilizing silica-based planar lightwave circuit (PLC) technology.

2 Circuit structure

The optical circuit of our proposed flat-passband filter is shown in Fig. 1. It consists of a multi-input AWG and a cascaded MZI structure connected to the AWG input waveguides. Reflecting the results in our previous report [4,5], we proposed two stages of cascaded MZIs to achieve a small chip size as well as sufficient flatness. To obtain an appropriate demultiplexing function, the FSR of the first stage MZI was set to half of the channel spacing of the AWG, and that of the second-stage MZIs was set to the same value as the channel spacing. The signals with four equally spaced frequencies $f_1, \ldots, f_4$ within a channel spacing are first divided by the first-stage MZI into two groups $f_1, f_3$ and $f_2, f_4$, and next divided by the second-stage MZIs into individual signals. The lower port of the upper second-stage MZI and the upper port of the lower second-stage MZI should cross each other so that the signals $f_1, \ldots, f_4$ are spatially arranged in this order at the input side of the AWG.

![Fig. 1 Optical circuit](image_url)

We found that insertion loss can be reduced if the input waveguides just before the input slab have narrower core and gap widths [5]. However, they lead to larger coupling between input waveguides. The coupling can cause non-zero chromatic dispersion due to the phase shift in coupled light. To reduce the coupling, we introduced curved waveguides, as shown in Fig. 1, that have a tapered core such that the core width gradually decreases from 3.5 µm to 1.5 µm as the core approaches the slab. The waveguide interval also gradually decreases from 12 µm to 4.8 µm to narrow the core and gap widths. By using this structure, the coupling can be reduced compared with constant core width of 1.5 µm because smaller bending radius is allowed while keeping the bending loss small. Moreover, dummy waveguides were...

448
arranged on both sides of four input waveguides to reduce chromatic dispersion by adjusting the phase of input field to the slab.

3 Demonstration
We demonstrated flat-passband filters with 100-GHz channel spacing using multi-input AWG and cascaded MZIs. The relative index difference $\Delta$ of 2.5% between the core and cladding was used to significantly reduce the chip size from a conventional $\Delta$ (0.8%). The minimum bending radius was 0.8 mm. Consequently, the typical chip size was 38.5 x 17 mm$^2$, which allows us to arrange seven chips on a 4-inch wafer. We fabricated chips with several different design parameters (the number of arrayed waveguides, input waveguide interval, and core widths of input/output waveguides at the edges of slabs).

The spectral responses measured by TE-polarized light are shown in Fig. 2(a) for different design parameters, and the simulation results are shown in Fig. 2(b). The passband shape for the measured results generally agrees well with that for the calculated one in each design. For example, ripples in the passband increased when we compare designs A and B, and the passband width was slightly widened when we compare designs A and D. The spectral responses for the central eight output ports of design D are shown in Fig. 3. We obtained very flat responses for all eight output ports. The 1-dB and 20-dB bandwidths were 0.645–0.658 nm and 0.944–0.960 nm, which correspond to figure-of-merit [2] values of 0.68–0.70. The minimum insertion loss was 5.7–5.8 dB. Comparing with the result of a normal-AWG input port that was not cascaded with MZIs, the increase in insertion loss due to the passband-flattening was estimated to be as small as 0.9–1.0 dB. The total insertion loss also contains fiber-to-chip transition loss of 3.4–3.7 dB that can be reduced by applying spot-size converters to the edges of the chip.

4 Conclusion
We demonstrated a multi/demultiplexer with a very flat passband and small excess loss, which consists of a multi-input AWG combined with a cascaded MZI structure using 2.5%-Δ silica waveguides. We obtained a very flat passband response (1-dB bandwidth of 0.645–0.658 nm and 20-dB bandwidth of 0.944–0.960 nm) with a loss penalty due to the passband flattening of 0.9–1.0 dB. The measured passband shape generally agreed with the simulation results.

Acknowledgments
This work belongs to “Photonic Network Project” which OITDA contracted with New Energy and Industrial Technology Development Organization (NEDO).

References
5 K. Maru et al., J. Lightwave Technol., to be published.
Fabrication and Evaluation of Tunable Band-Selection Interleaver Switch on Silicon Waveguide

Junya MATSUI†*, Soichiro HONDA†, Zhigang WU†, Katsuyuki UTAKA†,
Tomohiko EDURA‡, Masahide TOKUDA‡, Ken TSUTSUI‡, and Yasuo WADA‡

† School of Fundamental Science and Engineering, Waseda University
3-4-1 Okubo, Shinjuku-ku, Tokyo 169-8555, Japan
‡ Nano Technology Research Center, Waseda University, 513 Waseda Tsurumakicho, Shinjuku-ku, Tokyo 169-0041, Japan
Tel: +81-3-5286-3394, Fax: +81-3-5286-3394, E-mail:*willkommen@fuji.waseda.jp

Abstract
We fabricated a tunable band-selection interleaver using a Michelson Interferometer with Si gratings on silicon waveguides. As a result, we obtained a large bandwidth of about 4nm due to a deep grating and interleaver switching.

1. Introduction
With the increase of communication traffic by using wavelength division multiplexing (WDM), low-cost and high-performance filter devices attract attentions [1],[2]. A waveguide using a silicon-on-insulator (SOI) substrate has large index difference between core and cladding layers so that it can give compactness for bending and it is expected to be suitable for such purposes.

A Si Bragg grating is particularly an important element to realize functional devices [3]. When it is applied to a Michelson interferometer (MI) as reflectors a band-selection interleaver can be realized, which is useful for dense WDM (DWDM). For such devices, superior reflection, transmission characteristics and tuning are demanded.

We fabricated a large coupling-coefficient Si Bragg grating with EB exposure and Deep-RIE. Then an MMI coupler and waveguides were formed with photolithography and ICP-RIE. We evaluated the fundamental filtering and tuning characteristics using thermooptic (TO) effect as well as switching.

2. Device structure
A schematic structure of the device we fabricated is shown in Fig.1. Parameters of the sample are as follows. The thickness and width of the rib waveguide are 1.2 µm and 2.1 µm, respectively, which is so-called a middle mode-size waveguide[3]. As for the MMI coupler, the length is 325µm, and the width is 8µm. And the depth and pitch of the grating are 200nm and 228nm, respectively, and the grating length is 600 µm. A SEM view of the rib waveguide with the Bragg grating is shown in Fig.2. The coupling coefficient calculated by those parameters is very large to be about 200cm⁻¹. In addition, the electrodes were formed on one of the waveguide and on the Bragg gratings using Cr and Au.

The measurement was done, as follows. The light from an amplified spontaneous emission (ASE) light source went through a polarization controller (PC), and is incident into Port1. The incident light was split into two arm waveguides of the MI by a 3dB MMI coupler. Only a specific wavelength range was...
reflected by each grating. Reflected lights were interfered in the MMI coupler with a phase difference determined by a path difference of two MI waveguides, and it was observed from Port2. In addition, an electrode installed on a part of one of the waveguides, was able to adjust the phase of light by TO effect. Moreover, other electrodes were also equipped on each grating, and a center wavelength of the reflection band could be tuned. The output characteristics of transmitted and reflected lights were measured by a spectrum analyzer.

3. Measurement result and consideration

Fig.3 shows tuning characteristics of the transmitted lights from the grating for TE mode. The thin line shows the one for no applied voltage. Rather large bandwidth of about 4nm was obtained. The thick and light lines show those under applied voltages of 3V and 6V, respectively. Large tuning of about 18nm was realized when 6V was applied, in which the power consumption was about 3.5W.

In addition, a transmission contrast of about 20dB was attained. It is noted that the transmission contrast can be improved by making an extinction ratio of the input light polarization be higher.

Fig. 4 shows switching characteristics of the interleaver by modulating the phase of one arm waveguide of the MI. The dark line shows for no applied voltage, and a light line for an applied voltage of 1V. It is known that an FSR of the MI is expressed in the following expression.

\[ \text{FSR} = \frac{\lambda^2}{2n_{\text{eff}} \Delta L}, \]

where \( \lambda \) is a wavelength, \( n_{\text{eff}} \) is an effective index, and \( \Delta L \) is a path difference. The path difference was about 450\( \mu \)m, giving the FSR of 0.79nm. Actually the FSR was measured to be about 0.8nm, which is almost coincident with the designed one available for DWDM systems. In addition, the extinction ratio of about 18dB was obtained. Poor extinction characteristics for the band selection in the wavelength range shorter than 1548nm and longer than 1553nm, is attributed to residual sidelobes, and it can be improved by adopting apodization of the grating. [4]

In addition, the interleaver switching was confirmed by modulation the phase by \( \pi \) under 1V application, as show in Fig.4.

4. Conclusion

We realized a tunable band-selection interleaver using a Michelson interferometer and deep Bragg gratings on a Si waveguide, and transmission and reflection characteristics were evaluated. Large extinction ratio of about 20dB and tuning of about 18nm were confirmed. In addition, characteristics of about 0.8nm FSR and an extinction ratio of about 18dB lead to the expectation as an interleaver for DWDM systems. More compact device and improvement of the performances are under study.

References


Roughness reduction of Si waveguides by KrF excimer laser reformation

Eih-Zhe Liang
Department of CSIE, Diwan College of Management, Tainan, Taiwan, R.O.C.
Email: ezliang@mail.dwu.edu.tw

Shih-Che Hung, Ching-Fuh Lin
Graduate Institute of Electro-Optical Engineering, National Taiwan University, Taipei, Taiwan, R.O.C.

Abstract
Roughness reduction of Si waveguides by KrF excimer laser reformation is demonstrated. The root-mean-square roughness of the reactive-ion-etched Si surface is reduced from 13.95 nm to 0.239 nm with a laser intensity of 1.4 J/cm².

In Si photonics, fabrication of low-loss waveguides becomes more challenging as the waveguide size keeps shrinking for compactness [1]. The scattering loss of Si waveguides is proportional to the root-mean-square (RMS) roughness of the sidewall surface [2]. It is predicted to be less than 1 dB/cm if the RMS roughness is less than 1 nm [3]. Dry-etch processes to fabrication Si waveguides, however, create significant roughness. Therefore, post-etch processes to reduce roughness of the sidewall surface are indispensable for low-loss Si waveguides.

Many techniques are devised for roughness reduction of Si waveguides: hydrogen annealing [4], dry oxidation [5], and wet chemical etching [6]. Hydrogen annealing technique operates at 1100 °C and can reduce the RMS roughness to 0.11 nm [4]. Dry oxidation method also operates at 1100 °C and can reduce the RMS roughness to 0.5 nm [5]. These two methods have thermal budget concerns when Si photonics are to be integrated with Si electronics. IC foundries have constraints on wafer doping and flatness, which requires photonic structures to be fabricated after electronics. Therefore post-etch processes at high temperature may degrade the electronics. Wet chemical etching method can reduce the RMS roughness to 0.7 nm [6]. Oxidation methods, however, have a tradeoff between sacrificial oxide thickness and roughness reduction capability. Large roughness like ragged sidewalls by ICP-RIE is hardly reduced by these ways.

Here, we demonstrate KrF excimer laser reformation for roughness reduction of Si waveguides. It is capable of reducing the RMS roughness of reactive-ion-etched Si surface from 13.95 nm to 0.239 nm. This technique has no limitation on thermal budget during the integration of Si electronics and photonics as protective coatings can be used to prevent exposure of the electronics during laser illumination [7].

The principle of laser reformation is to illuminate the vertical sidewall of Si waveguides with a high energy pulse laser at an incident angle θ, as illustrated in Fig. 1. The KrF excimer laser has a wavelength of 248 nm, a pulse duration of 25 ns and a laser intensity of 1~2 J/cm². The absorbed light in Si converts to thermal energy. The illuminated surface Φ serves as a heat generation source and has different generation rates at the vertical sidewall and the ridge as the illumination intensity varies there. With an incident angle larger than 45°, the vertical sidewall has a larger generation rate than that of the ridge. With appropriate intensities and incident angles, only the vertical sidewall melts. The molten sidewall, denoted by Ω, reforms due to the surface tension of liquid Si. It is this surface tension that flattens the roughness. After the laser pulse is gone, Si recrystallizes back to the surface and roughness reduction is done.

Two notable steps are taken in laser reformation. First, the incident laser illuminates vertically and the illuminated Si wafer is placed at an incline angle. This configuration allows the gravity acts on the molten Si along the direction of incident laser. The molten Si is pulled toward the ridge rather than the substrate and results in a less deformed ridge profile. This configuration fixes the direction of laser illumination and therefore avoids a complicated lens setup. Second, the illuminated wafer is placed in a vacuum chamber at a base pressure of 10⁻⁶ torr. This suppresses surface oxidation and impurity in-diffusion, which results in less increase of surface recombinant velocity, about 25 cm/s.

Fig. 1 Illustration of laser reformation of Si waveguides.

Fig. 2 AFM scans of an as-etched Si surface (left) and a laser reformed surface (right).

Reactive-ion-etched Si surfaces are investigated before and after laser reformation. A rough Si surface is prepared by reactive-ion-etch with a mixture gas of SF₆
and CHF$_3$. The as-etched Si surface has a RMS roughness of 13.95 nm, as shown in Fig. 2. It reduces to 0.239 nm after illumination of KrF excimer laser with an intensity of 1.4 J/cm$^2$. The relation of residual RMS roughness with respect to the laser intensities and the number of pulses is shown in Fig. 3. Obviously, the smaller RMS roughness requires the higher intensity and it does not depend on the number of pulses. To have a RMS roughness less than 1 nm, intensity higher than 1.2 J/cm$^2$ must be used.

Fig. 3 The relation of residual RMS roughness of the as-etched Si surfaces by laser reformation with one pulse (square) and five pulses (circle).

Reactive-ion-etched Si ridge waveguides are examined before and after laser reformation. An as-fabricated ridge waveguide with a width of 400nm and a height of 500nm are shown in Fig. 4. The waveguide is etched with gas mixture of SF$_6$ and CHF$_3$, gas pressure of 10 mtorr and r.f. power of 50W [8]. Both of its sidewalls are illuminated by 5 pulses of KrF excimer laser. The illumination intensity is 1.4 J/cm$^2$ and the incident angle is 75$^\circ$ from the normal of the Si substrate. The sidewalls experience a laser intensity of 1.37 J/cm$^2$. The shape of the ridge waveguide is deformed to be like a dome. The degree of deformation is comparable with that of the hydrogen annealing method [4]. In design of Si waveguides, the dimension and morphology has less influence on the scattering loss than the sidewall roughness does. Therefore, the deformation caused by laser reformation is not a tradeoff of roughness reduction capability if the optical mode change due to laser reformation can be predicted.

Fig. 4 SEM cross-sectional views of the as-fabricated Si ridge waveguides before (left) and after laser reformation (right).

Fig. 5 shows the SEM top-views of as-fabricated Si ridge waveguides before and after laser reformation. It has a width of 500nm and a height of 500nm. Obviously, the laser reformed ridge waveguide is smoother than the as-fabricated one. The standard deviation of linewidth of the as-fabricated waveguide is 13.6 nm. It reduces to 3.0 nm in the laser reformed waveguide, by a worst-case estimation from the SEM photos.

Fig. 5 SEM top-views of the as-fabricated waveguide (left) and the laser reformed waveguide (right).

The roughness of sidewalls in Si waveguides fabricated by dry-etch processes comes mainly from two origins. First is the roughness due to the dry-etch processes. In most cases, it exhibits as the variation of the waveguide linewidth along the vertical direction, as shown in Fig. 4. Second is the roughness due to the pattern transfer of masks. It exhibits as the variation of the waveguide linewidth along the lateral direction, as shown in Fig. 5. Clearly, laser reformation can reduce both kind of roughness and produce smoother Si waveguides. The roughness reduction capability of laser reformation is up to 13.95 nm and the ratio of the RMS roughness before and after this technique is about 58.4.

In conclusion, roughness reduction of Si waveguides by KrF excimer laser reformation is investigated. The RMS roughness of the reactive-ion-etched Si surface can be reduced to 0.239nm. This technique can be used as a post-etch processes for low-loss Si waveguides.

Reference
A Wavelength and Converter Assignment Scheme for Decreasing Blocking Probability in Wavelength-Routed Networks

Yukinobu Fukushima, Takahiro Ooishi, and Tokumi Yokohira

The Graduate School of Natural Science and Technology, Okayama University, 3-1-1, Tsushimanaka, Okayama-shi, 700-8530, Japan
Tel: +81-86-251-8248  FAX: +81-86-251-8255  Email: {fukusima, takahiro, yokohira}@net.cne.okayama-u.ac.jp

Abstract—We propose a wavelength and converter assignment scheme for decreasing blocking probability in wavelength-routed networks. Our scheme avoids the competition of wavelength converter reservation by considering the wavelength converter usage history and the number of idle converters.

I. INTRODUCTION

Wavelength-routed networks consisting of reconfigurable wavelength routing nodes interconnected by wavelength-division multiplexed (WDM) fibers (Fig.1) are emerging as a promising candidate for high speed backbone networks. In wavelength-routed networks, a lightpath [1] is established between source and destination edge nodes to transmit data.

It is preferable for nodes to set-up or tear-down lightpaths in a distributed manner for scalability and reliability [2]. In addition, dynamic configuration of lightpaths is required for effective utilization of wavelength resources. A backward reservation protocol is used for distributed and dynamic control of wavelength-routed networks. When the protocol sets up a lightpath, it first gathers wavelength and wavelength converter availability information on the route with a control packet from the source node to the destination node. Next, at the destination node, it determines which idle wavelengths and converters to be assigned to the lightpath. Then it reserves the idle wavelengths and converters from the destination node to the source node.

The wavelength conversion improves lightpath blocking probability by eliminating the wavelength continuity constraint (i.e., the constraint that the same wavelength must be assigned to a lightpath on links along a route). However, because wavelength converter cost remains expensive in the near future, the number of wavelength converters deployed in the network is limited. Therefore, we need to realize as low blocking probability as possible with limited number of wavelength converters.

In [3], FLR (First-Longest lambda-Run), which is a wavelength and converter assignment scheme for a backward reservation protocol, is proposed. The FLR tries to decrease lightpath blocking probability by minimizing the number of converters assigned to a lightpath. However, because the FLR only uses the number of wavelength conversions and does not consider the competition of wavelength converter reservation among node-pairs, it does not always decrease lightpath blocking probability.

In this paper, we propose a wavelength and converter assignment scheme for decreasing blocking probability. Our scheme achieves this objective by 1) not interfering with other node-pair’s wavelength conversion based on converter usage history, and 2) selecting to perform wavelength conversion on intermediate nodes with more idle converters, when setting up a lightpath.

II. FLR ALGORITHM

The FLR [3] decreases blocking probability by minimizing the number of wavelength conversions in setting up a lightpath. Given 1) route of a lightpath and 2) wavelength and wavelength converter availability information on the route, the FLR determines wavelengths and converters to be assigned to the lightpath.

Although the FLR minimizes the number of wavelength conversions in setting up a lightpath, the minimizing policy does not always lead to decreasing lightpath blocking probability. In the above example, it used the last wavelength converter on node \( v_1 \), which may results in interfering with other node-pairs that need to perform wavelength conversion on node \( v_3 \) (e.g., node-pair \( (v_3, v_4) \)).
III. A WAVELENGTH AND CONVERTER ASSIGNMENT SCHEME FOR DECREASING BLOCKING PROBABILITY

A. Algorithm description

We propose a wavelength and converter assignment scheme that tries to decrease lightpath blocking probability by preventing node-pairs from competing for the wavelength converters on the same node. Our scheme achieves this by 1) not interfering with other node-pair’s wavelength conversion, and 2) selecting to perform wavelength conversion on intermediate nodes with more idle converters.

Given the same inputs 1) and 2) as the FLR, our scheme first calculates Cost(\(v_i\)), wavelength conversion cost of each intermediate node \(v_i\) along the route of the lightpath to be set up. Then our scheme determines idle wavelengths and converters to be assigned to the lightpath so that the sum of wavelength conversion costs is the minimum. If such wavelengths and converters are found, the lightpath is successfully set up, otherwise it is blocked.

In determining the idle wavelengths and converters with the minimum sum, we make the layer-k-graph [4] consisting of assignable wavelengths on the route and apply Dijkstra’s algorithm to it. The assignable wavelength resembles a lambda-run, but it may not be as long as possible because we do not always select the longest assignable wavelength.

We define the wavelength conversion cost of intermediate node \(v_i\) as follows:

\[
Cost(v_i) = \frac{U_i}{A_i},
\]

where \(U_i\) is the proportion of the node-pairs except the source-destination pair of the lightpath that used wavelength converters on node \(v_i\) in \(M\) latest entries of wavelength converter usage history, and \(A_i\) is the number of idle converters on node \(v_i\).

Figure 3 depicts an example of establishing a lightpath with our proposed scheme. The number of entries in wavelength converter usage history (\(M\)) is three. The wavelength converter usage history on node \(v_2\) consists of node-pairs (s, d) (v1, d), and (s, v3). This means that these node-pairs performed wavelength conversion on node \(v_2\) in the past. Similarly, the history of node \(v_3\) has (v2, d), (v2, v4), and (v2, v5). In this case, Cost(v1) = (2/3)/2 = 1/3, and Cost(v2) = (3/3)/1 = 1. By applying Dijkstra’s algorithm, our proposed scheme selects the following wavelengths and converters with the minimum cost: \(\lambda_1\), s on links \(e_s\) and \(e_3\), a wavelength converter on node \(v_2\), and \(\lambda_2\), s on links \(e_s\), \(e_4\), and \(e_5\). When there are multiple wavelengths with the same wavelength converter cost, our scheme selects the wavelength with the minimum index (i.e., First-Fit policy).

B. Simulation results

We show our simulation model in Table 1. Figure 4 describes lightpath blocking probability when the number (\(N\)) of converters deployed per node is changed. When \(N\) is 2, 3, and 4, our proposed scheme decreases blocking probability by 7–24 % compared with the FLR. On the other hand, when \(N\) is 1 or larger than 4, our scheme shows almost the same blocking probability as the FLR. When \(N\) is 1, because \(N\) is too small, generally there is little difference in blocking probability between the two schemes. On the other hand, when \(N\) is more than 4, because \(N\) is large and each node has redundant converters, the location of performing wavelength conversion does not affect the blocking probability. In addition, a wavelength assignment policy also does not affect the probability because the wavelength continuity constraint is eliminated by the enough conversion capability. Therefore, in this case, there is also little difference in blocking probability between the two schemes.

Under the actual operation of the network, as few wavelength converters as possible to achieve the near optimum performance will be deployed for cost reduction. Because our scheme offers a blocking probability reduction when the probability is near the optimum, that is, \(N\) is 3 or 4, our scheme is attractive.

Table 1 Simulation parameters

<table>
<thead>
<tr>
<th>Network model</th>
<th>12 node ring network</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wavelength number</td>
<td>16 wavelengths per fiber</td>
</tr>
<tr>
<td>Number of converters per node</td>
<td>Uniform</td>
</tr>
<tr>
<td>Converter type</td>
<td>Full-range converter</td>
</tr>
<tr>
<td>Lightpath request arrival</td>
<td>Poisson arrival with rate (\lambda) on every node-pair</td>
</tr>
<tr>
<td>Lightpath holding time</td>
<td>Exponential distribution with mean N/(\mu)</td>
</tr>
<tr>
<td>Traffic load</td>
<td>41, 45 Erlangs (=(\lambda / \mu))</td>
</tr>
<tr>
<td>Routing</td>
<td>LLR (Least-Loaded Routing)</td>
</tr>
</tbody>
</table>

Converter usage history size (\(M\))

![Figure 3: 5-hop physical route of a lightpath and wavelength conversion cost of node \(v_1\) and \(v_5\)](image)

IV. CONCLUSION

We proposed a wavelength and converter assignment scheme in wavelength-routed networks. Through the simulation experiments, we found that our scheme achieves a blocking probability reduction of about 7–24 % compared with the FLR scheme.

As a future research work, we plan to investigate the effect of wavelength converter usage history size on blocking probability.

REFERENCES


Experimental analysis of bidirectional WDM networks using two identical sets of wavelengths on a single fiber

H. Obara and M. Sakata

Electrical and Electronic Engineering Department, Akita University, 1-1 Tegata Gakuen, Akita, 010-8502 Japan
Phone +81-188892496, Fax. +81-188370406, Email obara@ee.akita-u.ac.jp

Abstract: Power penalty due to in-band crosstalk from Rayleigh backscattering in a new bidirectional WDM network is analyzed experimentally. Some discrepancy in power penalty between theory and experimental results is pointed out.

Introduction: A new bidirectional wavelength-division multiplexing (WDM) technique using two identical sets of wavelengths on a single fiber has recently emerged [1]. Its transmission performance with respect to power penalty was analyzed theoretically. However, no experimental analysis for verifying the theory is published to day. We present preliminary experimental result on the technique, followed by the outlines of a target network configuration and its theoretical performance analysis.

Network configuration: Fig. 1(a) exemplifies a multifiber WDM express network (MWEN) in a ring configuration [2]. There are four nodes (n0 - n3) and four unidirectional fibers (a - d). A fiber has its own destination node and originates a node adjacent to the destination. The network can be decomposed into unidirectional subsystems, one of which is shown in Fig. 1(b). Nodes other than the destination node along a fiber insert wavelengths through optical couplers, which are omitted in Fig. 1(a) for simplicity. Inserted wavelengths pass through intermediate nodes and are finally demultiplexed at the destination. As a result, the arrangement shown in Fig. 1(a) provides a full-mesh wavelength connection among the nodes.

In Fig. 1(b), we see that wavelength efficiency degrades unlike conventional WDM ADM- and XC-systems because wavelengths are not reused within a fiber. A bidirectional design shown in Fig. 1(c), however, reuses waves and can double the efficiency. If we substitute Fig. 1(c) for Fig. 1(a), the efficiency turns out to be nearly twice as high (or two unidirectional fibers, say a and b, are merged into a single bidirectional fiber noted by “a+b”). We provide optical edge filters (EF's) in every node in order to eliminate Rayleigh backscattered (RB) noise of identical wavelength, which severely degrades SNR.

Performance: Consider a cascade of fiber segments, including EF's and optical amplifiers (OA's) shown in Fig. 2. It corresponds to an optical transmission model of the bidirectional system shown in Fig. 1(c). We assume it has m segments of RB noise generation before coupling with counter-propagating wave at the origination node, where a pair of identical waves is launched in both directions. RB noise is generated in every fiber segment, and is amplified and attenuated by OA's with a gain G and EF's with a filter rejection D (i.e., reciprocal of filter gain). The net gain at a node is expressed as $\eta = G/D$. The total RB-signal beat noise at the origination node is given by

$$\sigma_{RB}^2 = \eta \frac{1 - \eta^m}{1 - \eta} \sigma_{RB-S}^2,$$

where $\sigma_{RB-S}^2$ denotes the mean square RB-signal beat noise generated in a fiber segment. $\sigma_{RB-S}^2$ is computed from the autocorrelation function (ACF) of beat noise between signal and RB noise.
which is given by

$$R_{RB} = R_{RB} \left( \frac{\sigma_{RN}^2}{\sigma_{RN0}^2} \right) \left( 1 - e^{-2\alpha L} \right),$$  

(2)

where $R_{RB}$, $\sigma_{RN}^2$, $\sigma_{RN0}^2$, denote the ACF of the electrical field of the input optical signal, RB coefficient, fiber attenuation coefficient, fiber length of a section, and ensemble averaging [1]. The power spectral density of RB beat noise is obtained by taking the Fourier transform of (2). For IM-DD systems, the total power of RB-signal beat noise at the origination node ($\sigma_{RN}^2$) relates to the amplitude of modulation signal ($I_m$), and is then expressed as [1]

$$\sigma_{RN}^2 = I_m^2 \cdot \sigma_{RN0}^2.$$  

(3)

Finally, we have the RB beat noise penalty as [1]

$$PP = -10 \log \left( \frac{Q_0^2 \cdot \sigma_{RN0}^2}{\sigma_{RN}^2} \right) \quad [\text{dB}],$$  

(4)

where $Q_0$ is the Q-factor corresponding to a required bit error rate (e.g., $Q_0=7.7$ for $BER=10^{-14}$).

To verify the theoretical results described above, laboratory experiments were undertaken. First of all we examined basic properties of RB component such as RB coefficient of a single-mode fiber, state/degree of polarization, and power spectrum density. Our experimental results agreed with previous research results. For example, the degree of polarization in different fiber segments adds in their power. This result suggested us to examine the RB power penalty by using a single fiber segment with an OA of a variable gain and a variable attenuator (ATTN), resulting in a simplified experimental set-up (Fig. 3).

Solid lines in Fig. 4 shows theoretical results of RB-induced power penalty (PP) as a function of $m$ and $\eta$ for a typical IM-DD system (i.e. $\alpha=0.25$ dB/km and $L=80$ km). As expected from (1), theoretical curves of PP diverges in the region of $\eta \geq 1$, while it converges elsewhere even if $m \to \infty$. Measured PP's were plotted in Fig.4, where we see some difference between experimental results and theoretical curves in relatively large PP region. A similar symptom was observed in experimental analysis of PP due to four-wave mixing noise [4], where no discussion for that reason was given. We guess such a difference was probably caused by mistuned threshold level of optical receiver in an SDH analyzer that was used in our bit error rate measurements. Recall that the beat noise between signal and RB noise in (2) is generated only for a mark of received optical signal. We assumed optimum threshold (or Q-factor) in our theoretical analysis, while the decision threshold of the SDH analyzer was likely set to the center of the received signal amplitude (we could not adjust the threshold level). Whatever reasons are there, we see that PP's remain less than 1-dB when $\eta<0.9$.

**Conclusion:** We have examined power penalty through experiments in a bidirectional WDM transmission utilizing two identical wavelengths in both directions for the first time. We observed some discrepancy in power penalty between theory and experimental results and discussed its possible cause. Further details of the discrepancy, however, are left for future study.

**References**


A Novel Star-Ring Optical Regional Network Architecture with Dynamic Wavelength/Waveband Broadcast and Select

Yueping Cai, Motoharu Matsuura, Naoto Kishi and Tetsuya Miki
The University of Electro-Communications
(1-5-1 Chofugaoka, Chofu-shi, Tokyo, Japan 182-8585
Tel: +81-424-90-5026  Fax: +81-424-90-7096  Email: cai@ice.uec.ac.jp)

Abstract
We propose a novel star-ring optical regional network architecture with dynamic wavelength/waveband broadcast and select. Data transmission is in star part while protection and control are in ring part. Its advantages are verified by evaluations.

1 Introduction
Photonic transport network using WDM (Wavelength Division Multiplexing) technology is the most promising solution to meet the explosive traffic growth over the internet nowadays. By using SC-MCS (Super Continuum Multi-Carrier Source), NTT has successfully demonstrated the ultra-dense WDM transmission with more than 1000 channels, each operating at 2.5 Gbps in a single fiber in the field experiment.[1] So how to make use of this kind of ultra-dense WDM technology to design the high performance and cost effective future photonic networks is an open question.

With the penetration of broadband internet access, the traffic collected from access networks is not only large in volume, but also yields highly bursty nature, which presents great challenges to current regional networks that are mostly based on legacy hierarchical SONET/SDH (Synchronous Optical Network /Synchronous Digital Hierarchy) ring networks. The access networks interface with the user premises and perform traffic grooming and distribution, while the optical regional networks aggregate the high tributary traffic (eg. 2.5 Gbps or 10 Gbps) from access networks and send them to the backbone network or the other access networks, so they are also called feeder network. [2]

To solve the problems of current SONET/SDH based regional networks, such as slow circuit provisioning, difficult to upgrade, inefficient for dynamic traffic and so on, we propose a novel star-ring optical regional network architecture using ultra-dense WDM technology mentioned above. How to dynamically broadcast, select, route, group and multiplex wavelength/waveband according to traffic demands will be our main research focus.

2 Novel Star-Ring Optical Regional Network
The fiber layer of our proposed network architecture has an overlay star-ring topology, with two RN(Regional Nodes): RNs connecting with each EN(Edge Node) in a star topology and RNr connecting with them in a ring topology respectively, as illustrated in Fig. 1, while the underlying cable layer is still in a classical ring topology.

Both RNs and RNr are connected to two PoPs (Point of Presence) of backbone network for protection purpose and they are also connected to each other with two fibers for communication between each other. EN grooms traffic from access networks and sends it into feeder network destined to other EN or RN. We divide the traffic from EN into two types: one is hubbed traffic which is from EN to RN; while the other is any-to-any traffic, which is from EN to EN inside the network.

Taking advantage of the novel star-ring topology, we propose a dynamic wavelength/waveband broadcast and select with bandwidth reservation and allocation scheme. The whole procedures can be demonstrated as shown in Fig. 2 for hubbed traffic. First, RNr will send out the reservation frame to each EN to collect their bandwidth requirements in the ring part network, after that, the RNr will allocate bandwidth (wavelength or waveband) to each EN according to their reservation information and write the results into allocation frame. Next, the allocation frame will loop in the ring part network again. Each EN
will configure itself to be ready for wavelength/ waveband selection, data transmission and receiving according to the allocation information. The allocation frame will also be sent to RNs for configuration of wavelength/waveband generation, distribution, routing, and grouping. After all finish, the RNs turns on the multi-carrier light source to generate multiple wavelengths and distribute them to all EN, and then each EN will select the wavelength/waveband allocated to it by tunable waveband filter and modulate it and send it to RNs, finally the RNs will route, group and multiplex the wavelengths or wavebands according to their destinations.

While the MHD of hubbed traffic in our proposed star-ring network $\bar{h}_{\text{h1}}$ is one hop, we can easily prove that $\bar{h}_{\text{h1}} \geq 1 = \bar{h}_{\text{w1}}$; then for any-to-any traffic, the MHD of Bi-ring network $\bar{h}_{\text{h2}}$ can be calculated by:

$$\bar{h}_{\text{h2}} = \frac{1}{N(N-1)} \sum_{i=1}^{N} \sum_{j=1, j \neq i}^{N} \min \{ |i - j|, N - |i - j| + 1 \}$$  \hspace{1cm} (2)

While $1 \leq i, j \leq N$, and $i, j$ is the index number of EN. The MHD of our proposed network $\bar{h}_{\text{w2}}$ can be calculated by:

$$\bar{h}_{\text{w2}} = \frac{1}{N(N-1)} \sum_{i=1}^{N} \sum_{j=1}^{N} \min \{ |i - j|, 2 \}$$  \hspace{1cm} (3)

Compare Eq.(2) with Eq.(3), we can find that when $N \geq 3$, $\bar{h}_{\text{h2}} \leq \bar{h}_{\text{w2}}$, and when N is large enough, the MHD of our proposal is very close to 2. Moreover, as N becomes larger, the advantage of our proposal is more obvious. Then let us consider about the comparison of network survivability between each other. We compare it in five basic failure types and the results can be simply summarized in Table. 1.

<table>
<thead>
<tr>
<th>Failure Type</th>
<th>Traditional Bi-Ring</th>
<th>Proposed Star-Ring</th>
</tr>
</thead>
<tbody>
<tr>
<td>RN</td>
<td>Able</td>
<td>Able</td>
</tr>
<tr>
<td>Single EN</td>
<td>Unable</td>
<td>Able</td>
</tr>
<tr>
<td>Multiple ENs</td>
<td>Unable</td>
<td>Able</td>
</tr>
<tr>
<td>Single Fiber</td>
<td>Able</td>
<td>Able</td>
</tr>
<tr>
<td>Multiple Fibers</td>
<td>Unable</td>
<td>Able</td>
</tr>
</tbody>
</table>

4 Conclusion

We have proposed a novel star-ring optical regional network architecture with dynamic wavelength/waveband broadcast and select based on bandwidth reservation and allocation. Shorter mean hop distance, better network reliability and more network flexibility can be achieved by our proposal.

References
Application of Two Upstream Wavelength In Fairness and Priority Environments

Jeauk Park, Byungchul Choi, Youngil Park
School of Electrical Engineering, Kookmin University, Seoul, 136-702, Korea email: ypark@kookmin.ac.kr

Abstract
When two wavelengths are used for PON ONUs, upstream bandwidth utilization can be made very efficient. If wavelengths are not evenly distributed, however, fairness among ONUs is violated. To deal with this problem, a fairness algorithm is suggested. Wavelengths asymmetry can be introduced on purpose to provide different priorities among ONUs. Algorithms and simulation results are presented for both cases.

1 Introduction
EPON is a very economic way to implement FTTH system. After its standardization by IEEE[1], deployment of EPON system is on the rise[2]. Current GEPON is specified to have 1 Gbps bandwidth for both upstream and downstream transmissions. However, the upstream utilization is not good due to guard time and MPCP protocol. Therefore, two-upstream-wavelength EPON system is being considered. ONUs are classified into two groups and different wavelengths are assigned. When different wavelengths are sent in turn, parts of them can be overlapped, obtaining higher bandwidth utilization.

If wavelengths are evenly distributed among ONUs, performance will be fair among ONUs. If this condition is not met, however, the ONU group with more ONUs in it will have lower performance. An algorithm to remove this unfairness is proposed in this study. The asymmetric wavelength assignment can be used by operators to serve different service level agreements (SLA). Performance difference among ONUs is analyzed in this priority environment.

2 Principles of the Proposed Algorithm
Fig. 1 illustrates the proposed two upstream wavelength PON (2W-PON) architecture. Each ONU is equipped with one of these wavelengths. When wavelengths are evenly distributed MPCP protocol orders that different wavelengths are sent in turn from ONU site with part of the packets from consecutive ONUs being overlapped in time. Packets are realigned by adjustment logic. Fig. 2 and Fig. 3 show the case when wavelength distribution is not even. For fairness each ONU is assigned once during one cycle in Fig. 2. Fig. 3 shows the opposite case, where multiple time slots are given to an ONU in one cycle to provide better priority.

3 Performance analysis
Performance is calculated for the proposed 2W-PON under several asymmetric conditions. Total 16 ONUs, 2 msec cycle time, 10 Mbytes ONU buffer size and self-similar traffic are assumed for the calculation [3]. Considering all sources of bandwidth loss including LD on/off time, receiver settling time, preamble, MPCP operation time and so forth, 35 μsec interval is assumed between consecutive ONUs using the same wavelength. Guard time is removed for consecutive ONUs transmitting different wavelengths. Fig. 4 shows that the throughput is reduced for asymmetric distribution in fairness condition, which is attributed to the increased guard time. Delay for 12:4 ratio shows sharp increase at load 0.4 while this value is shifted to load 0.6 for 8:8 ratio as in Fig. 5.

Throughput for priority condition is calculated separately for the two wavelengths. P#1 in Fig. 6 indicates the wavelength which services fewer ONUs while P#2 for more ONUs. Throughput in this figure is the total amount for the ONUs in each group. Throughput is proportional to the number of ONUs for
lower load. As load is increased, however, P#2 saturates while P#1 continues to increase. The added throughput of P#1 and P#2 decreases as asymmetry grows in priority condition. It is occurred by the reduced window size for each ONU since ONUs with P#1 are assigned more than once in one cycle and this value increases with higher asymmetry. Better performance with P#1 is more apparent in delay as illustrated in Fig. 7.

**Conclusion:** Similar throughput and delay are found in fairness condition although the total performance is degraded for asymmetric wavelength distribution. This asymmetry can be used for SLA when operated by a priority algorithm.

4 References
Traffic Grooming & Wavelength Assignment System over WWW for Regional WDM Optical IP Networks

Akira Ueno, Satoshi Kawase, Masahiro Taniue, Osanori Koyama, and Yutaka Katsuyama
Graduate School of Engineering, Osaka Prefecture University
1-1 Gakuen-cho, Naka-ku, Sakai, Osaka, 599-8531 JAPAN
akira@eis.osakafu-u.ac.jp, katsu@eis.osakafu-u.ac.jp

Abstract—A traffic grooming & wavelength assignment system was investigated and implemented for use in optical IP networks. The system performance was examined, and confirmed to be satisfactory.

I. INTRODUCTION

Recently, optical networks have been introduced commonly to satisfy the rapid increase of traffic demands. The WDM (Wavelength-Division Multiplexing) technologies provide more bandwidth per fiber, and are used in core backbone networks, metropolitan area networks, and regional wide area networks. In this paper, a design system is described, developed for regional WDM optical IP (Internet Protocol) networks, used mainly for business use as intranet, service provider networks, local government networks, campus networks, and so on. The system could design the network successfully by outputting lightpath connections and the wavelength number required to groom given traffic.

II. TRAFFIC GROOMING & WAVELENGTH ASSIGNMENT SYSTEM

Figure 1 shows an example of a network considered in this paper. Nodes represent user buildings, and are connected with single-mode (SM) fibers. One-fiber connection between nodes is called a fiber path. The lightwave connection with one of multiplexed wavelengths is called a lightpath [1]. Each node consists of a wavelength-routing device (WRD), optical transceivers with different wavelengths, and a layer-3 switch (L3SW). In this example of bi-directional dual fiber ring network, some lightpaths connect 6 nodes so as to satisfy the traffic demands between nodes. The wavelength \( \lambda_5 \) is allocated to bi-directional lightpaths between Nodes 1 and 5. The lightwave with \( \lambda_5 \) propagates through the fiber paths from Nodes 1 to 5 via Node 6, and the lightpaths pass through the WRD of Node 6. The WRD has also optical switches (SWs) in the fiber connections. By the SWs, the add/drop and passing-through lightpaths can be changed. This switching can change source and destination nodes of the lightpath, and therefore, the lightpaths can be reconfigured [2]. The system should output such lightpath connections as well as the optical components for each WRD, needed to groom given traffic.

To implement the system functions, a web-based configuration was adopted, as shown in Fig.2. When the system is used, administrators get access to a central server by a web browser. In the central server, a web server and a central application (AP) together with database are installed. The design functions are executed in 2 process engines: design and calculation engines. The design engine has functions for lightpath design and wavelength assignment as well as the lightpath reconfiguration design. All the calculations are made in the calculation engine.

Figure 3 show the lightpath design flow implemented in the system. The traffic matrix \( T=(t_{sd}) \) is input in the flow, giving the traffic \( t_{sd} \) from a source node \( s \) to a destination node \( d \) (sd). Firstly, lightpaths LP\(_J\) (\( J = 1 \) to \( N \)) are given in such a way to connect adjacent nodes to form a lightpath (LP) ring. Here, \( N \) denotes the node number. This LP ring is given as the initial lightpaths to
connect all the nodes. The given traffic $\tau_{sd}$ can be groomed directly by a lightpath $LP_{sd}$ when the nodes are adjacent. When the $sd$ nodes are not adjacent, the shortest lightpath $LP_{sd}$ should be found to allocate the traffic $\tau_{sd}$ in the system. Dijkstra algorithm was implemented to calculate the shortest lightpath $LP_{sd}$. By this process, the shortest multi-hop lightpath is found. The traffic $\tau_{sd}$ is allocated to all the lightpaths $LP_j$ composing $LP_{sd}$. The processes of finding the shortest lightpath $LP_{sd}$ and $\tau_{sd}$ allocation are made for all combinations of the $sd$ pairs. After that, all the values of the allocated traffic to each $LP_{sd}$ are summed up, and the total traffic $\tau_{sd}(J)$ groomed by the lightpath $LP_{sd}$ is obtained. And the maximum value $\tau_{sd}(J)$ is found among all the values of $\tau_{sd}(J)$. If $\tau_{sd}(J) < C_s$, the grooming is completed, and the obtained lightpath set is the designed result, where $C_s$ is a transmission capacity by one wavelength.

If $\tau_{sd}(J) > C_s$, groomed traffic is overflown, requiring more bandwidth. Therefore, one lightpath $LP_{K1}$ is added ($K=N+1-K_1$) to one of the unconnected $sd$ pairs. And the grooming process is re-executed to obtain the total traffic $\tau_{K1}(K_1)$. And the maximum value $\tau_{K1}(K_1)$ is determined in the same way as the first calculation. The grooming process is executed for all possible addition $LP_{K1}$, $LP_{K1+1}$, $LP_{K1+2}$, … to the untried $sd$ pairs, and the maximum value $\tau_{K1}(K_1)$ is determined in each trial addition of a lightpath. Then, if $\tau_{K1}(K_1) < C_s$, is true, the lightpath design is completed. Otherwise, another lightpath is added ($K_1=K_2=-K_1+1$), and the same design processes continue. When $\tau_{K1}(K_1) < C_s$, the lightpath design is completed and stopped. This lightpath design also provides the lightpath sets multiplexed in each fiber path. Therefore, wavelength assignment processes can be made by giving different wavelength to each lightpath in the fiber path.

Figure 4 shows an example of a lightpath figure designed under a condition of $C_s=1000Mbps$, which was displayed on a web browser by the system. Traffic values were input to the system for the design: when $(s,d)= (1,2)$ and $(s,d)= (s>1,1)$, $\tau_{sd}=450Mbps$, and otherwise $\tau_{sd}=100Mbps$. In this example, the bi-directional lightpaths with solid lines ($LP_1-LP_8$) were obtained for the given traffic. After that, one reserve lightpath was added by the administrator, and redesigned. The reserve LP9 is shown with brackets in Fig.4. Figure 5 shows the clockwise (CW) and counter clockwise (CCW) lightpaths multiplexed in the fiber paths $FP_{pr}$ together with the assigned wavelengths. For example, CW lightpath 2 (LP2) is accommodated in fiber paths $FP_{12}$ and $FP_{21}$, and CCW LP2 is accommodated in fiber paths $FP_{23}$ and $FP_{32}$, which is consistent with the result that LP2 connects nodes 1 and 2, as shown in Fig.4. It is found that different wavelengths are allocated in each fiber path, satisfying the physical requirement to WDM. As a result, 3 wavelengths are necessary to groom the given traffic, including the one for the reserve lightpath LP9.

III. CONCLUSION

A traffic grooming and wavelength assignment system over WWW has been designed and implemented for regional optical IP networks. The performance was examined by designing a network connecting 6 nodes so as to groom given traffic. The result clarified that the system could design the network successfully by outputting lightpath connections and the wavelength number required to groom the given traffic.

References


Figure 3 Lightpath design flow implemented in the system

![Figure 3](image)

Figure 4 Example of a lightpath figure designed and displayed on a web browser by the system

![Figure 4](image)

Figure 5 Lightpaths multiplexed in fiber paths and assigned wavelengths

![Figure 5](image)
Design Tool for Wavelength Routing Functions

Genichi Mouri¹, Kohei Okada², and Kimio Oguchi¹²

Information Networking Lab., ¹Faculty of Engineering, ²Graduate School of Engineering, SEIKEI University
3-3-1 Kichijoji-Kitamachi, Musashino, Tokyo, 180-8633, JAPAN
Tel : +81-422-37-3732, Fax : +81-422-37-3871, E-mail : *mouri@nifty.com **oguchi@st.seikei.ac.jp

Abstract - A design tool for wavelength routing functions that implements the Wavelength Transfer Matrix method is proposed. We implement the design tool program and verify its functionality to confirm its practical effectiveness.

1. Introduction

Current End-to-End connections in photonic networks are routed electrically with OE-E0 conversion. However, as the transmission data continues to increase, we are approaching the performance limits of electronic routing. Accordingly, the wavelength routing network is being researched and developed. Photonic networks use routing devices such as AWG (Arrayed Waveguide Grating) to determine the route of each WDM (Wavelength Division Multiplexing) channel.

Designing the wavelength routing network is extremely difficult if the number of wavelengths is large and the network configuration is complicated. The Wavelength Transfer Matrix (WTM) method was proposed by the author to facilitate network design. This method treats a network as a number of matrices.

In this paper, a new algorithm and tool to design wavelength routing network that uses the WTM method are proposed. Its correctness is verified in a network with an AWG as prototypical example of a wavelength routing network.

2. Wavelength Transfer Matrix (WTM) method²

A general routing device with wavelength routing functions is depicted in Fig.1. This device consists of n-input ports, m -output ports and a wavelength routing function (as described Matrix L). This method leads to signals (Matrix O) that are output from each output port using the routing function when WDM signals (Matrix I) are input to each input port.

Signals from the output port are identified by the following notations.

\[
O = L \cdot I \quad \lambda_k \cdot \lambda_i = \lambda_j \\
\lambda_k \cdot \lambda_i = 0 \quad (k \neq l)
\]

where each element of Matrix L is expressed by operator \( \lambda \) that has the feature identified above.

3. WTM of AWG

An AWG can route and multiplex individual wavelengths, as well as demultiplexing a WDM signal. Figure 2 shows a typical example of the wavelength routing function offered by a 4x4 AWG; the four nodes are input to the AWG ports, and then routed by wavelength. This wavelength routing functionality is described below using the WTM method.

\[
L = \begin{bmatrix}
A_1 & A_2 & A_3 & A_4 \\
A_2 & A_3 & A_4 & A_1 \\
A_3 & A_4 & A_1 & A_2 \\
A_4 & A_1 & A_2 & A_3 
\end{bmatrix} \quad I = \begin{bmatrix}
A_1 + A_2 + A_3 + A_4 \\
B_1 + B_2 + B_3 + B_4 \\
C_1 + C_2 + C_3 + C_4 \\
D_1 + D_2 + D_3 + D_4 
\end{bmatrix}
\]

where \( X_k \) denotes the signal of \( \lambda_k \) from node X.

\[
O = L \cdot I 
\]

The WTM method allows the destination of each multiplexed signal from Node A-D to be clearly read.

Fig.1. Routing device with wavelength routing

Fig.2. An example of 4x4 AWG routing
4. Schema of design tool

Schematic diagram of the proposed design tool is depicted in Fig.3. User accesses a GUI (Graphical User Interface) for locating any node, routing device, or fiber. This tool includes the proposed wavelength routing algorithm and databases that have parameters for each node, routing device, and fiber. Details of the databases and a description of routing algorithm are given in the following sections.

5. Databases

To treat the parameters of each node, routing device, and fiber as program variables, we set three databases: NodeParamDB, DeviceParamDB, and FiberParamDB. NodeParamDB holds the coordinates of each node location, the number of I/O ports and multiplexed signals, wavelengths of signals, node type (send / receive / send & receive) and time-slot. DeviceParamDB holds the coordinates of each device, the number of I/O ports, and WTM for each device. FiberParamDB holds the parameters on both ends of fiber, i.e. terminated or connected to the next device.

6. Routing algorithm

Figure 4 shows the flow chart of the routing algorithm. In order to start the routing process, the algorithm loads the signal parameters from the NodeParamDB as the wavelength of the signal is required to analyze the routing functions. Next, a node is selected to send signals on the pre-assigned time slot. The destination of the signal is decided by the fiber parameters in the FiberParamDB. The parameters of the device that the signal is headed for are loaded from the DeviceParamDB or NodeParamDB. If the next device is a routing device, it is loaded from DeviceParamDB again. In other cases, it is loaded from NodeParamDB, then, the signal is received, which terminates Process. If the device is the routing device such as AWG, ADM (Add/Drop Multiplexer), the algorithm runs the routing process that computes the route by the WTM method. The algorithm repeats the same routine until the signal reaches the end node.

7. Simulation of a network with an AWG

The design tool program was implemented (Visual C++) to verify the effectiveness of the proposed wavelength routing algorithm. Routing of an 8 x 8 AWG was simulated as an example as shown in Fig.5. Each node sends signals each with different 8 wavelengths and the signals are routed by the AWG as shown by color lines in the AWG depending on each wavelength in the upper of Fig.5. The results of the simulation also indicate signal transfer to each receive node with characteristics like A1 (signal of wavelength \( \lambda_1 \) from node A) - H8 (signal of wavelength \( \lambda_8 \) from node H) as shown in the bottom of Fig.5.

Comparing these simulation results to the results obtained in the actual AWG and calculated by WTM method verifies that proposed algorithm and design tool work correctly. Other examples such as the ADM application and the AWG with loop-backed fibers application indicate correctness of the proposed method.

8. Conclusion

A design tool for wavelength routing functions that uses the WTM (Wavelength Transfer Matrix) method was developed. According to simulation results, the proposed tool was verified to work correctly.

References


Filterless Optical Networks: A Unique and Novel Passive WAN Network Solution

Christine Tremblay, François Gagnon, Benoît Châtelain
Éric Bernier, Michel P. Bélanger

1: Université du Québec – ÉTS, 1100 Notre-Dame W. St, Montreal, QC Canada H3C 1K3, Tel (514) 396-8532, Fax (514) 396-8684, christine.tremblay@ele.etsmtl.ca
2: Nortel, 3500 Carling Avenue, Nepean, ON Canada K2H 8E Tel (613) 763-9654, Fax (613) 768-1140, ebern@nortel.com

Abstract
The concept of a Filterless optical network is introduced and validated. This passive WAN network solution is based on advanced transmission technologies and passive optical interconnections between nodes. The study looks at the cost reduction impact of this unique WAN architecture through a case study.

Introduction
The use of advanced modulation formats combined with electrical compensation technologies can permit significant changes on the architecture and management at the physical layer level of future optical networks through: Dispersion Compensating Module (DCM) - less optical line systems for up to 80 000 ps/nm dispersion compensation; tunable transceivers, an enabling technology for optical networks of broadcast & select type; increased system margin, which translates into improved flexibility for optical network design. This paper introduces the novel concept of a “Filterless” network architecture and looks at its cost reduction impact.

Filterless optical networks
Recent developments of optical wavelength selective switch (WSS) and optical cross-connects (OXC) and advances in the optimization methods for complex optical multi-branch optical system, have rendered possible the realization of long and ultra-long haul transparent networks. National scale all-optical networks are being deployed currently. But, in some cases, this first wave of deployment is limited by transmission impairments of optical subsystems. In current, optically agile wide area network (WAN) architectures, the agility is delivered at the network nodes. This agility is engendering additional system and network capital costs. In several cases, the additional capital costs are compensated by the operational benefit of the delivered agility.

In this paper we propose to examine an attempt to minimize the additional capital cost driven by the agility while maintaining (improving) the operational advantages of an agile network. We propose a passive network topology that eliminates or minimizes the active photonic reconfigurable component count in the optical line system. This alternate approach reduces the installed first cost (IFC) of the network by transferring the cost to the transmit/receive function at the terminals. Taking advantage of recent transmission technology breakthroughs such as electronic dispersion compensation, advanced modulation formats and tunable transceivers, the proposal takes on the premises that the need for agility can now be provided by wavelength tuning at the transmitter and wavelength discrimination at the receiver much like agility is achieved in radio networks.

The Filterless concept is a unique and novel passive WAN network solution based on the use of passive splitters and combiners for interconnecting fiber links. The resulting network does not require optically reconfigurable components as well as any coloured components, hence the name Filterless. The resulting network architecture can be expected to bring significant advantages such as lower cost, easiness of maintenance and reconfigurability, as well as good resilience and excellent multicast capabilities. The study looks at the cost reduction impact of deploying an optical network in which the photonic switching elements have been replaced by passive optical splitters and combiners.

Filterless network problem
A Filterless network is based on the construction of a set of physical optical links between all nodes of the network. Since all nodes are optically connected to each other, real-time delay sensitive broadcast and multicast transmissions are facilitated. The obtained Filterless physical topology depends on the splitters and combiners configuration at each node. As in a classical Routing and Wavelength Assignment (RWA) problem, the Filterless problem can be partitioned in two parts: (1) establishing the fiber connections; (2) routing and assigning the wavelengths according to the traffic demand.

Virtual topology design problem
A Filterless optical network can be represented by a graph $G = (V, E)$ with:
- $V$: set of $N$ network nodes;
- $E$: set of directed arcs connecting the nodes;
- Weight of the arcs representing the signal attenuation.

Table 1 summarizes the input parameters, the variables, the constraints and the optimization objective.

A fiber connection algorithm was developed for interconnecting the nodes using optical splitters and combiners without creating closed loops in order to avoid laser effects. Splitters and combiners were also used to add and drop wavelengths at each node. An algorithm was developed for routing and assigning the wavelengths according to the traffic demand between node pairs.
Table 1

<table>
<thead>
<tr>
<th>Input parameters</th>
<th>Network physical topology</th>
</tr>
</thead>
<tbody>
<tr>
<td>Variables</td>
<td>A assignment to nodes for a traffic stream launched on a given color</td>
</tr>
<tr>
<td>Constraints</td>
<td>Signal SNR vs distance</td>
</tr>
<tr>
<td>Objective</td>
<td>Minimize the overall network cost</td>
</tr>
</tbody>
</table>

Case study
A 7-node subset of German optical network was used to validate the Filterless network concept and to perform a comparative cost analysis with respect to opaque and photonic switching network topologies. The traffic matrix, expressed in number of 10 Gb/s wavelengths per link, can be found in Table 2.

Table 2

<table>
<thead>
<tr>
<th>Node</th>
<th>A</th>
<th>B</th>
<th>C</th>
<th>D</th>
<th>E</th>
<th>F</th>
<th>G</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td></td>
<td>6</td>
<td>3</td>
<td>2</td>
<td>3</td>
<td>5</td>
<td>2</td>
</tr>
<tr>
<td>B</td>
<td>6</td>
<td>-</td>
<td>6</td>
<td>3</td>
<td>6</td>
<td>8</td>
<td>4</td>
</tr>
<tr>
<td>C</td>
<td>3</td>
<td>6</td>
<td>-</td>
<td>2</td>
<td>3</td>
<td>4</td>
<td>2</td>
</tr>
<tr>
<td>D</td>
<td>2</td>
<td>3</td>
<td>2</td>
<td>-</td>
<td>2</td>
<td>2</td>
<td>1</td>
</tr>
<tr>
<td>E</td>
<td>3</td>
<td>6</td>
<td>3</td>
<td>2</td>
<td>-</td>
<td>4</td>
<td>2</td>
</tr>
<tr>
<td>F</td>
<td>5</td>
<td>9</td>
<td>4</td>
<td>2</td>
<td>4</td>
<td>-</td>
<td>3</td>
</tr>
<tr>
<td>G</td>
<td>2</td>
<td>4</td>
<td>2</td>
<td>1</td>
<td>2</td>
<td>3</td>
<td>-</td>
</tr>
</tbody>
</table>

Filterless solution example
A Filterless solution example on the 7-node reference network is shown in Figure 1. By using splitters and combiners, physical optical links are created between nodes. Splitters and combiners are also used at each node to add and drop wavelengths.

Once the fiber connection matrix determined, wavelengths were assigned to meet the traffic requirements between all node pairs. For cost comparison purposes, an opaque solution based on OEO regeneration at every node and an active photonic solution based on the use of an optical reconfiguration device were also considered.

Performance evaluation and comparison
The network cost was evaluated by computing the cost of extra components added to the network. The results are shown in Table 2. Although an active photonic network solution is already a very good improvement over an opaque network solution, a Filterless network solution is even more interesting with a further cost reduction by a factor of 47.

Table 2

<table>
<thead>
<tr>
<th>Network solution</th>
<th>Extra Components</th>
<th>Quantity</th>
<th>Unit cost (a.u.)</th>
<th>Total cost (a.u.)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Opaque Transceivers</td>
<td>130</td>
<td>1</td>
<td>130</td>
<td>130</td>
</tr>
<tr>
<td>Active photonic Optical reconfiguration device</td>
<td>9</td>
<td>2</td>
<td>18</td>
<td></td>
</tr>
<tr>
<td>Optical amplifiers</td>
<td>18</td>
<td>1</td>
<td>18</td>
<td>18</td>
</tr>
<tr>
<td>Filterless Optical splitters</td>
<td>38</td>
<td>0.02</td>
<td>0.76</td>
<td></td>
</tr>
</tbody>
</table>

In a Filterless network solution, all the traffic demands can be met with 148 wavelengths, corresponding to 296 transceivers. The added cost of the opaque network solution comes from the 130 extra regenerating transceivers while the added cost in an active photonic network comes from the use of 9 optical reconfiguration devices and associated 18 optical amplifiers (one per direction) to compensate for their insertion loss.

Conclusions
This paper presented a novel concept of Filterless optical network. The concept has been validated through a case study. This case study is intended for illustrative purposes only and might not take into account all factors affecting the results like deployment scenarios, actual growth rate or competitive pressure. A Filterless network is found to be more cost effective and more reliable at the expense of greater wavelength utilization. Consequently, the bandwidth/fiber efficiency of a Filterless network is lower than that of opaque and active photonic network. With the increasing availability of wavelengths (1000 to 2000 wavelengths per fiber forecasted in 5 to 7 years) the reduced bandwidth efficiency issue is not seen as a major issue. The presented solution examples were obtained from a broadcast traffic matrix. Because every node of the Filterless network is multicast-capable, the bandwidth efficiency is much higher in the case of multicast transmission.

References
We propose and demonstrate a novel wavelength monitoring method of NRZ signals using NRZ-to-PRZ conversion by $\pi$-Phase Shifted FBG. The dynamic range and the measurement range of the proposed method were about 23 dB and 0.2 nm.

1 Introduction

By rapid increase in the demand of the internet traffic, dense wavelength division multiplexing (DWDM) is highlighted because of its large channel capacity. From ITU-T recommendation, however, the allowable wavelength drift of the signal for DWDM with 50-GHz channel spacing should be less than 20 pm. Thus, the wavelength monitoring of the signal should be required to eliminate the power penalty from the wavelength drift. Several wavelength monitoring methods were reported [1,2]. However, the previous methods used the active device [1] or the pilot tone [2] to monitor the wavelength of the signal.

Pseudo-return-to-zero (PRZ) signal has the large clock component due to its RZ-like pattern. NRZ-to-PRZ conversion based on the optical delay interferometer (ODI) has attracted due to its easy implementation. Recently, we reported the NRZ-to-PRZ conversion method using $\pi$ phase shifted fiber Bragg grating ($\pi$-PSFBG), which acted as an ODI, for the clock extraction [4]. $\pi$-PSFBG, in which a single lightwave path is shared to generate an interferometric delay, has been recognized as an effective configuration to mitigate the environmental perturbations.

In this paper, we propose a novel wavelength monitoring method for NRZ signal using NRZ-to-PRZ conversion by $\pi$-PSFBG. The proposed method can monitor the wavelength of the NRZ signal without any system modification. The feasibility of the proposed monitoring method for the 10-Gb/s NRZ transmission is experimentally demonstrated.

2 Operation Principle

The $\pi$-PSFBG consists of two sub FBGs and the fixed phase shifted region. The two sub FBGs are serially placed with a gap. The NRZ signal is sent to the $\pi$-PSFBG through an optical circulator. Then a part of the incoming NRZ signal is reflected by the first FBG (FBG1) and another part of the NRZ signal is reflected by the second FBG (FBG2) with a certain time delay, which is determined by the length of the sub FBG and the gap [3]. The signal reflected by FBG2 has $\pi$ phase difference by using the fixed phase shifted region. These NRZ signals interfere with each other when receiving with a photo detector. If the time delay is less than the bit duration, the NRZ signals are converted into the PRZ signals, that is, only both the leading and trailing edges of the NRZ signals are clearly seen due to the out-of-phase between two NRZ signals reflected by FBG1 and FBG2. These PRZ signals have a large clock component.

If the phase difference is not matched to $\pi$, the signals between the edges remain. It means that the imperfect out-of-phase makes the worse PRZ signals, introducing the low clock component. The imperfection of the out-of-phase comes from the ODI perturbations and the wavelength drift of the signal. If the perturbations are well prevented, we can monitor the wavelength drift of the signal by measuring the clock component power of the PRZ signals.
3 Experiment and results

The experimental setup is shown in Fig. 1. The output of the tunable laser source was modulated with a 10-Gb/s pseudo random binary sequence of length of $2^{23}-1$ using a LiNbO$_3$ modulator. The modulated signal was sent to the π-PSFBG with the time delay of 40-ps via an optical circulator. The reflected signals from the π-PSFBG, which are the PRZ signals, were taken from the other port of the optical circulator. Then, the PRZ signals were detected by a photodetector and the detected signals were amplified by a low noise preamplifier. Finally the clock component power of the PRZ signals was measured by an electrical spectrum analyzer.

Optical waveforms of the NRZ and PRZ signals for the ‘101111100001100’ patterns were measured by the sampling scope with 20GHz bandwidth, as shown in Fig. 2 (a) and (b). The PRZ signals at both the leading and trailing edges of the NRZ pulses are clearly seen. Fig. 2 (c) and (d) show the electrical spectra of the NRZ and PRZ signals. As shown in Fig. 2 (c) and (d), the clock component was drastically increased.

Fig. 3 shows the relative clock component power according to the wavelength drift of the signal. The clock component power was the maximum for the zero wavelength drift, i.e. the perfect out-of-phase. As changing the wavelength, the phase difference between two NRZ signals reflected by the FBGs was not matched to π, introducing the lower clock component power. The clock component power variation has the periodic of 0.2 nm, which is same with the free spectral range (FSR) of the π-PSFBG. That is, the measurement range was 0.2 nm, which is determined by the FSR of π-PSFBG. The dynamic range of the proposed method was 23 dB.

4 Conclusion

We have proposed and experimentally demonstrated a novel wavelength monitoring method of NRZ signals using NRZ-to-PRZ conversion by π-PSFBG. As changing the wavelength of the signal, the phase difference was not matched to π, introducing bad PRZ signals with low clock component power. By measuring the clock component power of the PRZ signal, we can monitor the amount of the wavelength drift. The dynamic range and the measurement range of the proposed monitoring method were about 23 dB and 0.2 nm, respectively.

Fig. 1. Experiment setup. TLS: tunable laser source, PC: polarization controller, MZM: LiNbO$_3$ modulator, EDFA: erbium-doped fiber amplifier, PPG: pulse pattern generator, ESA: electrical spectrum analyzer, O/E: optical receiver, AMP: low noise electrical amplifier.

Fig. 2. (a) waveform of NRZ signals, (b) waveform of PRZ signals, (c) electrical spectrum of NRZ signals, (c) electrical spectrum of PRZ signals.

Fig. 3. Clock component power of PRZ signals according to the wavelength drift.

Acknowledgement

This work was partially supported by KOSEF through grant No. R01-2006-000-11088-0 from the Basic Program and BK-21.

Reference

A spatial BER contour for indoor visible light LAN and mobile networking

Swook Hann, Hyung-Sik Jang, Yun-Jung Choi, Jung-Hun Kim and Dong-Hwan Kim
Strategic Technology Group & Photonics Systems Lab., Korea Photonics Technology Institute
971-351, Wolchul-dong, Buk-gu, Gwangju 500-430, Republic of Korea
Tel: +82-62-605-9381, Fax: +82-62-605-9399, E-mail: swookhann@yahoo.com

Abstract
We propose and demonstrate a spatial bit error rate measurement for visible light LAN and mobile networking. The proposed method uses commercial LED array and a Si-pin diode, whose central wavelength is in the visible range. We show a 3-D measurement result of the proposed system using a spatial BER contour.

1 Introduction
A visible light communication is considered as a promising solution for the indoor eco-friendly wireless networks. Also the visible lighting of the low-cost LEDs will be spread with the LED promising markets. Furthermore, the LED lighting is hopeful with the energy efficiency and long-lived working time [1].

To apply the LED into the communication area, there are several merits as like: security against eavesdropping; high available bandwidth; no electro-magnetic interference; unregulated by any law, etc.

Up to now, a number of works have done with optical wireless communication systems using the white LED, IrDA and RC-LED [1-3]. However, those works cannot demonstrate a spatial bit-error rate (BER) contour measurement, always excepting the numerical analyses.

The BER contour is useful measurement to apply the indoor wireless LAN and mobile networking with the distributed LED ceilings. In order to take concrete the visible light communication, study of shadowing area should be considered under the wireless data transmission with the spatial BER measurement. To demonstrate the inter-symbol interference (ISI) due to an optical path difference in the indoor spatial environment is also useful step to realize the visible light communication systems. The LED array can be used not only for optical wireless LAN access points but also for indoor ceiling lights to illuminate room so that a shadow and a dark area may not be produced in view of vision and communication. To the best of the author’s knowledge, this experimental demonstration is first time.

2 Principle of Operation
The proposed BER contour demonstration is shown in Fig. 1. The LED ceiling spots are hanging on the Z-axis stage of height 1 m. The Si-pin detection circuit is mounted on the X-Y stages. The visible detector can scan the square the 1x1 m². To get the BER measurement, the measurement boards are designed by 10-Mbps by help of pulse pattern generator and error analyzer. The BER contour can be gotten at each spatial point of X-Y-Z with instant BER value. The BER values contain the ISI and the power distribution of the LED array over the 3-D space, a cubic of 1 m³.
3 Experimental Results and Discussion

The complete setup for the spatial BER measurement of visible light communication is shown in Fig.2. The LED array set, 2x7, and Si-pin diode circuits are prepared to get the spatial BER values. The line of sight (LOS) BER measurement of those circuits is shown in Fig.2(a). To demonstrate the visible light communication system, the bit-rate is conducted at 1Mbps – 10Mbps as shown in Fig.2(b). The visible LED light transfers the spatial data of PRBS 2^7-1 to the cubic indoor space. The visible detection aperture has a hemispherical lens with diameter of 14 mm and with Si-pin effective area of 150 mm^2. The directivity of the PD is 35° at half angle.

Figure 3 shows the BER spatial contour of the visible light communication systems. The spatial bit-rate of 10Mbps is also showed in Fig.3(a). The contour map depicts the shadow region of the indoor wires optical LAN. The oval region of the dark colors, 160x100 mm^2, is considered as the communication area at one site of LED array in Fig.3(b). However, the angle of the LED array, 0.055 sr, in our experimental setup, is marked of effective communication area with less than BER of 10^{-7}, therefore, the access points or the distributed LED array will be prepared for more wide wireless service in Fig.3(b). In the view of illumination engineer, the shadow region can be prevented with the large area LED array or a number of LED spots on the ceiling. To get a wide non-shadowing area under the fixed data-rate, the high luminous intensity will be installed on the ceiling. Also, to prevent the ISI from the diffused reflection of circumstance bounds, the noise filtering of receiver circuits or Non-OOK coding schemes are considered.

4 Conclusions

We proposed and experimentally demonstrated the feasibility of the spatial BER contour measurement of visible light communications for indoor wireless LAN and mobile networking. The BER spatial contour depicts the shadowing and the ISI effect of the LED illumination under the indoor wireless communication. In this works, the BER spatial contour is achieved at a cubic space.

Acknowledgements

The work was partially supported by a Grant No. 10023250, MOCIE of Korea.

References

Simulation Results on Transmission of 10Gb/s Optical MSK System with Narrowband Frequency Discrimination Receiver over 560 km Standard Single Mode Fiber without Dispersion Compensation

Thanh Liem Huynh, Thirukkumaran Sivahumaran, Le Nguyen Binh and Khee Khok Pang
Email: thanh.huynh@eng.monash.edu.au

Abstract — We achieve by simulation the results on transmission of 10Gb/s optical MSK system with non-coherent narrowband frequency discrimination receiver over a distance of up to 560 km SSMF without dispersion compensation.

I. INTRODUCTION

Advanced modulation formats with narrow spectrum and high dispersion tolerance limits for long-haul and metropolitan optical systems have recently caught extensive research effort. As one of the potential formats, minimum shift keying (MSK) has recently been studied and adapted for optical communications [1-4]. The reported receiver schemes in [2-4] utilizes Mach-Zehnder Delay Interferometer (MZDI) for non-coherent detection based on signal phase information which is severely distorted by the dispersive optical channel.

In this paper, we first propose a simple receiver configuration for non-coherent detection of optical MSK signals. The receiver is based on narrowband frequency discrimination and employs dual narrowband optical filters and an optical delay line (ODL). With use of the proposed receiver, we report the achievement of transmission of 10Gb/s optical MSK signals over 560 km stand single mode fiber (SSMF) without use of any dispersion compensation means. To the authors’ best knowledge, this reach has not been yet achieved without aid of pre or post optical/electronic equalizers.

The operational principles of the proposed receiver are described in section 2. The numerical results on transmission performance for 10Gb/s optical MSK system over various optical link distances up to 560 km or equivalently a residual dispersion tolerance of up to ±9520 ps/nm are reported in section 3.

II. OPTICAL MSK NARROWBAND FREQUENCY DISCRIMINATION RECEIVER

Fig.1 shows the block diagram of the proposed narrowband frequency discrimination receiver. Operational principles of the proposed receiver is based on frequency discrimination. The output of the optical channel is split into two branches and passed through two optical filters F1 and F2, one on each branch. The center frequencies and the bandwidths of the two filters F1 and F2 are selected so that F1 would capture most of the MSK signal only when a +1 is transmitted and F2 would capture most of the MSK signal only when a -1 is transmitted.

The optical MSK lightwaves are amplified with a low noise optical amplifier before being split into two optical paths. The USB and the LSB frequencies that correspond to logics ‘1’ and ‘0’ are discriminated by two narrowband optical discrimination filters F1 and F2 with their center frequencies located at f1 and f2. An ODL which is easily implemented in integrated optics is introduced to compensate the differential group delay given by 2πf0βL where f0 = |f1 - f2| = R/2; βL represents the group velocity delay parameter of the fiber and L is the fiber length. If the differential group delay is fully compensated, the optical lightwaves in two paths are expected to arrive at the photodiodes simultaneously. As shown in Fig. 1, the photodiodes are arranged in a balanced configuration. It is noted that the receiver sensitivity of the scheme can be controlled by adjusting the gain of the optical amplifier. By using dual narrowband optical filters, ASE noise induced by the optical amplifier is mostly filtered. Balanced eye-diagram is obtained at output of the electrical amplifier.

III. TRANSMISSION PERFORMANCE OF 10GB/S OPTICAL MSK SYSTEMS WITH NARROWBAND FILTER RECEIVER

Fig. 2 shows the simulation test-bed for the 10Gb/s optical MSK transmission system. The transmission system consists of a number of 80 km SSMF spans for varying the transmission distance from 0 km to 560 km SSMF. Therefore, various accumulative fiber chromatic dispersion values are created and reach up to 9520 ps/nm.
The input power into each span ($P_x$) was kept lower than the non-linear threshold power. The EDFA2 provides an optical gain to completely compensate for the attenuation of the signals after 80 km SSMF optical span. As shown in Fig. 2, the optical received power ($P_{Rx}$) is measured at the input of the proposed MSK receiver. In order to obtain the BER curve at a particular length of fiber, $P_{Rx}$ is varied by adjusting the attenuator. The key parameters used in the simulation are given in Table 1.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input power ($P_0$)</td>
<td>-3 dBm</td>
</tr>
<tr>
<td>Operating wavelength ($\lambda$)</td>
<td>1550 nm</td>
</tr>
<tr>
<td>Bit Rate ($R$)</td>
<td>10 Gb/s</td>
</tr>
<tr>
<td>SSMF fiber ($\beta_2$)</td>
<td>2.68e-26 or $</td>
</tr>
<tr>
<td>Attenuation ($\alpha$)</td>
<td>0.2 dB/km</td>
</tr>
<tr>
<td>Narrowband Gaussian filter</td>
<td>BT = 0.13</td>
</tr>
<tr>
<td>Constant Delay ($\tau$)</td>
<td>$</td>
</tr>
<tr>
<td>EDFA2: G</td>
<td>16 dB</td>
</tr>
<tr>
<td>Noise Figure</td>
<td>5 dB</td>
</tr>
<tr>
<td>$N_0$</td>
<td>$20 \text{ pA/}(\text{Hz})^{1/2}$</td>
</tr>
<tr>
<td>$d_{eq}$</td>
<td>20 pA/ (Hz)</td>
</tr>
</tbody>
</table>

The numerical results are obtained via Monte Carlo simulation. As shown in Fig. 2, the 1dB and 2 dB power penalties of the 10Gb/s optical MSK systems are obtained when the transmission length is 320 km and approximately 390 km SSMF respectively, which corresponds to residual dispersions of 5440 ps/nm and 6630 ps/nm respectively. Although suffering about 15dB penalty in receiver sensitivity for case of transmission over 560 km SSMF without dispersion compensation, the performance of optical MSK narrowband frequency discrimination receiver for 10Gb/s optical MSK system is still able to reach 1e-9 at the received optical power of -7dBm. Equivalently, a residual dispersion tolerance of up to $\pm 9520$ ps/nm is achieved.

### IV. Conclusion

In this paper, we have proposed a novel but simple receiver configuration for non-coherent detection of 10Gb/s optical MSK signals using narrowband optical frequency discrimination. The availability of narrowband filters [5] facilitates the practical implementation of our proposed detection scheme. The operational principles of the proposed receiver was explained in detail. We have achieved the non-coherent detection of 10Gb/s optical MSK signals over up to 560 km SSMF optical link for without dispersion compensation. The analysis and optimization process for the proposed receiver will be reported in another paper.

### REFERENCES


Performance Analysis of New Switch Architecture for Variable Packet Size

Mohd Shahril Salleh 1,2, Kaharudin Dimyati2
1. Photonic Network Unit, TM R&D, UPM-MTDC, 43400 Serdang, Selangor, Malaysia
2. Electrical Engineering Dept, University Malaya, 50603 Kuala Lumpur, Malaysia.

Abstract- This paper present a new contention resolution (switch) architecture, named as Packet Slicing Architecture (PSA) that not only has the capability to maximized packet throughput but also able to effectively manage variable packet size. Simulation shows that this architecture gives better performance than the conventional architecture.

I. INTRODUCTION

Contention resolution is still an unsolved problem in an optical packet switched network (OPS) [1], since there is no Optical Random Access Memory (O-RAM) successfully being invented thus far. Three different contention resolution strategies are commonly proposed in literature: Fiber Delay Line (FDL), deflection routing and wavelength conversion [2]. In an attempt to combat this problem, we proposed new switch architecture, Packet Slicing Architecture (PSA), to better manage packet contention within an OPS environment.

The architecture (Fig. 1) uses FDL to temporarily store the packet and subsequently resolve the contention. Based on [1], we categorized each incoming packets into 3 different groups (A, B and C). Group A for packet size below than 100 bytes, group B for packet size from 100 bytes up to 600 and group C for packet size more than 600 bytes.

PSA architecture consists of passive and active components and is controlled by an electronic control unit (ECU). The components are arrayed waveguide grating (AWG), Fiber Delay line (FDL) and semiconductor optical amplifier (SOA). The architecture is divided into three major interfaces:

- Input interface, consist of DEMUX that will demultiplex each incoming packet according to the incoming wavelength. SOA is used here to convert the wavelength for each packet by classes determined in the control unit. SOA 1 will handle packet from category A, SOA 2 for category B and SOA 3 for category C. After the packets has been classified, each packet will be routed into 3 different path using NxN AWGS.
- Switch Fabric; besides switching the packet on to the desired output port, it will also perform contention resolution. PSA consists of 3 NxN AWGR. AWGR1 will handle packet category A, AWGR2 will handle packet category B and AWGR3 packet category C. Each AWGR is equipped with different length of FDL for contention resolution.
- Output interface consists SOA and multiplexer. Using SOA 4, 5 and 6 in this part, we can convert packet wavelength to avoid packet collision and also amplify or regenerates the data (2R).

As has been exhibited by many research, traditional analyses of switch performance using Poisson or Gaussian arrival statistics is not anymore suitable for packet network. Instead, many proofs show that the nature of packet network traffic is indeed self similar in nature. Thus, in this paper, we will analyze the performance of our switch architecture under the influence of this traffic model. First, we will analyze how the architecture affects packet loss, and then a comparison between this architecture and conventional architecture is made.

2 TRAFFIC MODEL

Self similar means that the traffic had similar statistical properties at a range of milliseconds, seconds, hours or even weeks [3].

Self similar traffic is obtained by aggregating multiple sub-streams, each sub stream consist of alternating Pareto – distributed on/off periods. Pareto distribution is a heavy tailed distribution with the probability density function.

\[ f(x) = ab^x / x^{a+1}, x \geq b. \]
Using equation above, where $a$ is a shape parameter, and $b$ is a location parameter. Pareto distribution with $1 < a$ has a finite mean and $a < 2$ has an infinite variance. To generate Pareto-distributed values, the following formula $(X_{\text{Pareto}} = b/U^{1/a})$ may be used, where $U$ is a uniform random variable from $(0 \leq U \leq 1)$. Each sub-stream generated depend on traffic load and have the load generated by one sub-stream is measured as:

$$\lambda = E[on]/(E[on]+E[off]),$$

where $E[on]$ and $E[off]$ are expected lengths of on and off periods respectively. All sub-streams generate equal loads. The expected size of on period $(E[on])$ is specified externally. The expected size of off period $(E[off])$ is calculated based on the target load

$$\lambda : E[off] = E[on]*((1/\lambda - 1).$$

The total load generated by the traffic generator is equal the sum of loads generated by all sub-streams: $\lambda = \sum \lambda_i$.

To reduce packet loss, packets facing contention need to be buffered (Fig. 1). As a result, we can see in Fig. 4 that the inclusion of buffer reduces the packet loss. From this figure, 3 buffers is sufficient to reduce packet loss to below than 10% at any traffic load and by using up to 6 number of buffers we can nearly eliminate the packet loss problem in the system.

![Fig. 2 Self similar traffic packets generated at 1700 t](image)

Fig. 2 shows the traffic generated, with different packet size from 0 to 1500 bytes, using the self similar equations. Each packet has different wavelength with different number of destination address (output port).

### 3. PERFORMANCE ANALYSIS

We perform our analysis through simulation. Based on different offered load 0.1 to 0.9, we generate traffic stream with self similar (Hurst parameter equal to 0.6) behavior at each input port of PSA. Each port consists of 3 different packet streams which correspond to 3 different wavelengths.

![Figure 3: Probability of packet contention at different traffic load](image)

We start by looking at packet loss. Figure 3 shows the probability of packet loss in PSA at different offered traffic load. As anticipated, higher load will result in higher packet loss.

Having the proposed architecture tested over packet loss, next we will evaluate this architecture against conventional architecture. Conventional architecture will typically used fixed length FDL to cater for different size of packet [4]. It can be seen from Fig. 5 that the delay time experienced by congested packet is reduced more than half by using TSS architecture.

### 4. Conclusion

In this paper we described new switch architecture, the PSA. From the result we proved that this architecture capable of handle variable packet length (A, B and C) and more efficient as compared to conventional technique.

References:


Abstract—We present a signal quality monitoring method that uses a low bit-rate reference signal to monitor a 40 Gb/s data signal. The concept is demonstrated both through numerical simulations and experimental work.

I. INTRODUCTION

The development in optical transmission is moving towards higher capacity in terms of higher bit rates and denser wavelength division multiplexed (WDM) systems. At the same time, transmission distances are increased due to both optical amplification but also optically switched networks. Still, operators need to assure sufficient signal quality in the networks; a task that is becoming more and more complex with the advances mentioned above. Optical-signal-to-noise (OSNR) monitoring is a vital part of assuring signal quality [1-2] as most errors are caused by a too low OSNR, which is either due to very long transmission distances, or excess loss sources [3]. In noise limited systems, the noise will have a well defined effect on the BER so accurate signal quality estimations can be made using OSNR monitoring.

In terms of measuring the OSNR, coarse WDM systems offer the simple method of linear interpolation of the inter channel noise level, however dense WDM systems have overlapping channel side bands thus hiding the true noise level [4]. Therefore, monitoring the OSNR and thus signal quality of individual channels through traditional methods is not an easy task. Furthermore, with the introduction of optical switching various channels in a fiber can originate from different traveled paths in a network and thus have accumulated different amounts of noise.

This means that monitoring of each individual channel is required but not necessarily possible in a dense WDM system.

This paper presents a method for monitoring the quality of a high-speed data signal by correlating the data signal with a known low bit-rate reference signal. By polarization multiplexing the low bit-rate reference signal with the high-speed data signal we are able to monitor the signal quality of the otherwise unknown high-speed data as well as achieving early-warning information on any signal quality changes. By using polarization multiplexing we also assure that the reference signal and data signal will experience the same signal degrading effects. The work is supported by both numerical investigations and an experimental demonstration.

II. PRINCIPLE OF OPERATION AND EXPERIMENTAL SETUP

Both high and low bit-rate signals will experience a degradation of signal quality when noise is added to the signals, and degrade the receiver sensitivity. In order to correlate a low bit-rate reference signal with a high bit-rate data signal we use polarization multiplexing such that each signal will travel the same optically switched path and accumulate the same amount of noise. The principle is illustrated in Figure 1.

In this study we used a 2.5 Gb/s non-return-to-zero (NRZ) signal as the reference signal, and a 40 Gb/s NRZ data as the data signal. We used 2.5 Gb/s for practical reasons, a lower bit rate signal could also have be used. As the 40 Gb/s signal bandwidth and thus receiver bandwidth is 16 times (12 dB) larger than it is for the 2.5 Gb/s signal, there is be a 12 dB difference in receiver sensitivity for the two signals assuming equal OSNR for both signals. If a relative power difference is introduced between the signals the difference will no longer be 12 dB. Figure 2 illustrates what happens when the 2.5 Gb/s reference signal power is reduced by 10 dB.

![Figure 1: Figure showing the principle of using a polarization multiplexed reference signal (TxREF) for signal quality monitoring.](image1)

![Figure 2: Results from numerical simulation illustrating the effect of reducing the 2.5 Gb/s reference signal power by 10 dB.](image2)
Notice that the receiver sensitivity changes for both the reference and data signal are showed relative to the data signal OSNR. By assuring a power difference of 10 dB the 2.5 Gb/s reference signal can now be used to provide an early-warning signal for the 40 Gb/s data signal. An OSNR change from 35 dB to 30 dB would hardly be detectable in the 40 Gb/s signal quality but the 2.5 Gb/s reference signal will produce a stronger early-warning signal so proper action can be taken in due time. The concept was also demonstrated in an experimental setup as illustrated in Figure 3. A 1550 nm laser source is split 50/50 and two Mach-Zehnder (MZ) modulators are used to generate the 2.5 Gb/s NRZ reference signal and the 40 Gb/s NRZ data signal. Pseudo-random bit sequences (PRBS) of $2^7$-1 and $2^{31}$-1 bits were used for the 2.5 Gb/s and 40 Gb/s signal, respectively. Then, the two signals are combined in a polarization beam splitter (PBS), amplified and sent to a fiber span consisting of 44 km standard single mode fiber (SMF) and 8 km dispersion compensating fibre (DCF). The dispersion was 100% compensated at the signal wavelength. The OSNR is adjusted by using an ASE source, as shown in Figure 3.

A second PBS is used to separate the data and reference signal, which are evaluated using two bit-error-rate-test-sets (BERT). To ensure optimum input state of polarization into the second PBS, a polarization analyzer was used.

III. EXPERIMENTAL RESULTS

For simplicity, we carried out the experimental proof-of-concept demonstration on a single-channel system as illustrated in the setup shown in Figure 3.

The relative power difference between the 2.5 Gb/s reference signal and the 40 Gb/s data signal was adjusted such that the reference signal would trace the signal quality of the 40 Gb/s data signal during various OSNR changes. The signal quality was in these experiments quantified in terms of receiver sensitivity, i.e. optical power to the receiver at bit error rate (BER) $1.0 \times 10^{-9}$.

Figure 4 shows the results of the experiments. Clearly, the 2.5 Gb/s reference signal quality correlates with the 40 Gb/s data signal quality. In contrast to the early-warning concept, the reference signal in this case will provide on-the-fly information about the otherwise unknown 40 Gb/s signal quality. Figure 5 shows an experimental demonstration of the early-warning configuration. In this case the power of the 2.5 Gb/s reference channel has been decreased by further 10 dB compared to the case in

REFERENCES

Investigation of Performance Degradations of DPSK RZ/NRZ Signals Due to Non-Optimal Driving Voltage and MZDI Delay

Muhammad Haris1, Jianjun Yu1,2, and Gee-Kung Chang1

1) School of Electrical and Computer Engineering, Georgia Institute of Technology, Atlanta, GA 30318, USA
2) NEC Labs America, Princeton, NJ, 08540

Phone: 404-894-1917, Email: haris@gatech.edu

Abstract: We investigate the impact of imperfect driving voltage at the transmitter and bit-delay mismatch for the Mach-Zehnder interferometer at the receiver on 8 x 40Gb/s DPSK RZ/NRZ transmission systems with different duty cycles.

Introduction:
Recently, advanced modulation formats have been investigated for high bit-rate systems to extend the transmission distance and improve spectral efficiency. Differential-phase-shift keying (DPSK) has drawn a great deal of attention in high-speed long-haul transmission systems because of its 3-dB higher receiver sensitivity and strong fiber nonlinearity tolerance [1, 2]. In the modulation scheme of DPSK, the “0”s and “1”s are represented by the phase (0 or π). In the receiver, a delay interferometer (DI), whose differential delay is equal to one bit period, converts phase modulation to intensity modulation. The one bit period delay is used to guarantee the maximal interference between neighboring bits. But in the real systems, the difference between the phases of “1” or “0”s may not be a perfect “π” at the transmitter, also the differential delay is not exact one bit period. This study is aimed at the investigating the RZ/NRZ-DPSK transmission performance when the modulator at the transmitter is over or under driven and the delay of Mach-Zehnder Delay Interferometer (MZDI) at the receiver is not optimal.

Simulation setup and results
The investigated system setup is shown in Figure 1. The transmitter consists of one laser source followed by a dual-drive external LiNbO3 modulator. The modulator performs phase modulation and is driven at twice the switching voltage by a NRZ data stream in a push-pull configuration and biased at the transmission null point [3]. Eight channels are considered with data rate of 40Gb/s per channel and 50GHz channel spacing. The multiplexer and demultiplexer both consist of third order band-pass Bessel filter with 3-dB bandwidth of 50GHz. Each span has 100km of SSMF which is compensated by dispersion compensation fiber (DCF). The optical power launched into SMF is kept at 0dBm per channel and noise figure of EDFA is 4dB. Loss of SMF and DCF is 0.25 and 0.45dB/km.

Fig. 1. (a) RZ/NRZ- DPSK transmitter setup (b) System setup

Fig. 2. Q value versus modulation factor for 8 x 40Gb/s RZ-DPSK signals with 100-500km SMF-28 length a) 33% RZ-DPSK (b) 50% RZ-DPSK (c) 66% RZ-DPSK

For the generation of RZ-DPSK signals, additional MZ modulator driven by a sinusoidal wave is used as a pulse carver. Depending on the modulation frequency, bias position, and driving voltage, RZ-DPSK signals with various duty ratios are generated, including 33% RZ-DPSK, 50% RZ-DPSK, and 66% RZ-DPSK [4].
Fig. 2 shows RZ-DPSK signal performance with varying modulation factor for all three duty cycle ratios when the signals are transmitted over 100 to 500 km of SMF-28. It is clear that a small variation of switching voltage results in performance degradation, but all three duty cycle ratio signals undergo similar distortion once the modulation factor is varied from 0.5 to 1.5. BER is 10⁻⁹ when Q value is 6.

A DPSK receiver consists of an optical delay interferometer (DI) and a balanced receiver. The DI is a Mach-Zehnder filter with a differential one-bit delay, and converts the incoming phase-shift-keying signals into two complementary amplitude-shift-keying signals [4]. In reality, due to temperature variations or ageing, MZDI delay might not be optimal before reconstruction and may cause performance degradation. Fig. 3 presents the performance penalties caused by imperfect MZDI delay at the receiver. Results are plotted for all three duty cycle ratios when the signals are transmitted over one to five spans of SMF-28. 33% RZ-DPSK shows the best overall results under these conditions. It is obvious that greater the duty ratio results in the worst of the degradation. 66% RZ-DPSK is affected the most and BER of 10⁻⁹ can not be achieved even for shorter distances once MZDI is set for delays of 0.3 to 0.6 bit.

Similar performance comparison is also drawn for 8 x 40Gb/s NRZ-DPSK signals. Results are plotted in Fig. 4. NRZ signal seems more robust in case of inaccurate MZDI delay as compared to RZ pulse shaping for shorter fiber lengths as it can be seen in Fig. 4(b), but for long SMF over 500km its performance is similar to 50% RZ-DPSK.

Conclusions:
Numerical simulation for 8 x 40Gb/s DPSK RZ/NRZ signals with 0.8 b/s/Hz spectral efficiency has been demonstrated and their tolerance towards impairments caused due to improper driving voltage and MZDI delay was investigated for different duty cycle ratios. Results show that DPSK-RZ and DPSK-NRZ show similar performance limitations overall as far as driving voltage is concerned, but NRZ pulse signals are more robust to degradations caused by small MZDI delay values as compared to RZ pulse signals.

References
Abstract: We demonstrated 6 channel all-optical multicast using Raman-assisted FWM in dispersion-shifted fiber with wavelength control capability to select the multicast channels at 10 Gbit/s. All multicast signal channels comply with the 100 GHz spaced ITU grid.

1. Introduction

All-optical multicast is important for tomorrow’s optical IP networks where applications such as Video Conferencing (VC) and High Definition (HD) Internet TV can be realized without the need for O-E-O conversion at the lower layers. Optical multicast on data plane was proposed and demonstrated using various techniques [1-5]. These include the use of splitter-and-delivery (SAD) switch consists of optical power splitters and amplifiers [1], non-linear SOA-based interferometer [2], injection locking of FP laser [3], Cross Phase Modulation in a Dispersion-shift Fiber (DSF) [4] and four wave mixing (FWM) in DSF [5]. However all previous proposed schemes can only be used for fixed multicast in the data plane without the possibility of changing multicast group member with control information from control plane.

In an IP network such as Internet with multicast switch, group membership protocol information and multicast routing protocol are used to support IP multicasting across the IP network. This allows the switch to send data only to the hosts that belong to the group. Data plane lightpath connections need to be reconfigured based on the control plane information to meet this requirement.

In this paper, we propose the use of Raman-assisted FWM for its high conversion efficiency and wide bandwidth operation to demonstrate all-optical multicast (6 channels) with the capability to select the multicast channels based on wavelength controls to enable the implementation of the group membership protocol based all optical multicast.

2. Experiment and Results

Figure 1 shows the experiment setup for realizing selectable multicast using Raman-assisted FWM. The 10 Gb/s non-return-to-zero (NRZ) data signal at 1552.2 nm was generated by external modulation of a tunable laser with the power (measured at the port 1 of the circulator 1) of 5.13 dBm. The wavelength (and the power measured at the port 1 of the circulator 1) of the CW pumps generated by DFB lasers are 1549.8 nm (C1, 4.04 dBm), 1553 nm (C2, 2.58 dBm) and 1555.4 nm (C3, 2.85 dBm) respectively.

The CW pumps are used as control signals to select the multicast channels. Table 1 shows how the output channels are determined by status of the CW pumps. The wavelength separation between data signal and the CW pumps are multiples of 0.8 nm which is the recommended wavelength grid specified by International Telecommunication Union (ITU) Recommendations. All newly generated wavelengths therefore comply with the ITU grid. The data signal and the pumps were combined with a Wavelength Division Multiplexer (WDM) after passing through polarization controllers PC2, PC3, PC4 and PC5 respectively.

The combined signals were then launched into a 10-km DSF fiber (zero dispersion wavelength is 1538 nm) through a circulator (CIR1). The Raman pump at 1455 nm is coupled into the DSF through another circulator (CIR2) in the counter-propagating direction. The Raman pump power at the input of the DSF is 1.9 W. The Raman pump is the key of the experiment to compensate the low FWM efficiency and provide the power efficiency required for newly generated multicast channels. The newly generated signals were obtained from
the output ports of WDM2 and sent to the optical spectrum analyzer and the 50 GHz oscilloscope for optical spectrum and time domain analysis, respectively.

Figure 2 show the output spectrum of the FWM signal for three cases: (a) all 3 pumps are present (b) C2 and C0 pumps are present and (c) C2 and C1 pumps are present. The results agree with those given in table 1 and show how multicast channels are selected by the control wavelengths. In all three cases, both 4 and 6 multicast channels carries the same data as the original signal channel with output power larger than -20 dBm have been generated respectively through the use of Raman-assisted FWM. The optical parameters of the output channels and pumps in figure 2a are shown in Table 2. All multicast channels were spaced at multiples of ITU grid spacing of 0.8 nm.

Both channel power and phase are the determining factors in a FWM process. The operating wavelengths are selected to be near the DSF zero dispersion point in order to achieve the best phase matching to maximize the FWM efficiency [6]. The system involves a series of FWM processes where all pumps and channels contribute. The observation of suppression of channels at 1548.2nm (left next to channel 1) and 1557 nm (right next to channel 6) is simply part of the result of the interaction of all the FWM processes in the system.

The eye diagrams of the input and output data signals in each channel are shown in figure 3 when all 3 pumps are present. Good eye opening is shown. The change on eyes shape of the channels is due to the nonlinear effect on data where conversion efficiency is higher on level “1” than “0”. Meantime channel 3 and 4 has the best eye and 1 and 6 has the worst among them. It is because channel 3 and 4 are directly generated by FWM of data and C2 where both have the best power. Channel 2/5 is generated by the FWM that involves channel 3/4 and could be seen as a copy of channel 3/4. Similarly channel 1 and 6 could be seen as a copy of channel 2 and 5. Every copy from a copy degrades a bit in the quality and this may limit the number of acceptable channels given even if we put in more pumps. The extinction ratio (ER) is good (larger than 10 dB). The system potentially can be operated at higher bit rate.

<table>
<thead>
<tr>
<th>λ (nm)</th>
<th>Input Power (dBm)</th>
<th>Output Power (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ch. 1</td>
<td>1549.0</td>
<td>-19.183</td>
</tr>
<tr>
<td>C1 Pump</td>
<td>1549.8</td>
<td>+4.042</td>
</tr>
<tr>
<td>Ch. 2</td>
<td>1550.6</td>
<td>-17.397</td>
</tr>
<tr>
<td>Ch. 3</td>
<td>1551.4</td>
<td>-15.823</td>
</tr>
<tr>
<td>Ch. 4</td>
<td>1553.0</td>
<td>-17.825</td>
</tr>
<tr>
<td>Ch. 5</td>
<td>1554.6</td>
<td>-16.860</td>
</tr>
<tr>
<td>C2 Pump</td>
<td>1555.4</td>
<td>+2.852</td>
</tr>
<tr>
<td>Ch. 6</td>
<td>1556.2</td>
<td>-19.873</td>
</tr>
</tbody>
</table>

Table 2. Optical parameters of all channels

3. Conclusion

We proposed and demonstrated a 6 channel optical multicast system using Raman-assisted FWM in DSF. The system enables the selection of multicast channels by 3 wavelength control signals. The eye diagrams of the multicast signal channels are clear. The system can operate potentially at bit rates higher than 10Gb/s demonstrated here.

4. References

Fabrication and Characteristics of Broadband Cr-doped Fibers by Drawing Tower


1Institute of Electro-Optical Engineering, National Sun Yat-sen University, Kaohsiung, Taiwan
2Graduate Institute of Electro-Optical Engineering, National Taiwan University, Taipei, Taiwan
3LeHigh University, Bethlehem, USA

Abstract: The characteristics of Cr-doped fibers fabricated by drawing-tower-method are reported. The ASE spectrum shows a broadband emission of 1.2-1.6 μm. This indicates that the new Cr-doped fibers may be used as a broadband fiber amplifier.

1. Introduction

The breakthrough technology in water-free fiber fabrication has opened the possibility for using fiber bandwidths all the way from 1.3 to 1.6 μm. The usable spectral band for optical channels in a wavelength-division-multiplexing (WDM) system strongly depends on the gain bandwidth of the fiber amplifiers. For example, the erbium (Er)-doped fiber amplifier provides gain in the C band, the L band, and even the S band which amounts to a total usable bandwidth of 140 nm. Optical gain is also provided by the thulium (Tm)-doped amplifier in the S band and by the praseodymium (Pr)-doped amplifier in the O band. However, none of these fiber amplifiers can fully cover the 1.3 – 1.6 μm range by itself. It is interesting to develop a single fiber amplifier that can cover the whole spectral range of 1.3 – 1.6 μm emission.

The transition-metal-doped materials such as Ni2+ ions and Cr4+ ions [1] have shown broadband 1.3-1.6 μm emissions. Recently, a silica-clad Cr4+:YAG crystal fiber amplifier has been demonstrated by the use of a codrawing laser-heated pedestal growth (LHPG) method [2-3] and by a similar rod-in-tube (RIT) method using a drawing-tower technique [4,5]. The Cr4+-doped fiber amplifier may be employed in the whole 1.3-1.6 μm range. However, it is difficult to achieve a smaller core diameter by the LHPG method. The growth of the Cr4+-doped fibers is slow and the core diameter tends to show some variations along the lengths. Therefore, the integration of LHPG fabricated Cr4+-doped fibers with standard single-mode fibers is still not readily achievable. Due to non-circular shape and lower amplified spontaneous emission (ASE) power level of the Cr4+-doped fibers fabricated by a drawing-tower method, the improvement in the preform fabrication and the drawing process are necessary.

In this study, we fabricate and measure the Cr4+-doped fibers by employing a commercial drawing-tower which has a better core diameter uniformity for adiabatic transmission. This approach has the benefits of a better core diameter uniformity to facilitate splicing with the standard single-mode fiber and a broadband emission from 1.2 to 1.6 μm which has not realized by using currently available fiber amplifiers. This broadband fiber may be used as a new Cr-doped fiber amplifier that is capable of covering the entire low-loss window of silica fiber to further increase the transmission capacity of the WDM system for optical communication applications.

2. Fabrication

The Cr-doped YAG preform was fabricated using a rod-in-tube (RIT) method [6]. A silica tube with outside and inner diameters of 32 mm and 12 mm respectively was used. The silica tube was first tapered in one end to form a cone shape. Then the specimen was inserted with a Cr4+-doped YAG crystal rod to constitute the preform. The silica tube became the cladding when the assembled preform was drawn into fiber by using the drawing-tower method. Figure 1 shows a schematic diagram of a Cr4+-doped YAG preform. The diameter and the length of the Cr4+-doped YAG crystal were 3 mm and 5 cm, respectively. In order to keep the shape of the core to be a good circularity, the pressure inside the tube should be controlled during the process of the drawing-tower method.

A commercial drawing tower with a carbon-resistant furnace was used to fabricate the Cr-doped fibers. Due to the different thermal expansion coefficients between the Cr4+-doped YAG crystal and silica, the rate of temperature rise was kept gradation and slow. The preform was loaded into the center of the furnace first, and then the temperature was augmented from 1100 °C to 2200 °C with a ramp rate of 20°C/min. The drawing temperature was set at 2060°C when the first lump of perform was dropped. The drawing temperature was much higher than the softening point of silica of 1667°C and the melting point of the Cr-doped YAG crystal of 1970°C. The drawing speed was set at 60 m/min that was much lower than that of 1.0 km/min used in the drawing a conventional single-mode fiber. The Cr-doped fibers were coated with two layers of resin in order to be rolled up to the tray.

3. Measurements and results

One kilometer of the Cr-doped fibers have successfully been drawn by using a commercial drawing tower.
Figure 2(a) shows a fiber end face with a 26-μm-diameter core and a 125-μm-diameter outer cladding. Smaller core diameter of the Cr-doped fibers may be fabricated by using different diameter of the silica tube and Cr⁴⁺-doped YAG crystal rod. To measure the refractive index and Cr⁴⁺ fluorescence intensity profiles, a laser scanning confocal microscope with a 635-nm DFB laser and a 1064-nm Yb fiber laser was used. A 40x objective lens with a NA of 0.65 was used to achieve 1-μm lateral spatial resolution [3]. Figure 2(b) shows a refractive-index profile with \( n_{\text{core}} = 1.609 \) and an index difference \( \Delta = 9.5\% \). The core and cladding were formed by inter-diffusion between YAG and SiO₂, hence the refractive index of core was decreasing.

![Fig. 1. Cr⁴⁺-doped YAG preform.](image)

An X-ray diffraction pattern of a Cr-doped fiber is shown in Fig. 3. It showed that it was a mixture crystal in the core. However, the core of the Cr-doped fiber was not a single crystal and the sapphire was a nano crystals mixed in silica matrix. For the Cr-doped fibers fabricated by the drawing-tower in this study, both the core and the fiber diameters have good uniformity which is good for splicing with the standard single-mode fibers and broadband WDM couplers for lightwave communication system applications.

![Fig. 2. (a) Photograph of the polished end with a 26-μm-diameter core and (b) the refractive index profile of the Cr-doped fiber.](image)

![Fig. 3. The X-ray diffraction pattern of a Cr-doped fiber.](image)

It is well known that Cr ions have several oxidation states such as Cr³⁺, Cr⁵⁺, Cr⁶⁺, and Cr⁶⁺. Cr³⁺ is the major Cr oxidation state in YAG. The Cr³⁺:YAG fluorescence ranges from 0.65 to 0.75 μm [7], whereas the 1.2 to 1.6 μm optical spectra are dominated by Cr⁴⁺. Therefore, to measure the fluorescence spectrum of Cr-doped fibers, a 1064-nm Yb fiber laser was used as the light sources. Figure 4(a) shows the fluorescence spectrum of the Cr-doped fiber. The fluorescence spectrum of the Cr-doped fiber was excited by 1064-nm in wavelength with the initial power of 1.0, 1.5, 2.0 W. The Cr⁴⁺ fluorescence spectrum indicates a typical broadband emission from 1.2 to 1.6 μm The Cr⁴⁺ fluorescence spectrum shows a peak emission about 1.4 μm and a typical broadband emission from 1.2 to 1.6 μm. In Fig. 4(b), where the amplified spontaneous emission (ASE) spectrum of the Cr-doped fiber is through a 2.0-cm propagation length, the fluorescence spectrum shows a typical broadband emission from 1.2 to 1.6 μm with the power density about microwatt level. In order to obtain a level of milliwatt in fluorescence intensity, an improvement in the fabrication of the Cr-doped fibers is currently under investigation and will be presented.

![Fig4. (a) The fluorescence spectrum and (b) the ASE spectrum of the Cr-doped fibers.](image)

4. Conclusion
In summary, we have successfully fabricated and measured the Cr-doped fibers by using a commercial drawing-tower technique and a rod-in-tube method. The Cr-doped fibers had a 26-μm core diameter and a 125-μm cladding diameter. The refractive index was \( n_{\text{core}}=1.609 \) with \( \Delta = 9.5\% \). The Cr⁴⁺ fluorescence spectrum showed a broadband emission from 1.2 to 1.6 μm. The advantages of using the drawing tower to fabricate the Cr-doped fibers are to have a better control of the core diameter and the fiber uniformity. These are beneficial when integrated with the standard single-mode fibers and broadband WDM couplers for lightweight communication systems. This study makes it achievable to fabricate a new Cr-doped fiber which has potential for being used as a broadband fiber amplifier to cover the bandwidths in the whole low-loss window of silica fibers.

5. Acknowledgement
This work was partially supported by the Department of Industrial Technology under the Contract 95-EC-17-A-07-S1-025, the MOE Program of the Aim for the Top University Plan, and the National Science Council under the Contract NSC 95-2215-E-110-090.

6. References
Gain and Noise Figure Enhancement through C-band Signal Injection in a Double-Pass L-band EDFA

Kai-Ming Feng and Pi-Jen Tsai
Institute of Communication Engineering, National Tsing Hua University, 101, Section 2 Kuang Fu Road, Hsin-Chu, Taiwan, 300
kmfeng@ee.nthu.edu.tw

Abstract: A high-efficient novel double-pass L-band EDFA with a C-band signal injection configuration is proposed. More than 41-dB small-signal-gain, <6-dB NF from 1570-nm to 1600-nm and 35% PCE with -10-dBm input power at 1580-nm was achieved.

In the past decade, the transmission traffic in optical networks has tremendously increased due to the advent of WDM technologies and the growth of Internet business. One key enabling component that makes WDM technologies possible is the maturity of erbium-doped fiber amplifiers (EDFA). Since conventional C-band EDFA (1525 nm to 1565 nm) can not relieve the bandwidth thirst, many researches had been conducted to extend the usable bandwidth from C-band to either L-band (1570 nm to 1600 nm) [1] or S-band (1480 nm to 1510 nm) [2] by using pure-silica-based EDFA technologies.

A gain-shifted L-band EDFA can be pumped at either 980 nm or 1480 nm. However, erbium ions have relatively high absorption in C-band which makes it possible to pump an L-band EDFA by a C-band laser [3]. Since L-band EDFAs typically work at low average inversion level (< 40%) to shift the optical gain from C-band to L-band, which is far from the peak absorption/emission bands of erbium ions, the pump conversion efficiency (PCE) of an L-band EDFA is relatively low. To improve the PCE, several approaches have been proposed, such as double-passing the optical signals back and forth in the EDF to reuse the residual C-band amplified spontaneous emission (ASE) power [4] or by injecting C-band ASE [5] or C-band laser signal [6] into L-band EDFA as an auxiliary pump source. Using C-band laser signal injection in an L-band EDFA has a couple advantages: (1) the injected co-propagating C-band signal can suppress the strong backward C-band ASE in the forward pumping scheme to improve PCE and signal gain. (2) The injected C-band signal enhances the averaged population inversion along the fiber which reduces NF, and (3) in addition to 980 nm pump, the amplified C-band laser signal acts as a secondary pump source. As a result, the PCE, signal gain, and NF performance can be enhanced in an L-band EDFA with an auxiliary C-band pumping architecture.

Instead of using an additional short-length EDFA as an ASE injection source to improve L-band EDFA performance [5], we demonstrated a novel gain-flatten double-pass L-band EDFA with a single C-band laser signal injection configuration. Our numerical and experimental results show that low noise figures (< 6dB), high gains (> 40 dB) and flattened gain profile (< 1.5 dB) were achieved by such a configuration. In addition, high pump conversion efficiency of about 35% is also demonstrated with -10 dBm optical input power at 1580 nm.

Fig.1 (a) Schematic diagram of the proposed L-band EDFA, (b) Comparison of the simulation results of gain and noise figure

The proposed double-pass L-band EDFA with C-band laser injection configuration is displayed in Fig. 1(a). The EDF used in our numerical simulation and experiment is Corning PureMode® Er 1550C3 EDF. The parameters of this fiber are: NA = 0.23, mode-field diameter = 5.5μm at 1550 nm, peak absorptions are 3.62 dB/m and 5.64 dB/m at 976 nm and 1529 nm, respectively. The forward-pumped two-stage EDFA scheme consists of 30 m and 50 m of EDF in the first and second stages, respectively, to obtain a flat gain profile from 1570 nm to 1600 nm but still with sufficiently high gain. A 90/10 optical coupler is used to couple the L-band signal and C-band injection, respectively, into the L-band EDFA. An optical fiber mirror is placed after the second stage of EDFA to reflect the amplified signals and ASE back into the EDF to achieve double-pass function. An optical circulator (OC1) is used to route the input signal into EDFA and the amplified L-band signal out from the EDFA and an optical spectrum analyzer (OSA)
is applied to evaluate the performance.

In our numerical simulation, to make a comparison, we simulate three double-stage, double-pass L-band EDFA configurations: traditional scheme with no injection, with C-band ASE injection from a short-length EDFA [5] and the proposed C-band laser signal injection; all with the same EDF length and pump power. The performance is evaluated by optical gain and noise figure (NF) with -30 dBm optical input power for the three configurations. Fig. 1(b) shows the simulated small signal gains and NFs. The square, circle and triangle symbols represent the performance of the C-band ASE injection scheme, C-band laser signal injection scheme and traditional scheme with no injection, respectively. It is clear that the optical gain of the proposed C-band laser signal injected double-pass configuration exhibits more than 42 dB in the interested wavelength range (1570 nm ~ 1600 nm). However, the other two reported configurations have about 7 dB gain differentials below the new proposed scheme. The high gain performance can be attributed to the better PCE by the assisted C-band laser signal pumping scheme. On the other hand, better PCE contributes to the lower noise figure. As a result, the new proposed L-band EDFA exhibits the lowest NF, all less than 6 dB from 1570 nm to 1600 nm. The noise figure improvement is about 1 dB over the entire interested wavelength range.

The experimental setup follows the simulation configuration, as depicted in Fig. 1. Two pump lasers, with pump power of 42 mW and 61 mW, are used to forwardly pump the two EDF sections at 980 nm. An optical fiber mirror with insertion loss monotonically increased from 2.2 dB to 2.8 dB in the wavelength range between 1570 nm to 1600 nm is used to achieve double-pass function in this L-band EDFA. The wavelength of the injected C-band signal is set at 1550 nm with -10 dBm power before the 10/90 optical coupler. An external cavity tunable laser is used to sweep the wavelength and optical input power in L-band. The input L-band optical power is measured before the 10/90 optical coupler. The gain and NF of the amplifier are then analyzed by an optical spectrum analyzer after an optical circulator.

The output optical spectra of the proposed L-band EDFA with input signal power of -30 dBm and wavelengths ranged from 1570 nm to 1600 nm with 5 nm increment is shown in Fig. 2(a) with 0.1 nm resolution bandwidth on OSA. We can see that the signal gains are all above 41 dB in this wavelength range and gain variation is within 1.5 dB. Fig. 2(b) compares the gain and noise figure distribution as functions of wavelength between the numerical simulation and experimental results with signal input power at -30 dBm from 1570 nm to 1600 nm. The measured optical gains vary from 41.48 dB to 42.89 dB and NFs are from 6.21 dB down to 5.22 dB. It can be seen clearly that the experimental results, for both gain and NF, match the numerical simulations well. Since the insertion loss of the fiber mirror used in the experiment monotonically increases from 2.2 dB to 2.8 dB as wavelength increases from 1570 nm to 1600 nm, the maximum deviation between simulation and experimental results is about 2.8 dB in signal gain at 1600 nm. Fig. 3 shows the measured optical gain and NF as a function of input power at wavelength 1580 nm. From the figure, the small signal gain of this new approach can be as high as about 43 dB with input optical power below -30 dBm. The noise figures are less than 5.8 dB with less than -10 dBm optical input power. The optical output power with -10 dBm input power is about 15.6 dBm, which corresponds to about 35.2% PCE.

Fig. 3: Measured gain and NF as functions of input power at 1580 nm

To sum up, a high-efficient novel double-pass, C-band injected, L-band EDFA is proposed. More than 41-dB small signal gain, <6-dB NF from 1570-nm to 1600-nm and 35% PCE with -10-dBm input power at 1580-nm was achieved.

References
Large Effective Area Photonic Crystal Fibers
with Negative Dispersion and Ultra-Low Splicing Loss

Nguyen H. Hai\textsuperscript{a), Yoshinori Namihira\textsuperscript{b), Shubi Kaijage, Feroza Begum, Tatsuya Kinjo, S. M. Abdur Razzak, and Nianyu Zou}

Graduate school of Engineering and science, University of the Ryukyus, 1Senbaru,Nishihara, Okinawa, 903-0129, Japan

Email: a) k068659@eve.u-ryukyu.ac.jp , b) namihira@eee.u-ryukyu.ac.jp

Abstract
In this paper, we propose and demonstrate a novel type of PCF that has two cladding layers with Ge-Doped rods at the inner cladding. The authors numerically show that it is possible to design a single mode PCF with large negative dispersion in all telecommunication bands with both large effective area greater than 200\(\mu\text{m}^2\) over the whole wavelength above 1.2\(\mu\text{m}\) and ultra low splicing loss of \(\sim 0.0\text{dB}\) per fusion-splicing to a conventional SMF.

Key words: photonic crystal fiber (PCF), large effective area, Ge doped

1 Introduction
In this paper, we will investigate a novel approach to manufacturing large-mode area fibers, namely that of micro-structuring the core instead of the cladding. The idea is to dope small areas in the core region to an index contrast which is large enough to be reproducible with small relative error. If the individual doped areas are sufficiently small, the light will experience a refractive index given by the average between the doped and silica/air interfaces. This idea has been explored in connection with micro-structured fibers lasers in order to control the doping level of the active core, but not as yet as a direct route to manufacturing LMA fibers.

An example of such a structure is shown in the Fig. 1(b). This design will obviously have the endlessly single mode property of the PCFs with micro-structured cladding. The advantage of the micro-structured-core design is that it should be much simpler to fabricate, since it has fewer elements and bigger silica/air interfaces in the final structure.

Fig. 1 Silica/air PCF with micro-structured cladding (left), and the proposed PCF with two cladding layers with Ge-Doped microstructure core (right).

2 Structure and design of large mode area PCF (LMS-PCFs)

Fig. 1 shows cross section of conventional PCF and our proposed LMA-PCF. A conventional PCF has uniform air-holes of diameter \(d\) and pitch \(\Lambda\). The LMA-PCF has two cladding layers with different indices. One is the inner cladding with a 2\% Ge-Doped core of diameter of \(d_1\) and a pitch of \(\Lambda_1\), the other is the outer cladding with a hole diameter of \(d_2\) and a pitch of \(\Lambda_2\). Thus, by scaling up the physical size of the structure, the effective index
contrast between two cladding can be made arbitrarily small, and the core size can be scaled up to the limit set by the tolerable micro- and macro-bending losses. In fact, due to the strong wavelength dependence of the effective cladding index, the fibers can be single-mode at all wavelengths.

Fig. 2 show the effective area ($A_{\text{eff}}$) as a function of the wavelength of the proposed LMA-PCF, where ($\Lambda_1=1.9\mu m$, $d_1=0.7\mu m$, $\Lambda_2=6.5\mu m$, $d_2=1.56\mu m$). From the results, it is seen that $A_{\text{eff}}$ variation with -1%, +2% and +4% variation of all parameters. The $A_{\text{eff}}$ of more than 200 $\mu m^2$ can be obtained over the whole wavelength range above 1.2 $\mu m$ with large negative dispersion. Fig 2(b) shows that the dispersion tend to increase at the longer wavelengths and dispersion properties of proposed PCFs are insensitive with -1%, +2% and +4% variation of all parameters ($\Lambda_1=1.9\mu m$, $d_1=0.7\mu m$, $\Lambda_2=6.5\mu m$, $d_2=1.56\mu m$).

It can be seen from Fig 3 that although the field has confined mostly in Ge doped core region, noticeable penetration exits and spreads to the second core, thus leading to the large effective are of the proposed PCF. There is no leakage outside of the first innermost air-hole ring due to the contrast refractive index between the two cladding is very strong. From the experimental result, we can conclude that, by scale up and adjust the physical size of air-holes at the second cladding, larger effective area of the fiber can be achieved as well as the fiber can be single mode at all wavelength. Since the Modal Diameter-Petermann II is greater than 15$\mu m$ above 1.2$\mu m$ wavelength, the splice loss between the LMA-PCFs and conventional single mode fiber can be estimated to be 0dB utilizing overlap integral.

3 Conclusion

Novel photonic crystal fibers with two different air-holes and 2% Ge rod-core claddings geometries have been investigated, in order to design single-mode large mode area fibers. We show that the proposed LMA-PCFs could archive an effective area of more than 200$\mu m^2$ and with negative dispersion and no splicing losses. We believe that the proposed LMA-PCF will be beneficial for future ultra−broadband transmission application. The proposed LMA-PCF with flattened-dispersion is being investigated.
A Novel Distributed Fiber-Optic Sensor Based on a Mach-Zehnder Interferometer

Qizhen Sun, Deming Liu, Jian Wang

School of Optoelectronic Science and Engineering, Huazhong University of Science and Technology, Wuhan 430074, Hubei, P. R. China
Tel: +86 27 62486521; Fax: +86 27 87556188; E-mail: qzsun@mail.hust.edu.cn

Abstract: We report a distributed fiber-optic vibration sensor using a ring Mach-Zehnder interferometer (MZI) configuration. A spatial resolution better than 40m of single vibration is experimentally verified, and the feasibility of locating multiple vibrations is simulated.

1. Introduction

Distributed optical fiber sensing is to detect and locate non-concurrent external signals along a whole length of fiber. Because of their extremely high sensitivity, interferometer sensors are well known for measuring time-varying signals. Many distributed sensors applicable for intrusion detecting are based on the combinations of Sagnac and other interferometers, such as Sagnac/Mach-Zehnder [1], Sagnac/Michelson [2], Sagnac/Sagnac [3, 4], etc. All the techniques need complicated system configurations and signal processing methods, and Sagnac interferometer is not sensitive to low frequency vibrations.

In this paper, instead of Sagnac interferometer, a simple configuration based on a ring MZI is newly adopted, which two light-waves simultaneously counter-propagate in and operates as dual interferometers. It is experimentally demonstrated that the proposed system is able to detect both the frequency and position of single vibration. Moreover, the feasibility of separating and locating the multiple vibrations is simulated.

2. Experimental setup and operation principle

The experimental arrangement for distributed fiber-optic vibration sensing is schematically shown in Fig.1. The DFB-LD driven by direct current is served as the light source, with 1550nm wavelength, 0.25nm line width and 1mW power. In order to improve interference visibility, the double-core and single-mode fiber cable with 1.01Km length and 1.47 refractive index of the fiber core is employed for sensing. The outputs are detected by PIN-FET detectors, with 0.96A/W responsivity and 3dB bandwidth of 10MHz. The sampling rate of data acquisition card with synchronous working of four channels is set to 5Mbps. A piezo-electronic cylinder is contained in the sensing cable to simulate a vibration.

2.1 Locating principle of single vibration

A ring MZI consists of coupler C1, C4 and C5, and double core sensing cable. A light-wave from laser is split by C1 into the port1 of circulators C3 and C4, then respectively into C5 and C6 from the port2 of C3 and C4 with no attenuation. Next, the two light-waves propagate in the two legs of the MZI in counter directions [5]. If a vibration is occurred at point P, the phase signals induced by it will be transferred to detectors through couplers and circulators. For simplicity of analysis, the length of all lead-in fibers is neglected. The distance of

\[ L_{cw} = L - Z_m \]  
\[ L_{ccw} = Z_m \]  

Obviously, measurement of \( \Delta T \), with the digital cross-correlation calculation between the outputs of PD1 and PD4 [6], enables to determine the vibration position.

2.2 Locating scheme of multiple vibrations

The system should be able to separate and locate multiple vibrations for practical applications. With real-time fast Fourier transform (FFT) frequency analysis and filtering of detected signals, we design a locating...
scheme of multiple vibrations to automatically locate everyone in the frequency domain. Fig.2 shows the schematic diagram.

Testing of this arrangement, based on the location of a single sinusoidal vibration, was performed by setting the vibrating frequency to 215Hz. Fig.3 shows the results from 6 averages at each vibration point. The straight line through the experimental points of the graph of Fig.3, demonstrates the linearity of the sensor and the agreement with the theoretical response for a single vibration. Fig.4 shows the discrepancy between measured and actual positions at every measure sample, giving a spatial resolution of better than 40m.

3. Results and discussions

Fig.5 clearly displays the emulational results of multiple vibrations locating scheme with LabView simulation. A, B, C and D are four group emulations which two or three vibrations at different positions and of different frequencies occur simultaneously. As shown in Fig.5, the method can separately position every one of multiple vibrations, and the calculated values tally with the actual setting values.

4. Conclusion

The paper presented a novel distributed fiber-optic vibration sensor employing a simple ring MZI configuration. We have experimentally demonstrated that the sensor can measure both frequency and position of a single vibration in real time. We have proposed a separately locating scheme of multiple vibrations and simulated its feasibility. We believe that this system has excellent practical potential for intrusion detecting.

References

Asymmetrical Raman resonator for multi-wavelength Raman fiber laser

Young-Eun Im (1)(2), Swook Hann (2), Dong-Hwan Kim (2), and Chang-Soo Park (1)
(1) School of Information and Mechatronics, GIST, Gwangju, 500-712, South Korea
Tel: +82-62-970-3145 Fax: +82-62-970-2237, csp@gist.ac.kr
(2) Strategic technology group, Korea Photonics Technology Institute (KOPTI), Gwangju, South Korea, swoookhann@yahoo.com

Abstract
A novel Raman fiber laser emitting three output wavelengths has been implemented using an asymmetrical Raman resonator of broadband chirped fiber Bragg grating (CFBG) and Germano-silicate Raman fiber. It is a promising solution of cost-effective and user-friendly technique to flexibly manipulate as many as output wavelengths.

Introduction
Multi-wavelength Raman fiber lasers (RFLs) for Raman amplification is a significant technique to achieve a broad and flat amplification in the long-haul wavelength-division-multiplexing (WDM) transmission systems [1]. Stimulated Raman scattering (SRS) in optical fibers shows a powerful solution to amplify optical signals and to generate new laser wavelengths, multi-wavelength RFLs using Germano-silicate fiber (GDF) or Phosphor-silicate fiber (PDF) have attracted considerable interest. Several schemes of multi-wavelength Raman lasers have been presented so far: nested cavities with six output wavelengths [2,3], cascaded fiber end reflector [4]. These methods, however, require not only a number of reflectors but also a sophisticated skill to align components in free space for achieving a multi-wavelength RFL. We chose the broadband CFBG as a wideband reflector to construct simple Raman resonator and eventually to reduce a unit cost of RFL so it would be a significant advantage to have high power RFL with well-proven reliability and mass-productivity.

In this paper, a novel and cost-effective technique has been proposed to enhance the multi-wavelength RFL in the long-wavelength spectral range of 1400 nm band by using the broadband CFBG reflector, for the first time to the best knowledge of the authors.

Experiment result and Discussions
The three-output asymmetrical RFL is shown schematically in Fig. 1. Ytterbium double-clad fiber laser (Yb-DCFL) with the maximum power of 17.6 W at 1070.8 nm is used as a pump source. The gain medium, GDF has a 23 mol % of GeO₂, the Raman cross section of the fiber yields a high Raman gain coefficient of 22 dB/kmW at 1310 nm. The fiber has a refractive index difference, Δn = 0.032, LP11 mode has a cutoff around 900 nm, and a background loss of 1 dB/km at 1480 nm. Considering the Stokes shift of GDF, to build the triple RFL in 14xx nm, fourteen FBGs are needed. In Fig. 1, discrete FBGs (DFBGs) with high reflection values are a function of threshold power reduction for Raman stokes shift. Four successive Raman shifts, 1124, 1184, 1247, and 1320 nm are generated.

In order to reduce the number of reflectors, we configured asymmetrical Raman resonator using the broadband CFBG instead of HR DFBG in the reflection part. A fabricated CFBG is a role of assembly DFBGs, whose reflection and transmission spectra are shown in Fig. 2. The CFBG was written in the SMF-28 in this experiment over high reflectivity, 99.9% in the wavelength range, 1402 to 1463 nm. We have designed
CFBG which is equal to the center wavelength of fifth stokes wave and the output wavelength of RFL is aimed at the S and C band for telecommunications. The maximum gain for the last Stokes (at 1320 nm) is at 1400 nm. Consequently, 1400 nm can transfer efficiently part of its energy to the last output wavelength (at 1460 nm).

![Fig. 2 Measured reflection and transmission spectra of broadband CFBG.](image)

Based on above results, asymmetrical Raman resonators with three wavelength outputs RFL were experimented. Output wavelengths have been chosen for 1415, 1430, and 1445 nm, their relative reflection values are 35, 70, and 30 %, respectively. The measured RFL spectra are shown in Fig. 3 (a) by adjusting pump power of 9.98 W at 1400 nm. Three lasers, 1415, 1430, and 1445 nm were sequentially generated at a threshold power of 3.93, 4.47, and 8.36 W, respectively. A total output power of about 1.02 W at 1400 nm with 15.5 % of slope efficiency. Note that the output spectrum is a little different ASE level in the center of 1400 nm compared to the symmetrical Raman fiber laser due to the unbalanced reflection of broadband CFBG. Three lasers are competed with each other in the fixed cavities with dissimilar reflection in their output reflectors. Output characteristics of RFL with respect to the input pump power are presented in Fig. 3 (b). Note that 1430 and 1445 nm outputs are higher than that of the 1415 nm including slope efficiencies. Although this laser showed lower output and efficiency compared to the symmetrical Raman resonator, we are convinced that it will be improved the efficiency of laser through the refinement of CFBG.

![Fig. 3 (a) Asymmetrical three-wavelength RFL spectra at input pump power 9.98 W at 1400 nm (b) RFL characteristics.](image)

**Conclusions**

In summary, we proposed a novel asymmetrical Raman resonator to enhance the multi-wavelength RFL in the spectral range 1402 nm to 1463 nm by using the broadband CFBG. Total output power of laser can be obtained 1.02 W and the pump conversion efficiency of laser is calculated about 15.5 %. The proposed method can provide a strong potential to inherently enhance the laser characteristics and expand its applications.

**References**


Design of Dispersionless Fiber Bragg Grating Filters by using Lagrange Multiplier Constrained Optimization Algorithm

Cheng-Ling Lee¹*, Ray-Kuang Lee², and Shien-Kuei Liaw³

¹ Department of Electro-Optical Engineering National United University, Miaoli, 360, R. O. C.
E-mail: cherry@nuu.edu.tw Phone: 886-37-381732 Fax: 886-37-351575
² Institute of Photonics Technologies, National Tsing-Hua University, Hsinchu, 300, R. O. C.
³ Department of Electronic Engineering, National Taiwan University of Science and Technology, No. 43, Sec. 4, Keelung Rd, Taipei, 106, R. O. C.

Abstract
This paper presents a new synthesized method for designing dispersionless FBG filters. The method is based on a multiobjective Lagrange multiplier constrained optimization which can constrain various parameters of the designed filters for practical applications.

1. Introduction
In general, the layer-peeling (LP) method is commonly used for designing FBG filters. In this paper, a new optimization-based approach for synthesizing the narrowband dispersionless FBG (NBDL-FBG) filter for optical fiber communications is investigated. The approach is based on a simple and direct algorithm, the Lagrange multiplier constrained optimization (LMCO) method. The LMCO method has been proved to be very useful in designing optical pulse shapes to achieve various goals [1]. Recently, we have presented the LMCO algorithm for designing multichannel FBGs for DWDM [2]. In the present study, we further extend the proposed algorithm in such a way that it can handle synthesis problems involving multiobjective optimization with not only the reflection power spectrum has to be optimized but also the phase of the designed FBG filter should be optimized. This is particularly important for designing narrowband dispersionless FBGs for which both the reflectivity in whole spectrum and in-band dispersion spectrum have to meet the required performance.

2. Lagrange Multiplier Constrained Optimization (LMCO) Algorithm for Designing Dispersionless FBGs
The LMCO algorithm is based on the well-known coupled-mode equations for FBGs [3]. An objective functional was defined that needed to be minimized. The objective functional needed to variation with \( R \) and \( S \) (the forward / backward-modes in FBGs). Therefore, in the dispersionless FBG designing case, the complex Lagrange multipliers \( \mu \) and \( \delta \mu \) were introduced and the cost function should be defined as following:

\[
J = \frac{1}{2} \int_0^1 \left[ r(\lambda) - r_0(\lambda) \right]^2 d\lambda + \beta \int_0^1 |s(z)|^2 dz + \eta \int_0^1 r_0(\lambda) \cdot \left[ \rho(\lambda) - \rho_0(\lambda) \right]^2 d\lambda + \int_0^1 \int_0^1 \left[ \mu_k \right] \cdot \left[ \int \left( \frac{dR}{dz} - i\delta R - i\kappa S \right) dz + \int \left( \frac{dS}{dz} + i\delta S + i\kappa^* R \right) dz \right] \right] dzd\lambda \tag{1}
\]

Where \( r(\lambda) = |\rho(\lambda)|^2 = |S(\lambda)/R(\lambda)|^2 \) is the reflection power spectrum, \( \rho(\lambda) \) is reflection coefficient, \( \kappa(z) \) is coupling coefficient profile, \( L \) is grating length, and \( \beta \) and \( \eta \) are positive numbers acting as weighting parameters for the constrained control. From the equation (2), the deviations from reflection power \( r(\lambda) \) and targeted reflection power \( r_0(\lambda) \) spectrum are summed over all spectral points. The phase in stop-band is the only item that we concerned, the deviations from reflection coefficient \( \rho(\lambda) \) and targeted reflection coefficient \( \rho_0(\lambda) \) spectra are needed to multiply by targeted reflection power \( r_0(\lambda) = \rho_0(\lambda) \times \rho(\lambda)^* \) which acts like a filter for neglecting the contribution of phase spectral points in outside of stopband. In the defined cost function, Eq. (1), the \( \kappa(z) \) was kept real and used to shape an output reflection power \( r(\lambda) \) of a given reflection spectrum \( r(\lambda) \) and to minimize the reflection power spectra difference, in-band reflection coefficient spectra difference and the norm of the coupling coefficient profiles simultaneously. To minimize the cost functional \( J \cdot \) a variational method for Eq (1) was used through \( \mu \) and \( \delta \mu \). Finally, the algorithm was solved in a self-consistent way and evolve \( \kappa \) by the following equation:

\[
\kappa_{\text{new}}(z) = \kappa_{\text{old}}(z) - \alpha \cdot \frac{\delta J}{\delta \kappa} \tag{2}
\]

The optimization of LMCO method for the design of FBGs progressed until convergence was obtained through the iteration procedures.

3. Design Results and Discussions
In order to evaluate the effectiveness of the proposed algorithm for NBDL-FBG. The filters were designed by...
use of the LMCO algorithm described in the previous section with an initial Gaussian apodization profile and the target spectrum of complex reflection coefficient for linear phase response is set to be

\[
\rho_s(\lambda) = \sqrt{\alpha} \exp \left\{ - \left( \frac{\delta_s}{\delta_c} \right)^{10} \right\} \exp \left\{ - j \times 3 \pi \times \delta \right\}
\]  
(3)

In this study case, we find that the best value of \( \alpha \) is around \( 1 \times 10^5 \) and \( \eta \) is about 0.5, which can achieve an optimal and smoother convergence for the designed NBDL-FBG filters. The unconstrained synthesized results for LMCO-NBDL-FBG are shown in the following. Fig. 1 shows the synthesized reflectivity spectra of 0.2nm NBDL-FBG filters by using LMCO and LP methods. The designed reflection spectrum met well with the target spectrum for the grating length 4cm. The corresponding apodization profiles of the index modulation (\( \kappa \)) for the above designed NBDL-FBG filters with different methods are shown in Fig.2. The results show the apodization profile designed by LMCO method is more symmetrical than that designed by using LP method. The initial guess of the Gaussian apodization profile (dotted line) for the proposed algorithm is also appeared in the Fig.2. The in-band group delay and in-band dispersion profiles of the designed NBDL-FBG filters shown in Fig.1 are appeared in Fig.3. From Fig.3, it can be clearly seen that the deviation of the LMCO-based dispersion profile in 75% central region of stopband is smoother and lower than that of LP-based dispersion profile. That is because that the third part of the cost function Eq. (1) multiply by targeted reflection power \( r(\lambda) \) which about 75% central region of stopband is the maximum value \( r_0 \sim 1 \) with higher weighting.

4. Conclusions

In conclusion, a novel NBDL-FBG synthesis method based on the multiobjective Lagrange multiplier constrained optimization (LMCO) is presented. When compared to the existing results from the LP method, the designed LMCO-filters have slightly better performances and can constrain various parameters by adjusting the weighting parameters in the cost function for practical applications.

5. Acknowledgements

The authors would like to acknowledge the financial support of the National Science Council of Taiwan through its grant: NSC-95-2221-E-239-053.

References:

Effect of EDFA saturation cross talk in direct-detection schemes

P. S. Chan and H. K. Tsang
Department of Electronic Engineering, The Chinese University of Hong Kong, Shatin, Hong Kong, China.
pschan@ee.cuhk.edu.hk

Abstract: The effects of EDFA saturation crosstalk under different modulation schemes were studied. Our experiments show that reconfigurable networks employing DPSK will have less eye closure from saturation crosstalk compared with NRZ-OOK or QAM.

For multi-channel DWDM systems, both nonlinear phase noise and amplitude noise degrade signal-to-noise ratio (SNR) to different extents. Reconfigurable networks employing reconfigurable optical add-drop multiplexers (ROADM) can have another possible SNR degradation because of the changes in number of channels that affects saturated EDFA performance and produce dynamic gain tilt and gain transient for systems [1]. The effect of saturation-induced gain modulation is found to vanish at higher signal frequencies [2]. Unlike crosstalk from other channels, saturation crosstalk from EDFA cannot be eliminated by band-pass filter.

During channel add/drop in a long haul WDM network, dynamic gain fluctuations lead to transients in survival channels whose characteristics (rise time, power level, etc.) vary with input power, EDFA characteristics and wavelength [3]. Such transients may result in error bursts because of the limited dynamic range of optical receivers. While the compensation of gain transient and gain tilt has been addressed in on-off keying system [4], the effect of which is being studied using other modulation formats in this paper.

When different channels are multiplexed into EDFA and one channel is added/dropped, the add/drop channel introduces power transient in each of the EDFA in the transmission link as shown in Figure 1(a). Transient power fluctuations are also produced in the surviving channels as shown in Figure 1(b). The magnitude of the transient will vary with channel wavelength and input power, and long-term gain-tilt can be introduced after the transient. Depending on the number of EDFA cascaded, surviving channels suffers from BER reduction because of power fluctuation.

![Fig. 1. Effects of channel added/dropped on WDM network after passing through EDFA. (a) Add/drop channel induces transient along transmission link. (b) Transient power fluctuation on one of the surviving channels (red) experienced another channel is added or dropped/ (green).](image-url)

We compare three direct detection schemes namely on-off keying (OOK), direct detection differential phase shift keying (DD-DPSK) and quadrature-amplitude modulation (QAM) [5]. Figure 2 shows the experimental setup of 16 DWDM channels each with −20dBm output power. A total of −8.23dBm and 11.2dBm averaged power were measured before and after the EDFA, respectively. With near 20dB gain, 15 channels were add/dropped by modulating the output power under 100kHz using pulse pattern generator. The survival channel at 1541.62nm, with different modulation formats, was extracted for SNR analysis. Figure 3 shows the eye diagrams of surviving channel under different modulation schemes with and without saturation crosstalk induced by EDFA.
Fig. 2. Schematic representation of different modulation formats.

NRZ OOK
Back-to-back 15 channels add/drop

DD-DPSK
Back-to-back 15 channels add/drop

QAM
Back-to-back 15 channels add/drop

Fig. 3. Demodulated signal eye diagrams of different modulation formats with (right) EDFA saturation crosstalk.

The different eye openings in Fig.3 shows that DD-DPSK has an advantage of less amplitude noise which agrees with corresponding simulation results numerical crosstalk analysis [6]. Other references of optically amplified DWDM systems including FSK [7] will be discussed and simulations as well as experimental results on 10Gb/s transmission will be presented.

Acknowledgements: This work was funded by direct grant project 2050385.

References
Experimental Demonstration of Wavelength-Selective In-Fiber Optical Intensity Modulation

M. Rajabvand, F. Behnia and M. T. Fatehi
EE Department, Sharif University of Technology, Azadi St. Tehran, Iran, P.O. 11365-9363, Fax. +98 21 6602 3261.
Rajabvand@ee.sharif.ir

Abstract: In this paper we report on experiments with a fiber-based optical intensity modulator. It uses tunability of FBG in its transition band, to add the low speed labels on WDM signals in a wavelength-selective manner.

1. Introduction
The fiber Bragg gratings have the best potential for achieving any desired spectral characteristics in all-fiber geometry. This distinguishing feature has made the FBG a fundamental component in fiber optic communication and sensor systems [1]. When the spectral shaping flexibility combines with tunability of FBG in an optical device, the controllable and reconfigurable fiber-based systems emerge [2]. Such components are enabler of true all-optical networks, in which signal conversion from optical to electrical is avoided.

Traditionally, mechanical stresses are applied to FBG by piezoelectric transducers (PZT) or by other means to tune the Bragg wavelength. PZT displacement is coupled to the FBG to move its reflection band so that one desired optical channel would lie totally inside the reflection band (or totally outside of it). Recently, we have proposed a new operation mode of FBG, which focuses on its transition bands of spectra [3]. When the wavelength of optical signal is in the transition region and the tuning voltage is low enough to guarantee the small-signal operation, optical power at the output of FBG will follow the variations of the applied voltage. We have shown that the combined PZT and FBG act as a wavelength-selective modulator (WSM), and can be described analytically by the intensity modulation equation [3]:

\[
P_{\text{out}}(t; \lambda_c) = (B_0 - A_0 \eta V(t)) P_{\text{in}}(t; \lambda_c),
\]

where \(P_{\text{out}}(t)\) and \(P_{\text{in}}(t)\) are transmitted and incident optical signals, respectively; \(B_0\) is the transmission coefficient at the optical carrier wavelength (\(\lambda_c\)); \(A_0\) is determined from the slope of FBG spectra and PZT specifications and \(\eta\) is mechanical coupling coefficient between FBG and PZT.

In this paper, we report on practical implementation of the WSM using multilayer piezo actuators (MLP) and FBG, and compare the measurements results with theoretical predictions.

2. Experiment
Fig. 1 shows the experiment set up for investigating the operation of WSM. The DWDM DFB laser source consists of eight 20 mw modules with individually selectable center wavelengths from 1551 to 1560 nm. The center wavelength tolerance is ± 0.01 nm, and the tuning range of every laser module is ± 1 nm. A variable attenuator is placed to control the optical power level. The laser light is transmitted via a fiber optical coupler to FBG. The measured transmission power spectrum of FBG is shown in Fig. 1, as well. Its maximum reflectivity is 93 %, the center wavelength is 1556.08 nm, and FWHM is 1.78 nm. The two ends of FBG are glued to a MLP. Dimensions of MLP are 10×10×36 mm, its nominal displacement is 32 \(\mu\)m @ 100 v, and resonant frequency is about 45 kHz.

When an electrical signal \(V(t)\) from the function generator is applied to MLP, the FBG is axially pulled from each end. The actuator displacement causes a corresponding change in the FBG length and consequently grating period and index of refraction. As a result, the Bragg wavelength fluctuates with \(V(t)\), which in turn varies (modulates) the reflectivity of the FBG at wavelength of the optical carrier. We measure the transmitted and reflected optical power variations individually using two InGaAs analog photodiodes with responsivity of about 0.9 A/W @ 1550 nm. The measurements results are discussed in the next section.

Fig. 1 WSM experiment set up and measured transmission power spectra for FBG.
3. Results and discussion

Fig. 2(a) shows the oscilloscope traces of signals in CH.1 and CH.2, which measure the signal intensity transmitted through and reflected off the WSM, respectively. The phase difference $\pi$ between the CH.1 and CH.2 signals is due to complementary variations of the reflected and transmitted power ($P_{\text{ref}} = P_{\text{in}} - P_{\text{trans}}$). The optical signal wavelength is tuned near the upper edge band, on wavelength $\lambda_c = 1557.0$ nm, and experiences $1.5$ dB transmission loss. During the increasing segment of $V(t)$, the MLP displacement increases and pulls the fiber grating in length. Because of tensile strains, the Bragg wavelength shifts to upper wavelengths, and decreases the transmission coefficient at $\lambda_c$. Thus the transmitted signal is modulated in opposite phase with respect to $V(t)$, and the reflected signal experience co-phase modulation. If amplitude of $V(t)$ is quite large and/or $\lambda_c$ is chosen to be in neighborhood of extrema of FBG spectra, uniform ascending/descending of transmission coefficient within tuning range fails and nonlinearity effects appear. Fig. 2(b) and (c) show the nonlinearly modulated transmitted and reflected signals, respectively. In fact, the modulation coefficient $A_0$ in equation (1) does not remain constant but it will be time variant which goes to zero at the extrema.

4. Conclusion

In this paper, we considered the tunability of FBG in transition band and implemented an all-fiber optical intensity modulator. The measurement results confirmed the concept of wavelength-selective modulation and its theoretical model. An array of WSM can be used for all-optical modulation of low speed labels on WDM signals [3].

Acknowledgment

This research is supported in parts by Hi-Tech Industries Center of Iran.

References

Defect-Core Hexagonal-lattice Photonic Crystal Fibers with Flattened Chromatic Dispersion and Low Confinement Loss

Shubi Kaijage a), Yoshinori Namihira b), Nguyen H. Hai, Feroza Begum, S. M. Abdur Razzak, Tatsuya Kinjo, Jitsuryo Nakahodo and Nianyu Zou

Graduate School of Engineering and Science, University of the Ryukyus, 1 Senbaru, Nishihara, Okinawa, 903-0213 Japan
E-mail: a) k068455@eve.u-ryukyu.ac.jp , b) namihira@eee.u-ryukyu.ac.jp

Abstract
Defect-core hexagonal photonic crystal fibers (DC-HPCFs) with nearly zero ultra-flattened dispersion of $0 \pm 0.35 \text{ ps/(nm-km)}$, in the wavelength range of 1.397 µm to 1.675 µm exhibiting extremely low confinement loss has been presented.

Key words: Photonic crystal fiber (PCF), chromatic dispersion

1. Introduction
Photonic Crystal Fibers (PCFs) also known as holey fiber or microstructure fiber, have been the subject of much research interest in the recent years for its promising and attractive properties, like single mode operation in a wide band, large waveguide dispersion, nonlinearity, ultra-flattened dispersion and so many [1-5] whereby can not be obtained by conventional optical fibers.

Index guiding PCFs have microscopic array of air channels running along its length that makes the low index cladding around the undoped silica core. In this way, light can be confined and guided properly through the fiber by the mechanism of total internal reflection (TIR). Compared to the conventional fibers, PCFs have additional design parameters namely, air-hole diameter, air-hole rings and hole to hole spacing known as the pitch, and hence offer design flexibility for the guiding properties [5].

Chromatic dispersion and losses in fibers limits the data carrying capacity in optical communication systems. Chromatic dispersion causes light pulses to spread causing inter-symbol interference and becomes a critical issue as transmission rate exceeds 10 Gb/s. The same challenge exists in the design of PCFs too. So far, various index guiding hexagonal PCFs (H-PCFs) with remarkable dispersion and leakage properties [5]-[7] have been reported. By using the defect-core HPCF as shown in Fig.1 to control both dispersion and dispersion slope. We consider a core having multiple defects characterized by omitting air holes A and B as can be seen. From Fig.1, $d$ and $d_1$ are the air-hole diameters and $\Lambda$ is the pitch. The precise control of air-hole diameters and lattice constant (pitch) facilitate flattened chromatic dispersion and low confinement loss simultaneously.

2. DC-HPCFs design
In conventional HPCFs, the cladding is usually formed by air-holes with the same sizes arrayed in a triangular lattice. Chromatic dispersion profile can be engineered by varying the air-hole diameter and hole to hole spacing; however it is difficult to control dispersion and dispersion slope by using PCFs with the same air-hole diameter in cladding region for wide wavelength range. We propose the index guiding defect-core hexagonal PCF as shown in Fig.1 to control both dispersion and dispersion slope. We consider a core having multiple defects characterized by omitting air holes A and B as can be seen. From Fig.1, $d$ and $d_1$ are the air-hole diameters and $\Lambda$ is the pitch. The precise control of air-hole diameters and lattice constant (pitch) facilitate flattened chromatic dispersion and low confinement loss simultaneously.

3. Numerical simulation results
The chromatic dispersion of defect-core HPCF is calculated by [6];

$$D (\lambda) = -\frac{d^2}{c} \frac{\text{Re}[n_{\text{eff}}]}{\lambda^2}$$

With unit of $[\text{ps/(nm.km)}]$, where, $\lambda$ is the operating wavelength, $\text{Re} (n_{\text{eff}})$ is the real part of the refractive index, $n_{\text{eff}}$ and $c$ is the velocity of light in a vacuum.

Confinement loss is the light confinement ability within
the core region. The confinement loss, \( L_c \) is then obtained from the imaginary part of \( n_{\text{eff}} \) as follows [8]:
\[
L_c = 8.686 \cdot k_0 \cdot \text{Im}[n_{\text{eff}}]
\]  
with unit of dB/m, where \( k_0 = 2\pi/\lambda \) is the free space wave number.

And effective mode area, \( A_{\text{eff}} \) is calculated by [8]:
\[
A_{\text{eff}} = \frac{\int \int |E|^2 |dx\,dy|}{\int \int |E|^4 |dx\,dy|}
\]  

where \( E \) is the electric field derived by solving Maxwell’s equations.

Fig. 2 shows the chromatic dispersion for the optimized DC-HPCF as a function of optical wavelength. It can be clearly seen that for the optimized set of parameters and air-hole ring configurations a flattened dispersion of 0± 0.35 ps/(nm-km) in the wavelength range of 1397 nm to 1675 nm has been realized. The optimized parameters which have been obtained through numerical simulations are \( \Lambda = 1.6 \) µm, \( d = 1.056 \) µm, \( d_r = -0.6 \) µm. At 1550 nm dispersion of +0.32 ps/(nm-km) was obtained. Also typical variations of design parameter, \( d \) by ±1% results in a shift of chromatic dispersion curve of about ±1 ps/(nm-km) , giving an upper bound on the severity of the fabrication imperfection. Fig. 3 shows the relation of confinement loss and effective area with wavelength for the above-mentioned optimized design parameters. It can be observed the remarkably low confinement loss, less than 10^{-4} dB/km at the entire communication window has obtained. In addition, from Fig. 3, we can see the effective area, \( A_{\text{eff}} \) of about 10.79 µm² at 1.55 µm, which may be suitable for nonlinear applications.

4. Conclusion
With defect-core HPCF, ultra-flattened chromatic dispersion with low confinement loss was demonstrated. Through numerical simulation flattened dispersion of 0±0.35 ps/(nm-km) with low confinement loss of less than 10^{-4} dB/km over wavelength range of 1397 µm – 1.675 µm was obtained. So, the designed DC-HPCF can be potentially used where nearly zero ultra-flattened dispersion and low confinement losses characteristics are simultaneously needed. Also, birefringence behaviors of this described structure attract our interest, these works are under study.

References

Fig. 2 Dispersion properties of six rings DC-HPCF for the optimized design parameters, \( \Lambda = 1.6 \) µm, \( d = 1.056 \) µm, \( d_r = 0.6 \) µm and for ±1% variation of design parameter, \( d \).

Fig. 3 Confinement loss and effective area with wavelength for the optimized design parameters \( \Lambda = 1.6 \) µm, \( d = 1.056 \) µm, \( d_r = 0.6 \) µm.
A Simple Optical Power Limiter for 40 GHz Pulses based on SOA Saturation

G. Contestabile, M. Presi, R. Proietti, N. Calabretta and E. Ciarapica
Scuola Superiore Sant'Anna, Via Moruzzi 1, 56124, Pisa (Italy), tel. +390505492142, fax. +390505492194 email: contesta@ssssap.it

Abstract—We experimentally characterize a simple optical power limiter made by an SOA and a bandpass-filter working on 40GHz pulses. We find a very large (25-33dB) amplitude-modulation-reduction factor for input modulating frequencies in the range 100kHz-1GHz.

Future transmission networks will benefit of new circuits operating transparently in the optical domain. Semiconductor Optical Amplifiers (SOAs), which show high non-linearity in a compact device, are largely employed in the realization of all-optical functions [1]. Unfortunately, signal distortions due to the semiconductor dynamics force to adopt interferometric or differential architectures to work at multi-gigabit data rates [1]. On the other hand, in case of periodic signals the saturation effects can be usefully exploited without pattern distortions. One application is the all-optical power limiting function for short periodic pulses. An all-optical power limiter can be very useful in a number of applications where it is necessary to reduce the amplitude jitter of periodic pulses. As an example, it could be used to reduce spurious amplitude modulations in mode-locked lasers, [2] and to equalize clock signals in clock recovery circuits [3, 4]. Several techniques have been proposed for power limiting, but, they are generally complex [2], and/or require high optical powers [4]. On the contrary, the saturated amplification in SOAs exploited together with a band-pass filter is very effective in a simple optical circuit working at low input power. Here, we give a comprehensive quantitative characterization of the potentials of such a scheme for 40 GHz pulse trains.

The experimental set-up is reported in fig. 1. 6 ps-pulses were carved on an optical carrier at \( \lambda = 1551.5 \) nm by means of an electro absorption modulator driven by a 40 GHz electrical signal. An amplitude modulation was superimposed by an additional intensity modulator driven by a sinusoidal signal from a wide-band waveform generator. The modulation depth and the frequency of the additive signal were controlled by adjusting both the bias of the modulator and the frequency of the electrical signal. The two electrical waveform generators were not synchronized in order to avoid any fixed phase relation among them. This guarantees an effective amplitude modulation also for high frequency modulations that are comparable with the pulse repetition rate. The signal was then amplified by an erbium doped fiber amplifier and power controlled. The limiting circuit was composed by an optical isolator, an SOA and a 0.8 nm tunable band-pass filter. The SOA was a polarization-insensitive pigtailed device with about 28 dB small signal gain, around 200 ps recovery time and 6 dBm output saturation power at 200 mA driving current.

The output signal was analyzed by usual test instruments. The amplitude equalization process can be qualitatively described as a limiting effect due to the saturation of the SOA gain ruled by the semiconductor gain dynamics. Indeed when the gain of the SOA is saturated, input pulses with different power exceeding a certain threshold value, are practically amplified to the same limiting value (see fig.2a). The frequency response of this process is mainly governed by the gain dynamics, whilst the strength of the effect is related mostly to the gain compression. Indeed, by increasing the input power, the amplifier gain compression increases and the effect becomes more pronounced. At the same time, when short pulses propagate through the SOA, self phase modulation and intra-band four wave mixing lead to spectral spreading of the signal. Red chirp is generated in correspondence of gain depletion during the saturation whilst blue chirp corresponds to the gain recovery process [5]. The spectral spreading leads to pulse distortion that can be mitigated by the band-pass filter placed at the SOA output. The filter bandwidth should be large enough to select only part of the output spectrum. We found it has a twofold action both on shape and frequency response of the process (a detailed description of this will be included in the conference presentation). As it is described in fig.2b, depending on the repetition rates (e.g. \( R_1 < R_2 \)), the pulse sequence can enter the SOA before the gain fully recovers: in this case the recovered gain decreases at increasing the repetition rate (\( G_2 < G_1 \)). Nevertheless, due to the pulse periodicity, the technique works also with partial gain recovery without suffering any pattern effect.
For this reason, the circuit is effective also when operated with pulse repetition rates exceeding the inverse of the SOA recovery time. In this case the strength of the effect becomes weaker. However, this implies that pulse trains with repetition rate exceeding 100 GHz might be managed by this technique [6]. To quantify the amount of amplitude modulation reduction obtained by the circuit, we define the amplitude modulation reduction (AMR) parameter, as the input/output ratio (in dB) of the frequency components at the superimposed amplitude modulation (measured on an electrical spectrum analyzer). The frequency response of the AMR for the 40 GHz pulses is reported in fig.3a in case of 50% and 80% modulation depths, by using 5 dBm average input power, and without detuning of the output band-pass filter. We found an AMR in the range 25-33 dB from 100 kHz up to 1 GHz, a cut-off point at a few GHz, and, still an AMR flat response of around 10 dB for modulating frequencies up to 25 GHz. According to our qualitative physical description, the cut-off frequency of the response should be related to the inverse of the SOA recovery time. On the other hand, a certain amount of limiting effect can be expected for modulation frequencies exceeding the cut-off. In this case a weaker effect is due to the partial gain recovery obtained from ultrafast carrier-carrier and carrier-phonon scattering working at sub and picosecond time scale [6]. To visualize the large effect of the limiting effect, we report as an example in fig.3b the case of a pulse train modulated at 100 kHz with 80% modulation depth. It experiences around 27 dB AMR and is completely reshaped to a pattern free condition. We characterized the limiting effect also as a function of the modulation depth. The results are summarized in fig.4a. We fixed a 5 GHz modulating frequency and varied the modulation depth. The AMR response is nearly flat and starts to decline once exceeded the 80%. The efficiency of the effect reduces when the modulation depth is so large that the weakest pulses in the sequence approach the threshold level of the input-output transfer function (see fig.2a). To show this, in fig.4b we report an ultimate case of 86% modulated pulses at 5 GHz. The traces show that the limiting effect still occurs but that is not complete. Finally, in order to characterize the effect of the filter position on the shape of the output pulses and on the efficiency of the limiting effect, we performed a systematic study for a fixed over-modulation condition. The results are reported in fig.5.

REFERENCES


Fig. 3. a) AMR vs. Modulation Frequency; b) Typical complete waveform reshaping by limiting amplification.

Fig. 4. a) AMR vs. Modulation Depth for a 5 GHz modulating frequency. b) Partial waveform reshaping for 86% modulation at 5 GHz.

Fig. 5. AMR vs. filter position. In the insets, corresponding pulse shapes.
Low-Cost Undercut-Etching-Active-Region Method to Fabricate Laterally Tapered Active Waveguide for Electroabsorption Modulator and Spot-Size Converters Integration

F. Z. Lin, S. A. Tsai, T. H. Wu, J. S. You, and Y. J. Chiu

No 70, Lien Hai Rd., Kaohsiung, Taiwan R.O.C.
Institute of Electro-Optical Engineering, NSYSU
Email: p923050005@student.nsysu.edu.tw
Tel: +86-7-525-2000 ext 4472, Fax: +86-7-525-4499

Abstract

A simple, low cost and all-wet-etching process was proposed to fabricate electroabsorption modulator (EAM) integrated with spot-size converters (SSCs). The laterally tapered active waveguide in the SSC region was formed by selectively undercut-wet-etching. Over 20dB/volt modulation efficiency as well as -16dB fiber-to-fiber insertion loss (w/o anti-reflection coating) and 11GHz of -3dB bandwidth in EAM performance was obtained.

1. Introduction

Broadband, high-quality and low-cost optical network is needed to cater the rapid-growth in the data transmission. In supporting such optical network, high-speed, low optical loss and low-cost are thus the main criteria for designing optoelectronic devices, such as laser diode, photodiode or electroabsorption modulator (EAM) [1-2]. Spot-size converters (SSC) have been widely used in these devices due to its directly coupling to the cleaved fiber for reducing the package cost. High-speed electroabsorption modulators are one of the important optoelectronic devices for fulfilling the requirement of high bandwidth-efficiency product and low-chirp performance in fiber communication. The prices, however, are generally on the complicated fabrication processes and their integration, such as selective area etching, re-growth, and high-resolution photolithography [3-4].

In this paper, we report a novel SSC-integrated EAM using wet-etching processing and tapered undercut-etching-active-region waveguide (UEARW) [5]. As shown in figure 1, two SSCs are integrated with electroabsorption for the input and output optical ports. Due to the selective etching properties between InP (waveguide cladding layer material) and InGaAsP (quantum-well material for active region), the laterally tapered active waveguide (LTAW) can be defined by selective wet etching from large laterally tapered active waveguide core. The efficiency of the optical mode transfer can thus be controlled just by engineering the undercut etching processing, avoiding the complicated regrowth or precise lithograph processes.

2. Fabrication and Measurement
Figure 1 shows the schematic structure and the schematic pictures of undercut-wet etching active region. In fabricating the 350μm-long LTAW, an HBr-based etching solution incorporated with HCl-based solution is first used to define the top p-cladding with the widths of from 6μm to 8μm. An H2O2-base solution is then utilized to selectively undercut-etch the MQWs from InP material. The 400μm-long waveguide are served as the active waveguide of EAM with width of 8μm. After forming the active waveguide and LTAW, the passive waveguide is defined by aligning the top LTAW and wet-etched by HBr-based etching solution. 0dBm TE-polarized light at wavelengths of 1570nm is used to characterize the optical transmission against bias. The launched and received fibers are cleaved single mode fibers. As shown in Figure 2, the fiber-to-fiber insertion loss (transmission at 0 volt) is 16 dB and the extinction ratio can exceed 20 dB, indicating that larger mode field in underlying passive waveguide can be transferred efficiently to active waveguide (EAM waveguide). The mode transfer efficiency calculated by beam propagation method [6] is illustrated in Figure 3. About 80% mode transfer efficiency can be achieved in the SSC region. Figure 4 shows the high-speed electro-optical response of the device. The –3dB bandwidth is 11GHz.

3. Conclusion

Spot-size converter with tapered undercut-etching-active-region is easily fabricated, leading to high optical transmission. High-modulation efficiency of 20dB/V and 11GHz of –3dB bandwidth in EAM performance were also obtained from these devices.

4. Acknowledgements

The authors would like to thank the financial supports from the National Science Council, Taiwan (NSC 93-2215-E-110-018) and (NSC 93-2215-E-110-013), Technology Development Program for Academia (92-EC-17-A-07-S1-025) and “Aim for the Top University Plan Taiwan”.

5. References

Power equalization for SOA-based all-optic switch by optical control pulse optimization

Songnian Fu1, P. Shum1, Y.D.Gong2, Xia Li1, Huy Quoc Lam1
1. Network Technology Research Centre, Nanyang Technological University, Singapore, 637533
E-mail: songnian@ntu.edu.sg
2. Institute for InfoComm, Research. 21, Heng Mui Keng Terrace, Singapore, 119613

Abstract—We present a power equalization method for the SOA-based all-optic switch by optimizing the format of optical control pulse. By injecting a negative optical control pulse, the output power fluctuation of a packet can be effectively reduced to zero. Our method is verified by a 2.5Gbps 2×2 exchange-bypass optical switch experiment with 4.8dB Q-factor increase.

1. Introduction
All optical packet switching (OPS) network is a promising technique which is introduced to fully utilize the bandwidth provided by current optical network [1]. The 2×2 all-optic exchange-bypass switch is the basic unit for an OPS system. Considering when the packet contention happens during passing the switch, a solution is to build an optical buffer, which stores one of the packets during contention period. Later the buffer will release the stored packets when the switch output port is free [2].

Recently we have developed a semiconductor optical amplifier (SOA)-based dual-loop optical buffer (DLOB) with a 3×3 collinear fiber coupler. Compared with the existing configurations, our proposed DLOB is simple, compact and inherently stable. Though the packet pattern quality was maintained well after 150us storage, we observed that the optical power of buffered packet was 4.83dB lower than that of the original packet due to the SOA gain saturation [3]. As a result, it is harmful to apply DLOB to the all-optic switch design. In this paper, we present an effective power equalization method by optimizing the format of optical control pulse. As the experimental result, the Q-factor of the packet at the optic switch output port is improved by 4.82dB.

2. Configuration of all-optic switch
Once the optical packet enters the OPS node, only the packet head undergoes O/E conversion, thus the packet’s routing information can be determined. Since the packet payload remains in optical form from source to destination, it can be encoded at high bit-rate with various modulation formats [4]. We first demonstrate the implementation of a 2.5Gbps 2×2 exchange-bypass optic switch with the DLOB in our design, as shown in Fig. 1. The solid lines are light paths and the dashed lines denote electricity circuit connections. The packets are assumed to arrive in a synchronized fashion and up-stream packets have the lower service priority. Therefore the power of the packet arriving at the switch input port is split into two asymmetrical portions. The signal with smaller power will be used to implement clock recovery and head recognition function electronically. Simultaneously the other signal with the most of the optical power is delayed by a 360-meters single mode fiber and waits for the switching operation from network control unit (NCU). If the objective address in the packet head is equal to the address of the local node, the NCU will generate a down-stream signal to drop the delayed packet to port C; otherwise the NCU will forward the packet to the port D of the switch. And there might be a conflict between an up-stream packets and a forward packet. Thus, if the packet contention takes place at the switch output port, the up-stream packets will be buffered by our DLOB. The NCU will give out a ‘load’ signal to store the up-stream packet into DLOB, when the output port D is detected to be occupied by a forward packet. Once the transmission of this ongoing forward packet is completed, up-stream packet will be released from DLOB and appears at the port D of the switch.

3. Experimental results and discussions

Fig. 1: The scheme of an optic 2×2 exchange-bypass switch with the DLOB. (a) Principle of operation of the 2×2 exchange-bypass optic switch. (b) DLOB-based 2×2 exchange-bypass switch.
Fig. 2 The 2.5Gbps 2×2 exchange-bypass switch experimental demonstration. (a) Format of 2.5Gbps input packet signal. (b) Format of generated positive control pulse. (c) Format of generated negative control pulse. (d) The output packet signals under the positive operation mode. (e) The output packet signals under the negative operation mode.

In our experiment, two distributed feedback (DFB) laser modules with maximum output power of 10dBm at different wavelengths were used as the control signal (λc=1553.4nm) and the packet signal (λs=1556.56nm) light sources, respectively. The waveforms of input 2.5Gbps packet signal and correspondingly generated the positive and negative control pulse are plotted in Fig. 2 (a) and Fig. 2 (b-c). Every packet with 3μs length contains 936 bytes 215-1 pseudorandom binary sequence (PRBS) at 2.5GHz bit-rate. The width of optical control pulse is 3.2μs in order to cover the whole packet, as shown in Fig. 2 (b-c). The optical control pulse 1 is used to load the packet and the control pulse 2 is generated to release the packet. Taking the packet A, B and C with 19.8 μs cycles into account, the optical control pulses are only synchronous with the packet A. Therefore the packet A will be stored in the switch during packet contention, whereas the packet B and C are reflected to the output port D of optic switch directly without storage, which is shown in Fig. 2 (b). When we monitor the output waveform at port D of the optic switch, it is clear to see that the interval between packet A and B is reduced to 9.9μs due to the storage in Fig. 2 (d). And the power leakage of the packet A at the input position results from the SOA unbalanced directional gain. However with the positive control pulse injection, the output peak power of packet A is 4.83dB smaller than that of the packet B. It is harmful to the receiver of next optic switch. In order to solve this problem, we adjusted the usual positive optical control pulse to the negative one, as shown in Fig. 2 (c). The power of the injected negative optical control pulse is kept the same as that of the positive injection one; the only difference is the polarity of the control pulse. We can observe that the output power is equalized in Fig. 2 (e). When the packet contention happens, the eye diagram is measured at the port D of the optic switch in Fig. 3. The Q-factor of the output packet under the negative optical pulse injection is 4.8dB higher than that of the packet pulse under the positive optical pulse injection. We believe that this power equalization method can be also applied to other SOA cross phase modulation-based applications.

4. Conclusion

By changing the usual positive control pulse injection to the negative pulse injection, the output power of SOA-based 2×2 exchange-bypass optic switch could be equalized effectively and performance is improved significantly. The power fluctuation of the packets can be reduced from 4.83dB to 0dB, while the Q-factor is improved by 4.8dB.

Acknowledgements

This work is partially supported by the project M47040039 of Agency for Science, Technology and Research, Singapore and partially supported by Open Fund of Key Laboratory of Optical Communication and Lightwave Technologies, Beijing University of Posts and Telecommunications, Ministry of Education, P. R. China.

References

A Multiple-Operation-State Optoelectronic Switch

Der-Feng Guo\(^1\), Jung-Hui Tsai\(^2\), Ming-Yue Fu\(^1\)

\(^1\) Department of Electronic Engineering
Air Force Academy, P.O. Box 1449 Kangshan, Kaohsiung County 820, TAIWAN
TEL: 886-7-6264185; FAX: 886-7-6264185; Email: g6f6guo@ms5.hinet.net

\(^2\) Department of Electronic Engineering
National Kaohsiung Normal University, 116 Ho-ping 1st Road, Kaohsiung 802, TAIWAN
TEL: 886-7-7172930 Ext.7216; Email: jhtsai@nkucc.nknu.edu.tw

Optoelectronic switches have bistable states switchable by optical or electrical input. Minority-carrier storage might slow down the response speed of these devices. Furthermore, the bistable states in the device characteristics could not effectively reduce the number of elements and process step in actual applications. Therefore, it is promising to develop an optoelectronic switch free from minority-carrier storage and presenting multiple operation states.

For minimizing the minority-carrier storage, triangular-barrier structure is one of the promising candidates. Introducing the hole accumulation and sequential avalanche multiplications into the device operation is one of the best ways to achieve the multiple states. In this work, a Schottky-contact triangular-barrier optoelectronic switch (STOS) is proposed. The triangular barrier is formed by inserting an InGaAs p-type delta-doped (\(\delta p^+\)) quantum well into an n' - GaAs layer. A double S-shaped negative-differential-resistance (NDR) phenomenon is observed in the STOS characteristics due to the hole accumulation in the \(\delta p^+\) well and the sequential avalanche multiplications in the reverse-biased pn and metal-semiconductor junctions.

The layer sequence is: an n'-GaAs (3×10\(^9\)cm\(^{-3}\)) layer of 200nm thickness that is designated the collector layer, an n' -GaAs (2×10\(^9\)cm\(^{-3}\)) layer of 600nm thickness that is designated the active layer, an \(\delta n'\)Ga\(_{0.5}\)As layer of 15nm thickness whose center is inserted with a \(\delta p^+\) (5×10\(^9\)cm\(^{-3}\)) sheet, an n' - GaAs (2×10\(^9\)cm\(^{-3}\)) layer of 600nm thickness that is designated the barrier layer, and an n-GaAs (3×10\(^9\)cm\(^{-3}\)) layer of 200nm thickness that is the emitter layer.

Figure 1 is the band diagram of the STOS with a positive collector-to-emitter voltage \(V_{CE}\). Figures 2(a) and (b) show the I-V characteristics of the STOS under dark and illumination, respectively. Double S-shaped NDR performances are observed under both conditions. Under illumination, lower values of \(V_{CE}\) voltage are sufficient to achieve the switchings in the initial off-state (\(V_{ui}\)) and intermediate on-state (\(V_{sz}\)) as a result of the photogenerated holes accumulated in the \(\delta p^+\) well to lower the barrier. Also owing to the accumulated holes lowering the barrier, the switching currents in the initial off-state (\(I_{sz}\)) and intermediate on-state (\(I_{sz}\)) are larger under illumination. The smaller holding voltages and holding currents in the intermediate on-state (\(V_{ui},I_{ui}\)) and final on-state (\(V_{sz},I_{sz}\)) under illumination are due to the internal gains increased by the photogenerated holes accumulated in the \(\delta p^+\) well.

In addition to the switching characteristics controllable by the input-light, we can observe the optical switching by biasing the STOS to just below turn-on and then introducing light to turn the STOS on, or by biasing the STOS to just before turn-off and removing the light source to turn the STOS off. Figure 3 shows the I-V characteristics of the STOS with a superimposed load line for both the dark (solid line) and illumination (dotted line) conditions. In the dark, the load line intersects the characteristics at A only. This means that the STOS is stable in the initial off-state in the dark. In the illumination case, it is seen that the stable point is at B in the intermediate on-state and is at C in the final on-state. The STOS will make a transition from A to B or C promptly when light falls on the STOS. As the light intensity drops, the characteristics revert to the dark situation and the STOS switches back to A. Therefore, the STOS can be switched on and off with an optical input.

---

\(\delta\) This work was supported by the National Science Council of the Republic of China under Contracts NSC 95-2221-E-013-004 and NSC 95-2221-E-017-013.
Figure 1. Band diagram of the STOS with a positive collector-to-emitter voltage $V_{CE}$.

Figure 2. Experimental current-voltage characteristics of the STOS under (a) dark and (b) illumination.

Figure 3. Current-voltage characteristics of the STOS with a superimposed load line for both the dark (solid line) and illumination (dotted line) conditions.
Electrorefractive Effect in InGaAsP Five Layer Asymmetric Coupled Quantum Well (FACQW) for High Performance Optical Modulators

Masayasu Fukuoka, Takahiro Toya, Taro Arakawa, and Kunio Tada*
Graduate School of Engineering, Yokohama National University, 79-5 Tokiwadai, Hodogaya-ku, Yokohama 240-8501, Japan, Tel: +45-339-4143, Fax: +81-45-338-1157, Email: arakawa@ynu.ac.jp
* Kanazawa Institute of Technology

Abstract
A new InGaAsP FACQW for 1.55 μm was proposed and its electrorefractive (ER) effect was analyzed. The FACQW is expected to show large ER sensitivity |dn/dF|, and is promising structure for high-performance optical modulators and switches.

1 Introduction
For semiconductor optical modulators and switches such as Mach-Zehnder modulators, large electrorefractive (ER) index change Δn with a small absorption loss is necessary in quantum wells (QWs). A five-layer asymmetric coupled quantum well (FACQW)[1] is one of the very promising candidates for producing giant electrorefractive index change. For example, giant ER sensitivity |dn/dF| as large as 1.7 x 10⁻⁴ cm/kV was observed in GaAs/AlGaAs FACQWs [2]. For 1.55 μm wavelength region, InGaAs/InAlAs FACQWs was proposed and their ER characteristics were theoretically studied [3]. InGaAsP systems are also frequently utilized for lasers, and electroabsorption modulator, etc. as Al-free materials.

In this paper, we propose a new In₀.₉₅Ga₀.₀₅As₀.₁₁P₀.₈₉/In₀.₅₈Ga₀.₄₂As₀.₉₀P₀.₁₀ FACQW for 1.55μm wavelength operation. Its ER effect was theoretically studied with Luttinger-Kohn Hamiltonian and non-variational method.

2 Valence Band structure of InGaAsP FACQW
The proposed FACQW is composed of 18-monolayer (ML) In₀.₉₅Ga₀.₀₅As₀.₁₁P₀.₈₉ (lattice-matched to InP) QW (QW1) and 20ML(=6+14)-In₀.₉₅Ga₀.₄₂As₀.₉₀P₀.₁₀ QW (QW2) with In₀.₅₈Ga₀.₄₂As₀.₉₀P₀.₁₀ 5 ML barrier layer inserted for potential modification as shown in Fig.1. The two QWs couple with each other through a 7-ML InAlAs barrier layer at the center. The wavefunction of HH (heavy hole) 1 is distributed dominantly in QW1, and that of HH2 is distributed dominantly in QW2. On the other hand, the wavefunctions of E(electron)1 and E2 are distributed equally in both wells. As a result, when the electric field |F| is slightly increased, the exciton absorption strengths and binding energies of E1-HH1 and E2-HH2 remarkably decrease while those of E1HH2 and E2HH1 increase. This large change contributes to the very large absorption coefficient change Δα in the FACQW for transverse electric (TE) mode as shown in the next section. In addition, the wavefunction of each band with wave number k ≠ 0 also contribute to the ER effect of the FACQW, therefore their influences were included in the analyses.

We analyzed valence band structures by solving the Schrödinger equations using kp perturbation theory with 4x4 Luttinger-Kohn Hamiltonian[4].

Figure 2 shows dispersion of the valence band of the proposed InGaAsP FACQW. The Dispersion curves of an InGaAsP rectangular quantum well (RQW) are also shown for comparison. The ground hole levels of the FACQW and the RQW have almost the same energies, but hole subbands are densely formed in the FACQW due to the coupled quantum well structure. Therefore, the influence of higher subbands in the FACQW is more significant than that in RQWs.

Fig.1 Schematic diagram of InGaAsP FACQW for 1.55μm wavelength. Absorption edge is 1.45 μm.
3 Electrorefractive Characteristics of InGaAsP FACQW

The absorption coefficients of the FACQW under an electric field were calculated with the non-variational approach referred in Ref.[5]. In the calculation, the dispersions of the ground state and the first excited state were taken into account. Figure 3 shows the calculated absorption coefficient spectra under various electric fields. With the increase of the electric field, the peak for E2-HH1 and the combined peak of E1-LH1 and E1-HH2 increase. Through the Kramers-Kronig relation, the large absorption coefficient arising brings about a large electrorefractive index change in the transparent wavelength region[1]. The dependence of $\Delta n$ of the FACQW at 1.55$\mu$m (120 nm from the absorption edge) on an applied electric field was shown in Fig. 4. The $\Delta n$ of a 18ML-thick InGaAsP RQW is shown for comparison. As shown in the figure, the FACQW produces a large $dn/dF$ (approximately $4.6 \times 10^{-4}$ cm/kV at $F=35-45$ kV/cm) though the RQW shows a quadratic dependence on $F$. This $dn/dF$ value of the FACQW is 5-10 times as large as that of the InGaAsP RQW, and is as large as that of InGaAs FACQW[3]. In addition, this large $dn/dF$ value can be obtained over 100 nm wavelength range. These characteristics are very promising for low-voltage and wide-band optical modulators and switches. In particular, low-voltage operation is suitable for ultrahigh speed operation which is limited by a driver IC.

4 Conclusion

A new In$_{0.95}$Ga$_{0.05}$As$_{0.11}$P$_{0.89}$/In$_{0.58}$Ga$_{0.42}$As$_{0.90}$P$_{0.10}$ FACQW for 1.55$\mu$m wavelength region was proposed. Valence band structures of the FACQW were analyzed with Luttinger-Kohn Hamiltonian based on $k$-$p$ perturbation theory. Electrorefractive effect was calculated using non-variational method considering $k \neq \sigma$ dispersions. The InGaAsP FACQW is expected to show very large ER sensitivity $|dn/dF|$ in a wide transparency wavelength region far from the absorption edge. The FACQW is expected to realize ultra-wideband, ultrafast optical modulators and switches.

References

Wavelength Converter Using an Optical MEMS Switch

Jaemyoung Lee¹, Joon-Hak Bang²

(¹Korea Polytechnic University 2121 Jungwang Shihung Kyunggi Korea(ROK) 429-793, ²Electronics and Communications Research Institute 161 Gajeong-dong Yuseong-gu, Daejeon, Korea(ROK), 305-350, Tel: +82-31-8041-0483, Fax: +82-31-8041-0499, Email: lee@kpu.ac.kr)

Abstract

We propose a wavelength converter using an optical MEMS switch. The proposed scheme can convert a wavelength to another through the optical space switch.

1 Introduction

In order to cope with the recent explosive increase in IP traffic, an innovative network system that makes use of new technology is required. For efficient signal traffic, optical communications employ a virtual wavelength allocation scheme in allocating a wavelength on a signal, in which the wavelength of a signal can be changed on a link-by-link basis [1-3]. To realize the virtual wavelength allocation scheme, each node should be equipped with wavelength converters which require tunable sources.

Although performance of wavelength tunable light sources, such as a tuning range and tuning speed, has been considerably improved, current optical communications require better performance for WDM communications. The tuning range of light sources is not wide enough, and the overall tuning speed is limited by slow wavelength locking speed to stabilize a wavelength within a certain range. Wavelength conversion speed of a wavelength converter is determined not by the tuning speed of a tunable source but the locking speed, because the performance is inherently limited by the slowest constituent element.

The proposed scheme of wavelength conversion consists of M light sources and an optical space switch, as shown in Fig. 1. Each light source has a unique fixed wavelength. A signal wavelength is converted to the wavelength of a probe beam by conducting the probe beam to a wavelength converter through a space switch.

Thus, the switch speed determines the wavelength conversion speed of a signal. In the proposed scheme, the switching time of a probe beam wavelength from one wavelength to another can be less than 10 ms using a commercially available optical space switch. Since the proposed scheme uses wavelength-fixed light sources, it does not need the wavelength locking control after changing a probe beam wavelength through the space switch, and it can cover a wide wavelength conversion range by simply employing light sources with corresponding wavelengths.

2 Experiment for feasibility test

Since the proposed scheme provides different probe beam power depending on optical paths taken in an optical space switch from a light source to an input port of a wavelength converter due to insertion loss of the optical space switch, we investigated allowable optical power deviation at output ports of an optical space switch for wavelength conversions in cross gain modulation (XGM) with SOAs [4,5]. In experiment, directly modulated input signal at 1557.3 nm was converted to 1555.3 nm by XGM at a bit rate of 2.5 Gb/s with nonreturn-to–zero pseudorandom binary sequences (PRBS) of length 2²³⁻¹. The input signal powers were -1 and 3 dBm, and the optical power of the probe beam changed from −10 to 5 dBm. Figure 2 displays the measured power penalties of wavelength-converted signals at 1555.3 nm as a function of probe beam power, and shows that the allowable probe beam power fluctuation to afford a 2 dB deviation in a power penalty from the minimum power penalty is larger than ±4 dB. This result indicates that the maximum tolerable optical
power difference between the minimum and the maximum insertion loss of the switch is larger than ±4 dB. That is, the proposed scheme can afford 2 dB deviation in power penalty as long as insertion loss difference between the maximum and the minimum losses of a space switch is within about 8 dB.

3 Management of MEMS switch input/output ports

An optical signal passing through an optical space switch experiences an optical insertion loss. For the 8 × 8 optical MEMS switch we investigated, the insertion loss ranges from 0.70 to 2.30 dB at 1550 nm. Since the performance of our proposed scheme depends more on the insertion loss difference between maximum and minimum losses rather than on the maximum insertion loss for a space switch, we investigated the insertion loss differences for each output port. Table 2 shows the insertion loss difference between maximum and minimum losses for each output port. For the case of 1*, a signal propagates from the plane 2 to the plane 1, while a signal travels from the plane 1 to the plane 2 for the case of 2*. Case 1* reveals less insertion loss difference comparing with the case 2*. It means that case 1* is better suited to our proposed scheme, because it could generate less power penalty.

4 Conclusion

A simple wavelength conversion scheme is proposed, which can provide improved performance characteristics in stability, conversion range of wavelengths and channel spacing. Wavelength conversion experiment based on the cross-gain modulation shows that the proposed method allows an 8 dB difference in probe beam power for a 2 dB power penalty deviation. By properly setting the input and output planes for an optical space switch, this scheme can show less power penalty.

5 References

Expanding the Bandwidth of Slow and Fast Pulse Propagation in Coupled Micro-resonators

David D. Smith\textsuperscript{a,b} and Hongrok Chang\textsuperscript{b}

\textsuperscript{a}NASA Marshall Space Flight Center, EV43, ISHM and Sensors Branch, Spacecraft and Vehicle Systems Department Huntsville, AL 35812, USA, Phone: 256-544-7778, Fax: 256-544-8762, Email: david.d.smith@nasa.gov

\textsuperscript{b}University of Alabama in Huntsville, Department of Physics, 201B Optics Bldg., John Wright Drive, Huntsville, AL 35899, USA

Abstract—Coupled resonators exhibit coherence effects which can be exploited for the delay or advancement of pulses with minimal distortion. The bandwidth and normalized pulse delay are simultaneously enhanced by proper choice of the inter-resonator couplings.

I. INTRODUCTION

The strongly modified dispersion associated with whispering-gallery-mode resonances in coherently coupled micro-resonators can be used to slow or advance pulses of light with minimal loss and pulse distortion, via effects such as coupled-resonator induced transparency [1] and gain-assisted superluminality [2], analogous to effects in atomic systems [3, 4]. These structures are therefore promising for applications such as all-optical buffers and delay lines [5-8], differential sensing and laser gyroscopy [9-11], high fidelity image processing [12], and optical computing schemes [13], but tend to suffer due to their small bandwidth and normalized pulse time delay. Recently attention has focused on expanding the useable bandwidth of slow and fast light materials [7, 14, 15]. In this paper, we demonstrate that proper choice of the couplings between resonators can simultaneously increase bandwidth and pulse delay, resulting in devices that can better accommodate the high bit content and bandwidth demands of modern communications systems.

\begin{equation}
\tilde{\tau}_2(\phi, \phi_i) = \frac{r_2 - a_2 e^{i\phi}}{1 - r_2 a_2 e^{i\phi}} = r_2 \exp\left[i \phi^{\text{eff}}_2\right],
\end{equation}

where

\begin{equation}
\tilde{\tau}_1(\phi, \phi_i) = \frac{r_1 - a_1 e^{i\phi}}{1 - r_1 a_1 e^{i\phi}} = r_1 \exp\left[i \phi^{\text{eff}}_1\right]
\end{equation}

is the complex transmittivity through the first resonator, $\phi_j = 2\pi n_j L_j / \lambda_n$ are the single-pass phase-shifts, $r_j$ are the coupler reflection coefficients, $a_j = \exp(-\alpha_j L_j / 2)$ are the single-pass attenuation coefficients, $n_j$ are the refractive indices, $\alpha_j$ are the loss coefficients, $L_j$ are the circumferences of the resonators, and $j = 1, 2$ specifies the first (furthest from the excitation waveguide) or second (closest to the waveguide) resonator. We will assume the resonators have identical optical path lengths (they are co-resonant) so that we can drop the subscript from the single-pass phase shifts. The transmittance is given by $T_2(\phi) = \tilde{\tau}_2^2$.

If the complex transmittivity $\tilde{\tau}$ is expressed as a phasor, i.e., $\tilde{\tau} = r \cos\phi^{\text{eff}} \tilde{\tau} + \tau \sin\phi^{\text{eff}} \hat{\tau}$, then the response can be plotted on an Argand diagram. It can be shown that extremes in transmission occur when

\begin{equation}
\frac{d\tilde{\tau}}{d\phi} \frac{d\phi^{\text{eff}}}{d\phi} = 0,
\end{equation}

i.e., when $\tilde{\tau} \perp d\tilde{\tau}/d\phi$, while dispersion reversals occur when

\begin{equation}
\frac{d\phi^{\text{eff}}}{d\phi} \frac{d\tilde{\tau}}{d\phi} = 0 \text{ or } \infty,
\end{equation}

i.e., when $\tilde{\tau} \parallel d\tilde{\tau}/d\phi$, or when the structure is critically-coupled to the excitation waveguide, i.e., $T = 0$. The derivative $d\phi^{\text{eff}}/d\phi$ (the phase time) thus determines the amount of pulse advancement or delay that occurs. The difference in transmittance at $\phi = 0$ and at the extremes of Eq. 3, $\phi = \phi^*_m$, determines whether induced
transparency or induced absorption occurs on-resonance. The transition between the two effects occurs at 
\[ \Delta T = T_z(0) - T_z(\phi_{\text{crit}}) = 0, \]
which corresponds to
\[ \tau_z = \frac{2\rho_z (1-a_2'^2)}{2a_z \tau_z (1-a_2'^2) + a_z (1-a_2') (1+\tau_z^2)}. \]
(5)

At this value of the inter-resonator coupling, neither induced transparency nor induced absorption occur, and instead the transmission spectrum is flattened as shown in Fig. 2. Moreover, for two resonators there are two loops, rather than one, in the Argand diagram, and the inner loop forms an almost perfect circle, indicative of the flattened spectrum. One can readily verify that 
\[ \tau_z(0) = \tau_z(\phi_{\text{crit}}) \]
from this graph because \( \phi(\tau_z) \approx \pi \) for this particular case. The broadening of the spectrum is also accompanied by an increase in the time delay, as evidenced by the increased slope \( d\phi(\tau)/d\phi \) in the case of the coupled resonators.

![Fig. 2. Spectral response for a single resonator (dashed lines), and two coupled resonators (solid lines). Top: transmission Argand diagrams. The plots are parametric, starting on-resonance (\( \phi = 0 \)), as indicated by the dot, and proceeding counterclockwise around the loop over a range of 2\pi. Bottom: spectra of the transmission and phase.](image_url)

Finally, we note that this same principal can be used to broaden and flatten the spectrum when the resonators incorporate an amplifying medium. In this case, the dispersion is anomalous and a pulse advancement rather than delay occurs.

III. CONCLUSION

The response of two coupled resonators exhibits coherence phenomena such as induced transparency and absorption and gain-assisted superluminality in analogy with multilevel atomic systems, which can be exploited for the delay or advancement of pulses in optical systems with minimal pulse distortion. Proper choice of the inter-resonator coupling can simultaneously increase the bandwidth and the normalized pulse delay (or advancement) which is crucial to increasing the storage capacity of these devices for the resolution of data packet contention issues in communications systems.

ACKNOWLEDGMENTS

The authors acknowledge support from the Institutional Research and Development Fund at Marshall Space Flight Center, the United Negro College Fund, and the NASA Administrator’s Fellowship Program.

REFERENCES

Novel 2x2 Photonic Switch Based on Multimode Interference Effect
Yu Narita, Mitsuhiro Yasumoto, and Hiroyuki Tsuda
Department of Electronics and Electrical Engineering, Keio University,
3-14-1 Hiyoshi, Kouhoku-ku, Yokohama, Japan
Phone:+81-45-563-1151 Facsimile:+81-45-566-8529 E-mail:narita@tsud.elec.keio.ac.jp

Abstract—We propose a 2x2 photonic switch that uses the multimode interference effect (MMI) and analyze its properties by the finite difference beam propagation method (FD-BPM). We then consider the characteristics of several well-known matrix switch architectures based on our proposed 2x2 switches.

I. INTRODUCTION

Photonic switches using the multimode interference (MMI) effect have many potential advantages such as compactness, suitability for integration, low polarization and wavelength dependence, and good fabrication tolerance [1], [2]. However, conventional MMI photonic switches have large crosstalk and are bulky. In this paper, we have proposed a novel 2x2 MMI photonic switch with parallel electrodes in its multimode waveguide regions to improve these issues. In addition, the losses of matrix switches using the proposed 2x2 switches have been estimated.

II. Device Structure and Principle

The structure and operating principles of the proposed switch are shown in Fig. 1. It has electrodes in the MMI waveguide which cause an index reduction when a voltage is applied to them. The length of the MMI waveguide for the photonic switch is set to be 3Lg. As shown in Fig. 1(a), the photonic switch guides the light from port 1 to port 4 when no voltage is applied to electrodes 1 and 2 (cross state) [3]. Here, Lg is defined as follows

$$L_g = \frac{4n_r W_{out}^2}{3\lambda_0}.$$  

where \(\lambda_0\) is the wavelength, \(n_r\) is the core effective refractive index, and \(W_{out}\) is the effective width of the MMI waveguide. On the other hand, as shown in Fig. 1(b), the input light into port 1 is switched to port 3 (bar state) when a voltage is applied to electrodes 1 and 2 because the effective positions of the MMI waveguide side boundaries are formed just under electrodes 1 and 2. The distance between those electrodes, \(W_{in}\), satisfies the following equation:

$$L'_g = \frac{4n_r W_{in}^2}{3\lambda_0},$$

where \(L'_g\) is the length of the MMI region.

III. ANALYSIS

The fundamental switching characteristics of the MMI photonic switch were analyzed by the finite difference beam propagation method using the (3, 3) Pade approximation. The material that forms the switch is assumed to be a III/V compound semiconductor in order to attain high-speed switching and a large refractive index change. The plasma effect with carrier injection may be used. The calculation was carried out for light in TE mode, with a wavelength of 1550 nm. Figure 2 shows the parameters of the 2x2 MMI switch that was used for the analysis. Taper waveguides were used as the input.
and output waveguides to prevent any large-angle
diffraction of the light into the multimode waveguide
and to reduce light leakage at the electrodes. It was
assumed that the maximum refractive index change
that could be induced by applying a voltage, $\Delta n$, was
-0.03, and that the absorption loss was 5 cm$^{-1}$. The
length and width of the switch were 1982 and 15.0
$\mu$m, respectively. It should be noted that the device
size would be much smaller when the maximum
refractive index change is large. It is because that the
tapered input and output waveguides are not
necessary and the width and the length of the MMI
region can be reduced. Figure 3 shows the output
powers from port 2 and port 3 as a function of the
index change when the input power from port 1 was 0
dBM. It had insertion losses of 5.0 and 5.6 dB, and
cross-talk of -45.4 and -46.6 dB for the cross state
and the bar state, respectively. The insertion losses
and cross-talk are largely dependent on the width of the
MMI waveguide and the distance between the
electrodes rather than the length of the MMI
waveguide.

IV. MATRIX SWITCH ARCHITECTURE

The characteristics of a matrix switch fabricated
using our proposed 2x2 MMI photonic switches
should be considered for practical applications in a
photonic network. In this section, we discuss six
well-known matrix switch architectures: the optical
crossbar, the N-stage Planar, the double crossbar, the
Benes, the three-stage Clos, and the extended
baseline [4].

The insertion losses for the different architectures
are shown in Table 1. $\alpha$ is the insertion loss of the 2x2
switch in dB.

<table>
<thead>
<tr>
<th>Network</th>
<th>Insertion Loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>Crossbar</td>
<td>2(N-1)$\alpha$</td>
</tr>
<tr>
<td>N-Stage Planar</td>
<td>N$\alpha$</td>
</tr>
<tr>
<td>Double Crossbar</td>
<td>(N+1)$\alpha$</td>
</tr>
<tr>
<td>Benes</td>
<td>(2log$_2$N-1)$\alpha$</td>
</tr>
<tr>
<td>Three-Stage Clos</td>
<td>(2n+2m+2r-3)$\alpha$</td>
</tr>
<tr>
<td>Extended Baseline</td>
<td>(3log$_2$N-2)$\alpha$</td>
</tr>
</tbody>
</table>

Figure 4 shows the insertion losses when $\alpha = 5.6$
dB. According to Ref. [4], the maximum attenuation
allowed without amplification or regeneration was
assumed to be 30 dB. The maximum matrix size
should be 8x8, even with the Benes architecture, in
order to satisfy that condition.

V. Conclusion

We proposed a novel 2x2 MMI photonic switch. Analysis
of its properties shows that it could achieve very
low cross-talk of less than -40 dB. The insertion
loss characteristics of several matrix switches
composed of the proposed 2x2 switches were
compared, and it was shown that the maximum
matrix size was 8x8 for the proposed switch without
optical amplification.

REFERENCE

“Multimode Interference Photonic Switches (MIPS),”
675-681(2002).
“Optical Switch Based on Multimode Interference
[3] Lucas B. Soldano and Erik C. M.Pennings,
“Optical Multi-Mode Interference Devices Based on
Self-Imaging: Principles and Applications,” IEEE J
optical space switching systems,” IEEE Commun.
Micro-Spherical Refractive Index Variation Sensor Using High Refractive Index Layer

A. Kobayashi, J. Osawa, H. Okayama and H. Nakajima
Waseda University, 3-4-1 Okubo, Shinjuku-ku, Tokyo 169-8555 Japan
Tel: +81-3-5286-3223, Fax: +81-3-5287-7056, Email: koba@pic.phys.waseda.ac.jp

Abstract

We describe the effects of radius, layer thickness and refractive index on the sensor sensitivity when cylinder/sphere is coated with a thin high refractive index layer.

1. Introduction

Recently the Surface Plasmon Resonance (SPR) is the most commonly used optical sensing scheme to measure antigen-antibody reactions and other chemical reactions, due to its high sensitivity to the surface refractive index change. However, SPR requires precious metals to excite surface plasmon. In this report, to realize an optical sensor, we use an optical resonance of the surface guided light mode which propagates on a microsphere surface. The resonator responds effectively to the exterior environmental variation because it has high Q value and traps light inside the sphere much longer than other optical resonators. When the sphere is coated with a thin layer of higher refractive index [1] to confine the light, the resonator becomes more sensitive to the sample on the surface [2].

In this paper, we describe the effect of the radius, the layer thickness and effective index variations of the sphere coated by a high refractive index layer. The refractive index of high refractive index layer is varied within 1.5 ~ 3.5 and we select the sphere radius as a parameter. We also study the sensor sensitivity by the BPM simulation (a cylinder cross section).

2. Device design

An example of the device structure is shown in Fig. 1. This device is composed of a microsphere (n=1.46) and V groove base material (n=2.1). The incident light hits the surface of microsphere from a base material under total reflection condition. We observe the resonated light in the sphere which escapes from the base material. This structure enables us to observe only the resonated light while non-resonated light penetrates upward.

There are two methods to evaluate the sensor performance against the sample refractive index variation. One method is to measure the shift of peak wavelength and the other is observing the output light power change.

The former method is very useful because the effective index variation is directly related to the sensitivity. In this study, we have found that we can obtain higher sensitivity when the radius is smaller. The latter method observes the optical power change which can be tailored by designing the resonator structure. We can enhance the sensitivity over that can be obtained directly by the effective index variation. We can obtain higher sensitivity when the radius is larger by the latter than the former method.

3. Simulation

We have studied the effect of the radius and the high refractive index layer. We first calculated a rate $dn_{eff}/dn$, the ratio of the effective index variation over the sample refractive index variation, when the sample refractive index is changed by 0.005. Fig. 2 shows the relationship between the refractive index and the layer thickness at $\lambda=0.633\mu m$ wavelength and for radius of $R=10, 50$ or $100\mu m$. The fundamental mode is excited inside the cylinder/sphere.

In Fig 2 the variation rates of effective index as a function of the layer refractive index and thickness are shown by contour lines. When the radius is smaller, the effect of layer coating is larger.
However, if the radius is too small, there is a range where guided mode cannot be excited. We also confirmed that at larger radius, there was not much difference in effective index variations.

Fig. 3 shows the effects of high refractive index layer when observing the intensity change of output power. Incident light travels around the sphere circumference, and then we observe the light which is coupled out from the incident position. Since this process is repeated, the resonator’s function is similar to the Fabry-Perot resonator. The coupling efficiency is 70%, at $\lambda=0.633\mu m$ wavelength, for a radius of $R=100\mu m$, and the sample refractive index of $n=1.33$. The output power change of a microsphere coated with a 0.5$\mu m$ thick layer is shown in the figure. The refractive indices of the coated layer are $n_t=1.76$, 2.5 and 3.45. As expected, the output curve becomes shaper when the microsphere is coated; therefore it is reasonable to say that coating is effective. However, when the refractive index of the layer is small, the layer is needed to be thick to make appreciable difference on the output power response. Also in this case, we confirmed that there was not much change on output power at different radii and layer thicknesses.

4. Conclusion
In this paper, we have described the effects of the radius, layer refractive index and thickness dependency on the sensor sensitivity when cylinder/sphere is coated with a thin high refractive index layer. When observing the effective index variation, the sensitivity is higher at smaller radius. The output power change is more sensitive when the refractive index of the layer is higher and the radius is larger due to a longer resonator length.

References
Formation of MgO:LiNbO$_3$ domain-inverted grating for quasi-phase matching by voltage application with insulation layer cladding

Takashi Tsubouchi, Nobuyuki Horikawa, Masatoshi Fujimura and Toshiaki Suhara
Dept. Electron. Eng., Grad. School Eng., Osaka Univ.  2-1 Yamada-oka, Suita, Osaka, 565-0871, Japan
Phone:+81-6-6879-7772  Fax:+81-6-6879-7793  Email:ttsubouchi@ioe.eei.eng.osaka-u.ac.jp

Abstract: We demonstrate a new method for formation of domain-inverted gratings in MgO:LiNbO$_3$ for quasi-phase matched nonlinear-optic devices. It was found that domain-inverted gratings could be obtained by application of voltage pulses with SiO$_2$ insulation layer cladding, and quasi-phase matched second harmonic generation was demonstrated.

LiNbO$_3$ is often used for implementation of quasi-phase matched (QPM) nonlinear-optic (NLO) devices$^1$ because of the large NLO coefficients. However, photorefractive damage (optical damage)$^2$ caused by propagation of a short-wavelength beam may make stable operation rather difficult. An effective method is to use MgO-doped LiNbO$_3$ (MgO:LiNbO$_3$), which exhibits high resistance to the damage.$^3$ Several techniques for formation of ferroelectric-domain-inverted gratings for QPM in MgO:LiNbO$_3$ have been reported$^4-8$. However, the techniques have not been fully established. In this paper, we report experimental insights leading to a new method for QPM grating formation in MgO:LiNbO$_3$ by voltage application with insulation layer cladding.

We first tried to fabricate domain-inverted gratings using conventional electrodes. Z-cut LiNbO$_3$ crystals of 0.5mm thickness doped with 5mol% of Mg were used. Periodic and uniform metal-film electrodes were formed on the +Z and –Z faces, respectively. The crystal was heated to 170°C in silicone oil, and voltage pulses (typically 1.8kV) were applied. Inverted domains were observed only below the outer edges of the periodic electrode fingers, i.e., the obtained structure was domain-inverted dots (see Fig.3(b)) rather than a domain-inverted grating. A large leakage current$^8$ was observed during the voltage application. It seemed that electric resistivity became low in the domain-inverted dot regions. Formation of the low-resistivity (semi-conducting) domain-inverted shoots gives rise to electric field concentration and completion of domain-inverted dot formation. However, the resultant short circuit would have prevented effective application of electric field to the crystal. To examine the leakage current phenomenon, the change of leakage current after the domain inversion was measured by applying 100V probe pulses. The measured current $I$ was converted into the resistivity $\rho = \frac{SV}{dI}$ with the probe voltage $V = 100V$, crystal thickness $d = 0.5mm$ and inversion area $S = 1.5mm^2$. Figure 1 shows the dependence of resistivity of domain-inverted region on elapsed time after application of a 1.8kV inversion pulse. The resistivity of domain-inverted region recovered over time. This result suggests that the leakage current can be reduced by applying inversion pulses with long intervals.

To avoid formation of the short circuit through the inverted shoots, we inserted a thin insulation layer between the crystal and the uniform electrode. The flow of inversion current and leakage current charges the insulation layer capacitance and reduces the voltage across the crystal. However, effective voltage application can be repeated many times by using multiple pulses.

Figure 2 shows the setup for fabrication of domain-inverted gratings by voltage application with insulation layer cladding. As an insulation layer, a SiO$_2$ film (thickness $\sim$2μm) was deposited by RF sputtering on the –Z face of the crystal. A uniform metal(Au)-film electrode was formed on the SiO$_2$ film. On the +Z face, a periodic metal(Al)-film electrode of the period $\Lambda = 18\mu m$, finger width $w = 6\mu m$, electrode length 10mm and finger length 1mm was formed. The crystal was heated to 170°C in silicone oil, and voltage pulses were applied. Figure 3 (a) and (b) show the structure (after etching) formed on +Z face by application of 100 pulses

![Fig.1 The dependence of resistivity of domain-inverted region on the time after an inversion pulse application.](image-url)
of effective voltage across the crystal $V_{\text{eff}} = 1.4 \text{kV}$, pulse width $\tau = 0.25 \text{ms}$ and pulse period $T = 10 \text{s}$ with and without the insulation layer cladding, respectively. The inverted regions of (a) are longer than those of (b). Current observed during the voltage application with insulation layer was lower than that without insulation layer. The result shows that formation of the short circuit was avoided by the insulation layer.

We tried to increase more the length of domain-inverted lines by increasing $V_{\text{eff}}$, $\tau$ and pulse repetition. The attempt, however, was not successful, since the domains were extended not only in the length but also in the width even below the electrode finger gaps. It turned out that the situation could be improved by depositing another SiO$_2$ layer over the periodic electrode. Figure 4 shows the structure formed on the $+Z$ face by voltage application of 100 pulses of $V_{\text{eff}} = 1.4 \text{kV}$, $\tau = 100 \text{ms}$ and $T = 60 \text{s}$ with 2μm-thick SiO$_2$ layer on the $-Z$ face and 1μm-thick SiO$_2$ layer over the periodic electrode on the $+Z$ face. A periodic electrode with the same period $\Lambda = 18 \mu m$ and narrower finger width $w = 3 \mu m$ was used. High-quality domain-inverted gratings were obtained over the entire electrode area as shown in Fig.4(a). The cross section of the structure observed after dicing the crystal and etching is shown in Fig.4(b). Domain-inverted regions continuing from $+Z$ face to $-Z$ face were obtained.

A second harmonic generation (SHG) experiment with domain-inverted gratings of period $\Lambda = 18 \mu m$ and interaction length $L = 10 \text{mm}$ formed by this method was performed by using a tunable InGaAsP laser and an Er-doped fiber amplifier as a pump source. QPM-SHG was achieved at a pump wavelength of 1553nm, with FWHM bandwidth of 1.4nm which is close to the theoretical prediction of 1.2nm.

In conclusion, we have presented a new method for fabrication of domain-inverted gratings in MgO:LiNbO$_3$ by voltage application with insulation layer cladding. We obtained high-quality domain-inverted gratings suitable for quasi-phase matching in nonlinear-optic devices.

DMD Evaluation of Rectangular Optical Waveguide for 10Gbps transmission with Numerical Analysis

Yoshitsugu WAKAZONO, Atsushi SUZUKI, Daisuke NAGAO, Takaaki ISHIKAWA, Tomoyuki HINO, Yoichi HASHIMOTO, Hiroshi MASUDA, Shuji SUZUKI, Mitsuaki TAMURA, Teiichi SUZUKI, Katsuya KIKUCHI, Hiroshi NAKAGAWA, Masahiro AOYAGI and Takashi MIKAWA

National Institute of Advanced Industrial Science and Technology (AIST)
AIST Tsukuba Central 2, 1-1-1, Umezono, Tsukuba, Ibaraki, 305-8568, Japan
phone:+81-861-5080(ext55480), fax+81-862-6511, E-mail:yoshitsugu.wakazono@aist.go.jp

Abstract
We calculated DMD to evaluate the transmission performance of the rectangular optical waveguide represented by polymer waveguide, considering the coupling efficiency of every mode. We revealed that DMD was small enough to apply this kind of optical waveguide for 10Gbps transmission and DMD strongly depended on the broadening of excited modes in the waveguide.

1. Introduction
Optical interconnection for board-to-board and chip-to-chip is promising solution for high-density and high-speed data transmission inside box. The feasibility demonstrations of optical interconnection for 10Gbps transmission have recently been reported using step index polymer waveguide having rectangular core over 0.3-1m length [1][2]. These experimental studies indicate polymer waveguide is attractive as a transmission media at least for shorter interconnect regime around 1m.

The transmission characteristics of SI(Step Index)-profile waveguide has been roughly estimated by DMD (Differential Mode Delay) derived from the relative index difference between core and cladding. In this estimation, DMD is overestimated, resulting in the poorer bandwidth of the waveguide, because only the fastest mode and the slowest mode are considered. In this paper, we considered the coupling efficiency for every mode in the SI rectangular waveguide represented by polymer waveguide and we calculated DMD to evaluate the transmission performance rigorously.

2. DMD evaluation procedure
2-1 Theoretical model of electro-magnetic fields in optical waveguide
The electro-magnetic fields $E$ and $H$ in optical waveguide follow the Maxwell’s equation. In this study, the waveguide extend along z axis and the electro-magnetic fields are assumed to propagate as plain wave along z axis. Therefore, $E$ and $H$ are described as follows:

\[ E = \tilde{E}(x,y) \exp(j(\omega t - \beta z)) \]
\[ H = \tilde{H}(x,y) \exp(j(\omega t - \beta z)) \]

where $\omega$ is angular frequency ; and $\beta$ is propagation constant along z.

We solved the Maxwell’s equation and calculated the propagation constants with using the assumption proposed by Marcatili[3]. Figure 1 shows the cross-sectional view of a waveguide. According to the assumptions, the field components in region 1 vary sinusoidally in x and y direction; those in the 2 and 4 vary sinusoidally along x and exponentially along y, and those in region 3 and 5 vary sinusoidally along y and exponentially along x.

![Figure 1 Cross-sectional view of rectangular waveguide for analysis](image)

In case of the polarization along x, the electro-magnetic fields in the optical waveguide consist of $\tilde{E}_x$ and $\tilde{H}_y$, the magnetic field $\tilde{H}_y$ is described as follows:

\[ \tilde{H}_y(x,y) = \begin{cases} \cos(k_x - \phi) \cos(k_y - \psi)(\text{region 1}) \\ \cos(k_x - \phi) \exp(-\gamma y) \ |1 - a| \cos(k_y - \psi)(\text{region 2 and 4}) \\ \cos(k_x - \phi) \exp(-\gamma y) \ |1 - a| \cos(k_y - \psi)(\text{region 3 and 5}) \end{cases} \]

where $a$ is the half width of the optical waveguide, $k_x$ and $k_y$ are propagation constant along x and y, respectively. $\phi$ and $\psi$ are defined as follows:

\[ \phi = (p-1)\frac{\pi}{2} \cdots (p = 1,2,3\cdots) \]
\[ \psi = (q-1)\frac{\pi}{2} \cdots (q = 1,2,3\cdots) \]

2-2 DMD evaluation
DMD is evaluated with the group delay of each mode in the optical waveguide. In case that group delay of mode $(p, q)$ is $\tau_{pq}$, the DMD is defined as follows[4]:

\[ \text{DMD} = \sqrt{\sum_{p} \sum_{q} a_{pq} (\tau_{pq} - \langle \tau \rangle)^2} \]

where $a_{pq}$ is the coupling efficiency between incident wave and mode $(p, q)$ in optical waveguide ; and $\langle \tau \rangle$ is...
the average of group delay weighted with the distribution ratio of power in optical waveguide.

In this study, the incident wave is described as Gaussian and it is assumed to enter into the optical waveguide as plain wave at z=0.

3. Simulation results

Figure 2 shows the DMD as a function of spot size of incident wave (Gaussian wave). The cross section of optical waveguide was rectangular and its size was 40um x 40um, 60um x 60um and 80um x 80um. In this study, the refractive index of core was 1.5, the relative index difference was 2.0% and the wavelength of incident wave was 0.85um. As the spot size was enlarged, the DMD decreased drastically. Especially, the spot size was set to be between 8 to 18um, DMD was under 1.5ps/m when the core size was 60um x 60um.

As the size of optical waveguide was enlarged, DMD decreased on the whole because the relative scale of core size and spot size of incident wave determined the broadening of excited modes in the waveguide. Considering both DMD and alignment tolerance, large core size is better for 10Gbps transmission.

![Figure 2 DMD in rectangular waveguide as a function of spot size of incident wave (the relative index difference = 2.0%)](image)

We evaluated the coupling efficiency of every mode in optical waveguide to search for the cause of the change of DMD. Figure 3 shows the coupling efficiency of every mode in optical waveguide. In case that the spot size of incident wave was 1um, almost all modes were excited although the coupling efficiency of every mode was less than 0.01. In case that the spot size of incident wave was 10um, only a few modes were excited. Therefore, in the case of 10Gbps transmission with optical waveguide, the reduction of bandwidth caused by mode dispersion caused by mode dispersion was negligible when the length of optical waveguide is less than 1m.

According to the rough estimation, the DMD of Si(Step Index)-profile waveguide is formulated as $n\Delta/c$, where $n$ is the refractive index of core, $\Delta$ is the relative index difference between core and cladding, and $c$ is the velocity of light. DMD in our case was estimated 100ps/m with the above formulation. In comparison with our rigorous evaluation, the DMD of rectangular waveguide with the rough formulation was overestimated.

4. Conclusion

In this paper, we calculated DMD to evaluate the transmission performance of the rectangular optical waveguide quantitatively in consideration of the coupling efficiency of every mode. We revealed that DMD was small enough to apply optical waveguide for 10Gbps transmission at 0.85um wavelength when core size of optical waveguide was 40-80 um square and refractive index of core was 1.5 with refractive index difference of 2.0%. DMD strongly depended on the broadening of excited modes in the waveguide.

References

Cesaro means of Fourier series for designing interleaver filters with planar lightwave circuit-type lattice structure

Juan Zhang
School of Communication and Information Engineering, Shanghai University, Shanghai 200072, China
E-mail: juanzhang_zj@hotmail.com, Phone: 86-21-56332285

Abstract
A simple and effective algorithm, based on the Cesaro means of Fourier series, is proposed for designing Fourier flat-top interleavers. Structural parameters for a PLC-type lattice-form interleaver can be easily obtained with a required spectral response.

1. Introduction
Lattice-form interleavers are based on the interference principle of Mach-Zehnder structure. In practical applications, several lattices are often cascaded to obtain better performance. In terms of cost, reliability and integration, the most attractive configurations of lattice-form interleaver are those on planar lightwave circuit (PLC) designs.

To get optimized structural parameters is a key point in R&D of PLC-type lattice-form interleavers. It can be accomplished by a synthesis algorithm [1]. However, this algorithm is very complex and the waveform flatness is not good enough [2].

In this paper, A simple and effective method, based on Cesaro means of Fourier series, is proposed. All the structural parameters can be obtained by a one-step simple analysis, solving a simple equations-set and correspondence of structure parameters. As an example, a 25GHz 3-lattice-cascaded PLC-type interleaver is designed using this method.

2. Theoretical analysis
2.1 Fourier series of periodic rectangle function

An ideal periodical rectangle function can be expressed by infinite Fourier series. For the ideal periodical rectangle transmittance (as shown in figure 1) with \( \Delta f/\Delta f \) of 1/2, it can be expressed as

\[
T(f) = \frac{1}{2} + \sin\left(\frac{\Delta f}{\Delta f} \right) \cos(2\pi f/\Delta f) + \sin\left(\frac{3\Delta f}{\Delta f} \right) \cos(2\pi 3f/\Delta f) + \ldots
\]

\[
+ \sin\left(\frac{(2N-1)\Delta f}{\Delta f} \right) \cos\left(2\pi (2N-1)f/\Delta f \right) + \ldots
\]

The more the Fourier series, the closer the waveform to the ideal periodical rectangle function. In general, the term N needs up to be hundred.

2.2 Cesaro means of Fourier series

Cesaro means of Fourier series is a method that only a few Fourier series is needed for obtaining a relatively ideal waveform [3]. For the periodical rectangle spectral transmittance with \( \Delta f/\Delta f \) of 1/2, it can be expressed by the Cesaro means of Fourier series with 5 Fourier series as

\[
\tilde{T}(f) = \frac{1}{2} + \frac{2}{16} \cos(2\pi f/\Delta f) + \frac{9}{16} \cos(2\pi 3f/\Delta f) + \frac{3}{16} \cos(2\pi 5f/\Delta f)
\]

which is denoted as the solid line in Fig. 2. The dash line in Fig. 2 is the waveform with 7 Fourier series considered.

2.3 Flattening of spectral transmittance

The spectral transmittance of a birefringent interleaver can be described by Jones Matrix method:

\[
T(f) = T_0 + T_1 \cos(2\pi f/\Delta f) + T_2 \cos(2\pi 3f/\Delta f) + \ldots
\]

\[
+ T_n \cos(2\pi nf/\Delta f) + \ldots
\]

The coefficient value of each cosine series and the constant term in Eq. (3) can be obtained by comparing with the expression of Cesaro means of Fourier series of the rectangle function. The relationship of structural parameters and a structural parameter can be determined through analyzing the expression, the value of the constant term and even term coefficients of cosine series in Eq. (3). Other structural parameters can be directly determined through solving an equations-set, which is composed of the odd term coefficients of cosine series. On the basis of these structural parameters, the structural parameters of lattice-form interleaver can be derived by the mathematical equivalence relationship between structural parameters of the lattice-form and birefringent interleavers.

2.4 Equivalent structural parameters of PLC-type lattice-form interleaver

This mathematical equivalence relationship can be expressed as [4]:

Fig. 2 The waveform corresponding to Cesaro means with five and seven Fourier Series

Technical Digest, July 2007, Pacifico Yokohama
where \( \phi_i \) is the phase factor correspond to the coupling ratio of the \( i \)th coupler, \( \Phi_i \) is the phase delay of the \( i \)th phase delay line. \( \theta_i \) and \( \theta_j \) are, respectively, the azimuth angle of the \( i \)th crystal and polarizer at the output port. \( t_i \) is the phase delay of the \( i \)th crystal. It can be seen that when the structural parameters of one of the two types of interleaver are obtained, the structural parameters of the other can be figured out using Eq. (4) directly.

### 3. Design example

A 25GHz 3-lattice-cascaded PLC-type interleaver is designed as an example. According to the Section 2, we should solve the structural parameters of 3-crystal-cascaded birefringent interleaver firstly, and then the structural parameters of the 3-lattice-cascaded PLC-type interleaver can be obtained using Eq. (4).

The spectral transmittance of the birefringent interleaver calculated by Jones Matrix is expressed as

\[
T(f) = a_1 + a_2 \cos(t_1) + a_3 \cos(t_2) + a_4 \cos(t_3) + a_5 \cos(t_1 + t_2) + a_6 \cos(t_1 + t_3) + a_7 \cos(t_2 + t_3) + a_8 \cos(t_1 + t_2 + t_3)
\]

In order to obtain a desired spectral transmittance, the Eq. (5) should be equalized or approximated.

Comparing Eq. (5) (detailed expression of parameters \( a_{0-13} \) are not given here) with Eq. (2), we obtain that the coefficient of the first, the third, the fifth Fourier series and the constant term in the Eq. (5) should be \( 15/8\pi \), \( 3/8\pi \), \( 3/40\pi \) and 1/2 respectively. And the coefficient of the second, the fourth Fourier series in Eq. (5) equal to zero. In order to satisfy the requirement of the constant term and even series term even, the azimuth angle of the first crystal can be 45° and the ratio of the three crystal thickness can be 1:2:2. The azimuth angle of the second, the third crystal and the polarizer at the output port can be obtained by solving the equations-set:

\[
\begin{align*}
\phi_1 &= \theta_1, \\
\phi_2 &= \theta_2 - \theta_1, \\
\phi_3 &= \theta_3 - \theta_2, \\
\phi_4 &= \theta_4 - \theta_3, \\
\Phi_1 - t_1 &= 0.
\end{align*}
\]

(4)

Solving the Eq. (6), we obtain \( \theta_1=45° \), \( \theta_2=103.49° \), \( \theta_3=-95.59° \), \( \theta_4=0° \) and it is denoted as the curve C in Fig. 3. Curve D is the waveform with \( \theta_1=45° \), \( \theta_2=14.5° \), \( \theta_3=9.93° \), \( \theta_4=0° \) obtained in Ref. [6]. Curve B is the waveform with \( \theta_1=45° \), \( \theta_2=-14.5° \), \( \theta_3=9.93° \), \( \theta_4=0° \) obtained in Ref. [5]. Curve A is the waveform with \( \theta_1=50° \), \( \theta_2=70° \), \( \theta_3=10° \) obtained in Ref. [2].

It can be seen that the isolation in curve C is greater than curve A and B. The flat-top of the curve C is accord with the curve D well. Only isolation is a little lower than the curve D. It shows that the Curve C obtained by Cesaro means of Fourier series can satisfy design requirement well. The results above are obtained by Cesaro means of Fourier series with 5 Fourier series considered. The waveform can be optimized further when 7 Fourier series are considered. At last, the structural parameters of the 3-lattice-cascaded PLC-type interleaver can be obtained using Eq. (4): \( \phi_1=\pi/4 \), \( \phi_2=0.324\pi \), \( \phi_3=-96.165\pi \), \( \phi_4=0.531\pi \).

### 4. Conclusion

Based on the Cesaro means of Fourier series and the mathematical equivalence relationship between structural parameters of the lattice-form and the birefringent interleavers, a new method for designing planar lightwave circuit (PLC)-type lattice-form interleavers was proposed. All the structural parameters can be obtained by a one-step simple analysis, solving a simple equations-set and correspondence of structure parameters. As an example, a 25GHz 3-lattice-cascaded PLC-type interleaver is designed using this method. The parameters obtained can satisfy design requirement well.

### Acknowledgements

This work is partly supported by the Shanghai Leading Academic Discipline Project (NO.T0102) and the Shanghai Municipal Committee of Education (No.05Z4A5).

### References

High Extinction Ratio of Switched Packets by Two-Stage Four Wave Mixing in Highly Nonlinear Dispersion-Shifted Fiber

Rebecca W. L. Fung1, Henry K. Y. Cheung1, C. H. Kwok2 and Kenneth K. Y. Wong1
1: Department of Electrical & Electronic Engineering, The University of Hong Kong, Hong Kong.
2: Department of Engineering, The University of Cambridge, United Kingdom.
E-mail: wlfung@eee.hku.hk and kycheung@eee.hku.hk

Abstract: We demonstrate a simple approach to enhance the extinction ratio of survived packets with two-cascaded four-wave-mixing technique. Improvement of extinction ratio by >5dB is obtained with surviving packets remained at the same wavelength for propagation.

I. Introduction
All-optical packet switching allowing ultra-fast forwarding of data packets in the optical domain, has been exhaustively investigated as a way of overcoming the envisaged future limitations of electronic packet switching technology [1]. In order to fulfill the high demand of the fast-growing telecommunication networks, an optical gating is suggested to be employed for the packet switching to bypass the slow-response transition period of the electrical gating [2]. Packet switching application based on four wave mixing (FWM) techniques can achieve high switching speed up to 100Gb/s [3] but at the expense of the low extinction ratio (ER) between the processed packets and the residue of the dropped packets when a gating signal with finite ER is used. In this paper, we propose a simple approach by cascading two FWM processes to obtain high ER and allow packets to be forwarded at the original wavelength after each processing node.

Extinction ratio of the survived packets plays a crucial role in the performance of the optical packet switched networks [4]. The finite residual power of the dropped packets in the surviving channel may significantly degrade the network performance when other packets are added onto it in the successive nodes. The mechanism of the proposed optical packet switch is shown in Fig. 1. Incoming data packets are chosen to be dropped (at timeslots t2 and t3) in compliance with a gating signal at the switching node which serves as a control pump. The data packets A and C synchronize with the gate-on interval are converted to the idler wavelength after the 1st stage with a finite residual power remained at timeslots t2 and t3. After the 2nd stage, the packets are converted back to the original signal wavelength with enhanced suppression for the dropped packets.

We demonstrate an experimental validation of the proposed packet switch with a 40Gb/s return-to-zero (RZ) signal which undergoes the two-stage processes through a 1km and a 500m highly nonlinear dispersion-shifted fiber (HNL-DSF). Results show that higher than 5dB ER improvement of the dropped signal can be successfully achieved.

II. Experiment
Fig. 2 shows the experimental setup of the optical packet switch.

Fig. 1. Schematic illustration of the proposed optical packet switch with enhanced extinction ratio for the survived packets

Fig. 2. Experimental setup of the optical packet switch.

524
was generated at the original signal wavelength of 1554.35nm which was then extracted out as the final output with a TBPF. A fiber Bragg grating (FBG) was used to further suppress the pump power. Power of the surviving channel was amplified with EDFA 3 and the ASE was removed by applying another TBPF afterward. The filtered output was then sent to the optical spectrum analyzer (OSA) and digital communication analyzer (DCA) to display the spectra and the waveforms after the switching process, respectively.

III. Results and Discussion

![Fig. 3. Measured optical spectra of the pump (1556.91 nm), the signal (1554.35 nm) and the idler (1559.15 nm) at the output end of (a) 1 km fiber, (b) BSF and (c) 500 m fiber, respectively.](image)

![Fig. 4(a) Quasi-square-pulsed pump (b) 40Gb/s RZ input signal.](image)

![Fig. 5. The converted signal at 1559.15nm after the 1st stage of FWM – (a) rising edge (b) falling edge and (c) eye diagram.](image)

![Fig. 6. The converted idler at 1559.15nm after the 2nd stage of FWM – (a) rising edge (b) falling edge and (c) eye diagram.](image)

In practical wavelength division multiplexed (WDM) optical network, each demultiplexed channel links to its assigned processing node in which add/drop multiplexing or packet selection are usually performed. The proposed dual-stage FWM technique can accommodate the survived packets on the same optical wavelength after performing packet dropping at each node for further propagation without wavelength conversion.

IV. Conclusion

We have successfully demonstrated for a simple approach to enhance the ER of the survived packets in a 40Gb/s RZ signal, with a two cascaded FWM technique. An ER of over 24dB with improvement of 5dB is obtained with the surviving packets remained at the same wavelength for further propagation.

V. Acknowledgement

The work described in this paper was partially support by a grant from the Research Grants Council of the Hong Kong Special Administrative Region, China (Project No. HKU 7179/06E). The authors would also like to acknowledge Sumitomo Electric Industries for providing the HNL-DSF. The work described in this paper was partially support by a grant from the Research Grants Council of the Hong Kong Special Administrative Region, China (Project No. HKU 7179/06E). The authors would also like to acknowledge Sumitomo Electric Industries for providing the HNL-DSF.

VI. Reference

100G Ethernet for Packet Transport Networks

Andreas Kirstädtter
Nokia Siemens Networks, Otto-Hahn-Ring 6, D-81730 München, Germany
Tel.: +49-89-636-47484, Fax: +49-89-636-45814
andreas.kirstaedter@siemens.com

Abstract: This paper considers the functionality and standards required to enable carrier-grade core networks based on Ethernet-over-WDM. Possible Ethernet backbone network architectures will be discussed together with 100G transmission technologies.

1 Introduction

Backbone networks represent the top of the carriers’ network hierarchy connecting networks of different cities, regions, countries, or continents. The complexity of these technologies imposes substantial financial efforts on network operators, both in the area of Capital Expenditures (CAPEX) and Operational Expenditures (OPEX).

Ethernet claims to be a possible enabler of cost-efficient networks, as it is characterized by simplicity, flexibility, interoperability, and low cost. While Ethernet is traditionally a Local Area Network technology, continuous developments already enabled its deployment in Metropolitan Area Networks. Recent research and standardization efforts aim at speeding up Ethernet to 100 Gbit/s and at resolving scalability issues, thus supplying Ethernet with carrier-grade features for core networks.

2 Carrier-Grade Ethernet-Based Core Networks

2.1 Carrier-Grade Requirements

In order to be suited for core networks, Ethernet needs carrier-grade performance and functionality. It has to offer and implement the required Quality-of-Service (QoS) and has to enable traffic engineering to fine-tune the network flows. Furthermore, it has to provide fast and efficient resilience mechanisms to recover from link and network element failures and has to enable various Operation, Administration, and Maintenance (OAM) features for the configuration and monitoring of the network. Last but not least, it has to provide secure network operation. Additionally, a high degree of scalability is needed for handling different traffic types and for user separation inside the network. This scalability in terms of address space, maximum transmission speed, and maximum transmission distance becomes an important issue for the next Ethernet generation. E.g., multi-layer operation and optimization can only be used if facilitated by reasonable values of the maximum transmission distance.

At a closer look, it becomes visible that many of these required features are currently implemented repeatedly at different network layers. E.g., resilience mechanisms are found in the WDM layer and in an intermediate Sonet/SDH layer as well as in the packet layers above them. A cost-efficient network and protocol architecture therefore has to evaluate these functional redundancies between the layers very carefully.

2.2 Forwarding Technology and Scalability

The necessary scalability requires new approaches to packet switching and forwarding within meshed end-to-end Ethernet networks. Traditionally, within Ethernet networks the Spanning Tree Protocol (STP) calculates a single tree structure based on configurable IDs of switches, configurable port weights, and priorities to connect any switch with each other. Although loop-less forwarding is guaranteed with this mechanism, STP provides only one path between two locations and a MAC address learning of any equipment is performed at the switches.

However, in the case of combining large networks and adding hundreds of customer networks with an Ethernet-based core network the number of MAC addresses will grow rapidly. Thus, scalability can no longer be provided with current layer-2 approaches and a separation of networks or an additional hierarchy between them has to be introduced to allow a scalable forwarding of data.

Also, the use of a single tree structure providing only a single path between two locations prevents the use of efficient traffic engineering and resilience mechanisms. Thus, several connection-oriented forwarding techniques for carrier-grade Ethernet transport networks are currently under discussion at standardization bodies: VLAN Cross-Connect (VLAN-XC), Provider Backbone Transport (PBT), and Transport Multi-Protocol Label Switching (T-MPLS) [1].

3 Multi-Layer Operation and Optimization

Another important aspect in the area of scalability is the maximum transmission distance of Ethernet signals. Multilayer network grooming approaches are very attractive for the purpose of reducing unnecessary packet processing in intermediate nodes [2] as transit traffic is allowed to bypass intermediate nodes. Traffic between two network edge nodes can either be transported transparently in the optical domain or can be converted to the electrical domain to allow electrical grooming along the path. The effort spent on extending the signal reach of Ethernet signals is rewarded by equipment savings. E.g., for a typical German reference network topology a maximum optical transmission distance of 600km already enables port count savings of around 30% [3].
Next to IP services, VPN business services transfer increasing traffic and generate high revenues for network providers. In particular, Ethernet services (E-Line and E-LAN) are evolving. Today these layer-2 services are commonly transported via IP/MPLS tunnels. However, the complex functionalities and protocols of the IP layer are often not required to transport these pure layer-2 services. Native End-to-end Ethernet structures will arise where Ethernet business services will be transported on pure layer-2 infrastructures without the need of complex data transformations and changes in the functional layer structure. In continuation, also a merger between Ethernet-based packet transport and IP networking appears on the horizon further reducing superfluous functional redundancies between the single packet protocol layers – just in the sense of the current clean-slate thinking for the future Internet.

4 CAPEX and OPEX Performance

In order to calculate the total CAPEX of specific network architectures, future traffic loads, network device counts, and network device prices were estimated following a careful analysis of market data and price developments [3] for a German backbone - assuming a homogenous traffic growth rate of 40% per annum 2009-12. A shortest-path routing algorithm was then applied to determine the single link loads from which the number of switches, routers, and line card ports were finally obtained depending on the network architecture. The following generic network architectures were considered:

(a) IP/PoS-over-WDM: Label Edge Routers (LERs) and Label Switch Routers (LSRs) connected p2p via Packet-over-Sonet (PoS) links. 1+1 protection.

(b) IP/PoS-over-SDH-over-WDM: SDH grooming.

(c) IP/PoS-over-OXC-over-WDM: OXC grooming.

(d) IP/MPLS-over-Ethernet-over-WDM: MPLS-enabled Ethernet switches within the core. 1:1 protection.

(e) Ethernet-over-WDM: Native Ethernet switches both at edge and core. A small share (30%) of traffic requires IP routing at edge. 1:1 protection.

(f) Ethernet-over-WDM with service-level protection:

Only premium traffic (share set to 30%) is protected. SDH (b) and WDM (c) grooming provide cost reduction potential compared to the expensive PoS interfaces used in (a). In the MPLS-Ethernet case (d), still a considerable amount of CAPEX is related to LERs and their interfaces. A native 100 Gbit/s Ethernet over WDM network (e) enables higher savings – also at the edge. Applying a service-level differentiated protection scheme (f), the CAPEX can be reduced even further.

OPEX were evaluated via a process-oriented approach [4] for the network repair process since the impact of 100 Gbit/s Ethernet gets most visibly in this process category. For each of the network architectures described above, the OPEX were evaluated using availability figures [5] for the equipment and weighting the average repair time with the average salary of a field or point-of-presence technician. As a general result, in 100 Gbit/s Ethernet networks are more economical terms of OPEX due to the reduced device count (less switches and line cards). A service-level protection scheme further reduces the required network transport capacity, the network element count, and thus the related OPEX.

5 100Gbit/s Transmission Aspects

Ethernet transmission at speeds of 100 Gbit/s over long distances is very desirable in terms of architecture-related network cost. The transmission of high speed data rates above 100 Gbit/s is well understood although second degree (slope) chromatic dispersion has to be exactly compensated, birefringence effects become grave, and the signal-to-noise ratio of 100 Gbit/s signals is generally lower as fewer photons are transmitted per optical impulse. Additionally, recent trials demonstrated the ability to process electronically the required bit rates of 107Gbit/s [6]. Still, the major problem is to find efficient optic-electrical and especially electro-optical conversion techniques for these high speeds. Pure electrical solutions are preferable to handle the data at the transmitter and receiver.

6 Conclusions

In the past, Ethernet evolved from LAN into Metro areas covering speeds from 10 Mbit/s up to 10 Gbit/s. Next-generation Ethernet with transmission-speeds of 100 Gbit/s will facilitate cost-efficient Ethernet transport. As soon as carrier-grade issues like scalability, network resilience, QoS, and OAM of Ethernet-based core network architectures are solved we might see a complete Ethernet-over-WDM core-network infrastructure with resolved redundancies between the single layers.

7 Acknowledgements

The author wishes to thank A. Schmid-Egger for the input to the CAPEX and OPEX analysis. Special thanks go to the whole Ethernet-project team at Nokia Siemens Networks and the BMBF for financial support.

8 References

5. NOBEL Project, “Availability Model and OPEX related parameters”, to be published.
Application of asynchronous amplitude histogram method to monitoring of RZ-DQPSK signals

Bartłomiej Kozicki and Hidehiko Takara

NTT Network Innovation Laboratories, NTT Corporation, 1-1 Hikari-no-oka, Yokosuka, Kanagawa 239-0847 Japan
B. Kozicki: Currently at Osaka University, 2-1 Yamadaoka, Suita, Osaka, 565-0871 Japan
Tel: +81-46-859-8555, Fax: +81-46-859-5541, Email: takara.hidehiko@lab.ntt.co.jp

Abstract: We experimentally and numerically demonstrate monitoring of optical signal-to-noise ratio (OSNR) and chromatic dispersion (CD) in optical return-to-zero differential quadrature phase-shift keying (RZ-DQPSK) signals over a broad range of values.

Introduction: Optical communication networks are developing constantly. Following the increase in demand for transmission bandwidth, new technologies are introduced in order to take advantage of the optical fiber in a more efficient way. There are several approaches to increasing the volume of transmitted information per fiber. One way is by increasing the modulation bit rate. The currently prevailing 10 Gbit/s systems are gradually being upgraded or exchanged with 40 Gbit/s systems. Another approach is by increasing the number of channels per fiber through additional wavelengths. Finally, advanced spectrally efficient modulation formats are introduced in order to increase utilization of the occupied bandwidth.

RZ-DQPSK modulation format has been shown to perform well both in high-bit rate and high-spectral efficiency systems [1]. It is therefore considered a viable direction of evolution of optical networks. As the optical systems are operated at much tighter tolerance limits their performance must be strictly supervised. Traditionally, performance monitoring has been done electronically. Increasingly, however, the transmission is being carried out exclusively in the optical domain. In order to control the quality of optical signal, optical performance monitoring (OPM) methods have been suggested [2]. In the recent literature, several methods were investigated. Among the actively researched are optical spectrum monitoring, RF component monitoring, degree of polarization monitoring, monitoring using subcarriers and pilot tones, using data correlation and asynchronous amplitude histogram. These methods provide information about one or more physical parameter, like optical signal-to-noise ratio (OSNR) or accumulated chromatic dispersion (CD).

In this contribution we numerically and experimentally demonstrate monitoring of OSNR and CD of RZ-DQPSK signal using asynchronous amplitude histogram method. To our knowledge, this is the first demonstration of OPM of signals with quadrature modulation.

Principle: Asynchronous amplitude histogram monitoring [3] is a method which uses statistical properties of the optical waveform to analyze the quality of signal. The principle of operation is presented in Fig. 1.a. Optical waveform is sampled in an electroabsorption modulator (EAM) which is operated by ultra-short pulses at a low repetition rate. The samples are received by a low-speed photodiode (PD) and passed to an analogue-to-digital converter (A/D). The following analysis assembles the amplitude waveform to analyze the properties of the optical signal, optical performance monitoring (OPM) methods have been suggested [2]. In the recent literature, several methods were investigated. Among the actively researched are optical spectrum monitoring, RF component monitoring, degree of polarization monitoring, monitoring using subcarriers and pilot tones, using data correlation and asynchronous amplitude histogram. These methods provide information about one or more physical parameter, like optical signal-to-noise ratio (OSNR) or accumulated chromatic dispersion (CD).

In this contribution we numerically and experimentally demonstrate monitoring of OSNR and CD of RZ-DQPSK signal using asynchronous amplitude histogram method. To our knowledge, this is the first demonstration of OPM of signals with quadrature modulation.

Principle: Asynchronous amplitude histogram monitoring [3] is a method which uses statistical properties of the optical waveform to analyze the quality of signal. The principle of operation is presented in Fig. 1.a. Optical waveform is sampled in an electroabsorption modulator (EAM) which is operated by ultra-short pulses at a low repetition rate. The samples are received by a low-speed photodiode (PD) and passed to an analogue-to-digital converter (A/D). The following analysis assembles the amplitude

Principle: Asynchronous amplitude histogram monitoring [3] is a method which uses statistical properties of the optical waveform to analyze the quality of signal. The principle of operation is presented in Fig. 1.a. Optical waveform is sampled in an electroabsorption modulator (EAM) which is operated by ultra-short pulses at a low repetition rate. The samples are received by a low-speed photodiode (PD) and passed to an analogue-to-digital converter (A/D). The following analysis assembles the amplitude

Principle: Asynchronous amplitude histogram monitoring [3] is a method which uses statistical properties of the optical waveform to analyze the quality of signal. The principle of operation is presented in Fig. 1.a. Optical waveform is sampled in an electroabsorption modulator (EAM) which is operated by ultra-short pulses at a low repetition rate. The samples are received by a low-speed photodiode (PD) and passed to an analogue-to-digital converter (A/D). The following analysis assembles the amplitude

Principle: Asynchronous amplitude histogram monitoring [3] is a method which uses statistical properties of the optical waveform to analyze the quality of signal. The principle of operation is presented in Fig. 1.a. Optical waveform is sampled in an electroabsorption modulator (EAM) which is operated by ultra-short pulses at a low repetition rate. The samples are received by a low-speed photodiode (PD) and passed to an analogue-to-digital converter (A/D). The following analysis assembles the amplitude

\[
F_{\text{snr,rz}} = \frac{N_l}{N_m}
\]
spacing between histogram peaks $\mu_0$ and $\mu_1$. This is shown in Fig. 1.c. We define the dispersion parameter:

$$F_{\text{dis},rz} = \frac{(\mu_1 - \mu_0)}{\mu_m}$$

(2)

The precision of monitoring employing the $F_{\text{snr},rz}$ and $F_{\text{dis},rz}$ parameters is verified in simulation and experiment.

**Experiment and results:** The experimental setup used for monitoring RZ-DQPSK signal is depicted in Fig. 1.b. Signal is generated using serial arrangement of two dual-drive (DD) Mach-Zehnder modulators (MZM) and a phase modulator (PM). Residual CD is introduced through a span of standard single-mode fibre (SMF) or dispersion compensating fibre (DCF). The level of OSNR is adjusted by attenuating the signal power before optical amplifier. Another attenuator, following the bandpass filter, is used to maintain constant input power to the OPM device. Results of noise measurement are shown in Fig. 2. The noise parameter $F_{\text{snr},rz}$ is plotted as a function of OSNR (dashed line – simulation; dots – experiment). The insets show the waveform and histogram at low and high OSNR level. Good agreement between the simulation and experiment is achieved with resolution of 1 dB between OSNR of 17 and 30 dB.

Results of monitoring CD in a 20 Gbit/s RZ-DQPSK signal are shown in Fig. 3. Parameter $F_{\text{dis},rz}$ is plotted as a function of residual chromatic dispersion (dashed line – simulation; dots – experiment). The insets show the waveforms and histograms at corresponding points. The measurement ranges from -600 to +600 ps/nm, with 20 ps/nm resolution, without distinguishing the sign of dispersion.

The performance of $F_{\text{snr},rz}$ and $F_{\text{dis},rz}$ parameters is also analyzed in context of Q-factor. The results are plotted in Fig. 4. The noise parameter (solid line) is linearly proportional to Q. The dispersion parameter increases monotonically, however a calibration is required to obtain a direct relation to Q. These results prove that the asynchronous amplitude histogram OPM method can be employed in RZ-DQPSK systems. The measurements cover a wide range of expected values.

**Conclusion:** In this contribution we demonstrated for the first time optical performance monitoring of RZ-DQPSK signals. Asynchronous amplitude histogram OPM method allows for an accurate and repeatable measurement of noise and chromatic dispersion. The range covers the expected parameter scope.

**Acknowledgments:** The authors thank M. Jinno and Prof. Kitayama for their encouragement and fruitful discussions. This work was supported in part by the National Institute of Information and Communications Technology (NICT) of Japan.

**References:**


40 Gb/s All-Optical Multi-Wavelength Conversion via a Single SOA-MZI for WDM Wavelength Multicast

N. Yan1, T. Silveira2,3, A. Teixeira2, A. Ferreira2, E. Tangdiongga1, P. Monteiro2,3 and A.M.J Koonen1
1COBRA Institute, Faculty Elektrotechniek (PT), Eindhoven University of Technology, Postbus 513 5600 MB Eindhoven, Netherlands. Tel: +31 40 247 3672. Fax: +31 40 245 5197. Email: n.yan@tue.nl
2Instituto de Telecomunicações, 3180-193 Aveiro, Portugal
3Siemens Networks SA, D1 Research Department, 2720-093 Amadora, Portugal

Abstract—WDM wavelength multicast is demonstrated by all-optical multi-wavelength conversion. We report for the first time simultaneous 200 GHz spaced one-to-four conversion at 40 Gb/s using a commercial SOA-MZI with error free operation until $10^{-10}$.

I. INTRODUCTION

Multi-wavelength conversion (MWC) has attracted increasing interest in the last few years. It allows all-optical wavelength multicast by simultaneously converting data on one wavelength to several other wavelengths to be routed passively by optical waveguides, e.g. an optical arrayed waveguide grating. All the data information remains in the optical domain, which eliminates the necessity of employing multiple optic-electronic-optic (OEO) transponders, reduces the switching system and operational cost, lowers the blocking probability, and increases the optical network transparency, efficiency and effectiveness.

Many multi-wavelength converters (MWCR) have been reported. However, only a few of them can support high data rate up to 40 Gb/s: nonlinear fiber by four-wave mixing (FWM) or supercontinuum generation [1], electroabsorption modulator by cross absorption modulation [2], semiconductor optical amplifier (SOA) by nonlinear polarization switching [3] and SOA - Mach-Zehnder interferometer (MZI) by cross-phase modulation (XPM) [4]. Among these, XPM-based MWCRs using a SOA-MZI excel the others by offering a great combination of advantages [4, 5] such as satisfactory and leveled converted efficiency, simultaneous conversion of a considerable number of channels, wavelength flexibility, wide conversion bandwidth, commercial product availability, integration potential, compactness, low power budget, ability of converting both return-to-zero (RZ) and non-return-to-zero (NRZ) data, possible signal regeneration and noise suppression, and the possibility of using differential scheme for high data rates. However, to our knowledge, so far at 40 Gb/s only one-to-three MWC with minimum 300 GHz channel spacing and a large 1200 GHz detuning has been reported without bit error rate (BER) measurements using a SOA-MZI [4].

In this paper, simultaneous one-to-four MWC at 40 Gb/s with ITU 200 GHz channel spacing and 600 GHz detuning is demonstrated based on a commercial hybrid integrated SOA-MZI regenerator using a push-pull configuration. The BER performance of such a system was evaluated in all the four MWC channels against back-to-back reference and single wavelength conversion cases. Error free operation until $10^{-10}$ was achieved.

II. EXPERIMENTAL SETUP AND RESULTS

The experimental setup is shown in Fig. 1. The commercially available SOA-MZI wavelength converter was from a CIP twin regenerator for 10 Gb/s standard operation. ITU 200 GHz spaced wavelengths were deployed. The 40 Gb/s RZ data signal was generated by modulating a 40 GHz ultrafast optical clock (UOC) using 40 Gb/s pseudorandom bit sequence with pattern length of $2^{31}-1$. The 2-ps pulse source was tuned to 1557.36 nm. An erbium-doped fiber amplifier (EDFA) was used to compensate the power loss in the modulation process. Push-pull configuration was employed to overcome the speed limit of the SOA carrier dynamics for the MZI to perform 40 Gb/s MWC. For this purpose the data signal was tapped onto both of the SOA-MZI arms A and D using a 50/50 coupler, with the lower data path delayed by a variable optical delay line (VODL) to achieve the differential mode. The optical power for the data channel and the delayed data channel were 7.3 dBm and -3 dBm respectively. The push-pull delay was around 6.5 ps. Four continuous waves (CWs) at wavelength 1547.72 nm, 1549.32 nm, 1550.92 nm and 1552.52 nm were combined by a multiplexer and injected in the co-propagating direction into SOA-MZI port B.

![Fig. 1. Experimental setup for all-optical one-to-four MWC at 40 Gb/s using a commercial SOA-MZI regenerator](image-url)
non-inverted output under such experimental conditions. The converted channels were selected utilizing an optical filter with a -3 dB bandwidth of 130 GHz for the detection at the pre-amplified receiver.

Fig. 2 presents the output spectrum with all the four simultaneously converted eye diagrams as insets. The eye diagrams of all the MWC channels were clear and open. However, the outer channels, Ch1 and Ch4, had visible noisy one level. This had been observed during all the measurements of one-to-four conversion. From the spectrum we also observed FWM satellite signals at both sides of the converted channels from the SOA nonlinear effect. The out-of-band FWM byproducts were about 17 dB weaker or more than the desired converted channels. The measurements taken from the oscilloscope indicated an average extinction ratio (ER) of 9.61 dB for the four simultaneously converted channels, while the largest ER difference among the different channels was no greater than 0.61 dB.

Fig. 2. Output spectrum of simultaneous one-to-four MWC with converted eye diagrams as insets

Fig. 3 shows the BER as a function of optical signal-to-noise ratio (OSNR) at the photo detector (PD) input for the four simultaneously converted channels, where the back-to-back reference and a single wavelength conversion curve were also plotted for comparison. The -3 dB bandwidth of the optical filter was 1 nm. From the BER results, we observed 4.5 dB OSNR penalty at BER of $10^{-9}$ for the single wavelength conversion to 1550.92 nm. This is mostly due to the pulse broadening caused by the wavelength conversion. The one-to-four simultaneously converted channels demonstrated a minimum 1 dB OSNR penalty with regard to the single wavelength converted case. From the BER curves it seems that Ch1 and Ch4 also had some error floor, which contributed to the OSNR sensitivity variation of around 2.7 dB at BER of $10^{-9}$ from the central channels Ch2 and Ch3. The worse performance of the outer channels agreed with the indication from the eye diagram insets presented in Fig. 2, where they exhibited evidently more noise at the logical one level. In Fig. 3, the eye diagrams of the 40 Gb/s RZ back-to-back signal, single wavelength conversion, one-to-four MWC outer channel Ch1 and inner channel Ch3 are also shown as insets. The eye opening of each case from the oscilloscope complied well with the BER measurements obtained.

Fig. 3. BER of four simultaneously converted channels with respect to the back-to-back reference and single wavelength conversion

III. CONCLUSION

Single and simultaneous one-to-four MWC using a commercial SOA-MZI regenerator of 10 Gb/s standard operation speed was successfully demonstrated for 40 Gb/s conversion with differential configuration. For the first time to our knowledge, this was achieved at 40 Gb/s with error free operation until $10^{-10}$ at 200 GHz channel spacing and only 600 GHz detuning. Clear and open converted eye diagrams were obtained. The OSNR penalty at BER of $10^{-9}$ was 4.5 dB for the single wavelength conversion and around another 1 dB or more for the MWC. No assist light was employed [5]. We did not observe obvious performance limitations due to the number of MWC channels, therefore, we believe that more channels could be added if more laser sources were available, provided that they are placed inside the gain spectrum of the SOAs. Our results proved that a single SOA-MZI can be a promising candidate for all-optical wavelength multicast at 40 Gb/s.

ACKNOWLEDGMENT

This work has been supported by the European Commission through the Network of Excellence ePhoton/ONe+ project. Siemens Networks SA is also acknowledged for lending their lab facilities. Authors T. Silveira and A. Ferreira are thankful to the Portuguese FCT for grants. The authors would also like to thank the FCT Portuguese project POSC/EEA-CPS/61714/2004 -CONPAC.

REFERENCES

Generation-Free-Platform architecture with flexible physical and logical links using optical technology

S. Yanagimachi*, Y. Hidaka*, J Suzuki*, J. Higuchi*, T. Yoshikawa*, A. Iwata*
System Platforms Research Labs., NEC Corporation
1753 Shimonumabe Nakahara-ku, Kawasaki, 211-8666, Japan
E-mail: *s-yanagimachi@ab.jp.nec.com, *y-hidaka@bq.jp.nec.com, *j-suzuki@ax.jp.nec.com, *j-higuchi@ax.jp.nec.com, *y-yoshikawa@cd.jp.nec.com, *a-iwata@ah.jp.nec.com

1. Abstract
To achieve scalable flexible platform, we propose a stackable and modular system (GF-PF: Generation-Free-Platform) with flexible physical link using optical interconnection and logical link using our ExpEther technology.

2. Introduction
Recently, changes in business environment and process are rapidly. Especially, in the field of IT/NW, platforms needed to be replaced every two or three years to keep pace with innovations in technology and these replacements cause considerable capital investment. To solve this problem, we have already proposed a new concept, Generation Free Platform (GF-PF) [1]. The objective of our research is to pursue a scalable flexible platform that can correspond flexibly to rapid changes in business conditions. To achieve this scalable flexible platform requires modularity and reconfigurability of hardware resources and the application resources running on that hardware. In addition, system designs are required that can flexible interconnect resources and provide scalable expansion of those resources. In this paper, we describe the detailed hardware architecture of scalable flexible platform, GF-PF; especially, we focus on flexibility and reconfigurability of that platform that can be achieved by using of a stackable system and a modular system with flexible physical and logical link technologies.

3. GF-PF architecture
GF-PF is the platform which provides various services, using flexible interconnection of any resources among different shelves, and it has two main features.

First, the GF-PF logical service platform has a virtualization function that hides the complexity of hardware resources and the application resources running on that hardware. In addition, system designs are required that can flexible interconnect resources and provide scalable expansion of those resources. In this paper, we describe the detailed hardware architecture of scalable flexible platform, GF-PF; especially, we focus on flexibility and reconfigurability of that platform that can be achieved by using of a stackable system and a modular system with flexible physical and logical link technologies.

Second, the GF-PF physical service platform has following features: a stackable system that allows flexibly expanding physical resources, and a modular system that can flexibly reconfigure its hardware platform. Figure 1 shows the architectures of our system.

The followings are requirements of the interconnection for a scalable flexible platform.

- Switch capacity scalability
- Physical link scalability
- Logical link (protocol level connection over physical link) scalability and reconfigurability

4. Modular switch architecture
Several switch architectures, e.g., one-stage switching or multi-stage switching, have been widely used. Multi-stage switch architecture has been used for large-scale nodes because systems with one-stage switches lack the flexibility necessary for maximum switch port from the initial deployment. To achieve linear expansion of switch capacity, we use modularized multi-stage switch based on CLOS architecture where switches are connected with optical interconnect. The modular switch card is illustrated in Fig. 2. This switch card consists of a mother card with four switch modules on it. The switch module has an LSI chip surrounded by the four optical transceivers (PETIT) [2]. Since a switch capacity of 40-160 Gbps can be implemented in a single switch module and the switch card has four elements per switch module, the switch capacity of a single switch card is 160-640Gbps. In our design, three switch cards are used to achieve 480-1.92-Tbps system capacity.

5. Physical link technology (Optical interconnect)
Generally, several interconnection among physical resources is electrical. When electrical interconnection is used, transmission speed is limited because of crosstalk between adjacent wiring, attenuation of signals, and the reflection. And also the number of backplane layers is increased and the connector area is larger in proportion to the increase in the number of slots. Furthermore, electrical wiring corresponds to a particular target transmission speed so flexible system to an increase in transmission speed is difficult. To solve these problems and to achieve scalable physical link, we use optical interconnection among physical resources. Furthermore, using optical interconnection makes backplane redesign unnecessary up to 20Gbps (10Gbps is available now, and 20Gbps is under...
small optical transceiver

I/O B

PEB

˜

Optical Module(Petit)

˜

ExpressEther

PEB

PEB

 Angelo

VLAN 2

High density optical connector

Server C

I/O C

rzą

I/O C

VLAN 2

Server B

Performance of PCIe interface can be up to x16. Our optical interconnection system has following features.

- Small optical transceiver
- High density optical connector
- Small fiber assembly area

We use our ultra small optical module (PETIT) in the GF-PF. This module has \(4 \times 3.125\) Gbps or \(4 \times 10\) Gbps TX/RX channels. The LSI size (12 x 12mm) is much smaller than those of conventional optical modules (e.g., the size of SNAP12 is 45 x 14.5mm).

In addition, we use multi-fiber optical connectors. Their dimensions are 53 x 13 mm; it accommodates 96-optical fibers, and achieves a throughput of 1Tbps.

To avoid degraded insertion loss and ensure reliability, the bending radius of the conventional optical fiber is kept within 25-30 mm. To make small fiber assembly area, in the GF-PF, we use high-\(\Delta\) and low diameter MMF that can tolerate a bending radius within 5 mm. Figure 5 compares the data of insertion loss for both types of MMF. The insertion loss of high-\(\Delta\) MMF is the same as for conventional MMF. Figure 6 is a photograph of a fiber assembly on a service card using a high-\(\Delta\), low diameter MMF. Due to achieving fiber assembly with small bending radius fiber, fiber assembly area can be much smaller than that of a conventional design.

6. Logical link technology (ExpEther)

To achieve flexible logical link over physical link, we already proposed ExpEther [3] as Ethernet-based virtualization technology. It groups modularized resources interconnected by an Ethernet, and transports a PCI Express (PCIe) packet between the grouped modules by encapsulating it into an Ethernet frame. ExpEther, by utilizing an Ethernet technology, can achieve scalability and easy reconfigurability of hardware resources. The configuration of ExpEther is illustrated using Figure 7. A PCIe-to-Ethernet bridge (PEB) performs the encapsulation and decapsulation of a PCIe packet. A server and an I/O assigned to its server belong to the same VLAN. Note that the assignment of an I/O to a server can be flexibly changed by setting the VLAN of the I/O to that of the server.

Figure 8 shows an ExpEther card which performs a PCIe packet encapsulation on a server side. This card accommodates x4 PCIe interface (total throughput is 10Gbps). By using our ultra small optical module (PETIT) for 10G Ethernet optical ports, we achieved compact design and high performance of the ExpEther card where the maximum throughput of the PCIe interface can be up to x16.

7. Conclusion

We propose the GF-PF architecture with a stackable system and a modular system with optical interconnection (for physical link) and our ExpEther (for logical link). In this design, the system can continue to grow and reconfigure to correspond with rapid changes in business conditions. We also describe in detail the optical interconnection used in our GF-PF. By using ultra small optical transceiver (PETIT) and high-\(\Delta\), low diameter MMFs for fiber assembly, its area can be much smaller than that of a conventional design.

Acknowledgements

This work was partly supported by the Ministry of Internal Affairs and Communications (MIC).

Reference

Optical Packet Interconnect System Using High-Speed SOA Switch for Peta-scale Computing System

Kyosuke Sone, Yasuhiko Aoki, Goji Nakagawa, Yutaka Kai, Setsuo Yoshida, Yutaka Takita, Susumu Kinoshita and Hiroshi Onaka
Fujitsu Limited
4-1-1 Kamikodanaka, Nakahara-ku, Kawasaki 211-8588, Japan (Company mail No. L50)
Tel: +81-44-754-2643, Fax: +81-44-754-2640, E-mail:sone.kyousuke@jp.fujitsu.com

Abstract

We confirmed the optical packet switching operation of our developed 2×2 optical packet interconnect system for peta-scale computing. The interconnect system consists of an arbiter for system control, leaf switches with WDM optical ports for optical packet generation, and semiconductor optical amplifier (SOA) switches for high bandwidth and fast switching.

1 Introduction

Peta-scale computing system consisting of more than 10,000 CPUs is indispensable for various kinds of simulations of nano-scale devices, multi-physics problems, and so on. In the Peta-scale computing system, a high throughput (bandwidth) interconnect system which allows high speed CPUs to freely exchange data is requested [1], [2]. The interconnect system has to support the data granularity ranging from 64-byte short length to several mega-byte huge data with microsecond-order latency.

Previously, we have proposed the WDM optical packet interconnect system and reported the feasibility of a 100-Tbps-class throughput by multi-gate broadcast-and-select SOA switch architecture [3]. In this paper, we describe the switching characteristics in our proposed system with a developed 2×2 optical packet interconnect system.

2 Optical packet interconnect system for peta-scale computing system

Our proposed optical packet interconnect system is shown in Fig. 1. A leaf switch at each computing-node group has two functions, namely the switching of the intra-rack signals and the aggregation of the inter-rack signals with packet queues for high bandwidth data interconnections. When the queue for a certain destination is filled, a queue manager at the node group sends a connection request to an arbiter attached to the optical packet switch fabric for scheduling, and the arbiter replies with a grant signal to the manager for the packet forwarding. Subsequently, the aggregated signals are converted to optical packet signals and interconnected between two computing node groups by the optical switch under the control of the arbiter. A WDM packet switching scheme is used to increase the bandwidth per switching port and reduce the number of OE/EO modules and cables. This system has a timing adjustment function which enables to send the optical packet signal to the destination node during the opening of the optical switch based on the system master clock of the arbiter.

3 Demonstration of 2×2 optical packet interconnect system

To confirm the feasibility of our proposed interconnect system, we demonstrated its operation with

---

Fig.1 Optical packet interconnect system for peta-scale computing system
a 2×2 optical packet interconnect system. The system consists of two leaf switches, an arbiter, and a 2×2 optical switch subsystem as shown in Fig. 2. The optical switch is a broadcast-and-select type SOA gate switch. SOA has the fast switching characteristic (ns order) with a high extinction ratio (> 60 dB). The wavelengths of the data signal and the control signal are 1.55 µm and 1.3 µm, respectively, and they are coupled and separated by WDM couplers. The switching characteristic of a 2×2 optical switch subsystem is shown in Fig. 3. We obtained a high-speed switching (< 10 ns) at both rising and falling for all SOA switches. The detail system operation of the 2×2 optical packet interconnect system is shown Fig. 4. Upper 4 lines indicate control signals for 4 SOA switches. Bottom 2 lines show switching operation at out ports. In this experiment, we intentionally change the SOA gain, so that we can see all SOA operations. For example, at Out #1 there are 3 levels corresponding to SOA 2-1 on (highest level), SOA 1-1 on (middle level), and the both-off (bottom level) states. Signals are modulated at 10 Gb/s, however, those are averaged due to the small bandwidth of measurement instruments. We confirmed that 1.2 µs optical packet data was successfully switched. The amount of skew, which is the difference in arrival time from each leaf switch to the optical switch, can be suppressed to less than 6 ns by the timing adjustment in the interconnect system. We can further shorten the current guard time (~45 ns) to 20 ns by improving the optical switching speed and the timing adjustment granularity.

4 Conclusions

We have developed a 2×2 optical packet interconnect system using a high-speed SOA switch. We successfully demonstrated the operation of optical packet switching and confirmed the feasibility for peta-scale computing system. In future, we will enhance the number of switch ports up to 256.

Acknowledgements

This research was partly supported by the Ministry of Education, Culture, Sports, Science and Technology of Japan for the design of the optical packet switch architecture and the National Institute of Information and Communications Technology of Japan for the development of the high speed optical switch fabric.

References

3 H. Onaka et al., ECOC2006, Tu4.6.6 (2006)
A study on fault recovery of optical paths in photonic cross-connect systems

S. Yoshida, E. Horiuchi, Y. Baba, Y. Akiyama, K. Onohara, T. Mizuochi, and S. Seno
Information Technology R&D Center, Mitsubishi Electric Corporation, 5-1-1 Ofuna, Kamakura, 247-8501, Japan
Phone: +81-467-41-2430, Fax: +81-467-41-2419, email: Yoshida.Sota@eb.MitsubishiElectric.co.jp

Abstract
To enhance optical path reliability, a photonic cross-connect (PXC) supports a range of fault recovery mechanisms. In this paper, we report the demonstration of these functions, and also address the stable operation of optical paths.

1. Introduction
Broadband networks have been developing significantly in recent years, which causes ongoing expansion of the networks’ capacity. There have been many investigations into photonic cross-connect (PXC) to realize the next generation of large-capacity optical transport networks [1][2]. PXC is a form of transmission equipment that can set up and switch optical paths by using all-optical switches.

PXCs can be used as components of a large-capacity optical network, where the PXCs house client devices such as routers, and WDM equipment is connected between the PXCs [3]. In addition, it is possible to reduce operational costs by supporting GMPLS protocols. To increase network availability, a PXC supports a range of GMPLS-based protection and restoration mechanisms. Further, the PXC incorporates redundant optical switches in order to enhance its fault tolerance.

We describe a prototype PXC and demonstrate its fault recovery mechanisms. We also report a stable optical path operating regime that avoids extraneous fault detection and recovery.

2. Transport Failure Recovery
Once a PXC detects a failure on an optical path, its fault recovery procedure is initiated. Faults can be divided into two types: an equipment failure (e.g. optical switch failure) and transport failure (e.g. a fiber cut). In the following, we present the failure detection mechanisms and fault recovery procedures of our PXC prototype.

2.1. 1+1 redundant optical switches [4]

Fig.1 shows the data plane architecture of the PXC prototype. The input signal is split between 1+1 redundant optical switches using the 3dB coupler on OptIF card A, and one of the signals is selected as the output signal using the 2x1 optical switch on OptIF card B. There are two optical power monitors, one each on OptIF card A and OptIF card B, to check the signal power.

When an input signal is detected but no output signal is detected, the failure is considered to be an optical switch failure and is recovered using the optical switch redundancy. For example (see Fig.1), when a failure occurs at MATSW#0, the 2x1 optical switch selects the route that traverses MATSW#1, for which the recovery time is very short.

If there is no input signal, the failure is considered to be a transport failure and the GMPLS failure recovery function is initiated. Section 2.2 covers the GMPLS failure recovery function in detail.

2.2. GMPLS failure recovery

Several GMPLS-based protection and restoration mechanisms such as 1+1 protection and Full LSP re-routing to meet various service requirements have been discussed [5][6].

In the 1+1 protection, two (link/node/SRLG) disjoint paths are established and the same traffic is transmitted simultaneously over the two paths. When a path terminator detects a failure on the primary path it selects traffic from the back-up path. Fig.2 shows an example where the primary path is established from PXC#1 to PXC#3 via PXC#2, the back-up path is established from PXC#1 to PXC#3, and a failure is assumed on the primary path. When PXC#3 detects the failure, it selects the traffic from the back-up path via the optical 2x1 switch.

Unlike 1+1 protection, in the full LSP re-routing, the back-up path is not established until a failure occurs on the primary path. The benefit of this method is high resource utilization and robustness against multiple failures. The control plane sequence diagram shown in Fig. 3 outlines how the re-routing works. A re-routable path is established from PXC#1 to PXC#3 via PXC#2 and a failure is assumed on the primary path. Once PXC#3 detects the failure, PXC#3 will notify the failure to the path’s initiator (PXC#1) using a Notify message. After receiving the Notify message, PXC#1 sends a PathTear message to release the primary path, computes an alternative route not traversing the failed links, and establishes a re-routed path using PATH/RESV messages. Compared to 1+1 protection, re-routing recovery is slow due to the control plane signaling.

![Fig.1: Data plane architecture of PXC](image)

![Fig.2: 1+1 protection](image)
2.3. Issues with transport failure recovery

Executing GMPLS failure recovery functions and optical switch redundancy at the same time may lead to undesirable consequences for the network operators. For example (see Fig.3), let us assume that there is an optical switch failure at intermediate PXC#2. In this case, it is possible to recover by switching over to the back-up optical switch at PXC#2, but PXC#3 also detects the failure simultaneously. The network operators expect the path to be restored using the optical switch redundancy. However unnecessary re-routing is initiated at PXC#3, and as a result the recovery time is several hundred msec.

3. Introduction of switch-over guard timers

In order to avoid this conflict, we propose to use two guard timers to choose between the GMPLS failure recovery functions and optical switch redundancy switching. One is T_EQP that blocks optical switch redundancy switching, and the other is T_LOL that blocks the GMPLS failure recovery functions. Network operators can assign priorities to these recovery operations by adjusting the guard timers. With T the optical switch redundancy recovery time, the relationships between these timers can be written as follows:

- a) Giving priority to the optical switch redundancy
   \[ T_{LOL} > T_{EQP} + T \]  
- b) Giving priority to the GMPLS failure recovery functions
   \[ T_{LOL} < T_{EQP} + T \]  

4. Verification

To evaluate our proposal, we implemented these timers on the PXC prototype. Fig.4 shows the network configuration and its three PXC's. A re-routable path and a 1+1 protection path were established between PXC#1 and PXC#3. We measured the failure recovery time by using SDH analyzers in the following two scenarios:

- Scenario 1: a fiber cut between PXC#1 and PXC#2;
- Scenario 2: an optical switch failure at PXC#2.

Table 1 shows the guard timer configurations. For either type of path, T_EQP was set at 0ms. We estimated failure recovery time to be 20ms. T_LOL of the re-routing path was set to 30ms to fulfill (1) above. T_LOL of the 1+1 protection path was set at 0ms to fulfill (2).

Table 2 shows the result of recovery time. We confirmed that PXC#3 initiated the GMPLS failure recovery functions after each guard time had elapsed. The re-routing path was restored within 350ms, and the 1+1 protection path was restored within 20ms. The re-routing path was restored within 23ms by optical switch redundancy switching at PXC#2. We confirmed that PXC#3 also detected the failure but it was recovered within T_LOL, so re-routing was not initiated. Meanwhile PXC#3 had initiated 1+1 protection and its recovery time was 20ms.

5. Conclusion

In this paper, we demonstrate the fault recovery mechanisms of the prototype PXC. We also address operational stability and confirm that using two guard timers prevents extraneous failure recovery operations.

Reference

A Study of Equipment Protection for High Availability of Control and Data Planes in Photonic Cross-Connect

K. Onohara, Y. Akiyama, T. Mizuochi, T. Ichikawa, H. Sato, K. Okubo, E. Horiiuchi, S. Yoshida, and Y. Baba
Information Technology R&D Center, Mitsubishi Electric Corporation, 5-1-1 Ofuna, Kamakura, Kanagawa 247-8501, Japan
Phone: +81-467-41-2443, Fax: +81-467-41-2486, E-mail: Onohara.Kiyooshi@MitsubishiElectric.co.jp

Abstract A photonic cross-connect with 1+1 redundant supervisory cards and optical switches to minimize any system outage has been developed. We show the resulting improved protection time as evaluated on a live prototype.

1. Introduction
In order to help meet the network needs arising from the continuous growth of Internet traffic, photonic cross-connects (PXC) are under widespread investigation [1-3], as a PXC is able to provide bit-rate free and format-free switching. The switch structure and its control system are much simpler than those of a digital cross-connect due to its complete lack of O/E/O conversions. However, an optical switch failure would cause all the transmission channels through it to fail. Therefore, the reliability of the optical switches is an important factor in the stable operation of carrier-grade networks.

In this paper, we report a PXC prototype with 1+1 redundant supervisory cards and optical switches for high availability network operation. The protection was carried out without any disconnection of live traffic. The protection time of the 1+1 protected optical switches was 9.5 ms when seven paths were disconnected simultaneously due to a simulated optical switch failure. The resulting improved system outage is also discussed.

2. 1+1 Redundant Supervisory Systems
An uninterrupted supervisory system is mandatory for PXC based optical networking, to minimize the recovery time if a system failure occurs. Any failure in a supervisory card should be recovered rapidly via some form of redundant architecture. Moreover, any failure in a supervisory card and any failures in its related elements, e.g. the control bus, control lines, and the connections between the control and management planes, should not affect live traffic. Fig. 1 shows the architecture of the PXC’s 1+1 redundant supervisory system. Not only the supervisory cards, but also the control buses for the optical interface (I/F) cards and the control lines for the optical switches, have a 1+1 architecture. The connection between the control and management planes also has 1+1 redundancy.

(1) 1+1 redundant supervisory cards
The working supervisory card passes a range of signalling information and data to the standby card, e.g. the used and unused ports, optical powers and power loss thresholds at each input/output port of the optical switch, in addition to a range of alarms, via a high availability (HA) link. The HA link also has 1+1 redundancy. If the working card fails, the standby card is promptly activated. The conditions in the data plane are preserved and no live traffic is disrupted.

(2) 1+1 redundant control buses for optical I/F cards
The 1+1 redundant control buses are used to supervise the optical I/F cards, one being the working bus and the other being the standby. Each optical I/F card is informed which is the working bus via two specific signal lines connected to the supervisory cards. The standby supervisory card receives information from the optical I/F cards, but sends nothing unless the working bus has failed.

(3) 1+1 redundant control lines for optical switches
The 1+1 redundant optical switches are controlled by cross-connected 100BASE-T cables. If a failure occurs in the working control line, the supervisory card immediately checks whether the standby line can connect correctly prior to activating the standby control line, in order to avoid any unwanted session conflict.

(4) 1+1 redundant connections to control & management planes
The connection between the control and management planes also has 1+1 redundancy. When the standby supervisory card is activated, it starts to use the same IP address as the working card in order to conceal the protection action from the control and management planes.

We evaluated the protection using a live prototype PXC. We intentionally caused failures by generating forced error signals or disconnecting control lines by hand. In each case, the supervisory system was successfully restored without any unwanted conflict. The data planes were also maintained and no live traffic was disrupted.

Abstract Fig. 1: PXC architecture with multiple 1+1 redundancy.

3. 1+1 Redundant Optical Switches
Fig. 2 illustrates the connections between the optical I/F cards and optical switches with 128x128 input/output ports.
Although Fig. 2 depicts only one input/output port in each optical I/F card for ease of visualization, each optical I/F card in fact has sixteen input/output ports. We concentrate on the traffic flow through optical I/F #i, #j, and #k as follows: (1) the input optical signals are split by couplers at the optical I/Fs (#i and #j/#k) and launched into the working and standby optical switches; (2) the duplicate signals are switched and launched into each optical I/F (#j/#k and #i), where the optical 2x1 switch selects one of the redundant signals. Optical power monitors are fitted at both input and output ports in order to distinguish a loss of light from an optical switch failure.

In the case of bi-directional label switched paths (LSPs) through #i and #j or #i and #k, the supervisory card sends a command to the 1+1 optical switches and the optical 2x1 switches to establish the paths, i.e. route A or C as working path and route B or D as protection path. If the working optical switch fails, the optical 2x1 switches in optical I/Fs (#i and #j/#k) are released, and the paths are bridged to route B or D. In the case of 1+1 dedicated protection with LSPs from #i to #j and #k and vice versa, the supervisory card sets up routes A and D.

We evaluated the protection time for an optical switch failure using the live prototype PXCs. We intentionally caused seven simultaneous failures in the fiber connections between the optical I/F card and the working optical switch to demonstrate the effect of an optical switch failure. Fig. 3 shows the measured 1+1 protection times for the failure. The maximum protection time was 9.5 ms. The protection time increases progressively by port number. This is because the supervisory card queues the optical 2x1 switches one at a time.

We discuss here the protection time of the architecture developed. If a hardware failure occurs, it could take some time before the supervisory system detects and restores the fault. This may depend, for example, on the frequency with which signals are sent, on the speed of fault detection and system notification in the hardware, and on the time it takes for the system to gather all the fault information from the various signals, correlate it, and then wait to avoid instability. The protection time $T_{pro}$ is given by [4]

$$ T_{pro} = \tau_{det} + \tau_{hold} + \tau_{notify} + \tau_{rec\_ope(w)} + \tau_{rec\_ope(h/w)}, $$

where $\tau_{det}$, $\tau_{hold}$, $\tau_{notify}$, $\tau_{rec\_ope(w)}$, and $\tau_{rec\_ope(h/w)}$ are respectively hardware fault detection time, hardware hold-off time, software fault notification time, and software and hardware recovery times. Here, $\tau_{det}$, $\tau_{hold}$, and $\tau_{notify}$ were set to 0.1 ms, 0 ms, and 5 ms at a maximum. $\tau_{rec\_ope(w)}$ and $\tau_{rec\_ope(h/w)}$ take 0.4 ms and 4 ms, respectively.

4. Outage Estimation for 1+1 Redundant System

We discuss outage of the 1+1 redundancy architecture. Here we assume the respective failure rates (FITs) of the supervisory card and the optical I/F card to be approximately 200 and 100. These values are estimated from the service records of similar field proven cards. The optical switches’ FIT rate is an intentional worst estimate of 20,000. The mean times to repair (MTTRs) are estimated at about 4 hours for optical switch replacements and less than 1 sec for supervisory card protection. The data plane outages with and without redundancy are calculated to be 13 sec and 42 min per year. The control plane outages with and without redundancy are calculated to be respectively 1.8 ms and 25 sec per year. The 1+1 redundancy enables highly reliable operation for carrier-grade networks.

5. Conclusion

We have developed a 1+1 protection based redundancy system for a photonic cross-connect for high availability network operation. The protection of the redundant supervisory cards was carried out without any disturbance to live traffic, and the protection time of the 1+1 protected optical switches was 9.5 ms when seven paths were disconnected simultaneously due to a working optical switch failure.

References

Fast Generation of Orthogonal Periodic Polynomials and Its Application to Smooth Approximation of the Internet Traffic

Hiroshi Hasegawa†, Naoya Matsusue†, and Ken-ichi Sato†
† Dept. of Electrical Engineering and Computer Science, Nagoya University
Furo-cho, Chikusa-ku, Nagoya, 464-8603 Japan
E-mails: hasegawa@nuee.nagoya-u.ac.jp, n.matusu@echo.nuee.nagoya-u.ac.jp, sato@nuee.nagoya-u.ac.jp

Abstract—In this paper we propose a method to generate a sequence of periodic orthogonal polynomials over [0,1], where polynomial weighting function is defined over the interval. Low computational load and numerical stability are realized by using the Lanczos’s three-term recurrence relation to generate the sequence. Further improvement is provided by omitting numerical integration, and therefore, the proposed method requires only a small number of arithmetic operations. Finally the generated sequence is employed for least-squares smooth approximation of given traffic data having diel periodicity.

I. INTRODUCTION

There are several periodicities in the Internet traffic because the users are dominated by common cycles such as the diel periodicity. For efficient adaptive network control in traffic engineering, one of the most important requirements would be estimation of such a trend in the traffic. The estimation can be essentially realized by approximation to the traffic with a low-frequency component having some specified periodicities. The smoothness can be characterized by several measurements such as support in the Fourier domain. Unfortunately, imposing periodicity is not straightforward if we employ such a transform using given fixed bases in transformed domain. Thus we need a method to design bases of the set of all periodic smooth functions, where such a method directly provides least-squares approximation of given function efficiently.

In this paper, we introduce a periodic polynomial approximation of given function, defined and periodic over closed interval, where smoothness is realized by restricting the degree of approximating polynomial. In this case, orthogonal bases are straightforwardly derived by applying the well-known Gram-Schmidt’s orthogonalization procedure [1] to a sequence of periodic monomials with increasing degree. We show that it is possible to realize this approximation scheme computationally efficient and numerically stable. Firstly, we derive significant reduction of computational load and further numerical accuracy in generating the bases by introducing a three-term recurrence relation [1–3]. Indeed, this relation is employed in several areas, for example stable design of digital filters[4]. Secondly, we provide a simple equation that enables us to exactly compute each base with small number of arithmetic operations. Numerical experiment shows that smooth periodic approximation is derived for real traffic data.

II. PRELIMINARIES

Let the set of all real numbers and integers be \( \mathbb{R} \) and \( \mathbb{Z} \), respectively. For all functions \( f, g : [0,1] \rightarrow \mathbb{R} \), define their inner product by

\[
\langle f, g \rangle := \int_0^1 w(x)f(x)g(x)dx,
\]

where \( \int \cdot dx \) means Lebesgue integral and \( 0 < w(x) < \infty \) over \([0,1]\) is a weighting function. Define the induced norm by \( \|f\|^2 := \langle f, f \rangle \) for all bounded function \( f : [0,1] \rightarrow \mathbb{R} \). Then the set of all functions having bounded norm forms a Hilbert space. In this paper, we specially assume that \( w(x) \) is a polynomial \( \sum_{n=0}^{N} w_n x^n \) such that \( w(x) > 0 \) for all \( x \in [0,1] \). The set of all polynomials \( p \) such that \( p(0) = p(1) = 0 \) and whose degrees are equal or less than \( N \) is defined by

\[
\begin{align*}
P_N &= \left\{ \sum_{n=2}^{N} a_n x^{n-1}(x-1) =: \sum_{n=2}^{N} a_n u_n(x) \right\} \\
&= \text{span}\{u_n(x)\}_{n=2}^{N}
\end{align*}
\]

The set of polynomials \( \{u_n\}_{n=2}^{N} \) is linearly independent each other since an uniformly zero polynomial must be the constant zero itself. The following simple equation will be often used.

\[
\int_0^1 x^k(x-1)^2 dx = \frac{2}{(k+1)(k+2)(k+3)}
\]

The network traffic has certain periodicity for example diel periodicity. Here we assume that, for given traffic data \( \text{trf} : [0,1] \rightarrow \mathbb{R} \), there exists expected value \( e_{01} \in \mathbb{R} \) around the edge of the interval. The problem to be finally resolved is:

find \( p \in P_N \)

such that \( \| \text{trf} - e_{01} \| = \inf_{p \in P_N} \| \text{trf} - e_{01} \| \)

(2)

It is well-known that orthogonal bases of the subspace \( P_N \) is useful to resolve this least-squares approximation problem [5]. Such bases \( \{v_n\}_{n=2}^{N} \) can be easily derived through application of the Gram-Schmidt procedure [5] as follows: \( v_2 = u_2 \) and

\[
v_n := u_n - \sum_{k=2}^{n-1} \frac{\langle u_n, v_k \rangle}{\|v_k\|^2} v_k \quad (n \geq 3)
\]

However, the Gram-Schmidt procedure consists of iterative scheme that requires to compute inner products
with all derived orthogonal components. This implies that computational load rapidly becomes heavy as the degree \( N \) of \( P_N \) increases and accumulation of numerical error due to complicated computation.

III. SMOOTH APPROXIMATION OF TRAFFIC BASED ON FAST GENERATION OF ORTHOGONAL PERIODIC POLYNOMIALS BY THE LANCZOS’S THREE-TERM RECURRENCE RELATION

Multiplication of the variable \( x \) is a linear operator satisfying \((xf, g) = (f, xg)\) for any \( f, g \) (i.e. self-adjoint operator). This property and the linear independence of \((u_n)_{n\geq 2}\) allow us to introduce the Lanczos’s three-term recurrence relation[1–3]. With the relation, we have the orthogonal polynomials \((v_n)_{n\geq 2}\) through the following equations

\[
v_{n+1}(x) = (x - n)v_n(x) \beta_n v_{n-1}(x) \quad (n = 2, 3, \ldots),
\]

where \( v_2 = u_2, v_1 \equiv 0 \) and

\[
n := \langle xv_n, v_n \rangle / \|v_n\|^2, \quad \beta_n := \|v_n\|^2 / \|v_{n-1}\|^2 \quad (n = 2, 3, \ldots)
\]

Therefore, the original orthogonalization procedure in (3) is now reduced to that in (4). The modified formulation is known to be numerically stable and employed for digital filter design [4]. In addition to this simplification, we can derive further reduction of computational load. Note that

\[
wv_m u_n \in P_{W+m+n}, \quad xv_m u_n \in P_{W+m+n+1}
\]

implies that application of (1) to (5) gives \( n \) and \( \beta_n \) only with small number of arithmetic operations.

The proposed smooth approximation method is summarized as follows.

Algorithm 1: For given weighting function \( w \) and maximum degree \( N \), derive periodic orthogonal polynomials \((v_n)_{n=2}^N\) through (4). The optimal periodic polynomial \( p \) of (2) is given by

\[
p = \sum_{k=2}^{N} (\text{trf} \ e_{01}, v_k) v_k \|v_k\|^2^{-1}
\]

Then \( p + e_{01} \) is a smooth approximation of the traffic \( \text{trf} \).

IV. NUMERICAL EXPERIMENT

We employ “Leipzig.1” in NLANR PMA Project[6] as the traffic data to be approximated. The original data is converted into a step function where each step corresponds to total amount of traffic in 5 minutes. The converted data is shown in Fig. 2 as a rapidly fluctuating curve. The traffic around zero o’clock \( e_{01} \) is estimated through ensemble average for 4hours. We also assume an uniform weighting \( w \equiv 1 \) and the maximum degree \( N = 10 \). The proposed method in Algorithm 1 generates a set of orthogonal polynomials (Fig. 1) and provides a smooth approximation of the original traffic (smooth curve in Fig. 2).

ACKNOWLEDGMENT

This work is partially supported by NICT (National Institute of Information and Communications Technology).

REFERENCES

A Study of Optical Fiber Network Capacity Escalation due to Localized Extreme Traffic Increase

Hidenori TAGA
Institute of Electro-Optical Engineering, National Sun Yat-Sen University
No.70 Lien Hai Road, Kaohsiung 804 Taiwan R.O.C.
Tel. +886-7-525-2000, Fax +886-7-525-4499, e-mail hidenoritaga@mail.nsysu.edu.tw

Abstract
Optical fiber network capacity escalation along the traffic increase was studied. Due to localized extreme traffic increase, the required capacity of the local line might be affected significantly while whole network capacity is not affected.

Introduction
A vast traffic of the broadband internet has penetrated in the commercial communication system, and the optical fiber network has become the infrastructure of such system. As the traffic is still growing in a steady pace, the capacity of the network should be expanded regularly to keep the acceptable network performance. One of the most important measures of the performance is the congestion of the network, and it can be quantitatively evaluated by the number of packet losses or the packet loss ratio. The network operators need to maintain the service level of the network keeping the packet loss ratio to be smaller than the target level, and they should introduce additional capacity for the network when the recorded packet loss ratio is approaching the target level.

When the traffic in the network is growing in uniform, the network operators can expect the network congestions based on the historical trend, and they can estimate required additional capacity to reduce the congestions. On the other hand, service providers will launch new services such as movie broadcasting [1], and it will cause localized extreme traffic increase in the network. Such an event will impact on the network congestions, and the network operators might require more network capacity than they are planning.

In this paper, the escalation of the optical fiber network capacity was studied and compared with the uniform traffic increase and the localized traffic increase. A numerical simulator was used to evaluate the optical fiber network. The results showed that the localized traffic increase might not cause significant impact on the whole network capacity, but it could cause significant increase of the required local line capacity of the network.

System model
Figure 1 shows the optical fiber network model used for this study. The topology of this model is based on NSFNET backbone [2].

The traffic in the model was assuming the packet traffic. In the initial condition, the average traffic was set to be the same between any nodes. The number of packets was randomly generated, and its distribution was Gaussian. In addition, the packet length was also random, and its distribution was Gaussian. The routing of the traffic adopted the shortest path method, and two diversity routes were pre-assigned in case for the line congestion. The initial line capacities between the nodes were set to satisfy the entire traffic only by the shortest path.

A Monte-Carlo method was used to evaluate the network performance. Each simulation conducted 100,000 times random assignment of the traffic. In case when congestion was occurred, the traffic was automatically diverted to the pre-assigned routes. The packet loss would occur when all diversity routes were congested, and numbers of the lost packets were recorded to evaluate the packet loss ratio.

The escalation of the network capacity required to maintain the service level was evaluated using this simulation model. The traffic between the nodes was increased for each simulation, and it resulted in the increase of the packet loss ratio. When the packet loss ratio between certain nodes exceeded the target level, the capacities of the lines transferring the traffic between the nodes were automatically increased. The criterion of the packet loss rate was targeted to 5 x 10^-6 in this simulation. By repeating this process, the escalation of the network capacity along the increase of the traffic was obtained. To increase the capacity of the line, a granularity of the additional capacity was adopted in the simulation, because the actual network would introduce additional capacity by the wavelength basis, and it should have a certain granularity. In this
simulation the granularity was set to be the same as the initial traffic between the two nodes.

Results and discussions
The network capacity escalation was evaluated for the uniform traffic increase and the localized extreme traffic increase. For the uniform case, the traffics between the nodes were increased by 5% for each simulation. For the localized case, the traffic from and to a specific node was increased by 10% while the other traffics were increased by 3.8% in order to keep almost identical total network traffic with the uniform case. In addition, for the localized case, two different nodes were arranged as the specific node, and they were node number 6 and 1.

Figure 2 shows the increase of the total network capacity for the uniform traffic increase and the localized traffic increase. Horizontal axes show the time of the simulation, and the vertical axes show the relative capacity assuming the initial capacity as 1. As seen in the figure, there were not so significant differences between three cases.

Figure 2  Evolution of the total network capacity

Figure 3 shows the transition of the packet loss ratio for the entire network. As seen in the figure, the loss ratio repeats increase and decrease, because the simulation program automatically increased the line capacity when the loss ratio exceeded the target rate. For the uniform case, the loss ratio did not keep exceeding the target rate. On the other hand, it sometimes kept exceeding the target rate for the localized case. This implies that the increment of the line capacity for the localized case might not be enough for this simulation.

Even though the total network capacity did not differ so significantly for three cases, local line capacity showed the significant difference. Figure 4 (A) and (B) show the difference of the line capacity escalation for node 6 and node 1. In this figure, the capacity is normalized by the granularity of the capacity increase unit. As seen in the figure, the difference of the line capacity between the uniform case and the localized case for node 6 was relatively moderate, and it was 8% to 42% after 20 times of simulation. On the other hand, the difference was larger for node 1, and it exceeded 80%. These results showed that the localized extreme traffic increase could cause significant impact on the local line capacity escalation.

Figure 3 Transition of the packet loss ratio

Figure 4 Escalation of the local line capacity

Conclusions
Network capacity escalation due to the localized extreme traffic increase was evaluated through the simulation. The results showed that even when the total network capacity became almost identical, the local line capacity could suffer significant impact, and this might cause some problems upon the planning of the network operator, because the escalation is far beyond the expectation.

References
Prefetching Protocol Proxy with Optimal Mirror Selection and Burst Transmission

Tomohiro TSUJI, Junichiro HONMA, Sho SHIMIZU, Yutaka ARAKAWA, and Naoaki YAMANAKA
Department of Information and Computer Science, Faculty of Science and Technology
Keio University, 3-14-1 Hiyoshi, Kohoku-ku, Yokohama, 223-8522, Japan
Email: tsuji@yamanaka.ics.keio.ac.jp

Abstract—In this paper, we propose a new accelerated download mechanism for huge data such as rich contents that can reduce the download time by about seventy percent as compared with the conventional manual operation.

Index Terms—prefetching, mirror server selection, protocol conversion

I. INTRODUCTION

The rich contents such as movies and music transmitted over the Internet have been increasing due to the explosive growth of Web services and the bandwidth of network. Also, larger files are exchanged over the Internet with Peer to Peer application. Then, the flash crowd problem have surfaced as the number of users who access to the Internet [1]. When file size is large, it is difficult to predict the number of accesses in flash crowd from the number of accesses in ordinary times. Therefore, many load-balancing methods such to deliver the large files to a lot of people have been proposed.

In this paper, we propose a new accelerated download mechanism for large data stored in multiple mirrors. Our proposed scheme includes prefetching, automatic optimal mirror selection, and protocol conversion. In the proposed scheme, we measure the response time from the client to all mirror servers. This enables us to save the time to download contents. However considering the delivery of HDTV movie contents, the file size is equivalent to several GBytes. Therefore, the download time increases even if such scheme is used. Then, we apply prefetching technology to this problem. By applying prefetching, it can reduce the feeling download time. Moreover, we propose protocol conversion technology in order to improve the real download time.

In this research, we mount these systems as a proxy and evaluate their performance. Implementation result shows the proposed scheme can reduce the download time by about seventy percent as compared with the conventional manual operation.

II. CONVENTIONAL SCHEME

Several mirror server selection schemes have been proposed for delivering the large files to a lot of people. The First one is server initiation type such as DNS (Domain Name Server) Filter and Round Robin [2] where requests from client are distributed automatically at server side. The Second is client initiation type where mirror server list is given to the client and the client selects the optimal mirror server. The former is chiefly used for the Web server and the streaming server, and has the advantage of transparency, i.e. the client need not consider the mirror server. But the client’s request is not always forwarded to the nearest mirror server. And also when the domain is different, this method cannot be used. While, the latter is used to deliver the software in Sorceforge.net. If there are a lot of mirror servers, the links to each mirror server with country information are often shown to client. The client have to select the best mirror server based on past experience and knowledge, but the client’s request is not always forwarded to the nearest mirror server.

III. PROPOSED SCHEME

In our proposed system the clients are connected to the contents server via prefetching proxy server. And there are a lot of mirror servers which store large files. Our proposed system is composed of prefetching, automatic optimal mirror selection, and protocol conversion. We show that our proposed scheme can greatly reduce the download time.

A. Mirror Selection Scheme & Prefetching

We propose an automatic optimal mirror selection scheme which selects the nearest mirror server among a lot of mirror servers. Figure 1 shows the procedure of our proposed scheme. First, a client accesses a web server through a prefetching proxy server (procedure 1). Second, in response to client’s access, a prefetching proxy server analyzes the HTML document received from Web server, and extracts the links to all mirror servers (procedure 2). Third, a prefetching proxy server selects a download link with shortest response time (procedure 3). Fourth, a prefetching proxy server requests content to the mirror server with shortest response time. A prefetching proxy server prefetches the content and stores in cache (procedure 4). Fifth, if a client clicks the link, the content will be downloaded from prefetching proxy server instead of Web server (procedure 5). Our system can shorten...
the real download time by selecting the optimal mirror server. And also, we can reduce the feeling download time by prefetching the content. Well then, how to select an optimal mirror server is an important issue in our system. We examine two methods such as ping and netselect.

1) Ping: Ping is a program that diagnoses TCP/IP networks such as the Internet and Intranet. In this research, we select the mirror server based on the web access time measured by ping packets. Ping packets are transmitted to all mirror servers after prefetching proxy server extracts the link of all mirror servers, and RTT are measured. After that, prefetching proxy server selects the optimal mirror server.

2) netselect: Nsets is a command to select the nearby mirror server. To use netselect, we have to give the list of mirror servers as the argument. After we set the list, the scores of each mirror server are calculated. Then the mirror server which has the shortest web access time is selected automatically. The calculating formula of the score is as follows. In this research, we employ netselect to select the optimal mirror server of rich contents. The scores are calculated by computing RTT, the number of hops and the download time.

\[
\text{score} = \frac{\text{tmp} \times \text{Number of transmission packets}}{\text{Number of receiving packets}} + \frac{\text{(tmp} \times \text{Number of minimum hops})}{10}
\]

First, tmp is calculated by formula (1). Second, score is calculated by formula (2) using the result of formula (1). By the above formulas, Netselect selects the mirror server which have the minimum score [3].

B. Protocol conversion

Most of the Internet models is Server-Client model by HTTP. Because the request from the client to the contents server is very small signal, there is little possibility that a long delay is generated. But when the content size is very large, the delay increases. So, we propose the protocol conversion scheme to save the real download time. This indicates that HTTP requests are transparently alternated by other protocol such as P2P or FTP.

IV. Evaluation

We evaluate the mirror server selection time and the download time on our experimental system. Figure 2 shows the experimental system of our proposed scheme. There are seven clients connected with the prefetching proxy servers respectively. In this evaluation, since we need many large files distributed in a lot of world wide mirror servers, we use sourceforge.net as the contents server. The object comparison is the time that man downloads the file from the contents server described by HTML with a lot of mirror servers. After this, the total download time indicates the time which include the mirror server selection time and download time. Figure 3 shows the total download time versus the file size. We evaluate the total download time about the six cases. The performance of proposed scheme is greatly improved in total download time compared with the conventional scheme in all case. The conventional scheme fails to select the mirror server which has the shortest web access time when the number of mirror servers increase. However, our proposed scheme can select the optimal mirror server quickly even in that case and also prefetching the content. That is the reason we can reduce the total download time greatly compared with the conventional scheme.

V. Conclusion

In this paper, we have proposed a new accelerated download mechanism for huge data such as rich contents that can reduce the download time by about seventy percent as compared with the conventional manual operation.

REFERENCES


[3] APT HOWTO
An Algorithm for Resource Optimization of Consolidating Two Coexisting Networks

Z.C. Xie, Lian K. Chen, Raymond H.M. Leung, and Calvin C.K. Chan

Department of Information Engineering, The Chinese University of Hong Kong, Shatin, N. T., Hong Kong SAR, China
Tel: +852-2609-8479, Fax: +852-2603-5032, Email: zcxie6@ie.cuhk.edu.hk

Abstract

We proposed an algorithm to derive the minimum number of fiber links required for resource optimization in consolidating two coexisting networks provided that every two nodes in the two networks are bi-directionally connected.

1 Introduction

With the advancement of optical fiber technologies and the surge demand of internet bandwidth in the last decade, there are many optical networks deployed by different parties. They maybe overlapped extensively in one region. It is also envisaged that more fiber links can be saved when transmission links at high data rate (e.g. 40Gb/s or 100Gb/s link) [1] are employed. In the consolidation of two networks to achieve high utilization of network resources, by traffic grooming and rerouting, some of the links can be suspended. Though for the suspended links, the deployed fibers can not be reallocated, the operation cost of the regenerator site can be saved [2]. For instance, if there are two identical and colocated optical ring networks with $d$ nodes and links each, there are totally $4d$ links considering bi-directional communication. It can be shown that the required number of links can be reduced from $4d$ to $d$ by inserting two short interconnections between all the colocated nodes of the two ring networks, thus substantially reduce the link cost by a factor of 4. Based on this, we propose an algorithm to derive the minimum number of links required for arbitrary connected networks. Only networks that can be viewed as planar graphs are considered. A simple equation for the minimum number of links required is derived for certain networks, whereas for others that do not have an exact solution, an upper bound is given.

2 Problem formulation and the algorithm

We model the existing optical network as a connected planar graph and make the following assumptions:

i. There are two identical optical networks in one region.
ii. The cost of every link between two nodes is identical.
iii. The interconnections between two networks only occur at colocated nodes and cost much less than that of a single link, thus their cost is negligible.

The objective is to derive the minimum number of links required so that there is a path from an arbitrary node to all other nodes in the two networks in which all of the links are directed. Assume that all colocated nodes will be interconnected with interconnections. Thus all traffic to and from the nodes on the second network will go through the colocated interconnection and the original links on the second network can be saved.

In graph theory, bridge is an edge (link) whose removal disconnects a graph [3]. We divide the bridges into two types, namely TP-I bridge and TP-II bridge. A leaf is a vertex of degree 1. TP-I bridge is the link incident to a leaf. The other bridges are TP-II bridges. An articulation point (cut vertex) is a vertex whose removal disconnects the graph. The following states our notations.

$L_{\min} =$ Minimum number of links required.
$B =$ Number of bridges in a graph.
$A_i =$ Number of the articulation points with the removal of which the graph will be divided into $i$ subgraphs.

The following states our algorithm. As the two coexisting networks are identical, we concentrate on only one of the networks.

Step 1, remove all the TP-I bridges and the leaves connect to them. Then some of the TP-II bridges will become TP-I bridges. Remove them as stated in step 1 until no TP-I bridges exist.

Step 2, remove all the TP-II bridges. Denote $V$ as the number of vertices remained after this step.

Step 3, remove all the articulation points. Thus the graph is divided into several subgraphs.

Step 4, restore the articulation points to all the subgraphs.

Step 5, check whether the resultant subgraphs are Hamiltonian [4] or not. A Hamiltonian cycle is a cycle that visits each node exactly once. If all the resultant subgraphs are Hamiltonian, we can derive that:

$$L_{\min} = 2B + V + \sum_{i=2} A_i(i-1)$$ (1)

Proof: It is obvious that every bridge needs two links with opposite directions in order that the two end nodes of the bridge can reach each other. For a single articulation point that will divide a graph into $i$ subgraphs, the total vertices after step 4 should be $V + (i-1)$. As the resultant subgraphs are Hamiltonian, the number of links required is equal to the number of vertices. If there are some subgraphs that are not Hamiltonian, we concentrate on the non-Hamiltonian graphs and make the following definitions.

D2 node: Nodes of degree two.

Arm: A path consists of entirely D2 nodes and the connecting links plus the two end links connecting to the adjacent non-D2 nodes.

j-D2 arm: Arms with $j$ D2 nodes.
For example, in Figure 1, there is one 3-D2 arm and two 2-D2 arms in the non-Hamiltonian graph. We denote \( N_i \) as node \( i \) and \( l_{ij} \) as the link connecting node \( i \) and \( j \). One of the two 2-D2 arms consists of \( l_{1,2} \), \( N_2 \), \( l_{2,3} \), \( N_3 \) and \( l_{3,4} \). Then we go to step 6.

**Figure 1.** A non-Hamiltonian graph with three arms.

**Step 6,** find all the arms first, then remove those arms one at a time until all the subgraphs are Hamiltonian. Denote \( M_i \) as the number of \( j \)-D2 arms deleted, and \( V_{hi} \) as the number of remaining vertices in the resultant Hamiltonian graphs. We come to the following equation:

\[
L_{\text{min}} = 2B + \left[ \sum_{j=1}^{r} M_i (j+1) \right] + \sum_{i=2}^{n} A(i-1) \quad (2)
\]

**Proof:** All the \( j+1 \) links of the \( j \)-D2 arms are required in order that all the D2 nodes in the arm can reach other nodes and be reachable from other nodes. So we need at least \( V_{hi} + \sum_{j=1}^{r} M_i (j+1) \) links. On the other hand, if an arm with the end nodes, \( N_i \) and \( N_j \) (they can be same nodes) is added between two nodes, \( N_{\text{in}} \) and \( N_{\text{out}} \) of a Hamilton cycle, all the D2 nodes in the arm can reach \( N_{\text{in}} \) and thus all other nodes via the link \( N_{\text{in}} l_{ij} N_{\text{out}} \). And also, all the D2 nodes in the arm can be reached by other nodes via the link \( N_{\text{in}} l_{ij} N_{\text{out}} \). As for the graph in Figure 1, when we view the arm consists of \( l_{1,2} \), \( N_2 \), \( l_{2,3} \), \( N_3 \) and \( l_{3,4} \) as is added to the Hamiltonian cycle consists of nodes \( N_1 \), \( N_2 \), \( N_3 \), \( N_4 \), \( N_5 \), \( N_6 \), \( N_7 \), \( N_8 \), \( N_9 \) and \( N_{10} \), \( N_{11} \) and \( N_{12} \), \( N_{13} \), \( N_{14} \) and \( N_{15} \), \( N_{16} \) and \( N_{17} \), \( N_{18} \) and \( N_{19} \). Thus all of the subgraphs are Hamiltonian except the first one. We continue to step 6, delete the arm consists of \( N_2 \), \( l_{1,2} \) and \( l_{2,3} \). Thus all of the subgraphs are Hamiltonian and, \( M_{ij} = 1 \), \( V_{hi} = 15 \). So the minimum number of links required is:

\[
L_{\text{min}} = 2B + \left[ \sum_{j=1}^{r} M_i (j+1) \right] + \sum_{i=2}^{n} A(i-1) = 30
\]

Thus 72.2% of the links are reduced considering two networks. One possible result after link reduction is illustrated in Figure 2 as indicated by the 30 arrows.

**Figure 2.** Two coexisting identical networks with interconnections between all the colocated nodes.

For step 1, we delete \( N_5 \) and \( l_{2,6} \), \( N_{18} \) and \( l_{19,20} \), \( N_5 \) and \( l_{17,18} \). Then delete also \( N_{10} \) and \( l_{17,19} \). For step 2, we delete \( l_{5,7} \) (TP-II bridge) and \( B=5 \). \( V=16 \). After step 4, we derive five subgraphs, namely the subgraphs consist of nodes 1-2-3-4-5, 7-8-9-10-11, 11-12-13, 11-14-15 and 15-16-17 and \( A_{f}=1 \) \( (N_{13}) \), \( A_{f}=1 \) \( (N_{17}) \). All of the subgraphs are Hamiltonian except the first one. We continue to step 6, delete the arm consists of \( N_2 \), \( l_{1,2} \) and \( l_{2,3} \). Thus all of the subgraphs are Hamiltonian and, \( M_{ij} = 1 \), \( V_{hi} = 15 \). So the minimum number of links required is:

\[
L_{\text{min}} = 2B + \left[ \sum_{j=1}^{r} M_i (j+1) \right] + \sum_{i=2}^{n} A(i-1) = 30
\]

Thus 72.2% of the links are reduced considering two networks. One possible result after link reduction is illustrated in Figure 2 as indicated by the 30 arrows.

**4 Summary**

We investigated the minimum number of links required in consolidating two duplicated networks to make every two nodes bi-directionally connected. We proposed an algorithm to derive the minimum number of links and found that it is at least the number of the nodes or at least twice the number of the bridges in one of the networks. The algorithm can be extended easily for the consolidation of two non-identical networks.

This work is supported in part by HK CERG Grant CUHK411006.

**5 References**

3 Reinhard Diestel, Graph Theory, Third Edition, pp. 11.
4 Ronald Gould, Graph Theory, pp. 131-147.
IP over Optical Packet Switched Network

Hideaki Furukawa¹, Naoya Wada¹, Hiroaki Harai¹, Makoto Naruse¹, Hideki Otsuki¹,
Tetsuya Miyazaki¹, Katsuya Ikezawa², Akira Toyma², Naoki Itou², Hiroshi Shimizu², Hiroshi Fujinuma³,
Hatsushi Isiduka³, Gabriella Cincotti⁴, and Kenichi Kitayama⁵

¹: National Institute of Information and Communications Technology, 4–2–1, Koganei, Tokyo 184-8795, Japan. Phone:+81-42-327-5694, Fax:+81-42-327-7035, E-mail:furukawa@nict.go.jp
²: Yokogawa Electric Co., 2–9–32, Nakacho, Musashino, Tokyo 180-8750, Japan
³: NTT Electronics Co., 1841-1, Tsuruma, Machida, Tokyo 194-0004, Japan.
⁴: University Roma Tre, via della Vasca Navale 84, I-00146 Rome, Italy
⁵: Osaka University, 2-1 Yamadaoka, Suita, Osaka 565-0871, Japan

Abstract

We demonstrate 10-Gbit-Ethernet/80-Gbps-optical-packet transporting by employing developed IP-OP/OP-IP converters, packet-transceiver with instantaneous-locking (<1 ns), and multiple-label-processor with ultra-fast processing (<100 ps) to achieve 7.2-Gbps IP packets throughput with low packet-loss-rate (<10⁻⁶).

Introduction

Optical packet switching (OPS) is an essential technology toward over-40-Gbps packet-switched network. To connect such high-speed OPS technology to real Internet technology seamlessly, transmission of IP packets over optical packets (OP) has been achieved (e.g., 2.5 Gbps [1,2] and 40 Gbps [3]), and 40 Gbps electronic routers are available.

To connect IP network with over-40-Gbps OPS networks with effective way, we have developed novel interfaces (called as IP-OP or OP-IP converters) between 80 Gbit/s OPS networks and 10 Gb Ethernet (10GbE) and shown the specification [4]. In addition, we have demonstrated 10 Gbps IP packet transmission over 80 (8x x 10) Gbps WDM-based optical packets by using the novel IP-OP and OP-IP converters, an 8-channel array truly burst-mode packets transceiver, and multiple optical code (OC) label processing [5]. In this paper, the compilation of these works is presented.

IP over OPS and its key technologies

Our proposed IP networks are illustrated in Fig. 1. Some networks may consist of Ethernet technology and others may be SONET/SDH. OPS networks are connected to those networks. At the ingress edge node of the OPS network, the IP-OP converter generates an optical label according to a look-up table and the destination address of an IP packet and then encapsulates the IP packet into an optical packet. For example, IP packets of which destination addresses are 192.168.0.1 is given optical label “A”. At the egress edge node, the OP-IP converter decapsulates the IP packet from the optical packet and again encapsulates the IP packet into appropriate frame (e.g., 10GbE frame) according to a look-up table. In OPS networks addressed in this paper, we introduce colored optical packets by using WDM-based packet compression [6]. As shown in Fig. 1, each optical packet consists of multiple wavelengths payloads of which approximate speed is 10 Gbps. By encapsulating 10GbE frame into multiple wavelengths payloads, we can use full bandwidth of OPS networks and the fine granularity can be achieved.

Figure 2 shows the experimental setup of IP packet over OPS. At ingress nodes, IP-OP converter extracts IP packets from 10GbE frames and generates an OC label selection signal from the IP address after header processing. 10 Gbps IP packets are converted into 8-channel 10 Gbps payloads in the IP-OP converter and re-converted optical signals at 8 different wavelength by developed burst-mode transmitter. By WDM, the data rate of optical packet is 80 (8x x 10) Gbps and the payload length in time is one-eighth of original 10 Gbps packet. Providing eight 10G-Ethernet ports, we can use the full bandwidth of 80 Gbps OPS networks even in the edge nodes. Here, eight electro-absorption modulators (EAM) with distributed feedback (DFB) lasers at different wavelength are arrayed as an 8 channel E/O converter. To be compatible with burst-mode packets, reverse bias voltage applied to EAM portions in OPS system, optical packets are switched based on label information. For multiple optical label processing, we first introduce 200 Gchip/s multiple label encoder/decoder with an arrayed waveguide configuration. It can generate and...
recognize simultaneously sixteen 16-chip optical phase shift keying codes with low latency [7]. The processing rate is 13 Gpacket/s. In matched case, label processor outputs auto-correlation signal with high peak. This signal is used to control optical switch.

At egress node, each optical packet is demultiplexed into 8 wavelengths and received by 8-channel array burst-mode 3R-receivers. Instantaneous clock and payload data recovery for burst-mode optical packets can be realized in less than 1 ns. The skew among 8-channel electrical payloads is adjusted by a phase shifter array. Finally, OP-IP converter recombines and converts these payloads into IP packets on 10G-Ethernet frame again.

**Experimental results**

First, to show an output 8-bit label-signal from IP-OP converter, three IP packets with different IP addresses were transmitted. We can set an arbitrary 8-bit label-signal pattern according to destination address of IP packet. Figure 3 shows the experimental results of generated 8-bit label-signals [11111111], [00000010], and [10101010] according to IP address: 192.168.0.1, 192.168.0.2, and 192.168.0.3, respectively. It is confirmed that different 8-bit label-signals were output according to destination IP addresses. Next, IP packets over 10G-Ethernet frames were transmitted from a network analyzer. The frame length was 1500 byte. The transmitted rate was from 7.2 Gbps to 3.6 Mbps. The duration of one optical packet is fixed at about 514 ns by current specification of IP-OP converter. Figures 4(a), 4(b), and 4(c) show the packet pattern, the waveform measured by a sampling oscilloscope and the spectrum of optical packets @ 3.6 Gbps, respectively. The wavelength spacing is 100 GHz. Figure 4(d) is the eye-diagram of an optical payload at a transmitted rate of 7.2 Gbps. In Fig. 4(h), instantaneous recovery of clock (lower part) and reversed payload (upper part) @ 3.6 Gbps are shown. The frame loss rate was measured when the transmitted rate was changed. The measurement time is 60 s. Figure 4(i) shows that the frame loss rate is less than 10^-6 at 7.2 Gbps.

**Conclusions**

We have demonstrated the transmission of 10 Gbps IP packets over WDM-based 80 Gbps OPS. The packet loss rate under 10^-6 was achieved when the transmitted rate was higher than 3.6 Gbps. Details of experiments will be given in presentation.

**Acknowledgement:** The authors thanks to T. Makino, H.Sumimoto, and Y.Tomiyama.

**References**

Demonstration of a Semiconductor-based Multi-Wavelength Light Source and its Application to Optical Label Processing

Kensuke Okamoto¹, and Hiroyuki Ueno-hara¹²

1 Microsystem Research Center, P&I Laboratory, Tokyo Institute of Technology 4259 Nagatsuta, Midori-ku, Yokohama-shi, Kanagawa 226-8503, Japan TEL./FAX : +81(45)924-5038, email : unohara.h.aa@m.titech.ac.jp
2 JST-CREST, 1-32-12 Higashi, Shibuya-ku, Tokyo 150-0011, Japan

Abstract A semiconductor-based multi-wavelength light source for optical label processing scheme is demonstrated. Two cascaded electroabsorption modulators generated multi-wavelength optical spectrum, and coding could be achieved with one of sidebands filtered by a fiber Bragg grating.

1. Introduction

Optical label processing scheme is one of the most technical key points for realizing optical label switching techniques[1]-[5]. We have proposed and demonstrated time-domain separated wavelength division multiplexed (TS-WDM) label scheme[6] as shown in Fig.1. In this scheme, each bit of optical label is wavelength-multiplexed in the same time slot, and it is separated with payloads in time. Therefore, optical labels with slower bit rate than that for payloads does not cause network efficiency degradation. And wavelength bandwidth of optical labels can be shared with that of payloads, and this suppresses the degradation of wavelength utilization efficiency. However, discrete DFB lasers were used as a multi-wavelength light source in Ref.[6], and realization of the compact light source is desirable.

In this paper, we propose the optical label generator using semiconductor-based multi-wavelength light sources. Multi-wavelength generation and optical coding using one of the generated sideband will be presented.

2. Structure of Semiconductor-based Multi-Wavelength Light Source

Schematic structure of the optical label generator and label processor using the multi-wavelength light source is shown in Fig.2. Optical label is coded by separating each sideband emitted from the multi-wavelength light source using an arrayed waveguide grating (AWG) and gating each signal with optical gates such as electroabsorption modulators (EAM’s). In the label processor, each wavelength signal for optical label is separated with an AWG. After optical-to-electrical (OE) conversion, signal levels are recognized by comparators, and then the label pattern is processed by the address table. Slow electronics can be used because bit rate of optical label is smaller than that of payloads as shown in Fig.1. The multi-wavelength light source we propose is indicated in the inset of Fig.2. CW light emitted from a DFB laser is modulated with two-stage EAM’s by sinusoidal RF signals with frequency of the same as the bit rate of optical label. By intensity and phase modulation in the EAM, multi-sidebands are generated just like a comb generator.

Fig.1 Conceptual image of optical packet frame.
(a) Time-axis: Each bit of multi-wavelength optical label is wavelength-multiplexed in the same time slot, followed by payload.
(b) Wavelength-axis: Optical label is assigned within the payload bandwidth.
(c) Wavelength-axis: Payloads are wavelength-multiplexed.

Fig.2 Schematic structure of a multi-wavelength optical label generator, label processor, and a multi-wavelength light source.

Fig.3 Experimental setup of optical label generation using a semiconductor-based multi-wavelength light source, a fiber Bragg grating and an electroabsorption modulator.
with LiNbO$_3$ modulators. All components are made of semiconductor, and so integration in one chip and resultant compact size could be realized.

3. Experimental setup

Experimental setup for optical coding using multi-wavelength light source is shown in Fig.3. The 1st stage EAM was modulated with a RF synthesizer with frequency of 12.5GHz. The bias voltage and the RF voltage amplitude were -2.4V and 5.0V, respectively. In the 2nd stage EAM, the bias voltage of -2.0V and the RF voltage amplitude of 4.5V were used. In these experiments, generated multi-wavelength signal was injected into a fiber Bragg grating (FBG) with the center wavelength of 1551.6nm and the reflection bandwidth of 0.1nm instead of an AWG. The reflected signal was coded with an EAM with 1Gbps 2$^{31}$-1 PRBS format. After OE conversion, the waveforms were observed with a sampling oscilloscope and bit error rate (BER) was measured with a BER tester.

4. Experimental results

Optical spectrum of the generated multi-wavelength signal is indicated by the solid line in Fig.4(a). Peak to valley ratio of more than 20dB and the power difference between the center and 0.4nm-separated sideband of 15dB were observed. Optical spectrum after filtering 1551.6nm wavelength signal with a FBG is shown in the dotted line in the same figure. Side-mode suppression ratio of more than 30dB could be obtained, and this is large enough for optical coding.

Eye diagram of the coded signal is shown in Fig.4(b). In comparison, CW light was coded with an EAM and its eye diagram is indicated in Fig.4(c). In both cases, clear eye openings could be achieved.

BER measurement results are indicated in Fig.4(d) for both cases. We could see no power penalty between them, and this indicates that there is no noticeable degradation in coded signal using the proposed multi-wavelength light source.

In summary, we proposed and investigated the characteristics of semiconductor-based multi-wavelength light source consisting of a DFB laser and two-stage EAM’s. We could obtain multi-wavelength signals, and we confirmed that the coding signal with 1Gbps PRBS format had no power penalty and no noticeable degradation.

Acknowledgment

We would like to thank Prof. Emeritus K. Iga, Prof. K. Kobayashi, Prof. F. Koyama and Assoc. Prof. T. Miyamoto for their encouragements and discussions.

References


SubFrame-Based Slot Reservation Scheme for Minimizing Transmission Delay in Optical Slot Switching Network

Fumiko Uehara, Teruo Kasahara, Masahiro Hayashitani, Daisuke Ishii, Yutaka Arakawa, Satoru Okamoto, and Naoaki Yamanaka
Dept. of Information and Computer Science, Faculty of Science and Technology, Keio University, Hiyoshi 3-14-1, Kohoku-ku, Yokohama, Kanagawa 223-8522, Japan
Tel: +81-45-560-1263, Fax: +81-45-560-1262,
Email: uehara@yamanaka.ics.keio.ac.jp

Abstract—We propose SubFrame-based slot reservation scheme for minimizing the transmission delay in optical slot switching (OSS) network. The computer simulation shows that our proposed scheme can reduce the transmission delay about 25% by using SubFrames.

I. INTRODUCTION

Optical Slot Switching (OSS) [1] architecture is a new scheme for the efficient data transfer in optical network. In OSS network, all nodes are synchronized and share the fixed-length time-period named slot. Each slot carries multiple packets as shown in Figure 1 (c). To realize efficient data transfer, a source node reserves the next slot before data transfer. By reserving the next slot, a source node can transmit data without the effect of the present slot reservation. Now, although the problem of guard time between slots has been discussed earlier [2], the slot reservation in OSS network has not been discussed yet. In OSS network, a source node uses one fixed-length slot for transmitting any size of data. It causes no data transfer in a slot. In this paper, we focus on this problem and study the reservation scheme which uses one or more slots variably to achieve short transmission delay.

II. SLOT RESERVATION SCHEMES

We assume that a node in OSS network uses two or more slots for transmitting large-size data which is beyond the slot-length. Fig. 1 shows three slot reservation schemes. Fig. 1 (a) shows MultiSlot Reservation (MSR). MSR uses one frame for data transfer and reserves slots continuously. In Fig. 1, a frame consists of 6 slots. Data transfer time of MSR is within a little time. Fig. 1 (b) shows MultiFrame Reservation (MFR). MFR uses one slot-number and reserves slots one by one per one frame periodically. Transmission delay of MFR is longer than that of MSR. However, the number of clients in one frame is larger than that of MSR. We consider the performances of MSR and MFR, and propose SubFrame-based slot Reservation (SFR). Fig. 1 (c) shows proposed SFR. SFR uses MSR and MFR variably according to the data size and the request occurrence time. SFR uses 2 SubFrames named SubFrame 1 and SubFrame 2. In Fig. 1 (c), a SubFrame consists of 3 slots. In SubFrame 1, SFR reserves slots by MSR. In SubFrame 2, SFR reserves slots by MFR.

III. PROPOSED SCHEME

Fig. 1 shows an example of the three schemes. Fig. 1 (a) shows a MSR example. Client 1 requests 4 slots at slot 5 of Frame 1. Client 1 checks Frame 1. At slot 5, there remains only 2 vacant slots in Frame 1. So, Client 1 checks Frame 2. In Frame 2, there remains 6 vacant slots. So, Client 1 reserves slot 1 ∼ slot 4 of Frame 2. Client 2 requests 4 slots at slot 1 of Frame 2. At slot 1, there remains only 2 slots in Frame 2. So, Client 2 checks Frame 3. In Frame 3, there remains 6 vacant slots. So, Client 2 reserves slot 1 ∼ slot 4 of Frame 3. Fig. 1 (b) shows a MFR example. Client 1 requests 4 slots at slot 5 of Frame 1. Client 1 checks slot 5 ∼ slot 6 in Frame 1 ∼ Frame 6 and slot 1 ∼ slot 4 in Frame 2 ∼ Frame 7. Slot 5 of Frame 1 is the earliest available slot. So, Client 1 reserves slot 5 of Frame 1 ∼ Frame 4. Client 5 requests 3 slots at slot 1 of Frame 3. Client 5 checks slot slot 1 ∼ slot 6 in Frame 3 ∼ Frame 8. Slot 2 of Frame 3 is the earliest available slot. So, Client 5 reserves slot 2 in Frame 3 ∼ Frame 5. Fig. 1 (c) shows a SFR example. Clinet 1 requests 4 slots at slot 5 of Frame 1. Clinet 1 checks SubFrame 2 of Frame 1. Slot 5 of Frame 1 is the earliest available slot. The data size is beyond the SubFrame-length. So, Clinet 1 chooses MFR. So, Client 1 reserves slot 5 in Frame 1 ∼ Frame 4. Client 2 requests 4 slots at slot 1 of Frame 2. Client 2 checks SubFrame 1 of Frame 2. There remains 3 vacant slots. The data size is beyond the number of vacant slots. So, Client 2 reserves 3 slots in SubFrame 1 of Frame 2 and 1 slot in SubFrame 2 of Frame 2. Client 2 checks SubFrame 2 of Frame 2. Slot 4 of Frame 2 is the earliest available slot. So, Client 2 reserves slot 1 ∼ slot 4 of Frame 2. Client 3 requests 3 slots at slot 4 of Frame 2. Client 3 checks SubFrame 2 of Frame 2. Slot 6 is the earliest available slot. The data size is not beyond the SubFrame-length. So, Client 3 checks SubFrame 1 of Frame 3. There remains 3...
We compare the transmission delay of MSR, MFR, and proposed SFR by the computer simulation. Frame-length is 20 slots. SubFrame-length is 10 slots. Access speed is 10 Gbps. The slot period is set to 10 msec. The input traffic follows rectangular distribution (1 slot ~ 16 slots). Figure 2 shows the transmission delay of MSR, MFR, and SFR.

We assume that the load is the request occurrence rate per slot and that the transmission delay is the time from request occurrence to the end of data transfer (When the load is 0.1, the average slot utilization is 0.8). When the load is low, MSR achieves the shortest transmission delay. It is because MSR reserves all slots continuously. When the load is over 0.12, SFR reduces the transmission delay about 25% than MSR. It is because that SFR compare $Time_{MS}$ and $Time_{MF}$ and choose the earlier one, while MSR can’t find continuously vacant slot. MFR reserves any size of data periodically and makes the transmission delay longer.

V. CONCLUSION

In this paper, we have proposed SubFrame-based slot reservation which minimizes the transmission delay by using MSR and MFR efficiently. The computer simulation showed that our proposed slot can reservation reduce the transmission delay about 25% by using SubFrames.

REFERENCES


Analytic Model for Optical Packet Switch with Output Variable All-Optical Buffers under Asynchronous Variable-Length Packet Traffic

Hyunjoo Lee, Changho Yun, Wansu Lim, and Kiseon Kim
Department of Information and Communications, Gwangju Institute of Science and Technology, 1, Oryong-dong, Buk-gu, Gwangju, 500-712, Korea Tel: +82-62-970-2252 Fax: +82-62-970-2274 Email: mabewo@gist.ac.kr

Abstract—We propose an analytic model of the output-queued optical packet switching (OQ-OPS) router using variable all-optical buffers in an asynchronous and variable length packet network scenario, and measure the packet loss probability.

I. INTRODUCTION

Due to enormously increased data traffic, a next-generation switching system should be reliable, scalable, flexible and power-efficient. An optical packet switching (OPS) router is a strong candidate by applying wavelength division multiplexing (WDM) and optical-label switching (OLS) technologies [1].

In the OPS router, a contention resolution is a significant issue, because it has a great effect on the network performance in terms of packet loss and latency. The OPS resolves the contention in the wavelength, time, and space domains [1]. Among these three domains, the time domain solution using optical buffers is not only the simplest and most cost effective but also crucial to decide the OPS architecture [2]. Common technologies of optical buffering are based on fiber delay lines (FDLs). Several papers have shown the performance of various OPS architectures with FDL buffers through analysis and simulation under many network scenarios [2], [3], [4]. However, the FDLs can provide limited unscalable buffer capacity and coarse delay, because they have a discrete delay unit called granularity [5]. Recently, all-optical variable-length buffers are proposed by several novel technologies such as folded-path architecture [6], innovative "slow-light" all-optical buffer [7], and time-slot assignment (TSA) functions [8], etc. For a future switching system, buffering strategies applying the variable all-optical buffers might be required for better performance. Some papers propose several architectures applying all-optical variable buffers and simulate the performance [10], [11]. However, these works are limited to the simulation results, excluding analytic works even though analytic works are effective to measure the performance as well as design OSP architectures.

In this paper, we propose an analytic model of the basic and simple variable buffer-based OPS architecture, output-queued OPS (OQ-OPS) [10], and analyze packet loss probability (PLP) using the proposed model under the practical scenario which supports asynchronously arriving variable-size packets for the well-matched high-speed IP networks [9].

![Diagram](image-url)

Fig. 1. An illustration obtaining the waiting time, system time, and exemplifying a packet loss occurrence of \((2W + 1)^{th}\) packet.

II. ANALYTICAL MODEL OF OQ-OPS WITH VARIABLE ALL-OPTICAL BUFFER

The OQ-OPS architecture is composed of \(N\) input and output fibers and each fiber has \(W\) wavelengths. \(D\) is the number of output queues using variable all-optical buffers on each wavelength channel, so that the total number of output buffers per output port is \(DW\). \(c\) denotes each buffer’s maximum capacity, so that a total buffer capacity is \(C = D \cdot c\).

When a packet arrives at the router, a switching unit converts its wavelength to forward the desired fiber because of its routing characteristics depending on the wavelength. When a packet needs to be buffered for resolving contention, its wavelength is converted to a particular wavelength, because the variable buffer is engineered to delay the packet in a specific wavelength [10]. The variable buffer delays the packet until the desired output wavelength channel is available.

Fig. 1 describes behavior of the arrival packets. The first \(W\) packets do not need delay, because the wavelength channels are free. At \(t_{W + 1}\), \((W + 1)^{th}\) packet arrived, but this should be buffered at the \(1^{st}\) buffer since there are no free channel. We denote \(Z_{W + 1}\) as the waiting time of this packet. \(Z_{W + 1}\) can be calculated by subtracting its inter-arrival time between the \(1^{st}\) packet and the \((W + 1)^{th}\) packet from the \(1^{st}\) packet.
length as shown in Fig. 1. Until the $2W^{th}$ packet, no packet loss occurs. Upon arrival of the $(2W + 1)^{th}$ packet at $t_{2W+1}$, it should be buffered at the $2W^{th}$ buffer, because the $1^{st}$ buffer is occupied at that time. $Z_{2W+1}$ is $(t_{2W+1} + Z_{2W+1}) - (t_{2W+2} + t_{2W+3} + \ldots + t_{2W+1})$, where $t_{2W+1} + Z_{2W+1}$ is the system time of the $(W + 1)^{th}$ packet. However, as $Z_{2W+1} > c$, this packet cannot be buffered, so that a packet loss occurs.

For analytically modelling, we assume that the packet inter-arrival time is exponential distribution $f_r(t)$ with arrival rate $\lambda$, and the packet length is exponential distribution $f_c(v)$ with an average length of $\frac{1}{c}$. $\nu$ is equal to $\frac{1}{c}$ as the average service time of each channel. The traffic load from $W$-wavelength channels is assumed to be $\rho = \frac{\lambda}{\nu}$, because $W$ wavelengths can correspond to $W$ servers. Through these assumptions, we can take a $M/M/W/DW$ Markov chain model whose state is the number of packets in one output port. For state $i \geq W$, the transition rate from the $i^{th}$ to the $(i + 1)^{th}$ can be expressed as $\lambda(1 - \beta_i)$, where $\beta_i$ is the PLP at state $i$ in the case of $Z_i > c$. The service rate at state $i$ is $\mu_i = \min(i, W)\mu$ due to the limited number of $W$ servers with rate $\mu$.

To verify the analytic model of OQ-OPS, we consider the average packet loss probability $\beta$ as a performance measurement. To obtain $\beta$ as well as the state probability $P_i$, we need to derive $\beta$. Aforementioned, the packet loss occurs when the waiting time is greater than a buffer capacity. When all buffers are occupied, namely state $i = (D+1)W$; the packet loss also occurs. In the case of $Z_i > c$, the packet loss probability $\beta_i$ can be expressed as:

$$\beta_i = \int_c^\infty f_{Z_i}(z)dz, \quad (1)$$

where $f_{Z_i}(z)$ is the probability density function (pdf) of $Z_i$. In Fig. 1, $Z_i$ can be express as $S_i-(W-n)\cdot t^*$, where $S_i$ is system time and $i \geq W$. If $n$ packet losses have occurred previously, $Z_i$ is $S_i-(W-n)\cdot t^*$. Here the sum of $t^*$ is distributed according to an Erlang distribution, $f_{\Sigma t^*}(t) = \frac{\lambda^\nu}{\nu!}t^{\nu-1}e^{-\lambda t}$, where $n \geq 0$. Since $S_i$ and sum of $t^*$ are independent, $f_{Z_i}(z)$ is given by the convolution of $f_{S_i}(z)$ and $f_{\Sigma t^*}(\cdot)$ [4], where $f_{S_i}$ is the pdf of $S_i$. $f_{S_i}$ can also be obtained by the convolution of $f_{\nu_i}(\cdot)$ and $f_{Z_i}(\cdot)$ due to the sum of independent random variables, $\nu_i + Z_i$. For $i < W$, $S_i$ equals to $\nu_i$, because $Z_i$ is 0. By applying the values of $\beta_i$ solved by means of iterations, the state probabilities $P_i$ can be calculated by the queuing theory formula [2]. Finally, the average packet loss probability, $\beta$ is:

$$\beta = \sum_{i=W}^{(D+1)W} \beta_i P_i + P_{(D+1)W}. \quad (2)$$

III. NUMERICAL RESULTS

In this section, we measure the PLP through the proposed OQ-OPS analytic model. Fig. 2 shows the PLP versus $D$ for different numbers of wavelengths and total buffer capacities per wavelength channel with $\rho = 0.8$ and $\nu = 400$. The 3 solid lines show the case of employing 4 wavelength channels, and the 2 dotted lines shows that of employing 10 wavelength channels.

As the shown result, increasing the number of buffers does not always have better PLP. The reason is that each buffer capacity is $C/D$, so that the more buffers we use, the less each buffer capacity due to a fixed total buffer capacity $C$. If $C$ is enough large, the effect to increase buffers could be enhanced. Intuitively, when more wavelengths we use, we obtain much better PLP than using more buffers, because the wavelength means the server. Furthermore, the more number of wavelengths, the less effect to increase buffers. An optimum point can be different according to the average packet length, traffic load, buffer capacity and wavelength number.

IV. CONCLUSION

In this paper, we propose the analytic model of the OQ-OPS using variable all-optical buffers in an asynchronous network with variable length packet. Through this proposed model, we measure the PLP, and obtain an approximate optimum number of buffers. This model can be useful to analyze other performance measurements, and extend the OPS model as applying various algorithms and assumptions such as burst packet arrival.

ACKNOWLEDGMENT

This research was supported by the MIC, Korea, under the ITRC support program supervised by the IITA (IITA-2006-C1090-0603-0007)

REFERENCES


555
Throughput improvement of deflection routed networks with an all-optical packet scrambler

C.Y. Li and P. K. A. Wai

1Photons Research Centre and Department of Electronic and Information Engineering, The Hong Kong Polytechnic University, Hung Hom, Hong Kong

Phone: +852 2766-6231, fax: +852 2362-8439, email:{enli, enwai} @polyu.edu.hk

Abstract: The throughput of deflection-routed networks are often lower than the expected because of the strong correlation between packets which can be removed by an all-optical packet scrambler.

1. Introduction

Deflection routing is an important alternative to the use of buffers to resolve output contention in optical packet-switched networks [1, 2]. One of the shortcomings of deflection routed networks is that, because of the strong correlation of packets in different time slots, the system performance is very sensitive to traffic distribution [3, 4]. Traditionally, the deflection routing nodes are assumed to have sufficient computational power to resolve the problem. However, this assumption is not valid in the optical packet-switched networks because only simple optical logic devices are available for ultrafast optical signal processing [2]. To improve the system performance, a straightforward way is to scramble the packets to reduce the packet correlation. In the past, scrambling was not considered in deflection routed networks because of the additional delay incurred by scrambling, e.g., there will be a delay of at least two time slots to interchange a pair of packets. The scrambling induced delay will not be justified if each packet deflection only costs an extra delay of three or four time slots. In future optical networks, the propagation time between nodes can be much larger than the packet transmission time. The delay caused by the packet scrambling will be negligible if compared to the deflection extra delay. It will then be useful to use an all-optical packet scrambler to reduce the correlation between packets in different time slots.

2. All-optical packet scrambler

Figure 1 shows the proposed two-stage all-optical packet scrambler. The incoming packets are grouped into pairs and the packets in the pairs are reordered by the scrambler at random. The scrambler can be added to the inputs or outputs of the nodes of a deflection routed network. In Fig. 1, there are two modules SB 1 and SB 2 each of which has four switches (SW 1 to SW 4) and two sets of fiber delay lines (FDL 1 and FDL 2). Time is equally divided into time slots S0, S1, ... each of which has a duration of one packet transmission time. We assume that each FDL in Fig. 1 has delay time of exact one slot for the storage of one packet. The switching time of all switches in Fig. 1 is assumed to be negligible. The switches SW A and SW B are in connection setting (0 − 1) at the start of time slot S0 and change to (0 − 2) after the passing of two time slots, i.e., the start of S2. Assume that the scrambler is installed at an output O1 of a node N1. During the two time slots S0 and S1, switches SW 1 to SW 4 of module SB 1 are all in connection setting (0 − 1). In the mean time, two time slots of bits are sent into module SB 1 from node output O1 via switch SW A, and are completely stored in optical fiber delay lines FDL 1 and FDL 2 when time slot S2 starts. Note that the two time slots of bits can be data bits for packets or dummy bits for idle slots. At the start of time slot S2, the switches SW 1 to SW 4 take one of the connection settings (0 − 1) and (0 − 2) at random. If the connection setting (0 − 1) is taken, the bits in fiber delay line FDL 2 will move out via switches SW 3 and SW 4 to switch SW B during the time slot S2. When the bits in FDL 2 moves out, those in fiber delay line FDL 1 will move into FDL 2 via switches SW 2 and will be sent out via switches SW 3 and SW 4 in the time slot S3. If switches SW 1 to SW 4 take connection setting (0 − 2) at the start of time slot S2, the bits in fiber delay line FDL 1 will be sent out via the switches SW2, SW 4 and SW B. Those in fiber delay line FDL 2 will move into FDL 1 via switches SW 3 and SW 1, and will be moved out via switches SW 2, SW 4 and SW B similarly in time slot S3. In parallel to the transmissions in module SB 1, slots of bits from node output O2 continues to fill up the fiber delay lines in module SB 2 during time slots S2 and S3. The operations similar to that in module SB 1 are repeated in module SB 2 from time slots S2 to S4 (the switches SW A and SW B will be reset to the connection setting (0 − 1) at time slot S4). As the time slot cycle continues, the proposed set up in Fig. 1 will provide the...
required packet scrambling function. To construct larger scramblers, we can either follow the approach in Fig. 1 or use the two-stage packet scramblers as the building blocks.

3. Performance evaluation

Figure 2 shows the throughput performance of using the proposed packet scrambler on a 10×10 deflection-routed Manhattan Street Network (MSN) that supports multicast services [4, 5]. For an r×c MSN, there are r rows and c columns of nodes. We label the nodes from left to right, and top to bottom. The node in the upper left corner is labeled 1, and that in lower right corner is labeled N, i.e., N = r×c. In the simulations of Fig. 2, we use the same multicast tree as that in [4], Fig 3 : {(1 → 2 → 3 → 4 → 5 → 6 → 7 → 17 → 27), (1 → 11), (3 → 13), (5 → 15)} . Node 1 is the root node of the tree that generates new multicast packets. The multicast packets are then sent along the tree to the multicast destination nodes. At the branch nodes 1, 3, and 5, the multicast packets are duplicated and forwarded to the downstream nodes.

In the simulations, the node-to-node propagation time is 100 slot times. We assume that a node has at most one new arrival packet per time slot but can receive two packets simultaneously. Apart from the multicast traffic, there is also background unicast traffic. The probability that a new packet arrives at a node is the offered load. The unicast load offered to each node is set to 0.1. We also assume that a node sends unicast packets uniformly to each node in the network except itself. When a multicast packet fails an output contention, it is deflected and will have to return to the same deflection node to continue the multicast, i.e., the back-to-the-deflection-node (BDN) scheme in [4]. The deflection of a unicast packet, however, has no such limitation. As the result shown in [4], the correlation between packets in different time slots will be strong if the BDN routing is used. Consequently, the BDN routing scheme is not recommended even though it is comparatively easier for the all-optical implementation because of the worst throughput performance [4].

In Fig. 2, the solid curve is the throughput performance of the multicast packet routing without the proposed packet scrambler, i.e., the BDN routing scheme. The curves with circles and squares are those with the proposed 2-stage and 4-stage packet scramblers. From Fig. 2, we observe that there is significant improvement if the proposed packet scramblers are used. The throughput of the packet routing without scrambler has the maximum value 0.22 at loading of 0.4 and drops to 0.17 at loading of 0.9. In contrast, the throughput of the packet routing with 2-stage scrambler has the maximum value 0.32 at loading of 0.7 and drops slightly to 0.31 at loading of 0.9. The proposed packet scrambler not only increases the network throughput but also improves the system stability. We also find that there is very little advantage in using 4-stage instead of 2-stage scramblers. Thus large scramblers are not necessary in most cases.

4. Conclusion

In this paper, we propose an all-optical packet scrambler to reduce the correlation between packets in different time slots in deflection-routed networks such that the system throughput performance is improved. We propose a design of the two-stage packet scrambler. We can also use the two-stage packet scramblers to build larger scramblers such as 4-stage and 8-stage. From simulation results, we observe that the two-stage packet scramblers are sufficient for most of the situations.

Acknowledgement: This research is supported by a grant from The Hong Kong Polytechnic University.

References:

All-Optical NRZ-to-RZ Format Conversion based on Optical Parametric Amplifier

Henry K. Y. Cheung, Rebecca W. L. Fung, P. C. Chui, and Kenneth K. Y. Wong

Department of Electrical & Electronic Engineering, The University of Hong Kong, Pokfulam Road, Hong Kong.
E-mail: kycheung@eee.hku.hk

Abstract: We demonstrate a 10Gb/s modulation-format conversion from nonreturn-to-zero to return-to-zero using pulsed-pumping optical parametric amplifier. The relationship between the gain and the output pulsewidth is investigated. Power penalty is improved by >3dB after the conversion.

I. Introduction

The two standard data formats, return-to-zero (RZ) and nonreturn-to-zero (NRZ) have been highly recommended to be employed in all-optical networks with a combination of optical-time-division multiplexing (OTDM) and wavelength-division multiplexing (WDM). It is widely recognized that the NRZ data format is usually employed in WDM networks requires less bandwidth per channel, while the RZ data format can be adopted to increase the total transmission capacity in OTDM networks with a better receiver performance [1]. Previous research effort in exploring various all-optical techniques for NRZ-to-RZ conversion include injection locking of a Fabry-Perot laser diode (FPLD) [2], four-wave-mixing (FWM) in semiconductor optical amplifier (SOA) [3] and spectral filtering of an cross phase modulation (XPM) broadened signal spectrum [4]. However, the above schemes suffer from several constraints, for example, operating bit-rate is limited by the slow carrier recovery time of FPLD and SOA, phase information is lost in XPM with no gain. These problems can be overcome by employing fiber-based optical parametric amplifier (OPA). OPA-based conversion can be accompanied by its large net gain, flexible gain bandwidth and ultra-fast response time. Also, the signal phase information can be preserved. In addition, with the use of pulsed-pumping scheme, one can obtain an even wider gain spectrum due to its high peak pump power and a relatively short fiber with less severe zero-dispersion wavelength $\lambda_d$ fluctuation [5], [6].

In this paper, we utilize the merits of pulsed-pump OPA to convert NRZ signals to RZ counterparts over a wide gain spectrum. Relationship between parametric gain and pulsewidth is experimentally analyzed and measured of bit-error-rate (BER) has been recorded at both signal and idler wavelengths.

Fig. 1 demonstrates the operating principle of the data format converter based on OPA. The NRZ pulse train $\lambda_a$ is time-synchronized with the local RZ sinusoidal clock $\lambda_p$, which serves as the pulsed-pump. They are fed into the input port of the OPA. The NRZ signal is amplified only when the pump gates on. Due to the exponential dependence of the parametric gain on the pump power, pulse narrowing effect of OPA is realized such that the pulsewidths of the RZ signal and generated idler will be narrower than that of the clocked pump [7], [8].

Fig. 1. Operating principle of the NRZ-to-RZ data format converter based on pulsed-pump OPA. FWHM: full width at half maximum.

II. Experiment

The experimental configuration is shown in Fig. 2. The nonlinear media consisted of a 500m of highly nonlinear dispersion-shifted fiber (HN-DSF) with a $\lambda_d$ of 1554nm and $\gamma = 10.4W^{-2}km^{-1}$. The pulsed-pump $\lambda_p$ was a tunable laser source TLS1, which was chosen at 1556 nm and intensity modulated by a 10GHz sinusoidal pulse train with pulsewidth of 50ps through the amplitude modulator (AM). The erbium-doped fiber amplifier (EDFA 1) served as the preamplifier to a booster EDFA 2. A tunable bandpass filter (TBPF) was used to reduce amplified spontaneous emission (ASE) noise. A NRZ signal was provided by TLS2, which was tuned from 1525 to 1576nm and intensity modulated by a 10Gb/s, $2^7$-1 pseudo-random bit sequence (PRBS). The fiber delay line synchronized the pump with the signal in time domain. The polarization controller (PC 4) was adjusted to align the signal with the pump so as to maximize the parametric gain. The pump and the signal were then combined with the use of a coupler and fed to the HN-DSF. The average input power of the pump and the signal were 24.7 and -5dBm, respectively. The output spectrum was observed from the optical spectrum analyzer (OSA). A TBPF with a 3dB-bandwidth of 0.65nm was used to filter out either the output signal or
the idler. We then measured their waveforms and BER using the digital communication analyzer (DCA) and BER tester (BERT), respectively.

### III. Result and Discussion

Fig. 3(a) and (b) illustrate the NRZ signal at 1538nm and the time synchronized pump at 1556nm, respectively. The original NRZ signal (Fig. 3(a)) is converted to a RZ signal at its original wavelength and its corresponding idler is generated at 1574nm, depicted in Fig. 3(c) and (d), respectively. The signal gain is recorded to be 18dB. Their pulselwidths measured at FWHM are 30 and 29ps, respectively. The insets show their eye diagrams with extinction ratios (ERs) of 14.9 and 15.4dB, respectively. The relatively lower ER at the signal wavelength is due to the residual power of the NRZ signal.

![Fig. 3. Timing diagram of (a) NRZ signal, (b) sinusoidal modulated pump, (c) OPA converted RZ signal and (d) its idler.](image)

We then widely tune the signal wavelength from 1525 to 1575nm to investigate the conversion performance over the OPA spectrum. Fig. 4 shows the OPA gain can attain 10dB over the bandwidth of 50nm, with maximum gain over 18dB. The reason of higher gain peak measured on the longer wavelength side is owing to the pump induced Raman amplification [5]. The pulselwidths of both the signal and the idler are recorded and plotted versus signal wavelengths. The pulselwidths maintain between 28 to 34ps. Results show that narrower pulselwidth can be obtained in higher gain region (> 12dB) indicating OPA effect is the dominating factor for pulse compression.

Fig. 5(a) and (c) show the eye diagrams of signals tuned at 1534 and 1545nm, respectively, while those of the corresponding idlers are shown in Fig. 5(b) and (d). Their ERs are over 15dB, respectively, which indicating a stable performance of NRZ to RZ conversion with high ER can be maintained over a wide spectral range.

![Fig. 4. Parametric gain of the pulsed-pump OPA and measured pulselwidths of RZ signal and idler.](image)

![Fig. 5. Eye diagrams of (a) \( \lambda_s=1534\text{nm} \) and (b) its idler; (c) \( \lambda_s=1545\text{nm} \) and (d) its idler.](image)

To evaluate the performance of the proposed converter, the receiver sensitivities of the converted RZ signal (1538nm) and idler (1574nm) are measured and compared with that of the input NRZ signal. Fig. 6 shows that negative power penalty can be obtained, which is mainly due to the change of data format. At BER of 10^-9, the receiver sensitivities of the RZ signal and its idler are -12.5 and -13dBm, respectively, showing 3.3 and 3.8dB improvement of power penalties over the input NRZ signal. The higher sensitivity of idler should be attributed to its higher ER comparing with that of the signal. The receiver sensitivity gains are more or less the same by tuning wavelengths within the high OPA gain regions (>12dB) due to small variation in pulselwidths and ERs.

### IV. Conclusion

We have demonstrated a NRZ-to-RZ data format conversion based on OPA using clocked pump. 10Gb/s NRZ signals are successfully converted to RZ format with net gain as high as 18dB and conversion bandwidth over 50nm. Results show that OPA could be a promising solution for data format conversion in WDM-OTDM interface in ultra-fast optical communication network.

### V. Acknowledgement

The work described in this paper was partially support by a grant from the Research Grants Council of the Hong Kong Special Administrative Region, China (Project No. HKU 7179/06E). The authors would also like to acknowledge Sumitomo Electric Industries for providing the HNL-DSF.

### VI. References

13B1-2

All-Optical RZ-OOK to RZ-BSPK Conversion Using Cross Phase Modulation

Nattapol Kulswan1 and Pasu Kaewplung2

Department of Electrical Engineering, Faculty of Engineering, Chulalongkorn University, Bangkok, Thailand 10330 Tel: (662) 2186512 Fax: (662) 2518991
1N.Kulswan@gmail.com, 2Pasu.k@chula.ac.th

Abstract: We propose an all-optical on-off keying to binary-phase-shifted keying conversion using cross-phase modulation. Our simulation shows that the converted signal has error-free conversion with power penalty as small as 0.23 dB.

1. Introduction

Next generation long-haul transmission systems have been inclined to employ advance modulation formats based on optical phase-shifted keying. A format that is currently receiving a lot of interest is the differential phase-shift-keyed (DPSK) modulation because it gives 3-dB benefit [1] over the on-off keying (OOK). During the transition from OOK-based transmission to DPSK-based transmission, it is unavoidable to have both OOK-supported equipments and DPSK-supported equipments operate in the same system. This presents the necessity of some devices that can transparently and all-optically convert from OOK to DPSK and vice versa. There were many works had focused on the all-optical signal processing techniques that can diminish the expensive electrical-to-optical-to-electrical (EOE) equipments. However, there are rarely reports concerning the all-optical conversion techniques which can be served as an interconnection between OOK and DPSK systems. The most recent work [2] used the optical amplifier-based Mach-Zehnder Interferometer (SOA-MZI) and demonstrated an error-free conversion at 10 Gbit/s.

In this paper, we propose a new method to convert modulation format from OOK to binary-phase-shift keying (BPSK) by using the cross-phase modulation (XPM) in a highly nonlinear dispersion shifted fiber (HNL-DSF) [3]. Our simulation shows an error-free conversion at a bit rate of 20 Gbit/s.

2. XPM-based OOK-to-BPSK conversion

Fig. 1(a) shows a schematic diagram of XPM-based OOK-to-BPSK conversion in a single HNL-DSF. When we transmit two signals (data which is OOK at wavelength $\lambda_1$ and probe signal which is the continue pulses at wavelength $\lambda_2$) into HNL-DSF, the phase of the probe signal will be changed by the XPM according to the data signal. It is obvious that XPM in a single HNL-DSF can achieve data transfer from intensity of the input signal to phase of the output signal. However the output signal will have unequal intensity among signal bits because the four-wave mixing (FWM) simultaneously occurs with the XPM when both data signal bit and probe signal bit exist. The FWM effect transfers power of both data and probe signals to two new signals at different sideband frequencies. The inset figures in Fig. 1(a) show the spectrum of the data signal and probe at the input and output ends of the HNL-DSF, obtained by computer simulation.

For simulation setup, the average loss of HNL-DSF is 0.51 dB/km, the zero dispersion wavelength (ZDW) is 1550 nm with the dispersion slope of 0.032 ps/(nm km), the nonlinear coefficient is 20.4 (W km)$^{-1}$, and the length is 3.054 km. The wavelengths of the data and the probe signals are 1547.72 and 1552.52 nm, respectively. These wavelengths of the data and the probe signals are selected equally far from a zero dispersion wavelength in order to have identical group velocity for maximizing the XPM efficiency.

The input powers of the data and probe signals are 37.8 mW and 1 mW, respectively. We can observe clearly the strong FWM between data and probe signals from the spectrum of the output of HNL-DSF. This intensity fluctuation among signal bits will be transferred to phase fluctuation by the Kerr effect.

![Figure 1 OOK to BPSK conversion](image)

(b) Figure 1 OOK to BPSK conversion (a) Single HNL-DSF (b) Power equalization in a pair of HNL-DSFs

The Intensity fluctuation of the converted data using XPM in signal HNL-DSF can be eliminated by our proposed scheme using an assist probe signal as shown in Fig. 1(b). The probe is split in to two directions, one into fiber#1 which is similar to Fig. 1(a), the other is launched into fiber#2. Also into fiber#2, the assist probe signal at the same wavelength as the data signal is launched with optimal power in order to shift the phase of all probe signal bits by rad. When the probe signals from two fibers combine together at the output end of HNL-DSFs, the destructive phase interference ($\pi-0$) causes the reduction in the intensity of that bit, while the constructive phase interference ($\pi+\pi$ or $0-0$) causes the enhancement in the intensity of that bit. As a result we can achieve the equalization of all converted signal bits by using the optimal power of the assist probe signal in...
combination with the optimal attenuation which reduces power of the probe signal from fiber#2 to \( (P_{o2}-P_{o})/2 \) as shown in Fig. 1(b).

3. Simulation

We perform the numerical simulation in order to explore the performance of our proposed OOK-to-BPSK converter. In the simulation, the parameters are the same as used in section 2 except the power of the probe is increased to 2 mW. The signal is consisted of 2\(^{41}\)-1 bits pulse train whose data rate is 20 and 40 Gbit/s. The optimal attenuation used in the simulation is 21.8 dB. The performance of the converter is expressed in term of the \( Q \) factor obtained from back-to-back detection. Fig. 2(a) shows the \( Q \) factor as a function of received power of the back-to-back converted signal comparing with the \( Q \) factor obtained from the back-to-back detection of pure DPSK signal without format conversion. The result from Fig. 2 shows that, at the bit-error rate of \( 10^{-12} \) \( (Q = 6.9) \), the OOK-to-DPSK converted signal using our proposed scheme exhibits only 0.23-dB and 0.38-dB power penalty from the pure DPSK at bit rate 20 and 40 Gbps, respectively. Fig. 2(b) and (c) show the eye diagrams of the back-to-back detected converted signal and the back-to-back detected pure DPSK signal at received power of -37 dBm at bit rate 20 Gbps, while Fig. 2(d) and (e) show the eye diagrams of the converted signal and the pure DPSK signal at received powers -34 dBm at bit rate 40 Gbps.

Fig. 2 shows that, at the bit-error rate of \( 10^{-12} \) \( (Q = 6.9) \), the OOK-to-DPSK converted signal using our proposed scheme exhibits only 0.23-dB and 0.38-dB power penalty from the pure DPSK at bit rate 20 and 40 Gbps, respectively. Fig. 2(b) and (c) show the eye diagrams of the back-to-back detected converted signal and the back-to-back detected pure DPSK signal at received power of -37 dBm at bit rate 20 Gbps, while Fig. 2(d) and (e) show the eye diagrams of the converted signal and the pure DPSK signal at received powers -34 dBm at bit rate 40 Gbps.

Fig. 3 shows the dispersion tolerance of the converted signal at 20 Gbps in term of the power penalty resulted from an amount of dispersion added to the signal. From Fig. 3, the dispersion tolerance for the converted signal is significantly stronger than pure DPSK especially in an anomalous dispersion region. However, in normal dispersion region and the pure DPSK seems to yield higher tolerance than the converted signal. This is because the nonlinear positive chirp of XPM on the signal cancels with the linear negative chirp of the anomalous dispersion.

Fig. 3: Dispersion tolerance of the converted signal and the pure DPSK in term of power penalty.

Fig. 4 shows power penalty of the converted signal when the OOK signal power is deviated from the optimal value. At the power penalty of 1 dB, the appropriate range of power mismatch is -8.14 to 7.5 dB.

Fig. 4: Power penalty of the converted signal at receiver when the power of OOK signal is deviated from the optimal value.

4. Conclusion

We have demonstrated a novel OOK-to-BPSK format conversion using the XPM effect in a pair of HNL-DSFs. The simulation results demonstrated that received power penalties of the back-to-back detected OOK-to-BPSK-converted signal to the back-to-back detected pure BPSK are as low as 0.23 and 0.38 dB although the data rates are as high as 20 and 40 Gbit/s, respectively. We also showed that the converted signal exhibits sufficiently tolerance against the dispersion and the power mismatch.

Acknowledgment

This work was supported the cooperation project between department of electrical engineering and private sector for research and development.

References

**All-Optical NRZ-to-PRZ Converter Based on Cascaded Long-Period Fiber Gratings**

Sie-Wook Jeon¹, Tae-Young Kim¹, Masanori Hanawa², Youngjoo Chung¹, Chang-Soo Park¹

¹Department of Information and Communications, Gwangju Institute of Science and Technology, 1 Oryong-dong, Buk-Gu, Gwangju, 500-712, Republic of Korea, ²University of Yamanashi, 4-3-11 Takeda, Kofu, Yamanashi 400-8511, Japan, Tel/Fax: +82-62-970-2218, E-mail: csp@gist.ac.kr

**ABSTRACT**

We propose an all-optical NRZ-to-PRZ converter based on the cascaded long-period fiber gratings to enhance the clock component for clock signal extraction. By using the proposed converter, clock-to-modulation component ratio was improved about 48 dB.

**I. INTRODUCTION**

In fiber-optic systems, the clock extraction is essential for jitter reduction of the electrically converted signal. In conventional fiber-optic systems, the non return-to-zero (NRZ) data format has been used to decrease the bandwidth occupied by the modulation signal, however, this has a difficulty in extracting clock components because the clock-to-modulation component ratio (CMR) of the NRZ signal is quite low. Thereby, this NRZ signal has been converted to the pseudo-return-to-zero (PRZ) signal, which can give a better CMR due to its RZ-like pattern.

All optical NRZ-to-PRZ conversion, in this respect, has been attractive to the high speed optical communication over 40-Gb/s. Among the previous methods [1-3], the method based on π-phase shifted fiber Bragg grating (π-PSFBG) [3] is most likely in terms that it can be implemented more compactly and insensitive to the environmental perturbation. Only drawback comes from the noise caused by the multiple reflections between two gratings as a reward of low insertion loss.

In this paper, we propose a novel all-optical NRZ-to-PRZ converter based on the cascaded long-period fiber gratings (LPGs), which can be implemented as a transmitted form and, as a result, free from multiple reflections. Also, this method is more useful for very high speed modulation because it basically uses precise time delay due to the index difference between the core and cladding modes and is insensitive to the environmental perturbation because the coupled mode to the cladding undergoes the same perturbation with the core mode along the fiber [4].

**II. PRINCIPLE**

The schematic diagram of the proposed NRZ-to-PRZ converter is depicted in Fig. 1. The converter consists of two cascaded LPGs apart from each other with the fiber length corresponding to the time delay required. This could be relatively long and, instead, it has an advantage of spatial margin to accurately implement very small time delay. When the NRZ signal propagating along the core mode meets the first LPG (LPG1), then the signal is split
had a large CMR of 38 dB, showing an enhancement of about 48 dB.

IV. CONCLUSION

We have proposed an all-optical NRZ-to-PRZ converter based on the cascaded long-period fiber gratings to extract clock components from the NRZ signal. Its performance was experimentally demonstrated for 10-Gb/s NRZ system. The clock component was drastically enhanced about 48 dB in CMR using the proposed converter with 50-ps delay.

ACKNOWLEDGEMENT

This work is partially supported by KOSEF through grant No. R01-2006-000-11088-0 and BK-21

REFERENCES


FIGURES

Fig. 2. Measured optical spectrum and experiment setup.

![Fig. 2. Measured optical spectrum and experiment setup.](image)

Fig. 3. Measured waveforms and RF spectra of the NRZ and PRZ signals: (a) and (b): Waveforms for ‘11101100’ pattern, (c) and (d) RF spectra for the NRZ and PRZ signals, respectively.

![Fig. 3. Measured waveforms and RF spectra of the NRZ and PRZ signals.](image)
Abstract
We implement a scheme for 40Gbit/s OTDM to 4x10Gbit/s WDM conversion using a sequential wavelength pulse train and wavelength conversion via birefringence switching in a semiconductor optical amplifier.

1 Introduction
In future high data capacity photonic gateways [1] it may become advantageous to avoid unnecessary optical–electrical and electrical-optical conversions in order to minimize costs, reduce overall power consumption and have less space requirement. All-optical trans-multiplexing whereby high data rate optical time division multiplexed (OTDM) signals are all-optically converted to lower data rate WDM may therefore become necessary in future photonic gateways which interconnect long-haul OTDM networks to local area WDM networks [1-4]. Previous work on OTDM to WDM trans-multiplexing have included the use of supercontinuum generation and fiber dispersion for a sequential wavelength pulse train (SWPT) [2] or using an optically modulated electroabsorption modulator and fiber dispersion for the SWPT and an all-optical wavelength conversion in dispersion-shifted fiber[3].

In this paper we describe a simple implementation of OTDM to WDM trans-multiplexing using a semiconductor optical amplifier for the wavelength converter and supercontinuum generation and fiber dispersion for the SWPT. The basic operation of the OTDM to WDM trans-multiplexer relies on a four-wavelength pulse train each at 10Gbit/s and interleaved sequentially in time. The four wavelength channels thus combine to produce the SWPT running as a 40GHz clock for the wavelength converter as shown in Fig.1.

2 Experiment
The setup for 40Gbit/s to 10Gbit/s transmultiplexer is depicted in Fig.2. The fiber laser (Calmar) outputs 10GHz pulses with 1.5ps FWHM and central wavelength at 1558nm. One output is modulated by pseudorandom binary sequence (PRBS) intensity signal (2^23 -1) generated by 10G BERT and an 40Gbit/s OTDM signal is formed by combining four streams of this data with appropriate delays. The other output is first compressed to increase the spectral width. An arrayed waveguide grating with 200GHz channel spacing and 0.4nm 3dB bandwidth was used to filter out four wavelengths and combine them into a single fiber. The SWPT is formed by using 103m length of high dispersion fiber (DCF has dispersion of -165ps/nm • km) and SMF (single mode fiber,80m,17ps/nm • km) to introduce exactly 25ps delay between neighbor wavelength channels. The SWPT thus forms a 40GHz pulse train. A tunable optical delay line is used to synchronize the 40GHz SWPT with the 40Gbit/s OTDM signal which act as the probe and pump signal respectively in the wavelength converter. The control signal is launched into the SOA in the opposite direction to the probe through a 3-port optical circulator. The average power of the probe and pump signal at the input of the SOA
are -10dBm and +3dBm respectively. The SOA (Alcatel 1901) is biased at 190mA and the wavelength conversion is performed by birefringence switching[5]. At the output of the polarizer, a third AWG is used to separate the different WDM channels. The communication signal analyzer (CSA) could analyze waveforms of the output conversion signals. The eye diagrams and pulses are obtained with a 32-GHz photodetector on a 50-GHz-sampling oscilloscope.

![Diagram](image)

Fig.2 Experiment setup: DCF = dispersion compensating fiber, SMF = single mode fiber, SOA = semiconductor optical amplifier, CSA = communications signal analyzer, PC = polarization controller

### 3 Results and discussion

Fig.3(a) shows that the four wavelengths pulses are overlap before senting into dispersion fiber and Fig.3(b) shows the four wavelengths pulses are interleaved in time after the dispersion fiber. Fig.4(a) shows the eyediagram of 10Gbit/s signal. Fig.4(b) shows the eyediagram of 40Gbit/s signal which is multiplied from the 10Gbit/s source. Fig.4(c) shows the eyediagram of converted 10Gbit/s signal which is still an opening eyediagram. Fig.5 shows the BER measurement.

![Diagram](image)

Fig.3 Comparison of (a)undispersed four-wavelength pulses at input of DCF and (b)dispersed four-wavelength pulses at output of SMF

![Diagram](image)

Fig.4 Eye pattern of (a)10Gbit/s source (b) 40Gbit/s signal after multiplexer (c) converted 10Gbit/s signal

The power penalty in our experiment mainly comes from the reflected pump light at the SOA facet, which has the same wavelength as the probe light. And it is difficult to precisely determine the polarization state of input probe pulses. We also find that the switching efficiency is not very good for the counter propagation pump structure.

### 4 Conclusion

A 40Gbit/s to 10Gbit/s signal processing system has been demonstrated based on the serial-to-parallel conversion approach. This system uses four-sequential-wavelength pulse train each operating at 10Gbit/s and interleaved in time to achieve 40Gbit/s serial-to-parallel conversion. We may expect that higher number of converted channels may be obtained using shorter pulse width source and narrower spacing channel AWG. The system enables the high bit rate optical signal to be processed by relatively low cost electronics.

**Acknowledgements:** This work was funded by a direct grant project 2050385.

**Reference**

Performance of Wavelength Exchange in Anomalous-dispersion Region

Rebecca W. L. Fung, Henry K. Y. Cheung, and Kenneth K.Y. Wong
Department of Electrical & Electronic Engineering, The University of Hong Kong, Pokfulam Road, Hong Kong.
E-mail: wlfung@eee.hku.hk

Abstract: We demonstrate the 10Gb/s wavelength exchange with two pumps at the anomalous-dispersion region. Extinction ratio between converted signal and residual signal is ~20dB. Bit error rate of <10^{-9} is achieved with power penalty of ~2dB.

I. Introduction

Wavelength exchange (WE) relying on four-wave mixing (FWM) in highly nonlinear dispersion-shifted fibers (HNL-DSF) has been studied in recent research. Previously, simultaneous conversion of two signals can be achieved by a suitable choice of wavelengths of two pumps at the normal dispersion region and the two signals at the anomalous region with respect to the zero-dispersion frequency \( \omega_0 \) of a fiber (denoted as WE I) [1]. Past results showed that the pump induced Raman amplification introduces asymmetric power transfer that degrades the performance of the WE I process. Such performance degradation is particularly severe when the two pumps are arranged orthogonally at the normal dispersion region [2]. Therefore, we have proposed another configuration with two pumps at the anomalous dispersion regime (denoted as WE II), where the performance degradation caused by Raman gain can be eliminated [3]. In theory, no Raman gain is provided by pumps allocated at anomalous dispersion region, so it can predict power transfer asymmetry can be avoided in WE II [3]. With this arrangement, signals at the normal dispersion region exhibit symmetric power transfer characteristics and a nearly-complete wavelength exchange can be achieved. In this paper, we demonstrate an experimental validation of WE II with two 10Gb/s 2^7-1 pseudo-random bit sequence (PRBS) signals. The performance of the exchanger is quantified by the measurements of eye diagrams and bit error rate (BER).

II. Experiment

The experimental configuration is shown in Fig.1. The wavelength exchanger consisted of 1 km of HNL-DSF with a zero-dispersion wavelength \( \lambda_0 \) of 1540nm, a dispersion slope of 0.03ps/nm^2/km and a fiber nonlinearity coefficient \( \gamma \) of 12W^{-1}km^{-1}. The two tunable laser sources, TLS1 and TLS2, were set at 1549nm and 1554nm respectively, served as the two pumps. They were phase-modulated (PM) by a 10Gb/s 2^7-1 PRBS [4] to suppress stimulated Brillouin scattering (SBS). The erbium-doped fiber amplifier (EDFA 1) served as the preamplifier to a booster EDFA 2, with a maximum output power of 33dBm. The two tunable bandpass filters (TBPF) with 2nm bandwidth were inserted after EDFA 1 so as to filter out the two pumps separately and reduce amplified spontaneous emission (ASE) noise. Two polarization controller (PC 3 & PC 4) were used to control the state of polarization (SOP) of the two pumps such that orthogonal pump configuration can be maintained by minimizing the power of the spurious FWM components. The exchange efficiency can be higher with using orthogonal pump allocation [5]. The SOP of each pump was adjusted by PC 1 & PC 2 while PC 5 & PC 7 were adjusted to minimize the insertion losses to the PM and amplitude modulator (AM), respectively. Wavelengths of the tunable laser sources, TLS3 and TLS4, were chosen at 1534 and 1529nm, respectively, which were served as signals. They were intensity-modulated with a 10Gb/s 2^7-1 PRBS. A variable optical attenuator (VOA) was inserted after EDFA 2 to adjust the input pump powers. A 95/5 coupler combined 95% of the two pumps and 5% of the signal into the fiber. The output power from the fiber was sent to the optical spectrum analyzer (OSA) to display the spectrum after the exchange. The waveforms and BER were measured using the digital communication analyzer (DCA) and BER tester, respectively.

III. Results and Discussion

It has been discussed that pump-induced Raman amplification will cause asymmetric power transfer at the normal dispersion region. Therefore, orthogonal WE II is introduced to improve the symmetry of WE [3]. The experimental result of orthogonal WE II is shown in Fig. 2. The experimental result demonstrates that symmetric power transfer characteristics can be obtained from WE II such that the maximum power of the idler (generated
at 1529nm) can be attained together with the minimum power of the residue signal. The maximum normalized idler power is equal to unity in the WE process as no Raman gain is provided by the two pumps. Result shows that a nearly complete WE can be obtained.

During the practical WE, two signals placed at 1534 and 1529nm undergo exchanging process simultaneously. Under the process, the two corresponding idlers were generated at 1529 and 1534nm accordingly. Extinction ratio (ER) was defined as the power ratio between the maximum converted signal (idler) and the minimum residual signal [2]. Here, the ERs were measured to be ~20 dB in both cases after WE as shown in Fig. 3.

Both signals are modulated with 10Gb/s 2¹-1 PRBS. Figure 4(a) and (b) represent the original signal waveforms before WE observed at 1534 and 1529nm, respectively. The insets show their eye diagrams with Q-factor of 10dB and 10.4dB, respectively.

The exchanged signals after WE observed at 1534 and 1529nm are shown in Fig. 5(a) and (b), respectively. The two figures illustrate that the original signal at 1529nm is efficiently exchanged to its corresponding idler wavelength at 1534nm while the original signal at 1534 nm is exchanged to its idler wavelength at 1529nm, indicating that WE II is successfully achieved. The insets show their eye diagrams with Q-factor of 9.1dB and 9.5dB, respectively. However, the data after WE is noisy because the noisy mark level is caused by the EDFA ASE noise from the two pumps. It is suggested that the ASE noise of the pump can be suppressed by a fiber-Bragg grating (FBG) with a narrow bandwidth and a high suppression level [5].

To evaluate the performance of WE II, the receiver sensitivities of the two exchanged signals are measured and compared with those of their corresponding original signals. The measured BER curves are plotted in Fig. 6. At BER of 10⁻⁹, the receiver sensitivities of the signals at 1534 and 1529nm are -22.4 and -23dBm, respectively. The power penalties incurred in the exchanger are measured to be 2.3 and 1.8dB, respectively. This power penalty is mainly due to the phase noise introduced by the phase dithering of the pumps, and can be improved with the complementary phase dithering approach [6].

Fig. 2. WE transfer characteristics in WE II with signal at 1534nm and idler at 1529nm – signal power (●) and idler power (○) [3].

Fig. 3. Measured optical spectra after WE.

Fig. 4. Data stream observed (a) at 1534nm (b) at 1529nm before WE.

Fig. 5. The exchanged signal (a) at 1534nm (from 1529nm to 1534nm) (b) at 1529nm (from 1534nm to 1529nm) after WE.

IV. Conclusion

We have successfully demonstrated the performance of WE at the anomalous dispersion region. Results show that performance degradation caused by Raman gain is avoided so that a nearly-complete WE can be achieved. Extinction ratio between the converted signal and the residue is ~20dB with clear opening eyes observed. BER of <10⁻⁹ is achieved with power penalties of ~2dB.

V. Acknowledgement

The work described in this paper was partially support by a grant from the Research Grants Council of the Hong Kong Special Administrative Region, China (Project No. HKU 7179/06E).

VI. References

All-Optical XNOR Gate using Fiber Optical Parametric Amplifier
David Ming-Fai Lai, Bill Ping-Piu Kuo, and Kenneth Kin-Yip Wong
(Department of Electrical and Electronic Engineering, The University of Hong Kong, Pokfulam Road, Hong Kong. Tel: +852-2859-2698, Email: mflai@eee.hku.hk)

Abstract
An all-optical XOR gate based on four-wave mixing and cross gain modulation in an optical parametric amplifier is presented. Ultra-fast response time can be achieved due to the sole use of parametric effects.

1. Introduction
All-optical exclusive-OR (XOR) gates have been of high popularity due to its myriad of functions, including comparison of data patterns [1], data encryption/decryption, or parity checking [2]. These can also be accomplished by XNOR gates since the XNOR output is the negation of the XOR output; a detector detecting for 1’s at XOR output can easily be configured to detect 0’s at a XNOR output to provide the same functionality.

All-optical XNOR gate based on FWM and XGM has been implemented using a semiconductor optical amplifier (SOA) in Ref. [3]. However, its bit rate is limited by the SOA’s carrier recovery time in XGM process. On the other hand, fiber-based optical signal processing, including all-optical sampling by FWM effect [4] and inverted and non-inverted wavelength conversion by XGM effect [5], has been investigated previously because of ultra-fast response of nonlinear effect in optical fibers. Therefore, fiber-based all-optical logic gates should be capable of overcoming the speed limit of their SOA-based counterparts using similar effects. In this paper, we report an all-optical 10Gb/s non-return to zero (NRZ) XNOR gate utilizing XGM and FWM effects in fiber optical parametric amplifier (OPA).

2. Principle of Operation
The operation principle of producing a XNOR gate using a combination of AND, NOR, and OR gates is similar to that as shown in Ref. [3]. Except it required an extra probe input, which acted as the output of a NOR gate. In our design, the XGM output is obtained directly from the OPA pump, since it is possible to achieve strong pump depletion as demonstrated in Ref. [6]. The OPA pump depletes itself whenever one or more signals are present, because it transfers a majority of its power to the signal(s). Hence, if XGM effect is present in the pump, it is equivalent to a NOR operation on the two input signals. Furthermore, strong FWM effect occurring on the two signals will produce new FWM peaks, where the peaks closest to the two signals tend to be the strongest. Since the generation of these new peaks require the presence of both signals in the nonlinear medium, these peaks are essentially the AND output of the two signals. By combining the XGM and FWM products, this results in a XNOR output. Fig. 1 summarizes the operation principle.

Fig. 1. Operation principle of an all-optical XNOR gate.

3. Experimental Setup
The experimental setup is shown in Fig. 2. The nonlinear medium used was a spool of 1km highly nonlinear dispersion-shifted fiber (HNL-DSF) with nonlinear coefficient of $14 W^{-1} km^{-1}$ and zero dispersion wavelength of 1560nm. The pump wavelength was set at 1561.8nm, which was phase modulated with a 10Gb/s 2^{23}-1 pseudo-random binary sequence (PRBS) to suppress stimulated Brillouin scattering (SBS) [7]. It was then connected to an erbium doped fiber amplifier (EDFA1) to reach output power of 23dBm. The tunable band pass filter (TBPF1) significantly suppressed the amplified spontaneous emission (ASE) noise from the EDFA. After that, the pump was amplified to 27dBm at EDFA2 before entering the WDM coupler.

The experimental setup was shown in Fig. 2. The nonlinear medium used was a spool of 1km highly nonlinear dispersion-shifted fiber (HNL-DSF) with nonlinear coefficient of $14 W^{-1} km^{-1}$ and zero dispersion wavelength of 1560nm. The pump wavelength was set at 1561.8nm, which was phase modulated with a 10Gb/s 2^{23}-1 pseudo-random binary sequence (PRBS) to suppress stimulated Brillouin scattering (SBS) [7]. It was then connected to an erbium doped fiber amplifier (EDFA1) to reach output power of 23dBm. The tunable band pass filter (TBPF1) significantly suppressed the amplified spontaneous emission (ASE) noise from the EDFA. After that, the pump was amplified to 27dBm at EDFA2 before entering the WDM coupler.

The two signals’ wavelengths were set at 1566.9nm and 1568.5nm. They were amplitude modulated with an identical NRZ 10Gb/s bit sequence. The tunable delay
line delayed one of the signals such that they were unsynchronized by one bit. They were then coupled together using a 50/50 coupler before amplified by EDFA3 to a total output power of about 15dBm.

The amplified pump and signals were then coupled together using a WDM coupler with a cutoff wavelength of 1564nm. The combined waveform was then injected into the HNL-DSF. The output of the HNL-DSF was split by a 50/50 coupler, where one output branch filtered out the pump wavelength, while the other filtered out an idler produced by FWM from the two signals. The attenuators at each branch reduced optical power to prevent damage to the tunable band pass filters (TBPF2 & 3) and ensured the 1’s from the pump and the FWM peaks are of the same power. The optical delay line (ODL2) was used to compensate the path difference between the two branches. They were then recombined using another 50/50 coupler and the output was monitored from a digital communication analyzer (DCA).

4. Results and discussion

Figure 3 illustrates the inputs and outputs of the AND, NOR, and XNOR gates. It can be seen from the figure that the AND output produces a ‘1’ only when both the inputs are ‘1’. The NOR output produces a ‘1’ only when both inputs are ‘0’. The last coupler output acted as an OR gate, where it combines the output of the AND and NOR gates to generate the XNOR.

Figure 4 shows the eye diagram of the resultant XNOR gate. The extinction ratios of the XNOR, NOR, and AND outputs are about 11dB, 12dB, and 24dB, respectively. The extinction ratio of the output (XNOR) is dominated by the NOR gate because it is generally difficult to deplete the pump by 100% [6], leaving a small residual power at off-state. This can be improved by optimizing the phase-matching condition [6]. Note that as the output signals generally preserve the pulse shapes of input signals, it could be expected that this XNOR gate can support higher bit rate operation.

5. Conclusion

We have successfully demonstrated an all optical XNOR gate using a single stage OPA. The minimal distortion at the output reveals a possibility for higher bit rate operation. Since this XNOR gate is generated from NOR and AND gates, this device is capable of providing AND and NOR outputs simultaneously in addition to its normal XNOR output, which may be useful in simplifying implementation of compound logic gates.

6. Acknowledgement

The work described in this paper was partially support by a grant from the Research Grants Council of the Hong Kong Special Administrative Region, China (Project No. HKU 7179/06E). The authors would also like to acknowledge Sumitomo Electric Industries for providing the HNL-DSF.

7. References

Maximum Likelihood Sequence Estimation for Impairment Compensation in Advanced Modulation Formats

Jian Zhao, Lian K. Chen, and Calvin C.K. Chan
Department of Information Engineering, The Chinese University of Hong Kong, Shatin, N. T., Hong Kong SAR, China
Tel: +852-2609-8479, Fax: +852-2603-5032, Email: lkchen@ie.cuhk.edu.hk

Abstract
We review and propose novel designs of maximum likelihood sequence estimation (MLSE) for effective impairment compensation in advanced modulation formats. The performance and computation complexity of these MLSE structures are investigated and summarized.

1 Introduction
As the capacity of the transmission systems increases, many signal degradation effects, such as chromatic dispersion (CD) and polarization mode dispersion (PMD), become prominent and seriously degrade the performance of the optical communication systems. Maximum likelihood sequence estimation (MLSE) has recently attracted considerable interest for impairment compensation because of its significant cost saving and adaptive compensation capability required in future dynamic optical networks [1-11]. Advanced optical modulation format such as differential phase shift keying (DPSK) is an alternative method to extend the capacity and the transmission reach of the communication systems. However, few studies have been performed to extend the transmission reach by combining advanced modulation formats and MLSE [12-14]. The optimal design of MLSE structures for different advanced modulation format is different. In this paper, we will review our recent work on the design of novel MLSE structures with high impairment compensation performance and low cost for different advanced modulation formats, including DPSK, amplitude shift keying/DPSK (ASK/DPSK) orthogonal modulation, differential quaternary phase shift keying (DQPSK), and 4-ASK. After evaluating the performance and analyzing the complexity of the proposed schemes, we can then determine the most cost-effective solution for the given requirements of a transmission system.

2 Novel MLSE Structures for Advanced Modulation Formats
(a) Multi-chip DPSK MLSE for CD and PMD Compensation
DPSK is one of the most desirable formats for high-speed optical transmission due to its 3-dB optical signal to noise ratio (OSNR) sensitivity improvement and higher tolerance to fiber nonlinear effects compared to OOK format. However, although conventional MLSE is effective to extend the transmission reach of the OOK format, it provides limited performance improvement for the DPSK format [9-11].

In this paper, we review our recently proposed 3-chip DPSK MLSE for CD and PMD compensation in DPSK format. The proposed method exploits the phase difference between not only the adjacent optical bits but also the bits with one bit slot apart for sequence estimation of the DPSK data [12]. The results show that 3-chip DPSK MLSE significantly outperforms conventional 2-chip DPSK MLSE in CD and PMD compensation. It is shown that 3-chip DPSK MLSE can enhance the CD tolerance of 10-Gbit/s DPSK signal to 2.5 times of that by using 2-chip DPSK MLSE and can bound the penalty for 100-ps differential group delay (DGD) by 1.4 dB. We will further investigate 4-, 5-, and 6-chip DPSK MLSEs. We will show that these structures can provide further performance improvement but at the expense of the implementation complexity increase. We suggest that in practice, 3- or 4-chip DPSK MLSE is optimal in terms of performance and complexity.

(b) Joint MLSE (J-MLSE) and Decision-Feedback J-MLSE (DF-J-MLSE) for CD Compensation in ASK/DPSK Orthogonal Modulation Format
ASK/DPSK orthogonal modulation format is an attractive multi-bit per symbol modulation format to enable close channel spacing in DWDM transmission and to carry optical payload / label in optical networks concurrently [15-17]. However, despite many experimental demonstrations of this format in the applications, few studies on the design of electronic equalization devices for such format have been performed.

In [13], we determined the fundamental impairment mechanism in CD-limited ASK/DPSK orthogonal modulation. Based on the fundamental finding, we showed that conventional MLSEs which only consider intra sub-channel interference of the ASK and DPSK sub-channels separately fail to improve the overall CD tolerance of the ASK/DPSK signal. J-MLSE was proposed to exploit the correlation information between the detected ASK and DPSK signals and was shown to improve the CD tolerance of the ASK/DPSK signal significantly.

However, a J-MLSE has higher implementation complexity which is proportional to $2^{2(m+1)}$, whereas a conventional MLSE's complexity is proportional to $2^{m+1}$, where $m$ is the MLSE's or J-MLSE's memory length. Recently we are investigating a novel DF-J-MLSE that reduces the implementation complexity to the same as that of a conventional MLSE while preserving the overall CD tolerance the same as that of a J-MLSE.
3-Chip DQPSK MLSE for CD and PMD Compensation

DQPSK is an attractive multi-bit per symbol modulation format for high-speed optical transmission due to its spectral efficiency and higher tolerance to CD and PMD compared to the DPSK format [18]. To enhance the transmission reach of the DQPSK signal, the design of electronic equalizer was proposed [14]. It was shown that separate equalization of the two tributaries of the DQPSK signal provides limited CD tolerance improvement while J-MLSE can effectively improve the CD tolerance of the DQPSK signal. In this paper, we will show some preliminary results of our recent work on a novel 3-chip DQPSK J-MLSE. The method searches the most probable path through the trellis for data sequence estimation by exploiting the phase difference between not only the adjacent optical bits but also the bits with one bit slot apart. The scheme significantly outperforms conventional MLSE and J-MLSE in CD and PMD compensation while maintaining the implementation complexity comparable to that of a J-MLSE. We show that the 3-chip DQPSK J-MLSE provides twofold CD tolerance enhancement compared to a J-MLSE and exhibits negative penalty for 100-ps DGD at 10 Gsym/s.

4-ASK MLSE for CD Compensation in CD-Varying Optical Systems

4-ASK format is another cost-effective multi-bit per symbol modulation format and requires only one optical modulator and receiver for signal generation and detection [19-20]. It can also be coded and decoded all optically [20]. However, due to the increased number of levels, such format is sensitive to CD-induced ISI. In [21], we showed that the optimal level spacing of the 4-ASK signal changes with the CD values and improper level spacing design leads to significant CD tolerance reduction. As a result, level spacing optimization is difficult in CD-varying 4-ASK optical systems, in which the CD frequently changes due to the time-varying effects of the installed fibers and different routing paths. In [21], we proposed 4-ASK MLSE for signal detection. It was shown that the proposed method can effectively alleviate the sensitivity of CD tolerance to level spacing, therefore, relaxing the difficulty of level spacing optimization. By using 4-ASK MLSE, the CD tolerance of the 4-ASK signal is significantly enhanced by at least a factor of two.

3 Discussions and Summary

In summary, we have reviewed our recently proposed MLSE structures for different advanced modulation formats. The proposed schemes significantly outperform the existing schemes without much complexity increase, as shown by Table 1. Table 2 compares the performance of different advanced modulation formats under the proposed schemes. As a result, the most cost-effective solution given the requirement of a transmission system can be determined. For instance, for a 100-km range optical network, ASK/DPSK orthogonal modulation is the best modulation format because of its CD tolerance around 1500 ps/nm and low complexity, as shown in Table 2. (This work was supported in part by the Hong Kong Research Grants Council, Project No. 411006.)

### Table 1: Performance improvement of our recently proposed schemes with respect to the existing schemes

<table>
<thead>
<tr>
<th>Scheme</th>
<th>DPSK</th>
<th>ASK/DPSK</th>
<th>DQPSK</th>
<th>4-ASK</th>
</tr>
</thead>
<tbody>
<tr>
<td>Existing schemes</td>
<td>Conventional MLSE</td>
<td>No report</td>
<td>J-MLSE</td>
<td>No report</td>
</tr>
<tr>
<td>Proposed schemes</td>
<td>3-chip DQPSK MLSE</td>
<td>DF-J-MLSE</td>
<td>3-chip DQPSK J-MLSE</td>
<td>4-ASK MLSE</td>
</tr>
<tr>
<td>CD sensitivity reduction</td>
<td>0 – 0.7 dB</td>
<td>2.5 times</td>
<td>0 – 2.5 dB</td>
<td>0 – 2.5 dB</td>
</tr>
<tr>
<td>Power penalty of DGD (T)</td>
<td>1.4 dB vs 4.2 dB</td>
<td>NA</td>
<td>NA</td>
<td>NA</td>
</tr>
<tr>
<td>Receiver complexity</td>
<td>2 times</td>
<td>The same</td>
<td>2 times</td>
<td>The same</td>
</tr>
<tr>
<td>MLSE complexity</td>
<td>comparable</td>
<td>NA</td>
<td>comparable</td>
<td>NA</td>
</tr>
</tbody>
</table>

Note: CD tolerance is evaluated at 3-dB power penalty. $T$ is the time period for one bit slot.

### Table 2: Performance comparison of different advanced modulation formats under our recently proposed schemes

<table>
<thead>
<tr>
<th>Scheme</th>
<th>DPSK</th>
<th>ASK/DPSK</th>
<th>DQPSK</th>
<th>4-ASK</th>
</tr>
</thead>
<tbody>
<tr>
<td>Existing schemes</td>
<td>Conventional MLSE</td>
<td>EF-J-MLSE</td>
<td>3-chip DQPSK J-MLSE</td>
<td>4-ASK MLSE</td>
</tr>
<tr>
<td>CD tolerance at $E_b/N_0=20$ dB</td>
<td>1000 pulses</td>
<td>1000 pulses</td>
<td>3000 pulses</td>
<td>0 pulses</td>
</tr>
<tr>
<td>$E_b/N_0$ vs DGD (dB)</td>
<td>15 dB</td>
<td>15 dB</td>
<td>15 dB</td>
<td>15 dB</td>
</tr>
<tr>
<td>Receiver complexity</td>
<td>Two 20-Gbit receivers</td>
<td>Two 10-Gbit receivers</td>
<td>Four 10-Gbit receivers</td>
<td>One 10-Gbit receiver</td>
</tr>
<tr>
<td>MLSE complexity</td>
<td>20 Gb/s and $2^{-11}$</td>
<td>10 Gb/s and $2^{-10}$</td>
<td>10 Gb/s and $2^{-10}$</td>
<td>10 Gb/s and $2^{-10}$</td>
</tr>
</tbody>
</table>

Note: All formats are 20-Gbit/s. $E_b$ is the average received signal power in 100 ps. $N_0$ is the noise spectral power density. The performance is evaluated at BER of $10^{-4}$.

### References

Maximum Likelihood Sequence Estimation for Chromatic-Dispersion Compensation in 4-ASK Modulation Format

Jian Zhao, Lian-Kuan Chen, Chun-Kit Chan
Department of Information Engineering, The Chinese University of Hong Kong, Shatin, N. T., Hong Kong SAR, China

Abstract
Maximum likelihood sequence estimation is found to greatly relax the difficulty of level-spacing optimization in CD-varying 4-ASK optical networks and significantly enhance the CD tolerance by at least a factor of two.

1 Introduction
4 amplitude shift keying (4-ASK) is a promising cost-effective spectral-efficient modulation format to extend the transmission capacity. By operating at only half of the bit rate, 4-ASK not only improves the tolerance to chromatic dispersion (CD) compared to on off keying format [1-3], but also alleviates the speed limitation of electrical and optical components.

The CD in optical networks may change frequently due to the time-varying effects of the installed fibers and different routing paths. In the CD-varying optical systems, the level-spacing optimization of the 4-ASK signal is difficult. It is because the optimal level spacing changes with CD values, and improper level-spacing design will lead to significant CD tolerance reduction.

In this paper, we propose to use maximum likelihood sequence estimation (MLSE) for 4-ASK signal detection. It is shown that MLSE can effectively alleviate the sensitivity of CD tolerance to level spacing, therefore, relax the difficulty of level-spacing optimization. By using MLSE, the CD tolerance of the 4-ASK signal is significantly enhanced by at least a factor of two.

2 System Model

Fig. 1 shows the system model. 10-Gs/s 4-ary electrical signal is generated by combining two binary sources, A and B, using a power combiner. Attenuator is used before the source B to adjust the level spacing of the 4-ASK signal. A continuous wave light is modulated by the 4-ASK electrical signal using a Mach-Zehnder modulator (MZM). The generated optical 4-ASK signal is fed into a piece of fiber where CD is introduced. At the receiver, the signal is optically pre-amplified, filtered by a Gaussian-shaped optical bandpass filter (OBPF) and detected. After O-E conversion, the electrical 4-ASK signal is amplified, filtered by a 4th-order Bessel electrical filter (EF), sampled, and decoded conventionally or by MLSE. The optical and electrical filter bandwidths are optimized and are found to be 20 GHz and 8 GHz, respectively. Analog to digital (A/D) converter in MLSE has 5-bit resolution. MLSE is a 16-state machine and its metric, PM(b_k), is:

\[
PM(b_k) = PM(b_{k-1}) - \sum_j \log(p(I(t_j)|b_{k-m},...,b_j))
\]

For one sample per bit, \(t = (k-m/2)T\). For two samples per bit, \(t = (k-m/2)T\) or \((k-(m+1)/2)T\). \(b_k\) and \(p(I(t_j)|b_{k-m},...,b_j)\) are 4-ASK logical data and the probability of the received signal value at \(t=t_j\) given the logical data \(b_{k-m},...,b_j\), \(m\) is the memory length. The initial metric of MLSE is obtained using nonparametric histogram method. The performance is evaluated in terms of the required \(E_b/N_0\) (received photon number per bit) to achieve BER of 10^{-9} for analytical investigation and 10^{-2} for simulation. \(E_b\) and \(N_0\) are the optical average power in one bit slot and the preamplifier’s noise power spectral density.

3 Dependency of CD Tolerance on Level Spacing without MLSE

Karhunen Loeve expansion and saddlepoint approximation are used for the analysis at a BER of 10^{-9} in this section. Fig. 2 shows the \(\log_{10}(BER)\) versus \(E_b/N_0\) for both matched filter (circles) and practical filter in the adopted system (triangles). From the figure, it is shown that the 4-ASK signal by using practical filter has around 2-dB penalty compared to that by using matched filter. The required back-to-back \(E_b/N_0\) for the adopted system at
BER of $10^{-2}$ and $10^{-9}$ are around 22 dB and 26 dB, which will agree with the results in Fig. 4 & 5. Fig. 3 shows the optimal power for level ‘1’ and level ‘2’ versus CD, where the optical power is normalized by that of level ‘3’. From the figure, it is shown that optimal level spacing changes with CD values. Because the lower eyes of the 4-ASK signal is vulnerable to CD-induced ISI, higher powers are required for level ‘1’ and level ‘2’ as CD value increases. However, in practice, after the initial system design, the level spacing is usually fixed regardless of CD variations. Therefore, it is desirable to find out the CD tolerance under a fixed level spacing. Fig. 4 depicts the CD tolerance of 10-Gs/s 4-ASK signal with level spacing optimized at (i) 0 ps/nm (circles), (ii) 700 ps/nm (triangles), and (iii) every CD value (diamonds) (the ideal case). It is shown that case (i) is worse than case (ii) and (iii) for CD values larger than 400 ps/nm. For case (ii), the required $E_b/N_0$ for large CD value approaches to that with case (iii). However, compared to case (i), case (ii) has several dB’s penalty for small CD values. Without MLSE, the performance trade-off for small and large CD values complicates the level-spacing optimization in CD-varying 4-ASK optical networks.

4 MLSE for CD Compensation in 4-ASK Format

In this section, Monte Carlo simulations are performed to investigate the performance of MLSE for CD compensation in 4-ASK format at BER of $10^{-2}$. Fig. 5 shows the required $E_b/N_0$ versus CD for different level spacings without MLSE (dotted), with one-sample per bit MLSE (dashed), and with two-sample per bit MLSE (solid). Circles and triangles represent the level spacing optimized at 0 ps/nm and 700 ps/nm, respectively. From the figure, it is found that MLSE using one- or two-sample per bit can significantly enhance the CD tolerance of 4-ASK signal. For two-sample per bit MLSE, CD tolerance is improved by at least a factor of two for both level-spacing designs. Furthermore, with two-sample per bit MLSE, the back-to-back sensitivity of 4-ASK signal with level spacing optimized at 700 ps/nm is improved by 1 dB. At $E_b/N_0$ of 28 dB, the CD tolerances for the two different level-spacing designs are both around 2000 ps/nm, more than two times of that without MLSE. Therefore, by using two-sample per bit MLSE, the sensitivity of CD tolerance to level spacing is alleviated.

5 Conclusions

We have showed that optimal level spacing of 4-ASK signal changes with CD values and improper level-spacing design leads to significant CD tolerance reduction. As a result, level-spacing optimization is difficult in CD-varying 4-ASK optical systems. We propose MLSE for 4-ASK signal detection. It is found that MLSE can effectively alleviate the sensitivity of CD tolerance to the level spacing, thus relax the difficulty of level-spacing optimization. By using MLSE, the CD tolerance of 4-ASK format is significantly enhanced. At $E_b/N_0$ of 28 dB, MLSE can enhance the CD tolerance of the 4-ASK signal to 2000 ps/nm, more than two times of that without MLSE. Thus MLSE makes the cost-effective 4-ASK a promising format to extend the transmission capacity of optical networks. This work is supported in part by HK CERG Grant CUHK/411005.

6 References

Simultaneous compensation of PMD and CD in a 10.7Gb/s Field Trial based on the MLSE

Dirk Breuer, Armin Ehrhardt, Daniel Frietsch, Manuel Paul, Lars Schuerer
T-Systems Enterprise Services GmbH, Systems Integration, Projekt Unit Next Generation Broadband Networks
Guslarer Ufer 35, 10589 Berlin, Germany
Phone: +49 30 3497 4312, Fax: +49 30 3497 4956

Hamdi Oeruen, Timo Winkler von Mohrenfels
CoreOptics GmbH, Nordostpark 12-14, 90411 Nuremberg, Germany
Phone: +49 911 94151 804, Fax: +49 911 480 86 82
{hoeruen, timo}@coreoptics.com

Abstract—The performance of an electronic dispersion compensator based on the maximum likelihood sequence estimation (MLSE) is investigated in a 10.7Gb/s field trial under the simultaneous influence of polarization mode dispersion and chromatic dispersion.

I. INTRODUCTION
Polarization Mode Dispersion (PMD) is one of the major impairments in optical communication systems deployed with data rates of 40Gb/s and higher. This effect causes severe signal distortions even though newly installed fibers may have low average PMD values. Different methods have been proposed to combat this effect in the optical as well as in the electrical domain [1][2]. A further challenge exists for optical network systems in which old fibers have been deployed. Such relatively old fibers may exhibit fairly large average PMD values, which could even become the dominant impact if data rates of 10Gb/s will be transmitted through dispersion compensated metro and long-haul systems.

To mitigate the PMD problem, electronic dispersion compensators (EDC) seem to be quite promising as they are easy to implement and cost effective. Three types of EDC have been evolved in the near past. These are the feed forward equalizer (FFE), the decision feedback equalizer (DFE), and the maximum-likelihood sequence estimator (MLSE) [2][3][4]. Within these three EDC, the MLSE has the highest complexity but can be expected to deliver the best performance against impairments such as PMD and chromatic dispersion (CD).

In this paper, we present the results of a field trial using a commercial EDC based on the MLSE technology (CoreOptics). We demonstrate the EDC performance on two different field fiber configurations by measuring the bit error ratio (BER) in dependence on the optical signal to noise ratio (OSNR) under the simultaneous impact of PMD and CD. Detailed description of the EDC’s operation has already been reported in [5]. An investigation of the same EDC was performed in a laboratory environment to compensate for the joint impact of chromatic dispersion and first-order PMD [6].

II. EXPERIMENTAL SETUP
Fig. 1 and Fig. 2 show the laboratory and the field setups. An optical signal having the standard G.709 optical transport network (OTN) frame was generated by an optical bit error rate tester (BERT). The payload was a pseudo-random bit sequence (PRBS) of length $2^{31}-1$. Its operation wavelength was at 1556.82 nm with an output power of 0.5 dBm. Prior to launching into the field link configuration, an erbium-doped fiber amplifier (EDFA) and a variable optical attenuator (VOA 1) were used to adjust the input power into the field to 5.0 dBm.

At the receiver side, a noise source with a second attenuator (VOA 2) was applied and coupled via a 3 dB coupler to adjust the OSNR whose values were measured with an optical spectrum analyzer at a resolution bandwidth of 0.1 nm. The tunable band pass filter (TBF) was used to filter out spectral noise and had a full width at half maximum (FWHM) of 1.0 nm. The third attenuator (VOA 3) kept the input optical signal into the device under test (DUT) – the CoreOptics receiver – at a constant -10.0 dBm power level. Finally, the equalized bits from the DUT were then optically forwarded into the optical BERT to perform the BER measurements.

Fig. 1: Laboratory configuration
In our field trials, two different fiber configurations were used which we call Link A and Link B. The field fiber configuration was buried between the labs of T-Systems in Berlin, Germany in the mid 1980s. Link A was without inline optical amplifiers and in Link B two amplifiers were used. To reach both link configurations from the laboratory, 6.9 km of buried dispersion shifted fibers (DSF) with relatively low measured PMD and CD values were used; these fibers were buried in 1996.

In order to monitor the instantaneous DGD for each fiber length, a polarimeter was used with a tunable probe laser set to the operating wavelength of the actual optical signal. Prior to each measurement, the optical signal was turned off and the probe laser was turned on in order to analyze the DGD performance of the field fiber configuration. After the BER versus OSNR performance test was performed the same procedure was repeated so that the DGD values before and after each test were recorded. Furthermore, a manual polarization controller was inserted at the transmitter side to assure the same state of polarization (SOP) launch of both laser sources. The CD and the DGD values before and after the OSNR measurement for Link A (including 2xDSF) were 666.1 ps/nm, 93.3 ps and 92.9 ps and for Link B (including 4xDSF) 2639.9 ps/nm, 97.9 ps and 96.1 ps. The DGD values for both links are close to each other whereas the longer link has a higher overall CD.

III. RESULTS AND DISCUSSION

The MLSE performance was evaluated by measuring the pre and post-FEC error rates. In order to count possible burst errors especially at low BER during our post-FEC measurements, the duration of each measured BER was at least 10 minutes. The results for back-to-back (b2b), Link A (including 2xDSF) and Link B (including 4xDSF) are given in Fig. 3. The OSNR value for b2b at a post-FEC error rate of 1E-11 is 12.4 dB. This is rather a large value compared with previous results [5][6]. One reason could be the relatively low extinction ratio (ER), which was 10.6 dB. In typical applications the ER is at least 13 dB. Another reason could be the high noise figure of the booster used at the receiver. Fig. 3 further shows the eye diagrams after the optical signal having propagated through both links respectively. All eye diagrams were measured prior to the receiver input of the MLSE.

Link A shows an OSNR penalty of 2.75 dB and Link B a penalty of 3.75 dB at a post-FEC error rate of 1E-11 with respect to b2b. These fairly low penalties indicate the superior performance of the MLSE with a required assumption of an excellent clock and data recovery of it. Furthermore, one would expect a linear increase in the OSNR penalty with the joint impact of PMD and CD. However, as PMD being the major contribution of errors to the transmitted signal, the addition of significant CD would rather add a minor contribution of further penalty to the performance of the MLSE.

IV. CONCLUSION

We have demonstrated the performance of a commercial EDC in a field trial where it applies the MLSE based technology. The unit allows the simultaneous equalization of the detrimental effects of PMD and CD that can be seen in fiber configurations typically deployed in core networks, especially for fiber plants with relatively old fiber spans having high PMD values. This effective compensation would enable service providers to introduce a cost effective and robust solution in their transmission systems in the backbone and metro network. It would also give the opportunity to upgrade such networks to 10Gb/s data rate with minimal disruption and without expensive new system designs.

REFERENCES

Mitigating Sampling Phase Sensitivity of the MLSE by Overlapping Branch Metrics

F.N. Hauske (1), B. Lankl (1), E.-D. Schmidt (2)
1: University of the Federal Armed Forces, Munich, Institute for Communications Engineering, D-85577 Neubiberg, phone: +49-89-60043924, fax: +49-89-60043921, fabian.hauske@unibw.de
2: Siemens Networks GmbH & Co. KG, Hofmannstr. 51, D-81379 Munich, Germany

Abstract A new method, which applies overlapping branch metrics of adjacent symbols, mitigates sampling phase sensitivity towards a large window without performance loss and without increasing the sampling rate.

Introduction
Recently electrical signal processing (ESP) has been investigated into for different modulation formats [1][2] to relax system penalties induced by optical linear and nonlinear distortions. In addition electrical equalization can adaptively compensate for optical and electrical distortions at a low cost. There is common consensus that electrical low-pass filtering with over-sampled Maximum Likelihood Sequence Estimation (MLSE) yields near-optimum performance compared to feed-forward (FFE) or decision feedback equalizers (DFE) [3]. Most investigations and experiments focus on optical impairment mitigation avoiding challenges of clock recovery implementation and sampling phase sensitivity. The implementation of an electrical MLSE for a 10Gbit/s transmission system has already been demonstrated experimentally [4]. It implies a four-state Viterbi decoder using a non-parametric channel estimation based on histograms to compute the branch metrics, which are stored in lookup tables. At a sampling rate of 20Gsamples/s (2 samples/symbol), an enhanced clock recovery controls the sampling phase in the ADC. From [3] we know that over-sampling is less sensitive to sampling phase variations and the choice of the sampling phase than classical Nyquist sampling. In addition it was shown in [5] that an MLSE with oversampling performs better in terms of required OSNR.

In the following we propose an alternative computation of the branch metrics with overlapping branch metrics for adjacent sampling instants. This allows sampling phase variations within the duration of two symbols or more without decreasing the performance. At the same time we improve the back-to-back performance.

Metric Robust to the Sampling Phase
With a j-fold oversampling we receive j samples \( r_{x,i}(i) \), \( x=1,...,j \) for every symbol slot \( i \), which lasts one symbol duration \( T \). We define the sampling phase \( \phi \) as the deviation from the open eye located at \( T/2 \) for undistorted transmission. A negative \( \phi \) refers to a sampling phase preceding to \( T/2 \), a positive \( \phi \) refers to a sampling phase succeeding \( T/2 \). The sampling instants refer to \( \phi(1,1)=\phi \) in case of classical Nyquist sampling, to \( \phi(1,2)=\phi-T/4 \) and \( \phi(2,2)=\phi+T/4 \) for 2-fold over-sampling and to

\[
\begin{align*}
\hat{D} &= \arg \max_{s} \left[ \sum_{i} \sum_{x=1}^{j} \ln p(r_{x,i}(i)\mid S) \right] \\
&+ \sum_{i} \sum_{x=j/2+1}^{j} \ln p(r_{x,i}(i-1)\mid S) \\
&+ \sum_{i} \sum_{x=1}^{j/2} \ln p(r_{x,i}(i+1)\mid S)
\end{align*}
\]

Figure 1: Schematic contribution to branch metrics for 2s2PDF (circle) and 2s4PDF (box)

\( \phi(1,3)=\phi-T/3, \quad \phi(2,3)=\phi \quad \text{and} \quad \phi(3,3)=\phi+T/3 \) for 3-fold oversampling.

The Viterbi-Algorithm (VA), as an implementation of the MLSE principle, applies a state probabilistic model with conditioned probability density functions (PDF) \( p(r(i)\mid S) \) stored in a lookup table. The state transition \( S \) is composed of \( O \) leading symbols and \( P \) trailing symbols interfering with the actual symbol. Typically we build one lookup table for every sampling instant \( r_{x,i} \). In the following, we will refer to these cases as “1s1PDF” for \( j=1 \), “2s2PDF” for \( j=2 \) and “3s3PDF” for \( j=3 \). The VA estimates the most likely digital sequence \( b \) maximizing the path metric. For \( j=2 \) the path metric is exemplarily shown in Eq. (1) as it is known from [2]. The path metrics for \( j=1 \) and \( j=3 \) can be derived similarly.

Fig.1 makes clear that for one step \( i \) in the branch metric classical schemes only apply \( j \) samples within the same symbol duration. However, the grouping is arbitrary and could be shifted by one sampling instant as well. To avoid such a hard cutoff, we propose an new scheme, which applies 2-fold over-sampling with 4 PDFs called “2s4PDF”. Now every sampling instant \( \phi(1,2) \) and \( \phi(2,2) \) is employed by both adjacent branch metrics \( i-1 \) and \( i+1 \) as well, leading to the additional contributions of Eq. (2) and Eq. (3). Yet, the 2 PDFs referring to the overlapping sample look differently and contain different information.
Numerical Results
To evaluate the performance of the four sampling schemes we use Monte-Carlo simulations. A PRBS of 2^11 bit is sent with 10Gbit/s NRZ-OOK modulation (low launch power) over 100km of SSMF and a relevant DCF to adjust residual dispersion, simulated by split-step-Fourier-method, including distortions like chromatic dispersion and SPM. The noise loaded waveforms are fed into the receiver. The optical Gaussian BP filter (1st order, 25GHz), the photo diode and the electrical LP Bessel filter (20th order, 4GHz) are followed by a 4bit A/D-converter. For equalization we applied a 4-state Viterbi-Algorithm leading to 8 state transitions and PDFs represented in the lookup tables.

Fig.2 and Fig.3 show the sampling phase sensitivity for variations of ϕ in case of three different distortions by residual dispersion. For CD=0ps/nm Nyquist sampling is clearly limited to the open eye for optimum performance. Shifting the sampling phase to adjacent symbols still leads to satisfactory results because the Viterbi-Algorithm partly compensates such offset. Over-sampling clearly opens the window for the optimum sampling instant. 2-fold over-sampling suffers a small penalty if none of the samples is located at the open eye. The scheme of 2s4PDFs proves the widest window in all cases, even outperforming the 3-fold over-sampling.

The PDFs were histogrammatically estimated without tail extrapolation. Especially for low distortions this can lead to badly conditioned PDFs. Applying 4 PDFs in case of 2s4PDF, this error can accumulate leading to a variation in performance, as it can be seen in Fig.2. Enhanced channel acquisition algorithms and tail extrapolation of the PDFs would avoid such influence.

In case of an optimum choice of the sampling phase the BER performance vs. OSNR (0.1nm) is shown in Fig.4. For a BER of 10^-2 and below 2s4PDF never performs worse than 2s2PDF and almost always performs equal or better than 3s3PDF (not plotted in the figure).

Further simulations including strong non-linear distortions confirm the robust behaviour of 2s4PDF with respect to sampling phase sensitivity and with respect to BER performance as well.

Conclusion
We propose a new method with overlapping branch metrics, which applies a number of N lookup tables out of a j-fold over-sampling with N>j. Especially in case of distortion by high residual chromatic dispersion, where the phase recovery of the clock signal might be challenging, the method clearly mitigates sampling phase sensitivity. Applying 2-fold over-sampling the scheme 2s4PDF even outperforms 3-fold over-sampling. The method comes only with a slight increase in complexity and would be easy to be implemented in state of the art equalizers like the MLSE.

References
1 Haunstein et al., ECOC2004, paper Th1.5.1
2 Cavallari et al., OFC2004, paper TuG2
3 Haunstein et al., OFC2001, paper WAA4
4 Färbert et al., ECOC2004, paper Th4.1.5
5 Stojanovic et al., ITG FG 5.3.1, Workshop Nov. 2005
6 Foggi et al., JLT, 24, 3073-3087 (2006)
Abstract: Requirements and operation of an optical regenerator for photonic network are discussed. An optical decision gate is a key element to achieve fully transparent regenerators. All-optical regenerators using optical fiber switches demonstrate ultra-high speed regeneration.

Introduction

Optical regeneration is an essential technology for improving the flexibility of photonic networks [1]. The record for maximum transmission capacity through a single fiber has been increasing year after year by increasing the bandwidth of optical amplifiers and developing devices and system technologies [2,3]. Systematic limitation mainly comes from the degradation of the signal-to-noise ratio (S/N) and the signal quality due to fiber transmission, optical amplification, optical routing, etc. Optical regeneration is expected to provide a function to increase system margin in optical transmission and achieve flexible networks. All-optical regenerators are especially attractive for high-capacity networks, where optoelectronic regenerators will no longer be appropriate. Several methods of optical regeneration have been investigated over the last years [4-21]. Semiconductor based optical devices are expected to achieve compact, cost-effective optical regenerators. Fiber-based optical regenerators are attractive because they can provide an ultra-broad band and ultra-high speed operation at data rates sufficiently higher than 100 Gb/s [7,18,19]. Recent developments in nanophotonic technologies and silicon-photonics suggest the possibility of ultra-densely integrated optical regenerators in the future.

This paper describes requirements for an optical regenerator and shows the current status of our technical development.

Optical 2R/3R-regeneration

In an optical 2R-regenerator two functions are required, (i) reamplification and (ii) reshaping, whereas in a 3R-regenerator (iii) retiming function is required in addition to (i) and (ii). The operation of reamplification (1R-regeneration) is basically the same as that of linear amplification.

Waveform deterioration and noise associated with transmission should be reduced by the reshaping functionality (Fig. 1). The noise above could be either amplitude/phase noise or signal jitter. The reshaping functionality is typical in the 3R-regenerator and is usually achieved in two steps. First, by synchronizing a train of optical pulses (optical clock pulses) generated locally by a high-quality pulse source at a repetition frequency corresponding to the line rate of the data signal, and second, by switching thus generated optical clock pulses in the decision gate by the data signal. An optical clock recovery element is usually used for the retiming. For clock recovery, the most popular method is based on converting the optical signal to an electric signal for extracting the clock signal and then on generating the optical clock pulses using pulse lasers or high-speed optical modulators, but some all-optical retiming methods have also been proposed [22-24].

Several types of optical regenerator have been proposed and demonstrated. Those optical regenerators use optical soliton technologies [4-7], semiconductor-based interferometers [8-11], an electroabsorption modulator [12], saturable absorbers [13,14], fiber-based optical switches or limiter amplifiers [15-20], and spectrum broadening and optical filtering [7,21].

Transparent operation is mostly important for devices to be applied to optical regenerators. The performance of the regenerator should not depend on the bit rate, modulation format, pulse shape or data patterns. As shown in the previous demonstrations mentioned above, reshaping functionalities are usually achieved by nonlinear optical effects. The nonlinear optical effects used in the regenerator components should then be sufficiently faster than the data rate. Another important issue is that the wavelength of the regenerated signal should be the same as that of the input data signal in order to make the design of the transmission link and switching nodes simple [18,19].

Optical regeneration using fiber switch

Fiber-based optical switches provide ultra-broadband [17] and ultra-high speed optical signal processing at data rates sufficiently higher than 100 Gb/s [25-27].
Such ultra high-speed optical fiber switches are promising candidates for achieving transparent optical regenerators. A highly-nonlinear dispersion-shifted fiber (HNLF) [28, 29] is effective for a high-speed fiber switch because it provides an efficient third-order nonlinearity within a short length, which facilitates fine control of the dispersion.

Using an optical Kerr-switch as a decision gate, we demonstrated 160-Gb/s optical 3R-regenerating transmission [18]. The optical 3R-regenerator was comprised of the pulse shaper, the decision gate, the optical clock recovery, and the wavelength shifter. In addition to reduce the amplitude noise it also demonstrated the jitter reduction by combining pulse shaping and optical switching. Using optical 3R-regeneration, we achieved error-free operation in a circulating-loop transmission. The result showed that the quality and the wavelength of the regenerated signal was the same as that of the original signal, which is an essential prerequisite of the 3R-regenerator in transmission systems and networks. Using an optical parametrically amplified fiber switch showed in [27], we demonstrated amplitude noise suppression of both OOK and RZ-DPSK data signals by utilizing gain saturation of parametric amplification in an HNLF [19]. The switch successfully achieved amplitude noise suppression, thereby improving receiver sensitivity for both 160-Gb/s OOK and DPSK signals. The switch also preserved the wavelength of the input data signal and featured ultra-broadband and transparent operation. The results indicated that the phase noise was not affected during the switching process. Note that amplitude noise suppression could be effective to reduce signal deterioration of a phase-modulated signal due to amplitude-noise to phase-noise conversion in fiber transmission and demultiplexing. The regenerating effect could be further improved by combining the optical fiber switch with a saturable absorbing element.

Conclusion

The requirements and basic operation of optical regenerators were discussed. To achieve features such as reshaping, suppression of amplitude/phase noise and timing jitter, a fully transparent optical decision gate is necessary. Our recent results in optical regeneration using optical fiber switches were shown as candidates to achieve an all-optical regenerator with ultra-broadband and ultra-high speed performance.

Further improvement in nonlinearity and transparent features of the reshaping element would be the key issues in achieving optical regenerators. Recent developments in nano-optical devices [30,31] and silicon-photonic-based devices [32] demonstrate that they could realize breakthrough in achieving optical regeneration in future photonic networks.

Acknowledgements

The author acknowledges F. Futami and R. Okabe at Fujitsu Laboratories for their contribution to this work and discussion, and H.G. Weber, R. Ludwig and their colleagues at the Heinrich-Hertz Institut (HHI) for their collaboration and discussion.

References

11. B. Lavigne et al., OFC 2003 , PDP15
13. M. Gay et al., ECOC2006, Tu1.3.3.
20. M. Matsumoto et al., ECOC2006, Tu1.3.5.
23. K. Igarashi et al., ECOC2006 PDP Th4.4.2.
27. S. Watanabe et al., ECOC2004, PDP Th4.1.6.
29. T. Nakanishi et al., OFC2006, OTuH7.
32. S. Ayotte et al., OFC2007, PDP42.
DPSK Signal Restoration Using Four-Wave Mixing in a Dispersion-Flattened Highly Nonlinear Photonic Crystal Fiber

Mable P. Fok and Chester Shu
Department of Electronic Engineering and Center for Advanced Research in Photonics, The Chinese University of Hong Kong, Shatin, N. T., Hong Kong.
Tel: (852) 2609 8263 Fax: (852) 2603 5558 E-mail: mfpfok@ee.cuhk.edu.hk

Abstract
We demonstrate phase-noise reduction of a 10-Gb/s DPSK signal using pump-modulated four-wave mixing in a photonic crystal fiber. The extinction ratios at the decoder outputs are improved by 5 and 27 dB.

1 Introduction
Differential phase-shift keying (DPSK) data format has received much attention in its immunity to fiber nonlinearities [1]. However, phase noise of a DPSK signal will accumulate during transmission and processing of the signal, resulting in a degradation of the signal quality. In this work, we develop an all-optical approach to reduce the phase noise and to enhance the extinction ratio using pump-modulated four-wave mixing (FWM) [2]. The DPSK input acts as a pump source during the process. A dispersion-flattened, highly nonlinear photonic crystal fiber (PCF) is used as the FWM medium to provide a large optical bandwidth of operation [3]. The power penalty is improved by over 5 dB. Also, the extinction ratios are enhanced by 5 and 27 dB at the destructive and constructive ports of the decoder, respectively.

2 Principle and Experimental Setup
To optimize the extinction ratio for a DPSK signal, a \( \pi \) phase shift at bit “1” and a zero phase shift at bit “0” are required. The phase noise and the amplitude noise should also be minimized. Fig. 1(a) plots the calculated phase dependence of the decoded signal intensity for the case of destructive interference between the data bits. As shown by the dark curve, the output intensity is non-zero when the phase of bit “1” is smaller than \( \pi \). By applying pump-modulated FWM and filtering out the converted signal at a frequency of \( (\omega_s - \omega_{cw}) \) from the input signal, the phase of bit 1 will be doubled. Here, \( \omega_s \) and \( \omega_{cw} \) are the frequencies of the DPSK signal and the cw pump, respectively. The intensity of the decoded output is pushed towards the zero level within the shaded region, resulting in an extinction ratio enhancement. When the phase of bit “1” varies between 0.3 \( \pi \) and 0.7 \( \pi \), the normalized intensity output changes by 0.3 after FWM. The result compares favorably to an intensity change of 0.65 in the decoded output of the original signal. Hence, phase noise reduction can be obtained simultaneously with extinction ratio enhancement. The phase noise is converted into amplitude noise after the decoding process. Fig. 1(b) shows a reduction of the amplitude noise in the output when the phase of the bit “1” is set at 0.5 \( \pi \). The light curve shows an improvement over the dark curve as a result of phase noise doubling in the FWM process.

Fig. 1 Effect of four-wave mixing on (a) the normalized output intensity at destructive interference of the data bits. The extinction ratio is enhanced in the shaded region. (b) the normalized amplitude noise in the decoded output for an input with 0.5 \( \pi \) phase shift for bit “1”.

Our experimental setup is shown in Fig. 2. A 10-Gb/s NRZ-DPSK signal at 1555 nm is generated at the phase modulator (PM). The driving amplitude is chosen to yield a 0.5 \( \pi \) phase shift for bit “1”. Phase noise is also added to the signal by passing it through another sinusoidally
driven PM. A tunable laser source is combined with the DPSK signal and the combined light is amplified and launched to a 64-m PCF to introduce FWM. The PCF has a dispersion coefficient of -1.3 ps/(km·nm) at 1550 nm. The dispersion slope is \(-10^3\) ps/(km·nm\(^2\)) and the nonlinear coefficient is 11.2 (W·km\(^{-1}\)).

**3 Results and Discussion**

The eye diagram of the 1555-nm DPSK signal with added phase noise is measured by decoding the signal with a 50-ps delay interferometer (DI). The results are shown in Fig. 3. The insufficient phase shift for bit “1” and the addition of phase noise result in a poor signal quality. Fig. 3(a) depicts the eye diagram at the destructive port showing a semi-closed eye with an extinction ratio of 24 dB. The eye is completely closed at the constructive port as shown in Fig. 3(b).

By filtering out the converted wavelength at 1553 nm from the FWM output, phase noise reduction and extinction ratio enhancement are obtained and the results are shown in Fig. 3(c) and (d). The eyes are clearly open and extinction ratio improvements of 5 and 27 dB are obtained at the destructive and constructive ports, respectively. By tuning the cw light from 1545 to 1565 nm, the converted wavelength has been changed from 1565 to 1545 nm. Over the tuning range, the output extinction ratio varies by 1.5 dB and improvements in both the extinction ratio and the eye opening are obtained.

The bit-error rate measurement results are shown in Fig. 4. At the destructive port, it is clearly observed that the regenerated signal shows a significant improvement in the BER performance and the error floor is removed. At the constructive port, the BER of the degraded input signal is very large and cannot be determined. Instead, the BER of the DPSK input without added phase noise is plotted. The regenerated signal, even with the addition of phase noise, shows a significant improvement.

**4 Conclusion**

Phase noise reduction and extinction ratio enhancement have been demonstrated for a 10-Gb/s NRZ-DPSK signal using four-wave mixing in a dispersion-flattened and highly nonlinear PCF. The scheme has been experimentally demonstrated over a converted wavelength range of 20 nm. The eye opening is enhanced along with a significant improvement of the power penalty in the BER measurement.

**Acknowledgment**

The authors thank Crystal Fibre A/S for providing the PCF. This work is supported by the Research Grants Council of Hong Kong (Project CUHK 4184/04E).

**References**

Experimental Demonstration of Filter-Free Wavelength Conversion
Using Nonlinear Optical Loop Mirror at 10Gbit/s

Motomine KANNAN, Shoichiro ODA, and Akihiro MARUTA
Graduate School of Engineering, Osaka University
2-1 Yamada-oka, Suita, Osaka 565-0871, Japan
TEL : +81-6-6879-7696, FAX : +81-6-6879-7688
E-mail : maruta@comm.eng.osaka-u.ac.jp

Abstract : We propose a novel filter-free all-optical wavelength conversion scheme using nonlinear optical loop mirror. Error-free operation with a conversion bandwidth of 20nm at a power penalty of 2dB is experimentally achieved at 10Gbit/s.

Introduction
An all-optical wavelength conversion (AOWC) is indispensable subsystem to avoid a packet contention in a photonic network. Conventional AOWCs reported previously require an optical bandpass filter (OBPF) to select the wavelength-converted signal. When each packet needs to be converted to different wavelength, the tuning speed of the OBPF may limit the system performance. Therefore, the development of the filter-free AOWC is expected. Only a few papers, however, have been reported [1-3].

In this paper, we propose a novel filter-free AOWC using nonlinear optical loop mirror (NOLM). The error-free operation is experimentally demonstrated at 10 Gbit/s.

Principle of Filter-Free AOWC
Figure 1 shows the schematic diagram of the proposed filter-free AOWC, which is almost same as a conventional NOLM-based AOWC [4,5]. In this scheme, the polarization state of the signal light with a wavelength of is orthogonal with respect to the one of the probe light with . When the nonlinear phase shift of the clockwise probe induced by the signal is , the probe is switched to the transmission port. Passing through the polarizer which is aligned so that only the probe passes through, we obtain only the wavelength-converted signal with without filtering. Note that the effective nonlinearity of the fiber in the NOLM reduces to one-third, compared with the case that the polarization state of the signal and that of the probe is parallel.

Experimental Demonstration
The experimental setup is shown in Fig. 2. The electro-absorption modulator (EAM)-based pulse source generates a pulse train with a pulse width of 13ps and a wavelength of 1550nm at 10Gbit/s. The pulse train is data-modulated with a 10Gbit/s pseudo-random bit sequence (PRBS) and also phase-modulated with a sinusoidal signal at 500MHz to suppress stimulated Brillouin scattering (SBS). The amplified pulse train by an EDFA is used as the signal and coupled with the NOLM through a polarization beam splitter (PBS). The input peak power of the signal is 650mW. The probe light from a laser diode (LD) is amplified to 0dBm and launched into the NOLM. The average zero dispersion wavelength, dispersion slope, nonlinear coefficient, fiber loss, and length of the HNLF in the NOLM are 1556nm, 0.03ps/nm$^2$/km, 12W$^{-1}$km$^{-1}$, 1.7dB/km, and 500m, respectively. At the output of the NOLM, the eye diagrams and spectra are observed by an oscilloscope (OSC) and an optical spectrum analyzer (OSA). We also measure the bit error rate of the wavelength-converted signal.
The fluctuation of the relative polarization state between the input signal and the probe causes noise in the converted signal. The power penalty of the signal converted from 1550nm to 1551.5nm is 3.7dB compared with the back-to-back. This further penalty can be attributed to the interference between the converted signal and the original signal which is leaked through the polarizer. Note that we could not achieve the error-free conversion from 1550nm to 1551nm because of the further interference. Although the signal with 1550nm could not be converted to 1551nm, we successfully demonstrated the filter-free AOWC from 1545nm to 1565nm at 10Gbit/s.

Conclusions
We have proposed a novel filter-free AOWC using NOLM, in which the polarization state of the signal and that of the probe is set to be orthogonal. We have demonstrated the error-free conversion with a bandwidth of 20nm at 10Gbit/s. Although the bit rate of the signal is limited to 10Gbit/s due to the laboratory constraint, this scheme can be scaled to 160Gbit/s thanks to the ultra fast response time of Kerr nonlinearity.

References
160 Gb/s retiming using rectangular pulses generated using a superstructured fibre Bragg grating

L.K. Oxenløwe (1), F. Parmigiani (2), M. Galili (1), D. Zibar, A.T. Clausen (1), M. Ibsen (2), P. Petropoulos (2), D.J. Richardson (2) and P. Jeppesen (1)
1 : COM•DTU, Technical University of Denmark, Building 343, DK-2800 Kgs. Lyngby, Denmark, lo@com.dtu.dk
2 : ORC, University of Southampton, Highfield, Southampton, SO17 1BJ, United Kingdom, frp@orc.soton.ac.uk

Abstract A 160 Gb/s retiming scheme incorporating a superstructured fibre Bragg grating for pulse shaping is demonstrated. Retiming is performed in a single step without wavelength conversion, and a substantial bit error rate improvement is achieved.

Introduction
At high-speed serial data rates, timing jitter becomes a serious detrimental factor. Timing jitter of data pulses should in general be less than about 5% of the high repetition rate timeslot, but at bit rates of 160 Gbit/s (timeslot: 6.25 ps) and above, it gets increasingly difficult to find pulse sources that can fulfil these requirements, in particular after transmission. Therefore retiming in a regenerator is a crucial functionality for high-speed systems. Most optical regenerator schemes also imply wavelength conversion, though it is often desirable to maintain the wavelength, without extra wavelength conversion [1].

In this paper, we demonstrate a 160 Gb/s pulse retiming scheme, which maintains the original wavelength and retimes the data in a single step. The retiming scheme contains a polarisation-insensitive superstructured fibre Bragg grating (SSFBG) [2, 3] with a sinc-shaped transfer function, to shape the incoming data pulses into rectangular pulses, and a polarisation-rotating Kerr switch based on 200 m highly non-linear fibre (HNLF). Using this scheme, a 160 Gb/s data signal with an error floor limited by rms timing jitter, is successfully retimed to become error free.

Experimental procedure
The experimental set-up is shown in Figure 1.

Figure 1. 160 Gb/s retiming set-up.

The heart of the set-up is the SSFBG and the Kerr switch. The SSFBG is used to shape the 160 Gb/s data pulses into ~5 ps rectangular pulses [2]. These are then aligned in polarisation so as to be attenuated in the Kerr switch. A short clock pulse then rotates the polarisation of the part of the rectangular data pulse that overlaps with the clock pulse, thereby allowing this part of the original data pulse to be transmitted through the polariser. This configuration ensures that the switched retimed pulse maintains its original wavelength, since it is actually part of the original data pulse. That is to say, the retimed pulses are carved out by the short clock pulses, which have low timing jitter, and hence, the retimed data pulses adopt the low jitter of the clock pulses. The clock pulse being narrower than the flattened signal generated by the SSFBG ensures that the switching is insensitive to mistiming in the data signal.

For the data signal, a 10 GHz semiconductor tunable mode-locked laser (TMLL) is used – the wavelength is set to 1557 nm, the FWHM pulse width is 1.8 ps and the rms timing jitter is ~410 fs. These pulses are data modulated (MOD) with a 2^7-1 PRBS and subsequently multiplexed in a PRBS and polarisation maintaining fibre-based pulse-interleaving multiplexer (MUX). The data is amplified, injected into the SSFBG generating the rectangular pulses and, through a polarisation controller (PC), aligned at 90° to the Kerr switch polariser. For the clock pulses, an Erbium-glass mode-locked laser (ERGO) with low rms timing jitter (~210 fs) is used – the wavelength is 1544 nm and the FWHM is 1.3 ps when linearly compressed in dispersion compensating fibre (DCF). The pulses are synchronised to the data pulses via the same synthesiser in this back-to-back set-up, and are multiplexed in an additional fibre-based polarisation maintaining multiplexer. The 160 GHz clean pulse train is amplified and filtered and its state of polarisation is aligned at 45° to the Kerr switch polariser. The Kerr switch contains a 200 m HNLF with dispersion slope ~ 0.017 ps/nm/km, zero dispersion at 1551 nm, and is linearly compressed in dispersion compensating fibre (DCF). These fibre properties ensure negligible pulse-to-pulse walk-off and pulse broadening.

The 160 Gb/s retimed data pulses have adopted the clock FWHM and jitter (i.e. ~1.2 ps and ~250 fs, respectively). In order to perform BER characterisation, the retimed data signal is demultiplexed to 10 Gb/s in as non-linear optical loop mirror (NOLM) demultiplexer with a 50 m HNLF (slope 0.018 ps/nm/km, zero dispersion wavelength at 1554 nm, γ ~ 10.5 W⁻¹km⁻¹).

Figure 1. 160 Gb/s retiming set-up.
Experimental results
Figure 2 shows the response of the SSFBG to the input pulse from the TMLL – in this case at 10 GHz.

The input spectral FWHM is ~2 nm, and the output spectrum is clearly sinc-like. Note that the slight spectral asymmetry originates from a corresponding slight asymmetry in the SSFBG spectral response. The FWHM of the measured temporal pulse is ~5.7 ps, though the flanks are not as steep as expected (due to the limited width of the input spectrum). The temporal width is measured with a cross-correlator with a 600 fs sampling pulse, so the de-convoluted trace has slightly steeper flanks. The flat top of the pulse extends over 2 ps, which is enough to eliminate most of the ~410 fs rms timing jitter of the data pulses.

Figure 3 shows results before and after the retiming system at 160 Gb/s. The input to the Kerr switch is 1.3 ps clock pulses and 5.7 ps rectangular data pulses. The spectrum shows the 160 GHz tones on the sinc-like spectrum. The clock pulses are aligned to the nominal centre of the data pulses, and thus sample this part of the waveform. As only the central part of the data pulse is sampled, and the clock pulses are so short, the retimed pulses are only 1.2 ps wide, and the rms timing jitter is drastically reduced to ~250 fs (as measured at 40 Gb/s). Accordingly, the sampled data pulses have a broader spectrum but are still centred on the original wavelength.

Figure 4 shows the BER characteristics with and without retiming. All 16 channels are successfully retimed and error free operation is achieved for all of them. This is in sharp contrast to the case when retiming is not applied – in that case, the 410 fs rms jitter creates a severe error floor for all channels, and a BER of 1E-6 is the best that can be obtained (at the maximum receiver power, -20 dBm). This clearly reveals the benefit of the retimer. Among the retimed channels with an average sensitivity of -27 dBm, there is a 5 dB sensitivity spread, which is due to uneven multiplexed pulses in the data and clock arms, and to polarisation drifts in the used Kerr switch.

Conclusions
We have presented a retiming scheme and successfully demonstrated its use in a full 160 Gb/s experiment. The technique relies on linear pulse shaping in a SSFBG and subsequent sampling in a HNLF-based Kerr switch. The clock pulses used to gate the Kerr switch have low timing jitter, and this is adopted by the data signal, thus rendering the data signal error free, something which was not possible without the retiming.

Acknowledgements
To the Danish Research Council funded project UltraNet, and to the COST-291 programme.

References
All optical clock recovery at 10 GHz using a Fabry-Perot laser diode

Xiaohui Fang¹, P. K. A. Wai¹, Chao Lu¹, H. Y. Tam², and K. K. Qureshi¹

¹Photonics Research Center and Department of Electronic and Information Engineering
The Hong Kong Polytechnic University, Hung Hom, Kowloon, Hong Kong
²Photonics Research Center and Department of Electrical Engineering
The Hong Kong Polytechnic University, Hung Hom, Kowloon, Hong Kong

enxfang@eie.polyu.edu.hk

Abstract: We reported all optical clock recovery at 10 GHz using a Fabry-Perot laser diode. The time bandwidth of the recovered clock pulse is ~0.5 and the RMS timing jitter is 470 fs.

1. Introduction
Clock recovery is very important because it synchronizes the receiver decision circuitry with the incoming data stream at optical communication nodes in ultra high speed communication networks. Optical clock recovery have been demonstrated by filtering out the clock frequency from the data signal including tank circuit using passive Fabry-Perot (FP) filter [1], based on self-pulsating distribute feedback (DFB) laser [2], or via injection locking of a long cavity FP laser diode (LD) [3]. All these all-optical clock recovery methods are simple and low cost, but the clock signal can only be recovered at a fixed frequency corresponding to the free spectral range of the LD used.

In this paper, we present a novel, simple, and low cost method for all optical clock recovery based on the switching of two longitudinal modes in a dc biased multi-quantum-well (MQW) FP-LD. The MQW FP-LD is of model number OL5204n-120/P20 acquired from OKI Electric Industry Co. Ltd. The MQW FP-LD has the advantages of rapid response time and high output power when compared to the commonly used FP-LDs with double-channel planar buried InGaAsP heterostructure. The central wavelength of the FP-LD is ~1550 nm, the wavelength spacing is 0.5 nm, and the threshold current is 17.6 mA. The dc biased current is ~45 mA.

The FP-LD is external injection-locked by a 10 GHz RZ pseudorandom binary sequence (PRBS) at one of FP-LD modes, with length of $2^{11}$ bits, pulsewidth of 50 ps and intensity of ~3 dBm. The FP-LD is simultaneously self-seeded at another mode. The intensity of the externally injected signal is higher than that of the feedback signal. So self seeding can be realized only at those time slots when the external RZ sequence is not a “1”. By tuning the optical delay line (ODL) such that the repetition rate of the external RZ

2. Experiment setup and operation principle

Figure 1 shows the experiment setup for all-optical clock recovery based on the switching of two longitudinal modes in a dc biased FP-LD. The MQW FP-LD is of model number OL5204n-120/P20 acquired from OKI Electric Industry Co. Ltd. The MQW FP-LD has the advantages of rapid response time and high output power when compared to the commonly used FP-LDs with double-channel planar buried InGaAsP heterostructure. The central wavelength of the FP-LD is ~1550 nm, the wavelength spacing is 0.5 nm, and the threshold current is 17.6 mA. The dc biased current is ~45 mA.

The FP-LD is external injection-locked by a 10 GHz RZ pseudorandom binary sequence (PRBS) at one of FP-LD modes, with length of $2^{11}$ bits, pulsewidth of 50 ps and intensity of ~3 dBm. The FP-LD is simultaneously self-seeded at another mode. The intensity of the externally injected signal is higher than that of the feedback signal. So self seeding can be realized only at those time slots when the external RZ sequence is not a “1”. By tuning the optical delay line (ODL) such that the repetition rate of the external RZ
signal is integral multiples of the fundamental frequency of the external self-seeding loop, clock signals can be recovered at the output of self seeding loop as shown in Fig. 1. An EDFA is used to amplify the feedback signal, and a polarization controller (PC) is used to optimize the switching between the two injection locked modes

3. Results and discussion

Figure 2 shows the recovered clock spectrum (dark curve), free running spectrum of the dc biased FP-LD (blue dotted curve) and the reflection spectrum of a uniform grating (pink curve) of length of 2 cm. The center reflection wavelength of the grating is about 1551.54 nm which is ~0.1 nm detune from the free running mode at 1551.42 nm and the grating bandwidth is about 0.23 nm. When the input data signal injection-locked a longitudinal mode of the FP-LD at 1557.03 nm, all other modes is suppressed and their wavelengths are shifted toward the short wavelength side. The 0.1 nm detune between the reflection wavelength of the grating and mode at 1551.42 nm is then large enough to be outside the locking range at this time slot. In fig. 2, the recovered clock is at the same wavelength of 1551.54 nm, and it’s intensity is ~29dB higher than that of external injection locking at 1557.03 nm, other modes is suppressed and filtered out. We cannot get a good 10 GHz clock output if a zero detune between the reflection wavelength and the nearest longitudinal mode of the FP-LD is used. Figure 3 shows the output clock recovery spectrum at 10 GHz. The pulsewidth is 50 ps and the time bandwidth product then is 0.5.

We measured the timing jitter information by observing the RF spectrum at scan span of 100 KHz as shown in Fig. 4. The RMS timing jitter is determined to be 470 fs [4].

4 Conclusion

All optical clock recovery at 10 GHz based on the switching of two longitudinal modes of a dc biased Fabry–Perot laser diode is presented. The time bandwidth of the recovered clock signal is 0.5 and the RMS timing jitter is 470 fs. We find that appropriate detune between the reflection wavelength of the uniform grating and nearest FP-LD longitudinal mode results in a good clock pulse with low noise and low jitter. The proposed clock recovery system is simple and cost effective.

5 Acknowledgement

This work was supported by The Hong Kong Polytechnic University under Project G-YX79.

6 Reference

We experimentally demonstrate true FEC-encoded PRBS transmission at 20 Gbps using a live precoder for return-to-zero differential quadrature phase-shift keying, together with a block turbo code forward error correction with soft-decoding.

1. Introduction

Modulation formats have been a major concern for the high speed optical transmission systems that are expected to exhibit a larger transmission capacity through higher spectral efficiencies [1]. Return-to-zero differential quadrature phase-shift keying (RZ-DQPSK) is a promising format owing to its high spectral efficiency, larger tolerance to chromatic dispersion and polarization-mode-dispersion as compared to RZ-DPSK. On the other hand, forward error correction (FEC) is also indispensable for improving received BER performance. Recently, we have developed above two technologies, i.e. RZ-DQPSK [2] and block turbo code (BTC) FEC [3]. However, to date each technology challenge has been pursued independently, and no experimental demonstration has yet been reported of these in a true combination.

In this paper, we demonstrate actual BTC FEC encoded RZ-DQPSK transmission at 24.8 Gbps using a live evaluation circuit. Very large scale LSI technology and a novel differential precoder [4] make implementation of this combination possible.

2. Block Turbo Code FEC encoded DQPSK

DQPSK requires a differential precoder circuit to map the input information onto four phase states. A well-known serial differential precoder consists of a flip-flop with a feedback path. However, when the symbol rate exceeds 10 Gsymbols/s, propagation delay caused by the feedback path creates a bottleneck in a circuit implemented in currently available high-speed semiconductor technology, e.g. SiGe HBT. Therefore, most transmission experiments reported to date used either pre-calculated pseudo random binary sequence (PRBS) patterns or programmed data received without precoding. We designed a novel parallelized precoder for DQPSK with a parallel prefix network, which is the well known computer engineers’ adder circuit. The proposed precoder makes it possible simultaneously to reduce the circuit size at a rate of \( k \log_2 k \), and the required speed at a rate of \( (s \log_2 k)/k \) [4], where \( k \) is the parallel expansion number and \( s \) is the symbol rate. It allows for an implementation by standard CMOS process.

Concatenated codes based on hard decision and iterative decoding are widely used in submarine line terminals. One of the best performing second generation FECs shows an NCG of 9.4 dB with a redundancy of 25%. That appears in ITU-T G.975.1 as two orthogonally interleaved BCH codes. We have developed a superior FEC LSI using BTC with 3-bit soft decision that outperforms every existing FEC LSI. The codeword is based on a BCH(144,128) x BCH(256,239) product code having a minimum distance of 30. The bit rate for STM-64 is 12.4 Gb/s with a redundancy of 23.6%. The 0.13 \( \mu \)m CMOS technology enabled full integration into a circuit volume of approximately 16 M gates. The soft decision IC was manufactured in 0.2 \( \mu \)m SiGe BiCMOS with \( f_t = 120 \) GHz. Approximately 20,000 bipolar and 1,500 CMOS transistors were placed on the 9 x 9 mm chip.

![2x BTC FECs and Precoder FPGA](image)
The LSIs show an NCG of 10.1 dB at 12.4 Gb/s and provide a SONET/SDH compliant section and line processing for OC-192/STM-64. The input client signal is mapped in two ways. The G.709 FEC in the incoming OTN signal is decoded first, then the OTU2 overhead is extracted. In the case of an STM-64 client signal, the overhead is terminated and its performance is monitored. The client signal is then mapped onto the BTC framer with an OPU2/ODU2/OTU2v overhead. The encoded 12.4 Gb/s signal is output to the line side. At the receiver side, three 12.4 Gb/s signals, i.e., the hard decision information and the MSB and LSB confidence information from the soft decision IC, are decoded by the BTC decoder. The OTU2v/ODU2/OPU2 overhead is terminated and demapped by the BTC de-framer. The OTN overhead is generated before launching to the client side.

3. Experiment and discussion
We built the BTC FEC LSI and the parallel prefix precoder onto a live circuit board. The precoder was implemented in a commercially available FPGA, operating at only 97 Mb/s, with a parallel expansion $k$ of 128. The circuit board, shown in Fig. 1, was designed for a 24.8 Gb/s (=12.4 Gsymbols/s) RZ-DQPSK system.

![Fig.2 Experimental setup for BTC FEC encoded 20 Gb/s RZ-DQPSK using proposed precoder.](image)

Fig. 2 illustrates the experimental setup for true FEC encoded/decoded 20 Gb/s PRBS transmission. A PRBS31 at 10 Gb/s was FEC-encoded by the block turbo code LSI with FEC redundancy of 23.6%, then split and electrically decorrelated by the 1280 bit delay circuit in order to generate two independent tributaries. Each tributary was differentially precoded by the precoder, serialized by the 128:1 electrical MUX, then converted to 24.8 Gb/s RZ-DQPSK optical signals by a parallel Mach-Zehnder I/Q modulator followed by a pulse-carving AM modulator. The I/Q modulator was biased at its null points and driven with a peak-to-peak voltage of 2Vpk by the precoded data streams. After some meters of single-mode fiber transmission, the optically preamplified RZ-DQPSK signal was input into a Mach-Zehnder delay interferometer (MZDI) followed by a balanced photo-detector. One of the tributaries was selected by adjusting the phase of the MZDI. A clear eye opening was observed. The received signal was FEC-decoded by the FEC LSI. The BER of the decoded data was measured at 10 Gb/s.

The error correction performance was tested by changing the received power. The result is plotted in Fig. 3. Worse BER was successfully improved by BTC FEC. An output BER of better than $3 \times 10^{-12}$ was obtained when a received power of -42.3dBm. Taking the 23.6% bit rate increase into consideration, the net coding gain was 10.1 dB at the post-FEC BER of $10^{-13}$. This is fully the same error correction capability that was measured previously for RZ-OOK [3]. Narrow spectrum of RZ-DQPSK was observed as an inset of Fig. 3.

![Fig.3 Measured BER performance for 24.8 Gbps RZ-DQPSK with BTC FEC](image)

4. Conclusion
We have demonstrated two promising technologies, i.e. RZ-DQPSK with the differential precoder and block turbo code FEC. We showed experimentally the BER performance of the combination of RZ-DQPSK and BTC FEC. These two technologies will together open up new possibilities for robust capacity deployment in the very high speed transmission systems.

References
Abstract: We report experiments that elucidate the shift of the optimum dispersion compensation value in an optical fiber transmission system due to phase modulation in electroabsorption external modulator. As a countermeasure, we introduce the use of a lithium niobate (LiNbO₃) phase modulator.

I. INTRODUCTION

Chromatic dispersion of the optical fiber is a factor that can limit the transmission distance possible in an optical fiber transmission system [1-3]. Dispersion compensation fiber with negative dispersion is generally used to offset the chromatic dispersion. However, since the external intensity modulator varies the imaginary part of the refractive index, the real part will suffer some level of modulation due to the Kramers-Kronig (K-K) relation [1]. This causes phase modulation (PM) of the optical signal passing through the optical modulator together with intensity modulation (IM). The phase modulation in the external modulator can affect the deterioration caused by the chromatic dispersion of the transmitted signal.

This paper describes the deterioration of the transmitted signal due to phase modulation in the electroabsorption modulator and introduces a countermeasure, the use of a LiNbO₃ phase modulator.

II. SHIFT IN OPTIMUM COMPENSATION VALUE BY PHASE MODULATION IN THE ELECTROABSORPTION MODULATOR

Figure 1 shows the experimental setup used to measure the effect of phase modulation in the electroabsorption modulator (EAM). A 1558.98 nm coherent light wave was input to an EAM. In the EAM, the light wave was modulated by a 12 GHz signal. Here, the bias voltage of the EAM was set to -2.08 V, and optical modulation index (OMI) to 40%. The modulated light wave was passed through a single mode fiber (SMF) or dispersion compensation fiber (DCF) and detected by a photo diode. SMF yields positive dispersion (0~800 ps/nm), and DCF yields negative dispersion (0~ -800 ps/nm). Detected 12 GHz signal power was measured by a spectrum analyzer.

Figure 2 shows the dependency of detected 12 GHz signal power on chromatic dispersion. In the figure, the optimum dispersion compensation value shifts -200 ps/nm from 0 ps/nm. In the EAM, since the external intensity modulator varies the imaginary part of the refractive index due to the modulation signal, the real part will suffer some level of modulation according to the Kramers-Kronig (K-K) relation [1]. This causes phase modulation (PM) of the optical signal passing through that optical modulator along with intensity modulation (IM). Moreover, the intensities and phases of the optical sidebands vary due to interference by the modulation components. As a result, the optimum compensation value shifts from 0 ps/nm to -200 ps/nm.

III. THE MODULATION SECTION TO OFFSET THE SHIFT USING LiNbO₃ PHASE MODULATOR

We investigate the use of the LiNbO₃ phase modulator to offset the shift described above. First of all, we describe about the experimental setup of the modulation section to offset the shift in the EAM.

Figure 3 shows the configuration of the proposed modulation section. In this section, the...
intensity-modulated optical signal from the EAM is input to a LiNbO$_3$ phase modulator. The LiNbO$_3$ phase modulator modulates the phase of the input optical signal under the control of the electrical modulation signal input to the LiNbO$_3$ phase modulator. Here, the power and the phase of this electrical signal are set by an attenuator and a phase shifter. The optical signal from the LiNbO$_3$ phase modulator is input to SMFs or DCFs shown in Fig.1. Detected 12 GHz signal power is measured at spectrum analyzer. Wavelength of the optical signal is 1558.98 nm, bias voltage of the EAM is -2.08 V, respectively.

Next, we describe about measurement results. Figure 4 shows dependency of the shift of the optimum compensation value on relative power of signal input to the LiNbO$_3$ modulator. As input signal power increases, the optimum compensation value shifts in a positive direction.

Figure 5 shows the shift of the complete compensation value at four values of relative phase of the signal input to the LiNbO$_3$ modulator. The horizontal axis indicates the relative phase of the signal input to the LiNbO$_3$ modulator with the signal input to the EAM. Relative input power of the signal is set to -7 dB in the Fig.4. The shift changes by 150 ps/nm when relative phase of the signal changes by $\pi$ radian.

These results indicate that we can optimize the dispersion compensation value by varying the power and the phase of the signal input to the LiNbO$_3$ phase modulator.

**IV. CONCLUSION**

We clarified that the phase modulation experienced in an EAM shifts the optimum compensation value. We also showed how to optimize the dispersion compensation value with the use of a LiNbO$_3$ phase modulator.

**REFERENCES**

Adaptive Steepest-Descent-Feedback Control of Tunable Dispersion Compensators using A Three-Point Sampling Method in Time-Domain Waveforms

Ken Tanizawa and Akira Hirose
Department of Electronic Engineering, The University of Tokyo,
7-3-1 Hongo, Bunkyo-Ku, Tokyo 113-8656, Japan, e-mail: tanizawa@eis.t.u-tokyo.ac.jp

Abstract
We propose a three-point waveform-sampling method for the optimal adaptive control of tunable dispersion compensators in optical dynamic-routing networks. We succeeded in decreasing the residual dispersion drastically in 10 Gb/s transmission simulations.

1 Introduction
Adaptive dispersion compensation using a tunable dispersion compensator (TDC) is essential in optical dynamic routing networks and the ultra high speed transmission over 160 Gb/s per channel. We have proposed and demonstrated a high speed, low cost adaptive TDC control method [1]. The method is based on a feedback control in which the peak eye-opening value is used as a feedback signal and is maximized by the steepest descent method.

We achieved a high speed control by the proposed algorithm based on the steepest descent method. Also, the implementation cost of the method is lower than the dispersion measurement method [2] as we do not need costly measurement instruments.

However, we face the problem of residual dispersion in the previous method; the dispersion is not perfectly compensated for after the maximization of the peak eye-opening value. The reason is that only the peak eye-opening value is used as the feedback signal.

In this paper, we report a new control algorithm based on three-point sampling in time-domain waveforms. We conducted 10 Gb/s transmission simulations and succeeded decreasing the residual dispersion adequately.

2 Adaptive control of TDC using three-point sampling method
Fig.1 shows the optical dynamic routing networks with our proposed steepest-descent-feedback control method. The control algorithm is divided into three steps.

The first step, (i) is a calculation of error value, \( Er \), which is the difference between the received and the reference waveforms. The reference waveform which is a received waveform unaffected by the dispersion is measured and registered before transmission. The detailed definition of the \( Er \) is shown in Fig.2. In the previous method, we obtained the \( Er \) as the difference of the peak values between the received and reference waveforms as shown in Fig.2 (a). Fig.2 (b) shows the three-point sampling method proposed in this paper. The \( Er \) is calculated as the summation of the difference of three points between the received and reference waveforms.

\[
Er = \frac{1}{2} \sum_{n=1}^{N} (R_n - P_n)^2
\]

The new definition of \( Er \) is effective in decreasing the

Fig.1 Schematic diagram of optical dynamic-routing networks with proposed adaptive control method
residual dispersion as it represents the effect of the dispersion on the waveform accurately.

Step (ii) is a calculation of the partial derivative of $E_r$ for the steepest descent approach. In the following simulations, we adapted virtually imaged phased array (VIPA) [3] as a TDC. The partial derivative with respect to the VIPA’s control parameter, $S_{ps/nm}$, is approximated as

$$
\frac{\partial E_r}{\partial S} = \frac{1}{2} \sum_{n=1}^{w} (P_n - R_n) \frac{2P_n}{\partial S}
$$

(2)

$$
\frac{\partial P_n}{\partial S} = \frac{P_n^2}{T_{FWHM}} \sqrt{1 - P_n^2} \left( \frac{n^2 P_n^2}{8} - 1 \right) \exp \left( -n^2 P_n^2 \frac{16}{2\pi} \right)
$$

(3)

where $T_{FWHM}$ is the full width at half maximum of the transmitted signal, $\lambda$ is the center wavelength and $c$ is the speed of light.

The final step (iii) is an update of $S$ by the steepest descent method.

$$
S \Rightarrow S - \epsilon \frac{\partial E_r}{\partial S}
$$

(4)

where $\epsilon$ is an appropriate constant. We repeat these steps until the $E_r$ becomes small enough.

3 Transmission simulations at 10Gb/s

We conducted transmission simulations by OptiSystem [4] to confirm the effectiveness of the proposed method. Fig.3 shows the simulation model. The transmission speed was set at 10 Gb/s and the modulation format was nonreturn-to-zero. The transmission fiber was SMF and the dispersion was 18 ps/nm/km. We measured the BERs and eye-diagrams when the transmission route changed from Route1 (where the dispersion had been compensated for perfectly; $S = -2700$ ps/nm) to Route2.

Fig.4 shows the eye-diagrams after the compensation. The eye-opening after the compensation by using the three-point sampling method is wider than the one obtained using the peak sampling method. The BER after compensation was $3.7 \times 10^{-10}$ for the peak sampling method and $<10^{-12}$ for the three-point sampling method respectively. Fig.5 shows the compensating dispersion for every update of the VIPA. The residual dispersion after the compensation by the three-point sampling method is about half of the residual dispersion when we use the peak sampling method.

4 Conclusion

We have reported an optimal feedback control method of the TDC in this paper. This method is based on the steepest descent method and the three-point sampling in time-domain waveform. Using this method, we can implement high speed and low cost optimal adaptive dispersion compensation.

5 References


Proposal for Coordinate Transformed Electronic Pre-compensator and its Robustness to Bias Error

Takashi Sugihaara, Takashi Mizuochi, Hiroshi Kubo, and Katsuhiro Shimizu
Information Technology R&D Center, MITSUBISHI ELECTRIC CORPORATION
5-1-1 Ofuna, Kamakura, Kanagawa 247-8501, Japan.
Tel: +81-467-41-2441, Fax: +81-467-41-2486: E-mail: Sugihaara.Takashi@ak.MitsubishiElectric.co.jp

Abstract: We propose a novel electronic pre-compensation transmitter, the coordinates of whose I/Q channels are transformed in order to reduce excess loss. Robustness to bias error in its modulator is investigated numerically.

1. Introduction

Electronic pre-compensation for equalizing chromatic dispersion in optical fiber transmission systems has been rigorously investigated [1-5]. Compared to prior optical dispersion compensation methods, electronic pre-compensation is expected to reduce power consumption and equipment size, to enable adaptive equalization, and to increase the range of compensatable dispersion values. In the pre-compensator, an optical I/Q modulator based on a dual parallel Mach-Zehnder modulator (MZM) [6] is usually used to generate an arbitrary complex electric field. However, this conventional structure has a problem with unwanted excess loss. Meanwhile, the modulator bias of a dual parallel MZM is quite important for the establishment of good pre-compensation performance. However, tolerance to bias offsets has not been studied to date.

In this paper, we propose a novel pre-compensator architecture whose I/Q coordinates are transformed. This enables the excess insertion loss to be mitigated as compared to a conventional pre-compensator. We also present the tolerance to bias offset in a dual parallel MZM.

2. Coordinate Transformed Pre-compensator

Fig. 1 shows the schematic diagram of a conventional electronic pre-compensator. The input data sequence is converted to two orthogonal pre-distorted signals, i.e. a real (I-ch) and an imaginary (Q-ch) part. The signal processing is executed by inverse-convolution between the input data stream and the impulse response of the transfer function of the optical fiber link. When the pre-compensator generates a signal for 0 ps/nm dispersion, 3 dB excess loss occurs. This is because the imaginary part (Q-ch) does not exist for 0 ps/nm. To avoid the excess loss, we propose the coordinate transformation of the I/Q channels in the dual parallel MZM shown in Fig.2. Here, after the real and the imaginary parts are combined and the coordinates are rotated through angle $\theta$, a transformed I/Q signal is produced as shown in the equations below.

\[ D_r(t)' = \cos \theta \cdot D_r(t) + \sin \theta \cdot D_i(t), \]  
\[ D_i(t)' = -\sin \theta \cdot D_r(t) + \cos \theta \cdot D_i(t), \]  

where $D_r(t)$ and $D_i(t)$ are the original real and imaginary parts for pre-compensation.

Fig. 3 shows the complex electric fields of a single pulse transmitted from the conventional (a) and proposed (b) pre-compensator for a pre-compensation value of 0 ps/nm. In the conventional case, there is no Q-ch component. The light traveling along the Q-ch arm is lost when constructing the dual parallel MZM. This causes an unwanted 3dB excess loss as mentioned above.
In contrast, the coordinates are rotated by $\pi/4$ in the case of the proposed configuration as shown in Fig.3(b). Since the I-ch and Q-ch amplitudes become equal, no excess loss occurs.

Fig. 4 shows the calculated I-ch/Q-ch power ratio for the coordinate transformed pre-compensator. In this calculation, we assumed an NRZ, PRBS 9, 40 Gb/s data sequence. The DAC was assumed to work at 80 Gsample/s – 6 bits quantization, which corresponds to double over-sampling of the symbol rate. We treated the processor of the pre-compensator as an FIR (Finite Impulse Response) filter. The number of taps of the FIR filter was 65 symbols. The total dispersion being pre-compensated was varied from 0 to 2000 ps/nm. As shown in Fig. 4, $\theta=\pi/4$ results in the same output power in the I-ch and Q-ch regardless of the pre-compensation value. This result shows that the coordinate transformation scheme is effective for dispersion pre-compensation.

We calculated the excess loss against the rotation angle for a range of dispersions being pre-compensated, as shown in Fig. 5. The maximum improvement of 3dB was obtained when a rotation angle of $\pi/4$ is chosen for 0 ps/nm. The excess loss changed cyclically with the rotation angle.

![Fig.4. Q-ch/I-ch power ratio with coordinate transformation for various dispersions.](image)

![Fig.5. Excess loss vs. rotation angle for various dispersions.](image)

3. Tolerance to Bias Control Error

Three independent bias control units are necessary for I-ch MZM, Q-ch MZM and the relative phase between the I-ch and Q-ch. Because the different I-ch/Q-ch power ratio due to the coordinate transformation could change the tolerance to bias offsets, we investigated the tolerance to bias error with and without the coordinate transformation. Fig. 6 shows the eye opening penalty vs. bias errors in the I-ch and Q-ch, and the phase difference between the I-ch and Q-ch. The penalties are normalized to the minimum penalty for each $\theta$. In these calculations, the total amount of dispersion to be pre-compensated is 2000 ps/nm, and the DAC has 6 quantization bits. The rotation angles are 0 and $\pi/4$. The allowable bias error window for $\theta=\pi/4$ is 0.1 $\pi$ for an eye opening penalty of 0.5dB. This is comparable to the requirement of a conventional pre-compensator ($\theta=0$).

![Fig.6. Eye opening penalty vs. bias error: (a) I-ch MZM (b) Q-ch MZM (c) Phase difference between I-ch and Q-ch.](image)

4. Conclusion

We have proposed a novel electronic pre-compensator with coordinate transformation. The excess loss of the dual parallel MZM was improved by up to 3dB thanks to the $\pi/4$ axis rotation. We also show the tolerance to bias error of the dual parallel MZM. The results show that the axis rotation is effective in diminishing the insertion loss of the modulator, while retaining the same robustness as a conventional pre-compensator.

Acknowledgments

This work was in part supported by a project of the National Institute of Information and Communications Technology (NICT), as part of a program of the Ministry of Public Management, Home Affairs, Posts and Telecommunications of Japan.

References

Record PMD Mitigation of 11 ps for 43 Gb/s RZ-DPSK by Distributed Polarisation Scrambling

Axel Klekamp and Henning Bülow, Alcatel Research and Innovation
Alcatel-Lucent Deutschland AG, 70435 Stuttgart Holderaeckerstr. 35, 70499 Stuttgart, Germany
Axel.Klekamp@alcatel-lucent.de

Abstract
We experimentally demonstrate a record mean DGD tolerance improvement of 220% to 11ps for 43Gb/s RZ-DPSK by a cascade of 5 distributed polarisation scramblers and UFEC. This tolerance for RZ exceeds the NRZ value by 2ps.

Introduction
Polarization mode dispersion (PMD) is considered as a serious limitation for implementation of data rates of 43 Gb/s per channel and beyond in optical long-haul communication systems. For 10 Gb/s PMD mitigation by polarization scrambling at the transmitter only [1] or distributed fast polarization scrambling was proposed by numerical evaluations [2,3]. The basic principle of these schemes is that the scramblers accelerate the PMD statistics of the transmission fiber down to sub FEC frame duration (e.g. 12ps for OTU 2 FEC). Hence PMD-induced outage times are transformed to short error bursts and can be corrected by the FEC. Improvement by distributed fast scrambling was experimentally demonstrated with FEC at 10.7 Gb/s [4] in a loop experiment emulating 16 cascaded scramblers. Recently successful operation at 42.7 Gb/s [5] was shown in a straight-line experiment with 5 cascaded scramblers. Here, we experimentally investigate the mean DGD tolerance at 43 Gb/s RZ-DPSK modulation format by using 5 fast polarization scramblers and compare the results with NRZ-DPSK. A scrambling speed of 50MHz allowed to investigate the performance of FEC’s potential burst error correction capability.

Experimental set-up
Fig. 1 depicts the experimental set up for the measurement. At the transmitter the RZ-DPSK signal is generated by a LiNbO₃ Mach Zehnder modulator driven by a 43Gb/s DPSK pre-coder followed by an RZ carver for RZ-50% with bias at 3dB point and 2x√π/2 modulation with 43.08 GHz. The data signal of one of the four 10Gb/s tributaries is obtained by FEC encoding and decoding of a 2²³-1 PRBS sequence (UFEC with net coding gain of 8.6 dB). The other three 10Gb/s tributaries are loaded with 2²³-1 PRBS sequence. To emulate the transmission over a fiber link loaded with slowly drifting PMD, the 43Gb/s signal was sent over 5 differential group delay (DGD) elements, as illustrated in Fig.1. The first of the 5 DGD elements was a tunable DGD emulator and the other 4 elements were polarization maintaining fibres (PMF) of fixed length (e.g. 3.44 ps, 3.44 ps, 3.0 ps, and 2.5 ps). Each fast scrambler (SCR) consists of a LiNbO₃ polarization modulator with 3 sections of 3 differently oriented waveplates (see inset in Fig.1). The retardation of the waveplates is modulated via 3 electrodes. They are driven at slightly different frequencies of f, f+0.2 MHz, f+0.4 MHz. The first two scramblers could only be operated at 1 MHz, whereas the last three scramblers were driven at 50 MHz. It was proven by simulation that PMD induced distortion dynamics i.e. jitter and eye-opening penalty (EOP), is dominated by the fastest scrambling frequency in the line [5]. The frequency offset of 0.2 MHz was chosen to be faster than the FEC frame rate (approx. 85 kHz). In front of the tunable DGD-emulator a slow mechanical SCR is used to ensure worst case measurements of PMD effects. Receiver OSNR was varied by an attenuator and 1dB OSNR penalty is generally measured with respect to back-to-back.

In the case of DPSK the signal was also sent over a straight line of 4 spans of 80 km SMF as described in [5]. However, no impact of the fiber (SMF) on the mean DGD tolerance was observed and therefore the RZ measurements were performed without fiber link. The receiver comprises a 1-bit delay interferometer to demodulate the DPSK signal and a balanced detector in front of a SF15 deserialiser chip. The 4 tributaries at the deserialiser outputs which belong to the 10.7Gb/s FEC channel were multiplexed and launched into the FEC decoder. The measurements at post FEC-BER were generally performed over several minutes and some samples (at the 1 dB limit) were also measured over longer periods (up to 2 hours) to ensure no occurrence of uncorrectable error bursts.

Fig. 1: Lab set-up of 43Gb/s DPSK-NRZ/RZ transmission link with 5 fast pol. scramblers, 5 DGD emulators and enhanced FEC on one 10.7Gb/s tributary.
**Results and discussion**

Fig. 2 displays the OSNR penalty versus mean DGD (mean DGD = $\sqrt{(DGD_1^2 + \ldots + DGD_5^2)}$) measured at the pre-FEC BER of $10^{-3}$ (Fig. 2a, error free after FEC) and at post-FEC-BER $10^{-9}$ (Fig. 2b) for 5 scramblers running at 50 MHz. The mean DGD tolerance converges to the upper limit of the fiber PMD tolerance for distributed scrambling [6]. Since recent measurements [5] revealed that for the current set-up a pre-FEC BER of $10^{-3}$ is already a sufficient criteria for error-free post-FEC operations even for fast polarisation scrambling, OSNR penalty measurements at this pre-FEC BER give a good picture of the mean DGD tolerance and are shown in Fig. 2a).

**Summary**

A record PMD mitigation of 11 ps (47% of the bit period) by distributed fast polarization scrambling and forward error correction has been demonstrated for the modulation format RZ-DPSK at a bit-rate of 43Gb/s. This is an increase in mean PMD tolerance by a factor of 3.2 compared to the case without scrambling. Measurements with pre-FEC BER indicate that a FEC without burst error correction can achieve a mean DGD tolerance of 10.2 ps for RZ-DPSK.

**Acknowledgement**

This work was partially supported by the German BMBF in frame of the EIBONE/FLINTSTONE project.

**References**

[1] B. Wedding et al., OFC 2001, paper WAA1
Ultra-compact highly nonlinear fiber module technologies

M. Takahashi, Y. Mimura, J. Hiroishi, M. Tadakuma, R. Sugizaki, and T. Yagi
(Furukawa Electric Co., Ltd., 6, Yawata-Kaigandori, Ichihara, Chiba 290 JAPAN,
Tel: +81-436-42-1728, Fax: +81-436-42-9340, Email: takahashi.masanori@furukawa.co.jp)

Abstract
We summarize the state of the art down-sized HNLF modules and their applications. This paper provides the theoretical limits of reducing cladding diameter of HNLF. Also, ultra-compact HNLF modules and SC light generation using this module is introduced.

1 Introduction
Highly nonlinear fiber (HNLF)\(^1\) is a key component holding great promises in applications in the all-optical signal processing\(^2\),\(^3\), broadband light sources\(^4\), optical pulse compressors\(^5\), etc., which are essential for future high-speed transmission systems.

To produce commercially viable HNLF requires more compact modules. Previously it was possible, using 90μm HNLF, to develop a module size of a 3.5-inch floppy disk\(^6\), but any further reduction in module sizes would require even finer fiber. Since HNLF has an extremely small mode field diameter (MFD), it is advantageous in terms of down-sizing but any major reduction in cladding diameter would require investigation of the theoretical design.

In this paper, we report our study of theoretical limit of down-sizing of HNLF and characteristics of fabricated down-sized HNLF and HNLF modules. Furthermore, we introduce our experimental result of supercontinuum (SC) generation by the fabricated compact module.

2 Theoretical approach to down-sized design
Transmission in optical fibers occurs generally by utilizing the difference in refractive index between the core and the cladding to confine the light in the area of the core. The intensity of this light decreased exponentially as the distance from the core becomes greater, but when the cladding diameter (cladding thickness) is insufficient, the transmission mode reaches the interface between the cladding and the coating and large leakage loss occurs. Thus in reducing cladding diameter it is necessary to clarify the cladding diameter required with respect to the design of the core.

By means of simulation using the finite element method (FEM), we analyzed the relationship between cladding diameter and leakage loss. In this simulation we used the refractive index profile of the core actually adopter in HNLF and that used in ordinary SMF. For each core design, simulations were carried out for leakage loss when cladding diameter was changed with no change in core diameter.

Figure 1 shows the relationship between cladding diameter and leakage loss obtained by simulation, demonstrating that the thinner the cladding diameter the greater the leakage loss. Comparing HNLF and SMF we can see that even an HNLF with a cladding diameter of 40μm has lower loss than 125μm SMF.

Figure 2 shows the leakage loss under macro bending. It should be noticed that as the cladding becomes thinner and light confinement weaker, the ability to withstand bending changes. However, it can be seen that an HNLF having a core designed for extremely strong light confinement will be only slightly susceptible to the effects of bending, and even when the cladding is thinner the ability to withstand bending will not change to any degree.
3 Characteristics of down-sized HNLF

Table 1 shows the characteristics of one of our prototype down-sized HNLFs designed and fabricated based on study of down-sizing. Cladding diameter of this HNLF is 51μm and sufficient cladding thickness is realized. Coating diameter is 85μm by the effect of down-sized cladding, and volume of this HNLF is reduced to 1/10 comparing to conventional fibers with 250μm coating. The basic optical characteristics shown in Table 1 are similar to those for an HNLF with a cladding diameter of 125μm having the same core design.

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cladding diameter</td>
<td>51 μm</td>
</tr>
<tr>
<td>Coating diameter</td>
<td>85 μm</td>
</tr>
<tr>
<td>Dispersion</td>
<td>0.78 ps/nm/km</td>
</tr>
<tr>
<td>Dispersion slope</td>
<td>0.017 ps/nm²/km</td>
</tr>
<tr>
<td>( \lambda_0 ) (nm)</td>
<td>1610</td>
</tr>
<tr>
<td>( \gamma ) (SPM)</td>
<td>13.8 W/km⁻¹</td>
</tr>
<tr>
<td>Attenuation loss</td>
<td>1.14 dB/km</td>
</tr>
<tr>
<td>PMD</td>
<td>0.05 ps/km¹/²</td>
</tr>
</tbody>
</table>

Bobbin size can be reduced drastically by small cladding and coating diameter of HNLF. Figure 3 shows small bobbins for down-sized HNLF. 400m of HNLF can be accommodated in the coin sized bobbin.

4 SC generation by down-sized HNLF

We conducted SC light generation experiment using fabricated ultra-compact module achieved by means of down-sized HNLF. SC light generated in the normal dispersion domain of HNLF is low noise, and since it has a good SNR, it is promising for applications in optical signal processing or as a broadband light source.

SC light generation experiment was carried out using the fabricated 1000m HNLF module shown in figure 3. In this experiment, the repetition rate of the pulse source was set at 10GHz, the pulse width was 2.1ps. Figure 4 shows the SC spectrum at an input power of 100mW. The spectrum bandwidth at the points where power decreased 20dB form peak power was 32nm, and ripple around the pump wavelength was sufficiently small at about 3dB.

5 Summary

We introduced recently reported down-sized HNLF. Theoretical discussion of the limits of reducing cladding diameter and characteristics of ultra-compact HNLF modules realized by means of this theory are briefly reviewed. We also show the result of SC light generation experiment which is carried out using fabricated ultra-compact modules, and satisfactory characteristics are confirmed.

6 References

1) M. Takahashi et al., JLT 23, No.11, 2005.
3) S. Watanabe et al., OFC2003, PD 16
4) Y. Takushima et al., PTL 11, No.3, 1999.
7) M. Takahashi et al., ECOC2005, We3.4.2.
9) M. Takahashi et al., ECOC2006, Th.1.5.1.
Comparison of Wavelength Converters Composed of HNL-DSF and Bi-HNLF by a New Figure-of-Merit

Shinji Yamashita and Hiroshi Kuno
Department of Electronic Engineering, Faculty of Engineering, The University of Tokyo
7-3-1 Hongo, Bunkyo-Ku Tokyo, 113-8656, JAPAN
Tel.: +81-3-5841-6659, Fax.:+81-3-5841-6025
syama@sagnac.t.u-tokyo.ac.jp, hkuno@sagnac.t.u-tokyo.ac.jp

Abstract—We propose here a new non-linear fiber figure-of-merit: Applicative FOM, by adding in FWM conversion bandwidth. AFOM is conceivably a novel figure for evaluating non-linear fibers, since it takes in account the effects of fiber dispersion.

I. INTRODUCTION

Amid non-linear phenomena in optical fibers are often viewed as threats for crosstalks and thereby inhibitions of accurate and fast transmission, they are also expected to have great possibilities in all-optical processing devices. Wavelength converters based on the four-wave mixing (FWM) phenomenon are one of the most interesting non-linear optics based devices [1].

In FWM interaction, a fourth wavelength lightwave is newly generated through the fiber from three different wavelength input lightwaves. Especially, in degenerated FWM, where two of the three input lightwaves have the same wavelength, can be reunderstood as an interaction of two lightwaves generating a third one. In this paper, degenerated FWM will be considered in both theoretical and experimental treatments for simplicity.

With the development in non-linear optics, various highly non-linear optical fibers, such as highly non-linear DSFs (HNL-DSF) and Bismuth-oxide-glass based highly non-linear fibers (Bi-HNLF) have been proposed for effective candidates. However, there are of course, characteristic tradeoffs among these fibers, and hence the figure-of-merit (FOM) of these fibers in application to all-optical processing devices is a very interesting issue.

The most conventional FOM for non-linear fibers, \( FOM_1 \), to be called in this paper, is defined as

\[
FOM_1 = \gamma \cdot L_{\text{eff}},
\]

where \( \gamma \equiv 2\pi n_2/\lambda A_{\text{eff}} \) is the non-linear parameter, with \( n_2 \): non-linear refractive index and \( A_{\text{eff}} \): effective core area, and \( L_{\text{eff}} \) as the effective fiber length. In addition to this, for further accurate evaluation, a second FOM, \( FOM_2 \) just to call, defined as

\[
FOM_2 = \gamma L_{\text{eff}} \cdot P_{\text{SBS}},
\]

has been proposed, where \( P_{\text{SBS}} \) is the SBS threshold [2].

However, when applying non-linear fibers to optical devices, \( FOM_2 \) seems to be insufficient for accurate evaluation, since it does not take in account the fiber dispersion, which is another distinctive characteristic of non-linear fibers. Therefore in this paper, we introduce a new FOM, Applicative FOM (AFOM), to also take in consideration the fiber dispersion, and compare / analyze theoretically and experimentally with HNL-DSF and Bi-HNLF when applied to FWM-based wavelength converters.

II. THEORETICAL AND EXPERIMENTAL TREATMENT

It is important to take in account the fiber dispersion in evaluating applications of non-linear fibers, since they have distinctive dispersion characteristics, as shown in Table 1 [3][4]. Hence we define AFOM as

\[
AFOM = \gamma L_{\text{eff}} \cdot P_{\text{SBS}} \cdot \Delta \lambda,
\]

where \( \Delta \lambda \) is \(-1 \) dB FWM conversion efficiency bandwidth. \( \Delta \lambda \) is largely dependant to fiber dispersion, and is a figure that directly reflects FWM performance. Therefore, it can be said that a fiber with high \( \Delta \lambda \) has a high AFOM, which is rather a natural understanding.

Here, we need to discuss further about \( \Delta \lambda \). With the supposition of the uniformity of the zero dispersion wavelength of the fiber, \( \lambda_0 \), is uniform, FWM conversion efficiency \( \eta \) can be written as

\[
\eta = \frac{\sin^2(\Delta\beta L/2)}{(\Delta\beta L/2)^2},
\]

when the product of fiber loss and fiber length \( \alpha L \) is very small [5]. \( \Delta \beta \) is the phase mismatch between the pump and probe wave, and can be written as

\[
\Delta \beta = -\beta_2 \Delta \omega^2 - \frac{1}{12} \beta_4 \Delta \omega^4,
\]
Here the plus sign is applied when \( \beta \) and \( \eta \) (5) and where \( \beta \) λ tions. The calculation results (Theo. 2), where \( \Delta \lambda \) [dB], and therefore AFOM was obtained through simula-

Here the plus sign is applied when \( \beta \) (ω) ≥ 2.298/β/3L, and the minus sign is applied otherwise. Furthermore, as we can see in Table 1, the effect of \( \Delta \beta \) is too small to be considered as a value for Bi-HNLF, and hence \( \Delta \lambda \) can be simply written as

\[
\Delta \lambda = \frac{\lambda^2}{\pi c} \left\{ \frac{1}{\beta_4} \left( -6 \beta_2 (\omega) \pm 6 \sqrt{\beta_2^2 (\omega) \pm \frac{2.298}{3} \beta_4 L} \right) \right\}^{1/2}
\]

(6)

The important fact is that \( \Delta \lambda \) depends on the wavelengths of pump and probe, thus meaning that AFOM also has a pump wavelength (\( \lambda_p \)) dependency. Calculation results are shown in Fig. 1, as HNL-DSF (Theo. 1).

However, the usual case is that \( \lambda_0 \) is not uniform throughout the fiber; it fluctuates along the fiber, thus giving randomness on \( \Delta \beta \) [6]. This especially fits for the HNL-DSF we have examined, since it has a relatively long fiber length. If we suppose that HNL-DSF has two \( \lambda_0 \), \( \lambda_0 \) and \( \lambda_0 \), its FWM conversion efficiency can be written as [5]:

\[
\eta = \frac{\eta'}{\eta' (\Delta \beta_1 = \Delta \beta_2 = 0)}
\]

(8)

where \( \Delta \beta_1 \) is the phase mismatch from \( \lambda_0 \) and \( \lambda_0 \) respectively, and

\[
\eta' = \left\{ \exp \left[ (\alpha - i \Delta \beta_1) L_1 \right] - 1 \right\} + \exp \left[ (\alpha + i \Delta \beta_1) L_1 \right] \\
\cdot \left\{ \exp \left[ (\alpha - i \Delta \beta_2) L_2 \right] - 1 \right\}^2 \\
- \alpha + i \Delta \beta_2
\]

(9)

Here, it is uneasy to determine \( \Delta \lambda \) which satisfies \( \eta = -1 \) [dB], and therefore AFOM was obtained through simulations. The calculation results (Theo. 2), where \( \lambda_0 = 1570 \)

Fig. 1: Calc. & Measurement Results of AFOM.

III. CONCLUSION

We have introduced a new FOM, AFOM, in applying non-linear fibers to devices by adding the factor of fiber dispersion (\( \Delta \lambda \)) in addition to the conventional FOMs. The greatest success in doing this is that AFOM actually presents the pump wavelength dependence of the performance of non-linear fibers in applied devices.

In this certain case, HNL-DSF can be said to have very high wavelength conversion performance in the zero dispersion region, whereas Bi-HNLF has a wide range pump wavelength selectivity. Hence AFOM has enabled us to evaluate each fiber, depending on the wavelength region intended for conversion usage.

IV. REFERENCES

SC Spectrum Broadening with Tapered Photonic Crystal Fiber

Yusuke Fujii¹, Naoki Karasawa², Yuhei Matsubara², and Soichi Kobayashi²

¹Photonic Science Technology, Inc., 1-3-1 Kashiwadai-minami Chitose, Japan, Tel: +81-123-42-0575 , Fax: +81-123-42-0576, Email: t-suda@psti7.com ²Chitose Institute of Science and Technology, 066-8685 Chitose, Japan, Tel: +81-123-27-6052 , Fax: +81-123-27-6052, Email: s-koba@photon.chitose.ac.jp

Abstract
We fabricated tapered photonic crystal fibers (taper PCF) with a waist diameter in the range of 25–95 μm and a waist length up to 20 mm by using the melting and stretching method. We observed super-continuum (SC) light by launching femtosecond pulse light into the taper PCF. The SC spectrum width depended on the waist diameter and the injected pulse power. The most expanded SC spectrum we observed was from taper PCFs with waist diameters of 95 μm.

1. Introduction
Femtosecond SC light is attractive as a source for communications, sensors, and biomedical applications. Broadening of the SC spectra has been reported for both conventional thin fibers and photonic crystal fibers [1]. The photonic crystal fiber presents advantages of ease in handling and low insertion loss compared with the thin fiber. In this study, we discuss SC spectrum broadening characteristics of tapered photonic crystal fibers with various waist diameters.

2. Fabrication process for taper PCF
Fig.1 shows the system for fabricating taper PCFs. The system comprises a torch and two stages. We used a photonic crystal fiber from Mitsubishi Cable Industries, Ltd (DIAGUIDE PCF LFR-131). The photonic crystal fibers were stretched in the flame from oxygen and propane gases. The taper PCF consists of a tapered region where the outer diameter was decreased from 125 μm to several tens of μm over the waist region. Fig.2 describes the form of a taper PCF with waist length of about 20 mm and waist diameter of 25 μm.

In this study, we fabricated the taper PCFs with waist diameters of 25, 45, and 95 μm. Table 1 lists the core diameters, hole diameters and pitch between holes in the waist region of the taper PCFs.

Injection gas
- O₂ Gas
- Propane Gas

Fig. 1 System for fabricating tapered photonic crystal fibers

Fig. 2 Profile of the taper PCF with waist diameter of 25 μm and waist length of 20 mm.
Table 1. Waist region dimensions at waist region of the taper PCFs wherein 125 μm is a normal PCF.

<table>
<thead>
<tr>
<th>Fiber diameter</th>
<th>Core Diameter</th>
<th>Hole Diameter</th>
<th>Pitch</th>
</tr>
</thead>
<tbody>
<tr>
<td>125</td>
<td>3.7</td>
<td>3.2</td>
<td>2.0</td>
</tr>
<tr>
<td>95</td>
<td>2.8</td>
<td>2.4</td>
<td>1.5</td>
</tr>
<tr>
<td>45</td>
<td>1.2</td>
<td>1.0</td>
<td>0.6</td>
</tr>
<tr>
<td>25</td>
<td>0.7</td>
<td>0.6</td>
<td>0.4</td>
</tr>
</tbody>
</table>

(Unit: μm)

3. Results and Discussion

Fig. 3 illustrates the experimental setup used to measure spectrum characteristics of tapered PCFs. The laser system consists of a Ti:sapphire laser with about 100-fs pulse width, 1 kHz repetition rate and 650-mW average power. The spectra are taken with an ocean optics USB4000 miniature fiber optic spectrometer with a 200–1100 nm detector range.

Fig. 4 shows the intensity variation in the spectrum characteristics when the diameter of the tapered PCF was varied. The output spectrum was broadened with increasing diameter. The reason for the broadening is that the zero dispersion wavelength of the tapered PCF with waist diameter of 95 μm corresponded to the injected laser wavelength of 800 nm [2][3].

Fig. 5 shows spectrum characteristics when the average injection power of the Ti:sapphire laser was varied and the tapered PCF had a waist diameter of 95 μm. The broadened output spectrum was measured in the range of 560–820 nm at 6.0 mW of injection power.

4. Summary

In this paper, we proposed a new tapered PCF and described how we achieved spectrum broadening by injecting a femtosecond pulse into tapered PCFs. An optimum broadened spectrum was obtained over the wavelength range of 560–820 nm with a controlled 2.8-μm core and 95-μm outer diameter in the waist region of a tapered PCF.

Reference

Abstract: We demonstrate the Bragg soliton pulse compression in an exponentially decreasing fiber Bragg grating. Exponential profile is also approximated as six discrete uniform sections. Nearly transform-limited pulse is achieved in both cases.

1. Introduction
Generation of short pulses has always been of great scientific and technological interests. Optical pulse compression is an important technique to generate ultra short optical pulses which find many applications especially in ultra high bit-rate communication systems [1]. Soliton-effect and adiabatic pulse compression techniques have been used for pulse compressors. In soliton pulse compression technique, the compressed pulses typically suffer from significant pedestal generation. Adiabatic pulse compression technique has been used to generate a stable train of pedestal-free pulses but it requires long length of fiber. Optical periodic structures or photonic band gap (PBG) materials such as fiber Bragg gratings (FBGs) have large dispersion (six orders of magnitude larger) compared to silica fibers [1]. Hence, the soliton dynamics could be studied on length scales of centimeters. Moores suggested that chirped solitary waves can be compressed more efficiently if the dispersion decreases approximately exponentially [2]. Recently, self-similar analysis has been utilized to study linearly chirped pulses in fiber amplifiers [3]. In this work, we investigate the possibility of pedestal-free Bragg soliton pulse compression.

2. Pulse Propagation in FBG
Two theoretical models have been used to describe nonlinear pulse propagation in FBGs. The first model uses the nonlinear coupled mode (NLCM) equations which describe the coupling between forward and backward traveling modes [1]. The second model is based on the nonlinear Schrödinger (NLS) type equation which is reduced from the NLCM equations using the multiple scale analysis [1, 4, 5]

$$i \frac{\partial E}{\partial z} = - \beta_{gs}^2(z) \frac{\partial^2 E}{\partial t^2} + \Gamma^s \left| E \right|^2 E = 0,$$

where \( E(z,t) \) is the envelope of the Bloch wave associated with the grating, \( z \) is the distance variable, \( t \) is the time variable, \( \beta_{gs}^2 \) represents the dispersion of the grating, and \( \Gamma^s \) is the effective nonlinear coefficient. For Eq. (1), adiabatic Bragg soliton pulse compression has been discussed wherein the maximum compression factor \( 4 \) was achieved [4,5] and the pedestals generated is very small [4]. In what follows, using self-similar analysis, we show that one can achieve pedestal-free compression with maximum compression factor beyond the limit obtained by the adiabatic compression process. The self-similar solution to Eq. (1) is given by

$$E(z,t) = \frac{1}{\sqrt{1 - \alpha_{gs} D(z)}} \left[ \frac{t-t_0}{1 - \alpha_{gs} D(z)} \right] \exp[i \alpha_0 (z) + i \alpha(z)(t-t_0)^\gamma]$$

where \( \alpha_0(z) = \alpha_{gs} \frac{\lambda_0}{20} \int_0^z \beta_{gs}^2(z') dz' \), \( \alpha(z) = \frac{\alpha_{gs}}{1 - \alpha_{gs} D(z)} \), \( D(z) = \frac{2}{\alpha_{gs}} \beta_{gs}^2(z') dz' \), and \( \alpha_{gs}, \lambda_0 \) are the integration constants. The self-similar solution is possible if and only if the dispersion varies exponentially, i.e.

$$\beta_{gs}^2(z) = \beta_{gs}^2(0) \exp[-2 \alpha_{gs} \beta_{gs}^2(0) z].$$

The function \( R(\theta) \) obeys the equation

$$\frac{d^2 R}{d \theta^2} - n R + 2 \lambda_n R' = 0,$$

where the scaling variable \( \theta \) and the coefficient \( \lambda_n \) are given by

$$\theta = \frac{t-t_0}{1 - \alpha_{gs} D(z)},$$

and \( \lambda_n = -\frac{\Gamma^s}{\beta_{gs}^2(0)} \), respectively. Finally, the bright solitary wave is given by

$$E(z,t) = \frac{1}{\sqrt{\Gamma^s}} \frac{1}{t_0 (1 - \alpha_{gs} D(z))} \text{sech}\left( \frac{t-t_0}{t_0 (1 - \alpha_{gs} D(z))} \right) \exp[i \alpha_0 (z) + i \alpha(z)(t-t_0)^\gamma].$$

where the integration constant \( \lambda_n \) is equal to \( 1/t_0^2 \), and \( t_0 \) is the initial pulsewidth. Equations (2) and (3) are the key results of this work which state that efficient pedestal-free Bragg soliton pulse compression is possible using a nonlinear FBG with an exponentially decreasing dispersion profile. The pulse compression factor is given by \( t_0 \lambda_n(t) = \exp[2 \alpha_{gs} \beta_{gs}^2 z] \) which also equals to the ratio of the initial and final dispersion values.

3. Bragg Soliton Pulse Compression
Self-similar analysis reveals that the dispersion has to be decreased exponentially in the grating. However, such careful profiling of FBG dispersion is not easy in practice. Therefore, we considered FBGs with stepwise approximation (SWA) of the exponentially decreasing dispersion FBG (DDFBG). Stepwise decreasing dispersion could easily be realized by concatenated FBG sections with discrete uniform dispersions. Therefore, Bragg soliton compression studies have been carried out for both DDFBG and SWA. We assume a grating length $L = 6$ cm and the initial pulse width $t_0 = 10$ ps. The initial grating induced dispersion is $\beta_2^g = -33$ ps$^2$/cm and it decreases to $\beta_2^g(L) = -4.56$ ps$^2$/cm for a given initial chirp $\alpha_{20} = -0.005$ THz$^2$. The compression factor is ~7 in both cases. The exponentially DDFBG can be approximated as six discrete uniform sections. If $M$ section is used to approximate exponentially DDFBG, the constant dispersion value $b_i$ at the $i$-th section is given by

$$b_i = \int_{0}^{L/M} \beta_2^g \exp\left(-2 \alpha_{20} \beta_2^g z' \right) dz' / \left(M \right)$$

for both profiles. Figure (3) gives the broadening of the bandwidth during the Bragg soliton pulse compression for both profiles considered in this work.

Figure (4) represents the time-bandwidth product of the pulse during the compression process in exponential DDFBG and SWA profiles. Eventually, the compressed pulse goes to the transform limited case since the (normalized) chirp decreases during the compression process.

4. Conclusions

Bragg soliton pulse compression has been investigated for both exponentially dispersion decreasing FBG as well as stepwise approximation to exponentially dispersion decreasing FBG. No pedestals are observed in the exponential profile. Only a small amount of pedestals (1.12%) appeared when the exponentially decreasing FBG is approximated by concatenated FBG sections with discrete uniform dispersions. Nearly transform-limited pulse has been achieved in the both cases.

5. References

Spatial Evolution of Supercontinuum Generation along a Varying Dispersion Tapered Fiber

Zhaoyang Wang1†, Hiroyasu Sone2, Yasuhide Tsuji3, Masaaki Imai1, and Shinya Sato1

1 Department of Electrical and Electronic Engineering, Muroran Institute of Technology, 27-1 Mizumoto-cho, Muroran 050-8585, Japan
Tel: +81-143-46-5523, Fax: +81-143-46-5501, E-mail: s1461016@mmm.muroran-it.ac.jp
2 Department of Computer Science, Kitami Institute of Technology, 165 Koen-cho, Kitami 090-8507, Japan
3 Department of Electrical and Electronic Engineering Kitami Institute of Technology, 165 Koen-cho, Kitami 090-8507, Japan

Abstract: Ultrabroad supercontinuum spectrum generation in the anomalous- and normal-dispersion regimes of a narrow-waist tapered fiber is theoretically analyzed by taking into account of the dispersion coefficient and effective core area varying along taper transition region.

1. Introduction

Ultrabroad and high coherence supercontinuum (SC) spectra have been obtained using femtosecond laser pulses in tapered fibers [1], which have some unique advantages of strong zero-dispersion wavelength (ZDW) shift, plan physical structure, short length and homemade producing method [2-3]. In this paper, we theoretically discuss the femtosecond pulse propagation through the tapered fiber including the narrow-waist region and the transition regions with decreasing or increasing diameters. The spectral intensity and the unwrapped spectral phase are shown how they evolve along the propagation distance.

2. Theoretical Model

Figure 1 shows the analytical profile of a biconical tapered fiber with 4 mm long untapered regions, 15 mm long transition regions, and 75 mm long waist with a uniform diameter of 2.3 μm. This is similar to the one described by Teipel et al. [2], in which SC spectrum over 500-1100 nm was generated. All simulations presented here are made with the Corning SMF-28 fiber parameters. The transitional region is supposed to obey an exponentially decaying and expanding shape as an adiabatic process for single-mode propagation.

The propagation constant of the fundamental mode at each propagation distance is obtained by solving the vector wave equation using the finite element method with analytical boundary conditions [4]. Then we know that the group velocity dispersion (GVD) varies from normal-dispersion to anomalous-dispersion regime along the preceding transition fiber. The effective core area \( A_{\text{eff}} \) of the fundamental mode is also calculated by scalar field distributions for the guided mode. Figure 2 shows the \( A_{\text{eff}} \) as a function of fiber diameter at a wavelength of 800 nm. The variation of the \( A_{\text{eff}} \) is nonlinear because the power leaks from the core into cladding and then into the surrounding air when the fiber diameter decreases. The intensity distribution for the diameters of 2, 20, 50, and 125 μm was shown in the upper graphs of Fig. 2. For the fiber under test, \( A_{\text{eff}} \) is 2.4 μm² at the waist region. Thus, the important nonlinear parameter \( \gamma \), which is defined as \( \gamma = n_2 \omega / c A_{\text{eff}} \), increases to 98.2 W⁻¹ km⁻¹ for a nonlinear refractive index \( n_2 = 3.0 \times 10^{-20} \text{ m}^2/\text{W} \). This value is approximately 17 times larger than the one at the untapered region and the resulting spectral broadening is induced intensively in the nonlinear processes.

To describe the evolution of intense femtosecond pulses through the tapered fiber, we use the extended NLSE [5] which models enhanced nonlinear effects including SPM, self-steepening and stimulated Raman scattering (SRS). The generalized Raman scattering susceptibility is approximated as a harmonic oscillator model [6] for the molecular vibrations to the Lorentzian function.

![Fig.1. Structure of a tapered fiber.](image)

![Fig.2. Effective area and intensity distribution for different taper diameters at λ = 800 nm.](image)
3. Numerical Results

The NLSE is numerically solved by means of a split-step Fourier method [5] and the enlargement of the SC spectral width due to the combined effects of dispersions and nonlinearities can be evaluated quantitatively. We have assumed the input ultrashort pulse to be hyperbolic secant type with a center wavelength of 800 nm and repetition rate of 80 MHz. These parameters are typical of those used in the abovementioned experimental works [2].

Figure 3 shows the evolutions of (a) spectral intensity \((\text{Intensity} = (\text{Im}[A(z,f)])^2 + (\text{Re}[A(z,f)])^2)\) and (b) spectral phase \((\text{Phase} = \tan^{-1}(\text{Im}[A(z,f)]/\text{Re}[A(z,f)])\) at different distances for output average power of 100 mW and input pulse duration of 380 fs. Here, the phase is processed by unwrapping method in order to maintain a smooth phase shift exceeding over \(2\pi\) rad. The upper and lower abscissas of these figures denote a corresponding wavelength and frequency, respectively. It is evident in Fig. 3(a) that the spectra broaden slightly but almost symmetrically at the initial propagation stage. However, when the pulse propagates down the distance of 68 mm, the spectrum extends its widths dramatically and then remains almost unchanged. The output spectrum covers approximately one octave at the 20-dB level and exhibits a better agreement with the experimental than the simulation result reported in the same Ref. [2].

On the other hand, Fig. 3(b) shows that the corresponding phase evolves in a different way, i.e., it increases significantly after the point of 68 mm, which is called the “critical distance” in our recent paper [6]. This behavior can be understood by noting that the spectral components, which cover not only the anomalous regime but also the normal regime, are accumulated continuously as the SC spectrum propagates down the fiber [5]. It is worthy to note that the critical distance is varied with different dispersion of fiber and/or pulse conditions in addition to average power. In this regard, a highly sophisticated design of the tapered fiber should be taken into account in terms of the critical distance.

4. Conclusions

In summary, we have conducted numerical studies of ultrashort pulse propagation in varying dispersion tapered fibers with the anomalous- and normal-dispersion and effective core area varying along the taper. It is found that there are drastic phase variations in the output taper transition region. We have also pointed out that the critical distance in the tapered fiber is a useful parameter for estimating the SC spectrum and phase evolutions.

References

Recent Advances in the Fiber Optic Sensors Based on Stimulated Brillouin Scattering _ Invited paper
Xiaoyi Bao†, C. Zhang‡, I. F. Ozkan†, Magdi Mohareb§, W. Li§, F. Ravet†, L. Chen†
†Department of Physics, University of Ottawa, Canada
‡Department of Civil Engineering, University of Ottawa, Canada
*xbao@uottawa.ca

Abstract
The significant developments in distributed Brillouin sensors have made them as the excellent candidates for the structural health monitoring in identifying early cracks, deformation and prevention of the structural failure using carbon coated fibers.

The concept of using Brillouin scattering for distributed fiber optic sensing was first proposed in 1989 and it was termed Brillouin optical time domain analysis (BOTDA). This approach involved launching a short pump pulse into one end of the test fiber and a CW (continuous wave) probe beam into the other end. The pulse length is equivalent to the spatial resolution. The frequency difference between the two lasers could be set to a particular value corresponding to the Brillouin frequency of the optical fibers, and the CW probe would experience gain at locations along the fiber. The gain as a function of position along the fiber could thus be determined by the time dependence of the detected CW light. By measuring the time dependent CW signal over a wide range of frequency differences between the pump and probe, the Brillouin frequency for each fiber location could be determined. This allowed for the strain or temperature distribution along the entire fiber length to be established.

The initial application for the distributed sensor was the cable strain monitoring over tens of kilometers with the spatial resolution of meters. Because of the Brillouin frequency changes with both temperature and strain, the separation of temperature and strain was realized by two fibers in parallel, one is used to monitor temperature which is insensitive to the strain, such as cable fiber as reference, and the second fiber is to sense both temperature and strain, so the temperature information was subtracted from the reference fiber.

Because of the demanding of the civil structural strain monitoring the spatial resolution of centimeters are required due to the size of the structures being hundreds of meters. This was realized by phase locking of two frequency stabilized lasers (bandwidth of KHz) at the Brillouin frequency to a few Hz for the pulse of 1ns equivalent to 10cm for the strain or temperature monitoring of the concrete combined with high temperature and strain resolution due to the narrow Brillouin linewidth at the spatial resolution of 10cm with Brillouin spectrum width less than 50MHz.

To apply this sensor system to the civil structural application, we must solve the problem of spectrum de-convolution under non-uniform strain condition as shown in Fig. 1. There are three methods to deal with this problem: 1) multiple peaking fitting in correlation to the location information to identify the spectral strain event; 2) using spectrum broadening and asymmetric factors plus the Brillouin frequency shift, this method is simple and quick to identify the deformation or cracks in the structures; 3) data reconstruction to solve the couple wave equations.

To bring the distributed sensor to the field, we need to solve the problem of the temperature effect on the measured strains. The simultaneous temperature and strain measurement has been realized by the Brillouin spectrum width, Brillouin frequency shift, and Stokes power in polarization maintained fibers (PMF) with spatial resolution of 20cm as seen in table 1. It was also realized using photonic crystal fibre with two separated Brillouin peaks and different temperature and strain coefficients for the spatial resolution of 15cm, temperature and strain resolution of 1.5°C and 15µε, respectively.

In order to monitor the pipe buckling process using our distributed Brillouin sensor, we measured a pipe of 1,740 mm in length, 508 mm in diameter and 6.35 mm in wall thickness, which is under the combined action of internal pressure, axial tension force and bending moments. The pipes are instrumented with strain gauges and two distributed Brillouin fiber sensors. Fig. 2 represents one pipe buckling location and pipe dimension, and fig. 3 shows the local pipe deformation identification using the Brillouin spectrum broadening factor. The experimental result demonstrates the broadening factor is an effective method to identify early distortion of the structures. Both the magnitude and location of buckles and deformations sequence are successfully predicted using the new schemes. This work was using the carbon-coated fiber which was found to measure significantly higher compressive and tensile strains than polymer coated fibers.

The carbon coated fibre can be stretched to the strain of 3.8% which is much larger than the standard communication fibre with ~1.2% of the strains as shown in Fig. 4 and large temperature range of 300℃, shown in Fig. 5. The carbon coated fibre has proved to
be the ideal candidate for sensing under large strain range under high temperature environments.

Acknowledgement: The author would like to thank the OFS Labs for providing carbon coated fibers. The research is supported by ISIS Canada and NSERC.

Table 1: Uncertainty of Temperature and Strain Calculated with Measured Brillouin Frequency (F), Power (P), and Bandwidth (B) in PMF

<table>
<thead>
<tr>
<th>Property</th>
<th>RMDA</th>
<th>Bow Te</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>F-P</td>
<td>F-B</td>
</tr>
<tr>
<td>Uncertainty of temperature (°C)</td>
<td>8</td>
<td>38</td>
</tr>
<tr>
<td>Uncertainty of strain (µε)</td>
<td>155</td>
<td>135</td>
</tr>
<tr>
<td>Maximum error of 3T(°C)</td>
<td>10</td>
<td>98</td>
</tr>
<tr>
<td>Maximum error of 3ε (µε)</td>
<td>331</td>
<td>249</td>
</tr>
<tr>
<td>Tm (°C)</td>
<td>2</td>
<td>40</td>
</tr>
<tr>
<td>Tm (3ε (µε))</td>
<td>211</td>
<td>189</td>
</tr>
</tbody>
</table>

Fig. 1: The strain section normalized to the spatial resolution versus the bandwidth of the Brillouin spectrum.

Fig. 2: Buckling of Specimen 1 at the end of the test.

Fig. 3: Location of the pipe buckling: a) maximum broadening factor shows local deformation of the pipe; b) the nonlinear broadening factor represents starting of the pipe deformation in three different locations in a).

Fig. 4: Brillouin frequency shift versus tensile strain for carbon coated fibre.

Fig. 5: Temperature dependence of the change of Brillouin frequency shift of the carbon coated fiber.

References:
Study on Low Cost Double-fiber Model Sensor for Distributed Strain Monitoring

Chuantong Wang and Katsunori Shida
Advanced Systems Control Engineering Department, Graduate School of Saga University 1 Honjo-machi, Saga 840-8502, Japan
email: 04643001@edu.cc.saga-u.ac.jp, shida@cc.saga-u.ac.jp

Abstract—This paper reports a double-fiber model sensor (DFMS) for distributed strain monitoring. By using coupling light power between two multi-mode fibers, the sensor needs no sophisticated instruments and consequently has an extremely low cost.

I. INTRODUCTION

Distributed optical fiber sensors are important and usable in many fields. However, the fundamental limitation of conventional OTDR requires weak signal detection with great bandwidth. Consequently, the entire sensor systems are sophisticated and costly in many cases. FBG sensor provides a relatively low cost solution for quasi-distributed sensing which is not able to sense the places without bragg gratings. Double-fiber model distributed optical fiber sensor is new solution which is different from the current techniques [1, 2].

II. SENSOR SYSTEM

Double-fiber model sensor study is motivated by the low cost applications such as exceptional strain monitoring in civil structure. The scheme of double-fiber model distributed optical fiber sensor system is illustrated in Fig. 1. The sensor is composed of two multi-mode fibers which are defined as the active fiber and the passive fiber respectively. The active fiber has active light propagating within it while the passive fiber has no light in common state and is always ready to receive coupling light from the active fiber. Localized strain couples light from the active fiber to passive fiber. The output of the active fiber (V1) is used for strain value sensing while the double ended outputs of the passive fiber (V2, V3) are used for strain position sensing. The distributed sensing is based on the attenuation of the passive fiber.

The coupling light between two multi-mode fibers under effectiveness of strain can be detected by photodiodes with simple preamp circuit. Low-pass filters are used to promoting signal noise ratio. LED is used for light source and there are no special requirements for performance bandwidth. The data treatment after A/D converting is quite simple without complex calculation and analysis. Therefore, the sensor needs no sophisticated instruments. It can be realized with extremely low cost which is meaningful to wider applications of distributed optical fiber sensors. The signal detection circuits and light coupling device used in our study are shown in Fig. 2.

III. DISTRIBUTED SENSING PRINCIPLE

The distributed sensing is based on the attenuation of the passive fiber which has a greater attenuation coefficient than the active fiber. As illustrated in Fig.3, the localized strain makes light lost from the active fiber and couple to the passive fiber. The coupled light propagates in the passive fiber in both directions of forward and backward. The output of the active fiber (Pf) has dependency on the strain value applied on it. The outputs of the passive fiber have dependency on both of the strain value and strain position. By comparing the values of P and PB, the strain position can be determined with the restriction of detectable conditions.
The relationship of strain position and outputs of the passive fiber is shown in the equation (1) which is restricted by the detectable conditions of the sensor.

\[ x = \frac{1}{2} \left[ L + \frac{10}{\beta} \log_{10} \left( \frac{P_f}{P_b} \cdot \frac{k_b}{k_f} \right) \right] \]  

where, \( P_f \) is the emitting light intensity at forward end of passive fiber, \( P_b \) is the emitting light intensity at backward end of passive fiber, \( x \) is the position of localized strain, \( k_f \) is the fraction of light propagating forward in the passive fiber, \( k_b \) is the fraction of light propagating backward in passive fiber, \( \beta \) is the attenuation coefficient of passive fiber, \( L \) is the length of fibers.

IV. DYNAMIC RANGE OF SENSOR

Since the sensor use coupling light and attenuation of passive fiber for distributed sensing, there is no bandwidth requirement for both of light source and light power detection. The signals can be detected by photodiodes with simple preamp circuits and filters. At the bandwidth of about 1 Hz, the noise equivalent power (NEP) can easily reach the level of -100 dBm which make the sensor have great dynamic range and high signal to noise ratio (SNR). The optical power of backward coupling light in the passive fiber is shown in Fig. 4 with the provided parameters value, where \( P_b \) is light source power, \( \alpha \) and \( \beta \) are the attenuation coefficients of the active fiber and the passive fiber respectively, \( k_{eb} \) is the total effective backward coupling coefficient.

Since \( k_{eb} \) is practically in the range of \( 10^{-6} \) to \( 10^{-2} \), the sensor can achieve a quite high SNR in the level of several kilometers and have a dynamic range of about 60 dB.

V. EXPERIMENTS AND DISCUSSIONS

An experimental prototype sensor is developed to confirm the proposed method and theoretical analysis result. The comparison of experimental result and the theoretical curve predicted by equation (1) are shown in Fig. 5. The prototype sensor is designed at an experimental length of 4 m and surely can be far promoted.

The performance parameters of the proposed sensor are compared with other distributed optical fiber sensing techniques in Table 1. The proposed sensor shows the competitive advantage by the low cost character although it provides relative inferior performance at some aspects.

VI. CONCLUSION

The double-fibers model sensor provides a low cost solution to distributed strain monitoring applications where the first exceptional strain is concerned. By using the coupling light between two multi-mode fibers, the sensor realizes the distributed sensing without any sophisticated instruments. The total systems can be realized with extremely low cost. It is meaningful to the wider applications of optical fiber sensors where the cost-effectiveness is critical limited.

REFERENCES


Measurement of the Brillouin Gain/Phase Characteristics in Optical Fibers Using a Double-Modulation Technique of a Single Light Source

Kenichiro Tsuji, Jungmin Kim, Teppei Yamaguchi, Noriaki Onodera, and Masatoshi Saruwatari.
Department of Communications Engineering, National Defense Academy
1-10-20 Hashirimizu, Yokosuka 239-8686, Japan.
Phone: +81-46-841-3810(ext.3383), Fax: +81-46-844-5911, E-mail: kentsuji@nda.ac.jp

Abstract
We propose a simple measurement method of the Brillouin gain/phase characteristics in optical fibers based on the pump and probe technique using a doubly-modulated single light source. Our method is based on the phenomena that an intensity modulated optical waveform having the same order of the Brillouin bandwidth is distorted by the Brillouin phase and gain spectra. The availability of the method is studied analytically and experimentally.

Introduction
Stimulated Brillouin scattering (SBS) in optical fibers is one of dominant nonlinearities caused by an interaction between optical field and acoustic wave. While the SBS limits the optical power that can be transmitted in fiber-optic systems, it also can provide optical amplification with high efficiency [1]. Although the SBS has a high optical gain with a relatively low pump power, it is not suitable for signal amplification in optical communication system because of its very narrow gain bandwidth of several tens of MHz.

Recently, the large phase change accompanying with the narrow and high gain spectra of the SBS has been remarked from a viewpoint of all-optical signal processing. This large phase change causes the higher group index in optical fibers, and then it let incoming optical pulses transmit more slowly than its original speed [2]. This phenomenon is called “slow light” and is considered to apply for a light-controlled optical delay line or an optical buffer that may be indispensable for future all-optical packet routers. Accordingly, the precise measurements not only for the Brillouin gain spectra (BGS) but also for its phase spectra (BPS) are very important so as to fully understand the slow light.

In this paper, we present a simple method to estimate the Brillouin phase characteristics as well as the SBS gain profile in optical fibers. Our method utilizes the phenomena that the waveform of the intensity-modulated (IM) probe light at the modulation frequency comparable to the Brillouin bandwidth is distorted when it is amplified by the Brillouin gain with the large phase characteristics. The availability of the proposed method is analytically and experimentally investigated.

Principle of the Brillouin gain/phase measurement
Figure 1 shows the measurement setup of the Brillouin gain/phase measurement. Figure 2 shows the principle of the Brillouin gain/phase measurement. The Brillouin gain spectrum in optical fibers is assumed to have an ideal Lorenzian shape function and its phase spectrum is described by its derivative as shown in the Fig. 2. When the modulation frequency of the probe light $f_{IM}$ is the same order of the Brillouin bandwidth, each spectrum component of the IM probe light experiences different gain and large phase shift as well, according to the probe frequency location relative to the BGS. This situation changes the intensity and phase relationship between each IM spectrum.
component and gives rise to modification (distortion) of the output probe IM waveform. Since this waveform modification strongly reflects the phase characteristics of the SBS, we can estimate the phase characteristics by analyzing the output probe waveform. The gain characteristics can also be estimated in the same manner when the lower frequency IM modulation is applied.

To simply evaluate the waveform modification, we utilize the $f_{IM}$ component of the output probe waveform because it can be easily measured with an RF spectrum analyzer after O-E conversion. We have analyzed and measured the $f_{IM}$ component of the output probe waveform depending on the frequency difference between pump and probe by scanning $f_{PM}$, taking into account the Brillouin gain/phase characteristics.

### Numerical simulation results

Figure 3 (a) shows the assumed BGS and BPS used in the simulation that is calculated from the Lorentzian function. The gain bandwidth $\Delta G$ is set to 20 MHz. The $\beta$ shows the maximum phase shift of the BPS and $\Delta f_p$ is the frequency width between the extremal points of the BPS. Fig. 3 (b) and (c) show simulated results of the $f_{IM}$ component power after the gain and phase of IM probe light are changed by SBS, assuming $f_{IM}$=1 MHz and 30 MHz, respectively. The horizontal axis shows the relative probe frequency location to the Storks shift. Each simulated curve shows the case when the $\beta$ is assumed to 0, 0.8, 1.0, 1.2, 1.4 rad, respectively. The BGS can be obtained from the $f_{IM}$ component power profile of $f_{IM}$=1 MHz regardless of the maximum phase shift $\beta$. When the $f_{IM}$= 30 MHz, we can see the extremal points of the BPS causes the dip in the $f_{IM}$ component power profile and its depth has the correlation with the magnitude of the $\beta$. Therefore, if we measured the frequency location and depth of the dip, we can obtain the values of $\beta$ and $\Delta f_p$.

### Experimental results

Figure 4 shows the measured $f_{IM}$ component power as a function of $f_{PM}$ that corresponds to the frequency difference between the pump and probe lights, when we measured a conventional single mode fiber with the length of 1.3 km. The pump power is 40 mW and the $f_{IM}$ is (a) 1 MHz and (b) 30 MHz. The similar $f_{IM}$ spectrum distribution is observed as compared with the simulation results. From the results of $f_{IM}$=1 MHz, the $\Delta f_p$=12 MHz is obtained. From the results of $f_{IM}$=30 MHz, we can clearly see two dips of the $f_{IM}$ component power in both sides of its peak at 10.802 GHz, which corresponds to the Storks shift. The measured $\Delta f_p$ from these dips is about 9 MHz. From the comparison to the simulation results, the maximum phase shift $\beta$ of the sample SMF is estimated to 1.4 rad. In the measured result, however, the asymmetric characteristic is observed. This might be due to the slightly asymmetric BGS caused by the inhomogeneity along the fiber length.

### Conclusion

We have presented a new measurement method of the SBS gain and phase characteristics in optical fibers using the doubly-modulated single light source utilizing the SBS-induced waveform distortion of the IM probe light with peculiar frequency. We successfully demonstrated that the frequency width $\Delta f_p$ of the Brillouin phase spectra can be measured from the dips of the $f_{IM}$ component power profile and the maximum phase shift $\beta$ can be estimated by comparing with the simulation.

### References


---

**Fig.3** (a) Assumed Brillouin gain/phase spectra. (b)(c) Simulated $f_{IM}$ spectrum power profile.

**Fig.4** Measured $f_{IM}$ component power as a function of the relative probe frequency $f_{PM}$.
Measurement of Raman gain efficiency for an optical fiber cable installed in the field by using OTDR

Kyoichi Oro¹, Hiroki Hatada², Ikuo Yamashita³, Tetsuro Yabu⁴ and Masaharu Ohashi⁵
¹Kansai Electric Power, R&D Center, 3-11-20 Nakoji, Amagasaki, Hyogo, 661-0974 Japan
²Osaka Prefecture University, 1-1 Gakuen-cho, Sakai, Osaka, 599-8531 Japan
Tel: +81-6-6494-9704, Fax: +81-6-6498-7662, E-mail: oro@rdd.kepco.co.jp

Abstract:
The measuring technique of Raman gain efficiency along the optical fiber based on the OTDR is applied to the optical test fiber cable installed in the field. Raman gain efficiency of the test cable in the field can be successfully estimated.

1 Introduction
In order to realize the large capacity optical transmission systems, Raman amplification is one of the key technologies. By development of distributed Raman amplifiers by using the transmission line, it is important to estimate the Raman characteristics for designing a gain of the transmission line. Then, Raman gain coefficient measurement by launching pump light into the fiber has been demonstrated [1] and measurement techniques by using OTDR [2],[3] have been proposed. For further precise design of Raman amplifiers, it is necessary to estimate Raman gain coefficient distribution along the fiber length. For the purpose, a simple measurement technique using an OTDR has been proposed [4]. It is attractive scheme because it also seems to be useful for examining a degraded fiber from a transmission line composed of many fibers. However, our measuring method has not been investigated for the installed optical fiber cable in the field.

In this paper, Raman gain efficiency distribution of an optical fiber cable installed in the field is measured by using the present OTDR technique.

2 Theoretical background
Figure 1 shows the schematic setup for measuring the Raman gain characteristics in the optical fibers by using OTDR.

The pulsed signal and continuous wave pump light are launched into the test fiber and they co-propagated in the test optical fiber. The signal power $P_s$ and pump power $P_p$ in the fiber are described by the following coupled power equation if the pump power is un-depleted.

$$\frac{dP_s}{dz} = \frac{g\text{eff}}{A_{\text{eff}}} P_s P_p - \alpha_s P_s$$

(1)

$$\frac{dP_p}{dz} = -\alpha_p P_p$$

(2)

where $g_{\text{eff}}$ denotes the Raman gain coefficient. $\alpha_s$ and $\alpha_p$ are the attenuation coefficients of respective signal and pump wavelengths. $A_{\text{eff}}$ denotes the effective area, which corresponds to the overlapping area between pump and signal powers.

From the equation, the signal power at the position of $z$ can be obtained as

$$P_s = P_s(0) \exp \left[ \frac{g\text{eff} P_s(0)}{A_{\text{eff}}} L_{\text{eff}}(z) - \alpha_s z \right]$$

(3)

where $L_{\text{eff}}(z)$ denotes the effective fiber length.

The signal power $P_s(z)$ at the position of $z$ is reflected and it travels to the input direction. Backscattered light can be expressed as the product of signal power $P_s(z)$ and $B(z)\alpha$. $\alpha$ is the scattering coefficient and $B(z)$ is the capture ratio.

Then, the backscattered signal light $P_s(z)B(z)\alpha$ is amplified by the counter propagating pump light. Therefore, the backscattered signal power from the position of $z$ can be expressed as

$$P_s(z,P_p) = P_s(0)\alpha B(z) \exp \left[ \frac{2g\text{eff} P_s(0)}{A_{\text{eff}}} L_{\text{eff}}(z) - 2\alpha_s z \right]$$

(5)

On the contrary, when the pump light is off, the backscattered light $P_s(z,0)$ can be obtained by substituting $P_p(0)=0$ into (5) as

$$P_s(z,0) = P_s(0)\alpha B(z) \exp \left[ - 2\alpha_s z \right]$$

(6)

The backscattered signal power of OTDR, $S(z,P_p)$ [$=10\log \{P_s(z,P_p)\}$] can be expressed as

Figure 1 Schematic setup of the proposed Raman gain coefficient measurement method.
\[ S(z, P_p) = 10 \log[P(z, P_p)] \]
\[ = 10 \log[P_r(0)] + 10 \log[\alpha B(z)] + \frac{2g \alpha P_r(0)}{A_{eff}} L_{eff} (z) - 10 \log(e) - 2\alpha z 10 \log(e) \]  
\[ (7) \]

Therefore, Raman gain efficiency can be derived from the backscattered signal powers with two different kinds of pump power as
\[
g_R = \frac{1}{2P_r(0)L_{eff} (z) - 10 \log(e)} [S(z, P_p) - S(z,0)] \]  
\[ (8) \]

The Raman gain coefficient \(g_R\) can be estimated by multiplying the Raman gain efficiency \(g_R/A_{eff}\) by the effective area \(A_{eff}\). Raman gain distribution along the fiber can be estimated in the same manner.

3. Experimental results

We measured the Raman gain efficiency for an optical test fiber composed of four 3-km long single-mode fibers (SMFs) as shown in Fig.1. A wavelength of the OTDR (Anritsu MW9076) was 1550nm. The pulse width was 1\(\mu\)s and averaging time was 2 minutes. Fabry-Perot type laser with a center wavelength of 1462nm was used as a pump light and output power was 90mW. FWHM of the optical filter was 2.4nm. Optical loss of the fiber at the pump wavelength is assumed to be same as the loss of 1.45\(\mu\)m measured by the OTDR.

Figure 2 shows the experimental results. From the OTDR waveforms with and without pump light, it is found the backscattered signal is amplified by the pump light. Raman gain property along the test fiber is also shown in the middle of Fig.2. It is seen that almost uniform gain is obtained in each section. Estimated Raman gain efficiency as shown in Fig.2 is appropriated value for the SMFs and gain difference of each fiber are well estimated.

Figure 3 shows the schematic structure and experimental results for an installed optical fiber in the field. An optical fiber cable including 200 fibers is set on poles and forms 550m loop. 5.5-km long test fiber was made by splicing these fibers as shown in the figure. Setting the equipment as the same condition with Fig.2 and adding 5-km long dummy fiber to the test fiber, measurement was done. Experimental OTDR waveforms and Raman gain efficiency are shown in Fig.3. It is found that the Raman gain efficiencies of the fibers can be estimated successfully by using the present technique.

OTDR based equipment is easily acceptable to real telecom facilities and the measurement time and accuracy seem to be allowable for a practical use.

4. Conclusion

The measuring technique of Raman gain efficiency along the optical fiber based on the OTDR was proposed. Experiments were performed for the 12-km long optical fiber composed of four 3-km long SMFs and a 5.5-km long fiber where ten fibers in an installed outside fiber cable with a length of 550m were spliced. Raman gain efficiency distribution for the test fiber in the field was successfully estimated by the present method.

References
Equivalent Optical Circuit Synthesis using Polarimetric OTDR
For Simulating Two-Pump Optical Fibre Parametric Amplifiers
Takeshi Ozeki, Takuro Kanou, Kazuki Hayashi and Teruhiko Kudou
Department of Electrical Engineering, Faculty of Science and Technology, Sophia University,
7-1, Kioi-cho, Chiyoda-ku, Tokyo, 102-8554, Japan
t-ozeki@gentei.ee.sophia.ac.jp

Abstract
Equivalent optical circuit synthesized by polarimetric OTDR measurement is applied to simulate polarization gain spectrum of two-pump fiber optical parametric amplifier resulting better estimation of polarization gain spectrum but 5dB discrepancy in average gain.

Introduction
The orthogonal two-pump fibre optical parametric amplifier (FOPA) is expected as a useful polarization independent FOPA [1]. However, the polarization gain spectrum of FOPA with PMD=0.015 ps/√km was analysed severely to show the relatively flat gain spectrum with 10dB gain reduction using the PMD operator B · σ · δω [2]. Further, the Raman scattering in the FOPA is analysed to show fairly large polarization dependant gain and gain-tilt [3]. The equivalent optical circuit synthsissed using a polarimetric OTDR is successfully demonstrated to map polarization mode dispersion (PMD) distributed along fiber length [4,5]. However, it needs to be verified by demonstrations in various applications. Here, we demonstrate to regenerate the polarization gain spectrum of a two-pump FOPA by an equivalent optical circuit synthesized by polarimetric OTDR measurement, for the first time. Additionaly, the behaviors of polarization states of constituent waves along the fiber are visualized.

Experiments
Figure 1 shows our experimental setup for an FOPA with orthogonally polarized pump waves. A 620m highly nonlinear fiber (HNF) is used in our FOPA. The zero dispersion wavelength of the fiber is 1548.5 nm and its dispersion slope is 0.048 ps/nm²/km. The measured nonlinear parameter γ is 9.8/W·km. The pump distributed feedback (DFB) laser diodes were modulated by current at 100 kHz (50%) for Stimulated Brillouin Scattering (SBS) suppression. The pump power of 1531.5nm and 1567nm waves were adjusted to order of 500mW for each after an EDFA. Two-pump FOPA gain spectrum was measured for linearly polarized signal waves rotated by 0°, 30°, 60° and 90° at the input end, which are depicted by color coded lines of red, black, blue and magenda, respectively, in Fig.2. The gain variation depending on the input polarization states is about 3.5dB at shorter wavelength region. Two spikes in Fig.2 correspond to attenuated two pumps.

Numerical Simulation
The numerical simulation based on measured parameters without PMD considering six optical vector waves in the non-linear coupled equations of the two-pump FOPA, is depicted in Fig.2 comparing the measured polarization gain spectrum. The gain and bandwidth simulated almost fit well with the measured values. The polarization dependences are reduced to negligible level by considering six vector waves [3].

Simulation of FOPA using Equivalent optical Circuit determined by Polarimetric OTDR
Here, the transfer matrix $j_m$ of $m^{th}$ fibre segment with 0.5m length determined by the polarimetric OTDR measurement [4] is used for simulation based on coupled six vector wave non-linear equations. The phase shifts $\phi_m^k, \phi_m^o, \psi_m^k$ of $j_m$ at the optical frequency of $k$ for the signal, idler, and two pump waves are described in the following way for the 1st order PMD, for an example, $\phi_m^k = \phi_m^o \cdot \omega_k / \omega_o$, where $\phi_m^o$ is the phase shift determined by the polarimetric OTDR measurement, $\omega_o$ and $\omega_k$ are the optical frequencies of the OTDR using $\lambda_0 = 1550nm$ and $k$ wave, respectively. $\theta_m$ is the rotation angle of the fibre segment. Fig.3 shows the evolutions of the x-component of the pump, and the signal wave intensity of both polarizations and y-component, along the HNF. The PMD of the NHF is 0.064ps with almost linear accumulation proportional to fibre length as shown in Fig.3. It is clear that the signal wave polarization varies almost corresponding to the polarization state variations of the pump wave. The polarization gain spectrums simulated using the Jones matrix synthesized by the polarimetric OTDR measurement are shown in Fig.4. The average gain reduction by introducing PMD is about 10dB, which agrees with the discussion of ref. [2]. The polarization dependence of the gain spectrum simulated is about 0.8dB, which is smaller than the 3.5dB PDG measured.

**Discussion**

The polarization gain spectrum is simulated by the six pairs of nonlinear vector coupled equations using all measured parameters including the fiber Jones matrix, for the first time. It is said that the quantitative fitting of the polarization gain spectrum is not sufficient in details yet.

It suggests the simultaneous consideration of Raman scattering [3,6,7,8], zero dispersion wavelength mapping [9] and PMD mapping for better fitting accuracy. The gain tilts observed could be attributed also to the Raman scattering contribution in FWM process [3].

**Conclusions**

Equivalent optical circuit synthesized using polarimetric OTDR measurement is applied to simulate polarization gain spectrum of two-pump optical fiber parametric amplifier, resulting better estimation in polarization gain variation but 5dB discrepancy in average gain.

**References**

Recent Advances in the Design and Fabrication of High Channel-Count Fiber Bragg Gratings

Ming Li and *Hongpu Li
*Department of Electrical and Electronic Engineering, Shizuoka University, Johoku 3-5-1, Hamamatsu, 432-8561, Japan
Tel & Fax: 81-53-4781617
dhli@ipc.shizuoka.ac.jp

Yunlong Sheng
Optics, Center for Optics, Photonics, and Laser, Laval University, Quebec G1K7P4, Canada

Abstract
We review our recent developments for high channel-count fiber Bragg grating (FBG). An 81-channel FBG has been experimentally demonstrated. Moreover, a novel design method for a multi-channel FBG with non-identical inter-channel properties has been demonstrated.

1. Introduction
Recently high channel-count fiber Bragg grating (FBG), as one of the fiber-based promising solutions to the broad-band chromatic dispersion compensation, has attracted great interests. However, with increasing the number of wavelength-division-multiplex (WDM) channels to cover the full C-band, high channel-count FBG devices become extremely difficult to realize. To date, various methods have been used trying to make FBG based broad-band devices realizable. Among them, the phase-only sampling seems to be the most prospective one, because it can deduce the index change required for an 81-channel sampled FBGs to a practical level and makes the multi-channel gratings particularly suitable to be fabricated with the robust side-writing phase-mask technique. In this paper, we review our recent developments in the design and fabrication techniques for high channel-count fiber Bragg gratings (FBG). First, a continuous phase-only sampling function has been originally proposed. An 81-channel (spacing of 50 GHz) phase-only sampled linearly chirped FBG is designed and demonstrated. Second, based on a discrete layer peeling (DLP) method, a 9-channel linearly chirped FBG with non-identical channel-channel dispersion response has been experimentally demonstrated. Unlike the general multi-channel fiber Bragg gratings designed with either a sampling or the Talbot method, these gratings have nearly ideal flat-top spectrum in both the transmission and reflection.

2. High channel-count fiber Bragg gratings based on phase-only sampling functions
In a sampled FBG, the induced refractive index modulation \( \Delta n \) can be expressed as
\[
\Delta n(x) = \text{Re} \left( \frac{\Delta n_i(x)}{2} \cdot \exp \left[ i \phi(x) \right] \right) \cdot s(x),
\]
where \( \Delta n_i(x) \) is the maximum index modulation, \( x \) is the position along the grating, \( \Lambda \) is the central pitch of the grating, \( \phi(x) \) is the local phase for one channel grating which determines the dispersion of the grating. \( s(x) \) stands for a sampling function with period \( P \). We use a continuous phase-only sampling function with an analytical form:
\[
s(x) = \exp \left[ i \left( \sum \alpha_n \cos \left( 2 \pi x / P + \beta_n \right) \right) \right],
\]
where the number of terms in the series \( M \) is minimized, and two free parameters for each term \( \alpha_n \) and \( \beta_n \) are optimized such that the channel spectrum (Fourier spectrum of the sampling function) is flat over the band of interest. In order to obtain a series of \( \alpha_n \) and \( \beta_n \), we use the simulated annealing algorithm. Optimization criteria would be the uniformity of the desired channels and the so-called in-band diffraction efficiency. The cost function is defined as
\[
E(x) = \sum_{m=1}^{N} \left| S_m \left( \alpha, \alpha', \ldots, \alpha_n, \beta, \beta', \ldots, \beta_n \right) \right|^2 - \eta \left( 2N + 1 \right),
\]
where \( S_m \) is the \( m \)-th Fourier coefficient of the base sampling function \( s(x) \). \( \eta \) is the target diffraction efficiency for all the in-band channel of \( 2N + 1 \). The optimal results for an 81-channels is shown in Fig. 1. For convenience, the period of the sampling function is normalized to one. In order to obtain the high diffraction efficiency \( \eta \), 41 harmonic terms were selected. The diffraction efficiencies obtained with the simulated annealing algorithm is 84%, respectively. Non-uniformity of the channel intensities are less than 1.0%.

Fig. 1. 81-channel continuous phase-only sampling function and its FFT spectrum. (a) Phase distribution, and (b) channel spectrum.

To verify the accomplished designs, the sampling function is multiplied by a single-channel FBG. The reflection spectrum of the sampled FBG was computed with the transfer matrix method. The results for a linear chirped 81-channel with spacing 50 GHz, chromatic dispersion -1360 ps/nm are shown in Fig. 2. The reflection
spectrum show high intra- and inter-channel uniformity within nearly whole C-band. Moreover, to confirm the numerical results, several 81-channel FBGs were fabricated with the side-writing phase-mask technique. Figure 3 shows the one of the typical measured results. It can be seen that nearly identical 81 channel with a channel spacing of 50 GHz, useable bandwidth of 0.11 nm, and chromatic dispersion of about –1400 ps/nm have been obtained.

3. Advanced design of multi-channel fiber Bragg grating based on a layer-peeling method

Layer-peeling algorithm was first proposed by Feced and Skaar in synthesis of one-channel FBG, which could be used to efficiently and accurately reconstruct grating structure from any a given reflective spectrum. Based on DLP method, first, one needs a reflective spectrum used as an ideal target. Now we apply the peeling method to a multi-channel fiber Bragg grating (FBG) used as a simultaneously chromatic dispersion and dispersion slope compensator. Contrary to the design based on sampling function where each channel remains the same properties), each channel would incur a slightly different dispersion, and the target for 2N+1 channel grating are expressed as,

\[ r(\lambda) = \sum_{n} r_{n}(\lambda) = \left[ f(\lambda) \otimes \sum_{m} \delta(\lambda - m \Delta \lambda, \Phi(\lambda)) \right] \exp(-i\Phi(\lambda)) \]  

where \( f(\lambda) \) is the amplitude spectrum for the central channel, \( \Delta \lambda \) is the channel spacing, \( \Phi(\lambda) \) is the accumulated phase of the total channels, which can be expressed as

\[ \Phi(\lambda) = -c \sum_{n=-N}^{N} \int_{\lambda - N \Delta \lambda / 2}^{\lambda + N \Delta \lambda / 2} \frac{2\pi}{\lambda} D_{2}^{m}(\lambda - \lambda_{m}) d\lambda, \]  

where \( \lambda_{m} \) is the central wavelength for mth channel, and \( D_{2}^{m} \) is the chromatic dispersion for mth channel, defined by

\[ D_{2}^{m}(\lambda) = (D_{c} + D_{s} m \Delta \lambda_{m}), \quad m = -N, -(N-1), \ldots, N-1, N. \] 

where \( D_{c} \) and \( D_{s} \) is the chromatic dispersion and dispersion slope at the central wavelength. For convenient, we chose \( D_{c} \) and \( D_{s} \) as \(-1020 \) ps/nm, and 4.2 ps/nm2 respectively. The corresponding reconstructed grating amplitude and phase profile, the calculated reflection, transmission, and the dispersion spectrum are shown in Fig. 4(a), (b), (c), and (d), respectively. It can be seen that a set of 9 channels with almost identical transmission and the wavelength spacing of 0.8 nm are generated. The dispersions are about \(-1020 \) mm/ps in central channel and ranges from \(-1006 \) ps/nm to \(-1034 \) ps/nm through the whole 9 channels. It is also found that the maximum index modulation required for a FBG with 9 channels is approximately 4 times (rather than 3 times) higher than that required for the single-channel grating.

4. Conclusion

In this paper, we propose the phase-only sampling which results in the maximum diffraction efficiency. We experimentally demonstrated an 81-channel FBG for nearly whole C band dispersion compensation which is fabricated with a novel diffraction pre-compensated phase mask. The grating specifications obtained agree well with the theoretical design. Moreover, we have presented a novel method for multi-channel FBG design, which enables us to design any kind of multi-channel FBGs where the spectral response of each channel could be either identical or non-identical. This design method can be easily extended to design any kinds of multi-channel FBGs covering the whole C- or L- band in DWDM system.

References

Broadband Long Period Grating on Hollow Optical Fiber with Femtosecond Laser Pulses

W. Ha\(^1\), Y. Jung\(^2\), J. Kim\(^3\), W. Shin\(^3\), I. Sohn\(^3\), D. Ko\(^3\), J. Lee\(^3\), and K. Oh\(^1\)

\(^1\) Institute of Physics and Applied Physics, Yonsei University 134 Shinchon-dong, Seodaemun-gu, Seoul 120-749, Republic of Korea, Tel: 82-2-2123-5608, Fax: 82-2-392-1592, E-mail: koh@yonsei.ac.kr

\(^2\) Department of Information and Communications, GIST 1 Oryong-dong, Buk-gu, Gwangju, 500-712, Republic of Korea

\(^3\) Advanced Photonics Research Institute, GIST 1 Oryong-dong, Buk-gu, Gwangju 500-712, Republic of Korea

Abstract

We have fabricated a broadband long period grating on hollow optical fiber by using femtosecond laser pulses. The filter showed a low insertion loss of 1.7 dB and very broad FWHM of 160 nm.

1 Introduction

In order to accommodate flexible wavelength manipulation in WDM optical communication systems, passive and dynamic channel selective all-fiber filters are in great demand [1]. In recent years, all-fiber band rejection filter with extremely narrow bandwidth less than 0.8 nm and broad bandwidth over 100 nm filters have been investigated with various optical fibers and fabrication schemes.

Recently, authors have reported efficient and broad-band mode conversion in a hollow optical fiber (HOF), where the spectral bandwidth was over 160 nm [1, 2]. Broadband coupling has been attributed to unique mode-beating dispersion in HOF, which is composed of central air-hole, ring core, and cladding. The coupling strength, however, was limited in the previous report due to weak acousto-optic interaction and subsequent index modulation.

In this paper we tried to fabricate long period grating (LPG) on HOF by using femtosecond laser pulses to mechanically inscribe slots in order to obtain even stronger mode coupling and wide bandwidth. We report detailed fabrication process and transmission characteristics were thoroughly investigated.

2 Experiment

![Fig.1. Schematic diagram of the LPG on HOF.](image)

The proposed LPG structure is schematically illustrated in Fig.1. The cladding of HOF is corrugated by femtosecond laser irradiation to the appropriate dimensions. The HOF used in this experiment had dimensions of 6 \(\mu m\) air-hole diameter and an outer diameter of 125 \(\mu m\) [2].

Performance of the Ti:sapphire laser used in this experiment is given follow; wavelength of 785.5 nm, pulse duration of 184 fs, 1 kHz repetition rate, average power of 3.59 mW, and N.A. value of 0.42 with x50 objective lens. A one-meter-long HOF was carved by point-by-point exposure; width of 150 \(\mu m\), depth of 30 \(\mu m\), and period of 550 \(\mu m\) with 20 repeat times.

Because pristine HOF is sensitive to mechanical stress, the corrugated sample was found to be highly sensitive against mechanical bends. Fabricated HOF-LPG pictures are shown in Fig.2 and Fig.3, where corrugation structures are clearly shown. The surface roughness depends on laser power and irradiation time.
For the fabricated HOF-LPGs, we have experimentally measured the output transmission as a function of wavelength with a white light source and optical spectrum analyzer, as shown in Fig.4. Note that the HOF-LPG was placed in a straight fiber holder without any bending while the spectrum was measured, in order to eliminate any bending effect. The filter showed an insertion loss of 1.7 dB and FWHM of 160 nm in the wavelength range of 1630 nm. Coupling strength more than 8 dB was achieved for the band rejection peaks generated in the wavelength range shown in Fig.4. Note that this coupling efficiency is more than 3 dB enhancement compared with prior acousto-optic tunable filter (AOTF) [3].

Generally, LPGs based on conventional single mode fiber (SMF) provide band rejection in a spectral bandwidth about 10 nm. On the while in the HOF, the central air-hole and the ring core result in significantly different mode beating dispersion between the core mode guide along the ring structure and the cladding modes. As experimentally observed in the case of AOTFs [3], the phase matching condition for the LP01_core-LP0_m clad anti-symmetric mode coupling occurs in a broad spectral range, so that HOF-AOTF can provide much broader band of rejection than SMF.

In this experiment of HOF-LPG, we are interested in symmetric coupling, LP01_core-LP0_m clad, to confirm the broadband nature of HOF mode coupling. As we could observe in Fig.4, very wide band rejection was confirmed for LP01_core-LP0_m clad, as well, which shows good agreement with a prior report.

Note that surface roughness and flatness affect the transmission with the spectral noise due to modal interference.

In conclusion, we successfully fabricated a broadband (160 nm) LPG on HOF by using femtosecond laser pulses. The filter showed an insertion loss of 1.7 dB and coupling strength of 8 dB.

3 Acknowledgement
This work was supported in part by the KOSEF (Program Nos. R01-2006-000-11277-0, R15-2004-024-00000-0), and the Science and Technological Cooperation program between and from MOST.

4 References
Ultrasonic Hydrophone Based on Etched Distributed-Bragg-Reflector Fiber Laser

Li-Yang Shao\textsuperscript{1,2}, Sien-Ting Lau\textsuperscript{3}, Xinyong Dong\textsuperscript{1}, Hwa-Yaw Tam\textsuperscript{1}

Helen Lai Wah Chan\textsuperscript{3} and A.Ping Zhang\textsuperscript{2}

1. Photonics Research Centre, Department of Electrical Engineering, The Hong Kong Polytechnic University, Hung Hom, Kowloon, Hong Kong SAR, China
2. Centre for Optical and Electromagnetic Research, Zhejiang University, Hangzhou, Zhejiang, China.
3. Department of Applied Physics, The Hong Kong Polytechnic University, Hung Hom, Kowloon, Hong Kong SAR, China. Email: eelyshao@polyu.edu.hk

Abstract We demonstrate a new type of medical ultrasonic hydrophone based on an etched distributed-Bragg-reflector fiber laser. High-frequency responses of the hydrophone in the range of 12MHz to 50MHz are experimentally characterized.

Introduction

High frequency ultrasonic imaging has been developed for various diagnostic or therapeutic medical applications, e.g. dermatology, ophthalmology, and intravascular imaging [1]. The polyvinylidene fluoride (PVDF) based hydrophone is a popular method for measuring ultrasonic field [2, 3]. However, this type of hydrophone has inherent limitations in characterizing high-frequency transducers due to the effect of spatial averaging. Fiber-optic ultrasonic hydrophones have considerable advantages over conventional PVDF hydrophones, such as immunity to EMI, small size, ease of fabrication and multiplexing capability. Several fiber-optic sensing schemes were reported, including interferometric technique [4], polarimetric technique [5], and fiber grating sensors [6]. In a recent paper, Guan \textit{et al.} proposed an ultrasonic hydrophone based on dual-polarization distributed-Bragg-reflector (DBR) fiber laser [7]. Ultrasonic waves incident on the laser cavity modulate the beat frequency of the two orthogonal polarization modes of the fiber grating laser. The frequency and amplitude of the incident ultrasound are determined by measuring the amplitude and frequency of the sidebands and main peak of the polarization mode beat (PMB) with a photodetector (PD) and a radio-frequency (RF) spectrum analyzer. This technique is able to measure high frequency ultrasound. However, the sensitivity of the sensor is not enough if the frequency of ultrasound is beyond about 20 MHz.

In this letter, we employed a wet-etching technique to reduce the diameter of a dual-polarization DBR fiber laser to improve sensor responses. Its operation principle is similar to the reported DBR fiber laser hydrophone [7]. However, the performance of hydrophone is greatly improved as showed by experimental results.

Experiment and results

A pair of 1551-nm Bragg gratings, spaced at 17.5 mm apart, was written in the Er/Yb codoped fiber to construct the DBR laser cavity. The DBR fiber laser was mounted by a Teflon frame and immersed into an acid bath with 24\% HF concentration at constant temperature of 30°C. The fiber cladding diameters were decreased proportionally to etching time, which etching rate was measured to be around 1 \( \mu \text{m/min} \). Figure 1 shows the configuration of ultrasound sensing experiment with etched DBR fiber laser. With several optical components (WDM, ISO, Polarizer), the PMB signal of the laser output at the PD and its frequency was measured as 939MHz by using an RF spectrum analyzer. Figure 2 shows the measured PMB signal spectra of fiber lasers with different fiber cladding diameters. It is of interest to note that the change of PMB frequency with different cladding diameters is very linear, and a reduction of cladding diameter of 1 micron increases PMB frequency of 1.24MHz. This provides a precise approach of controlling the beat frequency. In the experiment, a focused broadband
transducer with center frequency of 30 MHz and focal length of 14 mm was used to generate the ultrasonic wave. The fiber laser was mounted at the focus of the transducer immersed in a water tank. Figure 3 shows the frequency response of the DBR laser based hydrophone. The inset of figure 3 shows the PMB signal spectrum (fiber cladding diameter is 68 μm), when the ultrasonic transducer was driven by a voltage of 4V at 40 MHz. As expected, sidebands appeared as the DBR fiber laser was subjected to the ultrasound. One can employ the amplitude difference between the main peak and the sidebands to determine the ultrasonic pressure. With a limitation of instrument, we can only see two peaks in the first two sensors. One is relatively broad, whereas the other is very sharp (corresponding to the half-wavelength and one-wavelength resonance of the fiber, respectively). The frequency of the sharp peak is approximately twice of that of the broader one. The peak shifted to high frequency and became flattening when the fiber cladding diameter is reduced. Comparing the two hydrophones with cladding diameter of 125 μm and 68 μm, the sensitivities are enhanced by up to 10 dB ~30 dB in the range of 30 MHz to 42 MHz. For the fiber sensor with 68 μm diameter, a flat response is obtained in the frequency ranged from 12 MHz to 32 MHz.

Conclusions
We have proposed and experimentally demonstrated an etched DBR fiber laser based hydrophone. By using a wet etching technique, the frequency response and sensitivity of the sensor have been improved significantly. Furthermore, with the smaller diameter, it is believed that this proposed hydrophone could provide a good spatial resolution in the beam profile measurement. The wet etching technique offers an effective method to tailor the performance of the DBR fiber laser hydrophone for medical ultrasonic applications.

Acknowledgement
The authors would like to thank the funding support from the Natural Science Foundation of China (Grant No: 60607011), The Hong Kong Polytechnic University-Zhejiang University Joint Research Project G-U224 and the Research Grant Council under Project No. PolyU 5198/03E.

References
All-Fiber Tunable Band Rejection Filter based on Helicoidal Long-Period Fiber Grating Pair


Advanced Photonics Research Institute(APRI), GIST, 1 Oryong-dong, Buk-gu, Gwangju, Korea
Tel: +82-62-970-3343, Fax: +82-62-970-3419, Email: swj6290@gist.ac.kr

*Dept. of Physics, Yonsei University, 134 Shinchon-dong, Seodaemun-gu, Seoul, 120-749, Korea
Tel: +82-2-2123-5608, Fax: +82-2-392-1592, Email: koh@yonsei.ac.kr

Abstract: We propose a new type of all-fiber bandwidth tunable rejection filter using cascaded helicoidal long-period fiber gratings in single-mode optical fiber and report controllable broadband rejection characteristic with low insertion and polarization dependent loss.

A helically structured long-period fiber grating (LPFG) was firstly demonstrated with the wiring technique for a two-mode fiber spatial-mode coupler [1]. Recently, the helical LPFG whose helical index modulation was obtained by releasing the residual stress of a pristine fiber, was fabricated by rotating the fiber under a continuous single side CO2 laser beam exposure and showed relatively small polarization dependent loss compared with that of a conventional LPFG. Also, the novel peak shift was demonstrated when applied co-directional and contra-directional torsions resulted in shorter and longer wavelength shifts, respectively [2]. These novel characteristics of helical LPFGs can be used for a torque sensor and tunable filters. In this paper, a bandwidth tunable all-fiber band rejection filter (BRF) is proposed with serially cascaded helicoidal LPFGs (HLPFG) by using new fabrication technique and characteristics of the proposed device are experimentally investigated.

![Fig. 1. Schematic of the proposed bandwidth tunable all-fiber BRF based on HLPFG pair.](image)

Compared with conventional optical fiber gratings, helically structured optical fiber gratings are characterized by the refractive index modulation of a helical structure such as a screw. HLPFGs are based on the eccentricity between the fiber core and the cladding from helicoidal structure [3,4]. When a standard single-mode optical fiber is twisted, its core follows a helicoidal path inside the cladding which produces a significant periodic index change along the fiber. Therefore, a helicoidal fiber is a kind of LPFG and its optical properties are determined by the grating period. The transmission spectrum of a HLPFG displays several dips that correspond to coupling of light from the core mode to the cladding modes. The spectral property of the proposed HLPFG is similar to conventional LPFGs. The schematic of the proposed bandwidth tunable all-fiber BRF is shown in Fig. 1. The proposed device is composed of two serially formed HLPFGs which have almost same resonance with different direction of helix in clockwise and counter-clockwise, respectively. As the co-directional or contra-directional torsion applied to the helix of each HLPFG can effectively reduce or enlarge the helical pitch, the overlap of rejection bandwidth of the cascaded HLPFGs can be adjusted by applying directional torsions to the helix of HLPFGs. By rotating one end of cascaded HLPFGs in a clockwise direction, the resonances of the clockwise HLPFG and the counter-clockwise HLPFG move to shorter and longer wavelength, respectively. In a counter-clockwise direction, the resonances of HLPFGs move conversely. The bandwidth tunable all-fiber BRF was realized with this novel resonant peak shifts of HLPFGs against the applied co-directional and contra-directional torsions. The helicoidal structures with the clear periodic index change were built up by CO2 laser beam irradiation onto an single-mode optical fiber while the fiber twisted and moved with a constant speed along the optical fiber axis. For precise control of the spectral characteristics and reduction of the transmission loss of HLPFGs, all fabrication parameters such as the rotation speed of fiber, moving speed of base stage, and the power of CO2 laser were optimized. The stabilized CO2 laser beam was focused on the fiber with power of 10 W. While two rotating fiber holders simultaneously rotated in opposite directions at a speed of 1.4 ° /sec, the actuator translated the fiber along its axis with a speed of 10 μm/sec. Ten helical pitches were formed in two section of a conventional single mode fiber (SMF-28, Corning) with opposite direction of helix in period of about 1300 μm.

The measured spectral responses of the fabricated clockwise and counter-clockwise HLPFG depending on the co-directional and contra-directional torsions are shown in Fig. 2(a) and Fig. 2(b). As the rotation angle was varied from -360 to 360 ° with 90 degrees intervals, the transmission spectra of HLPFGs were measured. In Fig. 2(a) and Fig. 2(b), initial resonance peaks of clockwise HLPFG (cHLPFG) and counter-clockwise
The measured wavelength shifts of the resonance peaks decrease or increase monotonically as the rotation angle increases depending on the direction of helix in HLPFGs. The magnified optical image and transmission characteristics of the fabricated bandwidth tunable all-fiber BRF are shown in Fig. 3. In Fig. 3(a), two helically formed periodic structures with different direction of helix are clearly shown. In order to investigate the tunable functionality of rejection bandwidth, the clockwise-directional or counter-clockwise directional torsion were applied to one end of the fabricated HLPFG pair. The measured spectral rejection bandwidth changes of proposed device are shown in Fig. 3(b). The rejection bandwidth of fabricated bandwidth tunable all-fiber BRF was centered near 1585 nm. For the rotation angle variations from -720 to 2160° with 180 degrees intervals, the transmission spectra and the calculated bandwidth tunability of fabricated bandwidth tunable all-fiber BRF were measured and shown in Fig. 3(c). From the measured results, the initial rejection bandwidths at the rejection level of 5, 10 and 20 dB were increased up to 28.9 nm, 20.57 nm and 14.79 nm, respectively. The maximum insertion loss of proposed rejection bandwidth tunable filter was kept under 1.3 dB and showed a dramatically low PDL value less than ~ 1.5 dB. This significantly low PDL is originated from the azimuthally uniform index modulation of the overall helicoidal structure of the fabricated HLPFG.

References
A Cascadable Approach for Widely Tunable Optical Delay of Phase-Modulated Signals

Mable P. Fok and Chester Shu
Department of Electronic Engineering and Center for Advanced Research in Photonics,
The Chinese University of Hong Kong, Shatin, N. T., Hong Kong.
Tel: (852) 2609 8263 Fax: (852) 2603 5558 E-mail: mpfok@ee.cuhk.edu.hk

Abstract
A cascadable fiber-optic delay line is demonstrated with a maximum delay of 800ps. The approach uses four-wave-mixing wavelength conversion in 32-cm bismuth-oxide highly nonlinear fiber and group velocity dispersion in wide-bandwidth chirped fiber Bragg gratings.

1 Introduction
Tunable optical delay is of importance in many areas including optical communications, optical signal processing, and optical control of phased array antennas. Recently, many approaches have been demonstrated to realize an electrically or optically tunable delay line. Examples are the recirculation of light in a waveguide/ring resonator [1], the slowing of light by changing the gain in a nonlinear medium [2], and the shifting of wavelength followed by propagation in a dispersive fiber [3]. Different characteristics can be obtained from the above approaches in terms of the maximum delay, the delay continuity, the signal degradation, and the compactness.

In this paper, we describe a cascadable optical delay line using four-wave mixing (FWM) wavelength conversion in a 32-cm bismuth-oxide nonlinear fiber (Bi-NLF) and group velocity dispersion (GVD) in wide-bandwidth chirped fiber Bragg gratings (CFBGs). The short Bi-NLF has an extremely large nonlinear coefficient of \( \sim 1100 \text{ W}^{-1}\text{km}^{-1} \) at 1550 nm and exhibits a high threshold for stimulated Brillouin scattering [4]. Since four-wave mixing is transparent to the bit rate and modulation format, the approach works well for both amplitude-modulated and phase-modulated signals. With the use of CFBGs, there is no need to incorporate a long dispersive fiber in the setup. Hence, our approach is compact, stable, and has a low latency. The cascadable approach allows the maximum delay to be increased by simply adding more delay stages. We study the performance of the tunable delay is using a 10-Gb/s differential phase-shift keying (DPSK) signal. The effect of cascading the delay stages is also analyzed.

2 Principle and Experimental Setup
Figure 1 shows the experimental setup of our approach.

A 10-Gb/s RZ-DPSK signal is generated by pulse carving of a 1554-nm CW laser followed by phase modulation with a $2^{31}-1$ PRBS. The 10-Gb/s optical signal is combined with a CW tunable laser using a 3-dB coupler. The combined light is then amplified by an erbium-doped fiber amplifier and is subsequently fed to a 32-cm Bi-NLF to introduce FWM. The converted wavelength is filtered out using a 0.4 nm optical bandpass filter. The 3-dB conversion bandwidth is 12 nm and the SBS threshold is over 30 dBm in our experiment. The delay...
stage consists of an optical circulator and a CFBG. The CFBG has a dispersion of $\sim 20$ ps/nm, a reflectivity of 97%, and a 3-dB bandwidth of 30 nm. The converted wavelength falls within the $\sim 30$ nm reflection band of the CFBG. The wavelength can be adjusted by tuning the pump wavelength. When the wavelength is changed, the component will be reflected at a different position along the CFBG. Thus, a variable time delay can be obtained. To increase the tunable delay range, another dispersion stage can be added in the setup. At the output port, a 100-ps delay interferometer is used as a DPSK decoder.

3 Results and Discussion

The 10-Gb/s input signal is fixed at 1554 nm while the pump wavelength is tuned from 1548 to 1558 nm to achieve a variable optical delay. With FWM in the Bi-NLF, the converted wavelength is obtained over the range of 1542 to 1562 nm and lies in the reflection band of the CFBG. The amount of delay depends on the converted wavelength. The change in delay is determined by the product of the wavelength shift and the GVD of the CFBG. With a single delay stage, the measured delay-to-pump wavelength shift ratio is 40 ps/nm while with two delay stages, the ratio is doubled to 80 ps/nm.

To study the performance of the tunable delay on 10-Gb/s DPSK signal, eye diagrams are measured and the results are shown in Fig. 2. Fig. 2(a)-(c) show the decoded input signal and the delayed outputs using one and two delay stages. A widely opened eye is observed in all the three cases.

The 10-Gb/s BER performance has been measured for the input signal, the single-stage output, and the two-stage output at different optical delays. The results are shown in Fig. 3. A power penalty of less than 2.5 dB is obtained over the tuning range. The power penalty difference between single-stage and two-stage delays is less than 0.5 dB, showing practical application of our cascadable approach. In certain application, the wavelength of the delayed output needs to be restored and wavelength reconversion can be performed using FWM again. The cascadability, the enhanced SBS threshold, the reduced latency, the compactness, and the stability of our approach are attractive for practical use of the tunable delay line for different amplitude and phase-modulated data formats.

4 Conclusion

An optically controlled tunable delay scheme has been demonstrated using FWM wavelength conversion in a 32-cm Bi-NLF together with GVD in cascaded CFBGs. The Bi-NLF offers a very large nonlinearity and gives rise to significant FWM over a 32-cm fiber segment. With the use of two CFBGs, a delay range up to 800 ps has been experimentally demonstrated. A 10-Gb/s DPSK signal has been delayed by different amounts and a power penalty of less than 2.5 dB is obtained.

Acknowledgment

The authors would like to thank Dr. Sugimoto and Dr. Ohara of Asahi Glass Co., Ltd. in providing the highly nonlinear bismuth oxide fiber. This work is supported by the Research Grants Council of Hong Kong (Project CUHK 4184/04E).

References

Polarization Dependency Reduction on Long Period Fiber Grating by Side Loading

Tomonori Kondo, Masanori Hanawa, Kazuhiko Nakamura
University of Yamanashi, 4-3-11, Takeda, Kofu, Yamanashi 400-8511, Japan
Tel/Fax: +81-55-220-8683, E-mail: hanawa@yamanashi.ac.jp

Abstract
A novel method to reduce polarization dependency on long period fiber grating (LPFG) by a side loading is proposed. The polarization dependent shift of the mode-coupling wavelength is reduced from 1.5nm to 0.2nm.

1 Introduction
LPFG has been widely used as optical signal processing device like a gain flatting filter for EDFA [1]. In general, a LPFG is fabricated by exposing an UV beam to one side of a photo sensitive fiber. Such one side-exposed LPFG has some amount of birefringence caused by its asymmetrical refractive index distribution [2,3]. The birefringence results in polarization dependencies such as polarization dependent loss or polarization dependent shift of the mode-coupling wavelength. To use LPFGs as optical signal processing device, it is important to reduce the polarization dependencies, especially in applications for high-speed communications.

In this paper, a simple birefringence reduction method, by side loading on LPFGs, is proposed. Both the numerical and experimental results clearly show that the proposed method can drastically reduce the polarization dependencies of LPFGs regardless of its simplicity.

2 Polarization dependency reduction by side loading
A mode-coupling wavelength $\lambda_p$ of an LPFG is given as follows,

$$\lambda_p = \Delta n_{ave} \Lambda,$$

(1)

where $\Delta n_{ave} = n_{co} - n_{cl}$ is the average effective index difference between the core and cladding modes, $\Lambda$ is the grating period. In general, LPFG writing process, an UV beam has been exposed to one side of a photosensitive fiber and it results in the asymmetrical refractive index distribution in the fiber core [2]. Here, it is assumed that the UV beam comes from the $x$ axial direction. The UV-induced index change is the greatest at the core-cladding interface on the $x$-axis and it decreases exponentially toward the opposite side of the fiber core. Therefore, the core index $n_{co}$ for the $x$-polarized light is greater than that for the $y$-polarized light, resulting in the previously mentioned polarization dependencies.

Fig.1 shows refractive index changes induced by side loading for both $x$- and $y$-polarized lights. Because they are symmetrical from side to side and up and down, only quarters of them are shown here. As you can see in Fig.1, when a side loading is applied on the LPFG from the $x$ axial direction, the refractive index change in the cladding for the $x$-polarized light is greater than that for the $y$-polarized light. Namely, the side loading relatively increases the cladding index $n_{cl}$ for the $x$-polarized light, and $\Delta n_{ave}$ for two orthogonally polarized lights approach each other, thus resulting in the reduced polarization dependencies.

3 Numerical analysis
The refractive index distributions in the optical fiber under the presence of the side loading were numerically analyzed by FEM. Fig.2 shows the index distributions for the $x$- and $y$-polarized lights. To obtain $n_{co}$ and $n_{cl}$ for two orthogonally polarized lights, the refractive index distributions on the two orthogonal axes, $x$- and $y$-axis, are averaged geometrically. Those averaged core and
cladding indices are used in the coupled mode analysis of LPGs. By solving the coupled mode equations, the changes in the coupling wavelength between the core and cladding modes by side loading were numerically analyzed.

4 Proof-in-principle experiment
To demonstrate the validity of the proposed method experimentally, a proof-in-principle experiment was performed. Fig.3 shows the experimental setup.

The LPFG used, in the experiments, was made by one-side exposure of the Kr-F excimer laser, and it is loaded by some sash weights as shown in Fig.3. To load from immediately above the LPFG, a piece of the SMF was put in parallel to the LPFG. The LPFG and the SMF were sandwiched by two aluminum plates and the sash weights were put on the upper aluminum plate. The upper aluminum plate was 200g in weight and the sash weight was changed in every 500g from 0g to 2,500g. The polarization state of the linearly polarized ASE was controlled by the polarization controller so that the ASE having different polarization angle could be launched into the DUT. Then, the transmission spectra for the two orthogonally polarized lights were measured using the OSA.

Fig.4 shows the mode-coupling wavelength difference for two orthogonally polarized lights vs. weight on the LPFG. In the figure, the solid line shows the numerical result and the symbols show the experimental results. The tendency of both the numerical and experimental results is almost the same. It has almost 1.5nm difference without any weight. According to the increasing weight, the wavelength difference is reduced. It is almost zero at 1.2kg and increases again for more weight.

The insets show the spectra, (a) without any loading and (b) 1.2kg loading, for every 5 degree of the polarization angle. In the inset (b), there is still some amount (0.2nm) of the polarization dependency regardless of the zero wavelength difference for two orthogonally polarized lights. By the 1.2kg loading, \( \Delta n_{\text{ave}} \) becomes the same only for the two orthogonally polarized lights. For other polarization angles, there is still small amount of \( \Delta n_{\text{ave}} \) difference, thus resulting in the polarization dependency. It is clearly found that the amount of the polarization dependency, however, is drastically reduced compared with the inset (a).

5 Conclusion
The polarization dependency reduction method on LPFG by a side loading was proposed. Although the tested LPFG had a 1.5nm of the mode-coupling wavelength difference in the load free condition, it was reduced to 0.2nm by the side loading. Applications of this technique for gain flattening filters are further works.

References
Femtosecond Pulse Generation by Nonlinear Polarization Rotation in a Bismuth-Based Er-Doped Fiber Laser

Noriyasu Tarumi, Yong-Won Song, and Shinji Yamashita
Dept. of Electronic Engineering, University of Tokyo, Tokyo 113-8656, Japan
Tel: +81-3-5841-6783, Fax: +81-3-5841-6025, Email: tarumi@sagnac.t.u-tokyo.ac.jp

Abstract
We realize a femtosecond all-fiber passively mode-locked laser using polarization rotation in a single nonlinear/gain medium, bismuth-based Er-doped fiber that features broad gain bandwidth along with high nonlinearity. Analysis on the pulsed output is presented.

1. Introduction
Many schemes to realize mode-locked lasers have been developed with their own advantages to meet the specific needs of the state-of-the-art optical technologies. Among them, passive mode locking by nonlinear polarization rotation in a nonlinear intracavity component has been spotlighted with its attractions including the short pulse generation as well as the simplified laser cavity structure [1]. So far, many attempts for achieving better pulse quality hire highly nonlinear fiber (HNLF). However, they need a separated active gain medium in addition to the nonlinear medium, so that the realized structure of the laser cavity is voluminous and costly. Recently, even if a mode-locked laser based on phosphate-based fiber is demonstrated [2], it also has some drawbacks including the limited gain bandwidth. For an improved pulse quality, it is highly expected to have a novel intracavity component that has both the high nonlinearity and the broader gain bandwidth compared with the conventional gain medium. In this work, we employ a Bi-based Er-doped fiber (Bi-EDF) as combined gain and nonlinear medium for the passive pulse formation by the nonlinear polarization rotation. Resultant full width at half maximum (FWHM) of the pulsed output spectrum is 15.6 nm, and the repetition rate is 5.21 MHz. The average output power is 8.57 dBm.

2. Experiment
Fig. 1 shows the configuration of the passively mode-locked laser operated by the nonlinear polarization rotation in Bi-EDF for sub-picosecond pulsed output. The Bi-EDF was pumped by 1480 nm laser diodes that provide an optimum pumping condition for the Bi-based amplifiers [3]. The Er concentration and the nonlinear coefficient of the Bi-EDF are 6,500 ppm and 501/W/km, respectively. The length of the Bi-EDF is 75 cm. The Bi-EDF is fusion spliced to the silica fibers that have high numerical aperture (NA) in order for the mode field matching. The NA is 0.2. WDM-isolators guarantee the uni-directional operation of the laser as well as the pump coupling. Two polarization controllers (PCs) are employed to control the states of polarization (SOPs) with respect to the polarizer and the active gain medium. The polarizer used has the extinction ratio of 30 dB. For the intra-cavity dispersion management, 18m of single mode fiber is inserted. The laser output is given by the 90/10 coupler and the isolator that cuts out the deleterious reflection back from the end facet of the output port.

3. Results and Discussion
As can be seen in Fig. 2, the Bi-EDF provides a wide gain bandwidth ranging from 1520 to 1580 nm, thereby ensures the shorter pulse realization as well as the wide
tuning range of the center wavelength. With the optimized SOPs, the pulses start to be formed with the pump current of 300 mA.

![Gain spectrum of Bi-EDF](image)

Fig. 2. Gain spectrum of Bi-EDF.

![Output spectrum of our pulsed laser](image)

Fig. 3. Output spectrum of our pulsed laser with FWHM of 15.6 nm.

Fig. 3 illustrates the optical spectrum of our pulsed laser. FWHM is estimated to 15.6 nm. Assuming the sech² transform-limit pulse shape, the temporal pulse width calculated is 165 fs. However, the second-harmonic generation (SHG) autocorrelation trace measurement shows the actual pulse width of 260 fs meaning that our pulsed output is highly chirped (see Fig. 4) [4]. The time-bandwidth product of the laser output is 0.494. The center wavelength of our pulsed output is 1567 nm.

![Autocorrelation trace](image)

Fig. 4. Autocorrelation trace of our pulsed laser with inferred pulse width of 260 fs.

In conclusion, we realized an enhanced femtosecond mode-locked laser passively operated by nonlinear polarization rotation in a high nonlinear gain medium. The combination of the nonlinear and gain media provides not only the structural simplification but also robust optical nonlinear functionality. Our scheme is expected to be one of the strong tools to generate the femtosecond optical pulses in the future competitive field.

References

10 GHz regeneratively mode-locked SOA fiber ring laser and its linewidth characteristics

Masato Yoshida, Atsushi Ono, and Masataka Nakazawa
Research Institute of Electrical Communication, Tohoku University, 2-1-1 Katahira, Aoba-ku, Sendai, Miyagi 980-8577, Japan
Tel: +81-22-217-5525, Fax: +81-22-217-5524, E-mail: masato@riece.tohoku.ac.jp

Abstract: We newly constructed a 10 GHz regeneratively mode-locked SOA fiber ring laser and measured its linewidth by changing the cavity length from 3.4 to 171 m. The linewidth was proportional to the inverse square root of the cavity length.

Introduction: Mode-locked lasers operating in the GHz region have been receiving a lot of attention with respect to their application to both ultrahigh-speed optical transmission and optical metrology including optical comb generation and opto-microwave oscillators [1]. An important characteristic of any laser used for these optical metrology applications is its longitudinal-mode linewidth. The linewidth of a cw laser was derived by Schawlow and Townes and is well known to be proportional to the inverse square of the cavity Q value [2]. Several experimental results concerning the relationship between the linewidth and the cavity Q value in mode-locked lasers have also been reported [3], [4]. Haneda et al. showed that the cavity Q value is a dominant linewidth parameter by using 10- and 40-GHz mode-locked laser diodes [3]. Yilmaz et al. showed that the linewidth is proportional to the inverse of the cavity length by using a 10-GHz mode-locked external linear cavity semiconductor laser [4]. In these reports, the lasers have short cavity lengths of less than 1 m. On the other hand, there have been no reports on the linewidth dependence on the cavity length in a mode-locked laser with a long cavity length exceeding several tens of meters.

In this paper we describe the first measurement, to our knowledge, of the linewidth of a mode-locked fiber ring laser as a function of the cavity length from 3.4 to 171 m. We discuss in detail the relationship between the linewidth and the cavity length, and indicate the limitation factor of the linewidth in a laser with a long cavity length.

Experimental Results: Figure 1 shows the structure of our mode-locked semiconductor optical amplifier (SOA) fiber ring laser, which was first reported by Kim et al. [5]. An SOA (InGaAsP) is used as a gain medium to realize a fiber ring laser with a short cavity length of several meters. The laser consists of an SOA, a 30 % output coupler, a polarization dependent isolator, a LiNbO3 (LN) intensity modulator, and an optical bandpass filter with a bandwidth of 3 nm. All the fibers in the cavity are polarization maintaining to prevent any polarization fluctuation. To change the cavity length, polarization-maintaining dispersion-shifted fibers (PM-DSFs) with lengths of 3.1, 48, and 168 m are installed in the cavity. In this laser, a regenerative mode-locking technique is adopted to achieve a stable pulse operation for a long period [6].

Figure 2 shows the output pulse characteristics of the laser. The threshold current for laser oscillation is 120 mA and the typical output power is 0.21 mW for a driving current of 160 mA. Figure 2(a-1) and (b-1) show the autocorrelation trace and optical spectrum when the cavity length is 3.4 m (without additional DSF). The pulse width is estimated to be 8.5 ps assuming a Gaussian pulse shape. A frequency chirp is generated by the gain-saturation-induced refraction index changes in the SOA [7]. By compensating for the chirp with a conventional single-mode fiber, we reduce this pulse width to 4.4 ps. The spectral width is 0.75 nm, which corresponds to a time-bandwidth product of 0.41. This indicates that the output pulse is close to a transform-limited Gaussian pulse. Figure 2(c-1) is a 10 GHz clock spectrum with many sidebands measured with an electrical spectrum analyzer. These sidebands correspond to supermode noise. When the cavity length is increased to 51 m by installing a 48 m-long DSF in the cavity, the pulse width is estimated to be 6.8 ps assuming a sech pulse shape from Fig. 2(a-2). After chirp compensation, the pulse width is reduced to 4.0 ps. The spectral width is 0.74 nm from Fig. 2(b-2), corresponding to a time-bandwidth product of 0.37. The supermode noise is suppressed by 65 dB compared with the peak value of the clock signal as shown in Fig. 2(c-2). This can be achieved with a self-phase modulation (SPM) effect generated in the DSF and the optical filter installed in the cavity [8]. When the cavity length is 171 m with a 168 m-long DSF, the suppression ratio of the supermode noise reaches as high as 70 dB and the pulse width is narrowed to 3.6 ps by a high-pass filter effect in the SOA [9] and the soliton effect [6].

A delayed self-heterodyne detection method is used to measure the longitudinal-mode linewidth of the laser [10]. For the present measurement we used an 80 km-long fiber delay line, which corresponds to a measurement resolution

Fig. 1. 10 GHz regeneratively mode-locked SOA fiber ring laser.
This suggests that an extrinsic noise source is dominant over the observed linewidth of this laser. We attribute this to the frequency jitter of the detected beat note induced mainly by thermal or acoustic cavity length fluctuations of the laser. In this case, the actual laser linewidth is 1/2 times the FWHM of the observed beat signal and is estimated to be 1.7 kHz for a cavity length of 171 m.

References:
Stable Pulse Generation from a Rational Harmonic Mode-Locked Fiber Ring Laser Using Carrier-Suppressed Return-To-Zero Modulation Format

Shingo Yamanaka and Joji Maeda
Department of Electrical Engineering, Faculty of Science and Technology, Tokyo University of Science,
2641 Yamazaki, Noda, Chiba, 278-8510, Japan
Phone: +81-4-7124-1501 (ext. 3725), e-mail: j7306653@ed.noda.tus.ac.jp

Abstract:
Using carrier-suppressed return-to-zero modulation format, we find stable pulse generation from a rational harmonic mode-locked fiber ring laser, the repetition rate of which is 20 GHz with the rational harmonic order of two.

1. Introduction
Optical time division multiplexing (OTDM) used for high speed optical communications needs an optical short pulse source with a high and stable repetition rate. Rational harmonic mode-locked fiber ring lasers can generate optical short pulses at a high repetition rate using a moderate modulation frequency [1]. However, the rational harmonic mode-locking has a problem of the pulse amplitude fluctuation, the origins of which are categorized into two classes. One is the fluctuation that forms a fixed pattern repeating in each modulation cycle. The other is the stochastic fluctuation, caused either by the change of the cavity length due to the temperature drift or by the competition between pairs of phase-locked modes. In particular, the patterned fluctuation is unique to rational harmonic mode locking, and should be suppressed using a special technique [2, 3].

We experimentally demonstrate the generation of a stable pulse train from a rational harmonic mode-locked laser using carrier suppressed return-to-zero (CS-RZ) format as a modulation waveform [4]. The most stable pulse is generated at the rational harmonic order of two achieving 20GHz repetition rate.

2. Pulse generation and modulation format
Let us consider a ring cavity with the cavity length of L. The fundamental frequency of the cavity, \( f_c \), is given as
\[
f_c = c/(2\pi n_{\text{eff}}),
\]
where \( c \) is the velocity of light in vacuum, \( n_{\text{eff}} \) is the effective refractive index of the cavity. Harmonic mode-locking can be achieved if the modulation frequency \( f_m \) is an integer multiple of \( f_c \), i.e.,
\[
f_m = n f_c,
\]
where \( n \) is an integer called harmonic order. This modulation frequency is equal to the pulse repetition rate. If we detune the modulation frequency by a fraction of \( f_c \), i.e.,
\[
f'_m = (n + \varphi) f_c,
\]
where \( p \) is an integer, then we can achieve rational harmonic mode-locking. The integer \( p \) is called rational harmonic order. This frequency detuning introduces time shift to pulses against the modulation waveform, and the self-consistency of the cavity field is achieved after \( p \) round-trips. In the laser output, we find \( p \) pulses within one modulation period, thus the repetition rate is
\[
f'_m = (pn + 1) f_c. \tag{4}
\]
Carrier suppressed return-to-zero (CS-RZ) is the signal format, in which the phase of adjacent symbols alternates between zero and \( \pi \). To obtain the CS-RZ format, we should bias the LN intensity modulator at the minimum transmission point (point x in Fig. 1 (a)), and keep the voltage swing within twice of \( V_c \), the half wavelength voltage of the modulator. In Fig. 1 (b), we show an oscilloscope trace of the CS-RZ signal obtained from 10 GHz RF signal, where we use the same LN modulator as that used in the laser experiment described below.

3. Experimental set-up
The experimental set-up is shown in Fig. 2. The laser cavity in this system consists of an LN intensity modulator, the insertion loss of which is 2.7 dB (EO-space, 12.5 Gb/s, zero-chirp), a variable delay line, an amplifier, an optical band pass filter and a 10 % output coupler. The amplifier consists of a 20 m bi-directionally pumped erbium-doped fiber amplifier and a semiconductor optical amplifier (COVEGA, BOA567). All components are featured polarization maintaining. The laser output is split in two arms. One is directed to an optical spectrum analyzer, and the other is directed to a high speed photo-detector (bandwidth \( \geq 50\text{GHz} \)).
connected to a sampling oscilloscope. An optical pre-amplifier (Furukawa, ErFA1215) is employed in the latter arm to obtain a clear view in the oscilloscope.

To achieve rational harmonic mode-locking, we manipulate the modulation frequency around 10 GHz with a step of 10 kHz. As an optical band pass filter, we tested four filters, #1: bandwidth (BW) 3nm (OPTQUEST, 0507C067), #2: BW 1nm (OPTQUEST, T030908A), #3: BW 1nm (OPTQUEST, T030115A), #4: BW 1nm (KOSHIN, IFOS-1565B-1-1-PY-FS).

The stability of the output pulses is evaluated from oscilloscope traces by measuring $V_{\text{av}}$ and $T_{\text{mod}}$, the standard deviation of the pulse peak voltage and that of the pulse leading edge time, respectively. We introduce a figure of stability $\sigma$ defined as

$$\sigma = \sqrt{\left(V_{\text{av}}/V_0\right)^2 + \left(T_{\text{mod}}/T_0\right)^2},$$

where $V_0$ is the average voltage, $T_0$ is the modulation period ($=100\mu s$). To obtain the statistics, oscilloscope traces are accumulated during 5 seconds.

4. Results

As a driving waveform of $p=2$ rational harmonic mode-locked laser, we compare an ordinary sinusoidal format and the CS-RZ format in terms of the quality of the pulse train. Oscilloscope traces and optical spectrum are shown in Fig. 3, where (a) and (b) are output waveform of the sinusoidally driven laser and that of the CS-RZ driven laser, respectively, and (c) and (d) are spectrum corresponding to (a) and (b), respectively. In these experiments we used the filter #4. The patterned fluctuation apparent in Fig. 3 (a) is relaxed in Fig. 3 (b). This tendency holds regardless of the type of the band pass filter. Whereas spectral peaks in Fig. 3(b) are spaced by 20 GHz, those in Fig. 3(d) are spaced by 10 GHz. This fact suggests the phase alternation in the output of the CS-RZ driven laser.

To examine the stochastic fluctuation of the CS-RZ driven laser, we show in Fig. 4 the stability of the harmonic ($p=1$) and rational harmonic ($p=2, 3, 4, 5$) mode-locking for each band pass filter, where the stability figure $\sigma$ is calculated using eq. (5). The smallest figures appear for $p=1$ with filter #1 and for $p=2$ with filter #4, where the figures are of the similar value. Thus, we can conclude that CS-RZ driven rational harmonic mode-locked lasers at harmonic order of two can show the similar stability as do ordinary harmonic mode-locked lasers.

5. Conclusion

We have experimentally demonstrated the generation of a stable pulse train in a rational harmonic mode-locked laser by using carrier suppressed return-to-zero (CS-RZ) format as a modulation waveform. Comparing the observed waveforms and spectrum, we have found the most stabilized pulse train for CS-RZ modulation format at rational harmonic order of two.

References


Pulse dropout and subharmonic locking in an active mode-locked birefringent fiber laser

Huy Quoc Lam 1, P. Shum 1, Le Nguyen Binh 2, Y.D. Gong 3, Ming Tang 1, Songnian Fu 1

1 Network Technology Research Centre, Nanyang Technological University, Singapore 637553; Tel: 65-67904682; e-mail: huylq@pmail.ntu.edu.sg
2 Electrical and Electronic Engineering Department, Monash University, Clayton, Melbourne, Australia VIC3800
3 Institute for InfoComm Research, 21 Heng Mui Keng Terrace, Singapore 119613

Abstract
We report the phenomenon of subharmonic locking of an active mode-locked birefringence fiber laser. The beating between two sets of modes in the birefringent cavity causes pulse dropping and forms the pulse pattern of 0101 or 00010001, which results in the reduction of pulse repetition rate to a half or a quarter of the modulation frequency. This phenomenon demonstrates the dynamic evolutionary property of the active mode-locked fiber laser.

1. Introduction
Active mode locking has attracted widespread attention in recent years due to its potential of producing short and high repetition rate pulse sequence, low timing jitter and ease of synchronization to a stable electronic clock signal [1, 2]. Although tremendous investigations have been performed to explore mode locking mechanisms and applications for various mode-locked fiber lasers [3, 4], the dynamic behavior of active mode-locked fiber laser is still not fully understood.

Recently, we reported the generation of dual amplitude pulses in an active mode-locked fiber laser within a birefringent cavity [5]. Different to normal mode-locked pulses with identical amplitude and polarization state, pulses polarized on both the X and Y axes simultaneously exist in the output pulse train. In this paper, we further investigate the dynamic property of the active mode-locked fiber laser in which the pulse dropout leading to subharmonic locking is presented. The demonstrated phenomenon is useful for understanding the polarization dynamic of active mode-locked fiber laser and for developing highly stable light source for future optical networks or sensor applications.

2. Experimental setup
Fig 1 shows the experimental setup of an active mode-locked fiber laser with a birefringence cavity. The laser has a ring configuration with 15 m of polarization maintaining Erbium-doped fiber (PM-EDF) serving as a gain medium. The birefringence of the cavity is mainly caused by this PM-EDF. The PM-EDF is pumped by a 980 nm laser diode through a WDM coupler. An isolator integrated in the WDM ensures unidirectional lasing. A thin-film 1.2 nm tunable optical bandpass filter is used for tuning the lasing wavelength. A LiNbO3 Mach-Zehnder intensity modulator (MZIM) is used to provide a periodical loss in the ring and hence force the laser to be mode locked. The modulator is driven by a microwave signal extracted from a signal generator. The polarization controller (PC) is used to adjust the polarization state of the lightwave signal traveling in the ring. The optical signal in the ring is coupled to the output port through a 70:30 coupler.

3. Results and discussion
The total length of the cavity is about 29.5 m which corresponds to a fundamental frequency $f_R$ of 6.923 MHz. The modulator is biased at the quadrature point. The central wavelength of the filter is 1550 nm. The modulation frequency is 2.997668 GHz which corresponds to the resonance of the 433rd harmonic. When the polarization controller is adjusted at an appropriate position, we observed on the oscilloscope a well defined mode-locked pulse train as shown in Fig...
2a. Starting from this setting, we recorded the evolution of the pulse train while rotating the middle plate of the polarization controller.

Remarkable frames extracted from the video file are presented in figs 2a – 2d. It can be seen from fig 2b that underfoot appear at every alternative pulse position as the polarization controller is rotated. Pulses in the pulse train is no longer unified but divided into two groups, denoted group A and B respectively. Pulses in group B are gradually dropped out when the PC is rotated. And finally, all pulses of group B are dropped and the pulse train consists of pulses from group A only (see fig 2c). The repetition rate of the pulse train is half of the modulation frequency, subharmonic locking occurs. It is noted that the amplitude of the pulse in this case is nearly double that of the normal locking case since the energy of the dropped pulses are transferred to the remained pulses. Further rotating of the PC results in dropping of pulses in group A as shown in fig 2d.

![Oscilloscope traces](a) full rate pulse train with clear pedestal, (b) pulses of group B are dropped in some slots, (c) all group-B pulses are dropped leaving a subharmonic mode-locked pulse with repetition rate at half of modulation frequency, (d) pulses of group A are dropped as further rotating of the polarization controller.

Pulse dropout and subharmonic locking were also observed when the modulation frequency increased to 6.009125 GHz. However, the pulse pattern in this case is ABCDABCD. The pulses are also gradually dropped as the PC is rotated. The evolution of the pulse train is as follow: ABCDABCD, _BCD_BCD, _CD_CD_, _D_. This results in subharmonic locking of the laser to a quarter of the modulation frequency.

The phenomenon can be explained due to the beating between different modes simultaneously oscillating in the laser cavity. In a birefringence cavity, there are two cavities, one with a fundamental frequency of \( f_{RX} \), the other with \( f_{RY} \). Thus there are two sets of modes oscillating inside the laser cavity. With proper setting of the PC, those two sets may oscillate independently or mutually couple energy from one to another and beat with each other inside the cavity. The independent oscillating of the two sets causes the dual polarization states locking, which has been reported in [5]. In contrast, the beating of the two sets may change the pattern of the pulse train through pulse dropping and causes the laser operate at repetition rate lower than the modulation frequency, i.e. subharmonic locking.

4. Conclusions

By employing a birefringent ring cavity and proper controlling the polarization state of the light inside the laser, we have generated a pulse train with a repetition rate at half or a quarter of the modulation frequency. This subharmonic locking comes from the pulses dropout due to beating between modes simultaneously oscillating in the cavity.

Acknowledgement

This work is partially supported by the Agency for Science, Technology and Research (A*STAR), Singapore, and Open Fund of Key Laboratory of Optical Communication and Lightwave Technologies, Beijing University of Posts and Telecommunications, Ministry of Education, P. R. China.

References

Development of a small size of Erbium doped optical fiber amplifier

Keiichi Hara  Shigeru Shikii  Shusei Aoki  Yasuaki Tamura  Hiroshi Takano
Optohub. Co., Ltd, 305 Live Tower, 7-6-8 Bessho, Minami-Ku Saitama-City, Japan
Phone: +81-48-844-8541, Fax: +81-48-844-8902, Email: hara@optohub.com

Abstract: We have developed a small size of 45x70x12 mm and +18dBm high output power erbium doped optical fiber amplifier module with two pump lasers and a control electronics circuit.

1. Introduction

Erbium doped optical fiber amplifiers (EDFAs) have been penetrated optical communication systems in the past 10 years. Multi-channel EDFAs have provided dramatically progress of optical communication systems, especially long haul terrestrial systems and undersea systems. Single channel EDFAs are applied to metro optical communication systems and CATV systems as well. Recently the metro systems are migrating from 1 or 2.5Gb/s to 10Gb/s. The 1 or 2.5 Gb/s systems can reach to around 80km without optical amplifiers. However the 10Gb/s systems require optical amplifier(s) to reach more than several 10km. Then recently developments of 40Gb/s commercial systems have been started. The 40Gb/s systems also require single channel optical amplifiers for transmitter and/or receiver sides. Of course for CATV systems single channel EDFAs are required as well. The multi-source agreement (MSA) type of EDFAs module, whose size is 70x90x12 mm, has been already expand to mainly for single channel applications. The MSA compliant EDFA module does not have control electronics, but the MSA size of EDFA modules with control electronics have been developed and provided by several vendors. Then according to requirement of smaller-sized transmission equipment, smaller-size Edfa module is required as well, and around half MSA sized EDFA module has been already developed and started to provide [1]- [5]. As for the half sized EDFA modules, however the characteristics and/or functions are restricted due to very small size. For example the output power is limited to around +15dBm and some EDFA modules do not have input power monitoring or control electronics.

In this paper, we describe a high output power of +18dBm and small size of half MSA size (45x70x12 mm) EDFA module incorporating two uncooled pump LDs and control electronics with full function.

2. Er doped optical fiber amplifier design and characteristics

The schematic diagram of the half size EDFA module configuration is shown in Fig.1. The EDFA module incorporates two pump LDs for bidirectional pumping scheme, small size hybrid optical components and a control electronics circuit. We develop the hybrid optical components shown at inside of dash line in Fig. 1, which incorporate a tap coupler, a photo detector, an optical isolator and WDM coupler for signal and pump. The size of hybrid optical component is 15x8.5x6mm. The functions of the control electronics are “automatic constant output signal power control”, “input and output signal power monitors and degradation alarms”, “pump
LD current alarm”, “EDFA module case temperature alarm” and” RS232 serial communication”. The RS232 interface functions are “set of output signal power”, “set of alarms threshold” and “read of monitor parameters values”. The electronics circuit consists of analog circuit for drive of LDs and a general-purpose microprocessor for monitors, alarms and RS232 communication. Those configuration and functions of the half MSA size EDFA module are same as a conventional MSA size EDFA module. For the package type of pump LDs uncooled mini-DIL and coaxial type are available, and for wavelengths of pump 980nm and 1480nm are available. Only in the case of bidirectional pump by using two 980nm LDs the remnant pump power should be paid attention because the remnant pump power come into the opposite side LD and may damages LD chip and make LD spectrum or power unstable. As for the EDF, conventional Si-EDF is adapted and the length is around 3m.We evaluate the half size EDFA module with 980nm forward pump of 150mW and 1480nm backward pump of 70mW, the output signal power of +18dBm is obtained at 0dBm the input signal power. The other characteristics and functions are shown in table 1. Fig. 2 shows the photograph of the half size EDFA module.

![Photograph of the EDFA module (Size: 45 × 70 × 12mm)](image)

### Table 1 Characteristics of the half MSA size EDFA module with 980nm forward and 1480nm backward pump

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Characteristics</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of pump LD</td>
<td>2</td>
<td>Uncooled type</td>
</tr>
<tr>
<td>Pump wavelength</td>
<td>980nm and 1480nm</td>
<td>980nm forward and 1480nm backward pumping</td>
</tr>
<tr>
<td>Signal wavelength</td>
<td>1530nm-1565nm</td>
<td></td>
</tr>
<tr>
<td>Output signal power</td>
<td>+18dBm</td>
<td>At 0dBm input signal power</td>
</tr>
<tr>
<td>Noise figure</td>
<td>5dB</td>
<td></td>
</tr>
<tr>
<td>Power supply</td>
<td>3.3V or 5V</td>
<td></td>
</tr>
<tr>
<td>Size</td>
<td>45 × 70 × 12mm</td>
<td></td>
</tr>
<tr>
<td>Alarm functions</td>
<td>Loss of input signal, Loss of output signal, LD current alarm, EDFA case temperature alarm</td>
<td></td>
</tr>
<tr>
<td>Monitor functions</td>
<td>Input and output signal power, LD current, EDFA case temperature</td>
<td></td>
</tr>
<tr>
<td>Control functions</td>
<td>Constant output signal power control, Pump LD on/off</td>
<td></td>
</tr>
<tr>
<td>Serial communication interface</td>
<td>Set of output signal power and alarms threshold, read of monitor parameters</td>
<td>RS-232</td>
</tr>
</tbody>
</table>

3. Conclusion

We have developed the half MSA size of EDFA module, which incorporate two uncooled pump LDs for high output power of +18dBm, full function of input and output monitors and alarms enough for single channel applications, and control electronics with RS232 serial communication interface.

### References

This Paper has been withdrawn.
Dispersion parameter and fiber length measurements technique over multi-wavelength bands

Young Cheol Kim, and Hyun Deok Kim
(School of Electrical Engineering and Computer Science, Kyungpook National University
1370 Sankyuk-dong, Puk-gu, Daegu, 702-701 South Korea, Tel: +82-53-950-7578, Email: hyundkim@ee.knu.ac.kr)

Abstract
We demonstrate a simultaneous measurements technique of the fiber length and the dispersion parameter by using low-cost Fabry-Perot lasers. It enables us to measure the dispersion parameter over multiple wavelength bands at the same time.

1 Introduction
The measurements of the dispersion parameter and the length of an optical fiber have become more important since the up-to-date high-capacity transmission systems employing the dispersion management and the distributed amplification techniques require accurate values of the dispersion parameter and the length. The measurements of the dispersion parameter and the length are usually performed independently using different instruments. Various measurement techniques have been developed for the dispersion measurement while optical-time-domain-reflectometry (OTDR) is dominantly used for the length measurement.[1-3]

We have demonstrated a simultaneous measurements technique of the dispersion parameter and the fiber length by using a low-cost Fabry-Perot laser diode (F-P LD) to decrease the measurement cost and the time required.[3-5]. However, the measurement wavelength range of the previous technique is limited by the gain spectra of the F-P LD and can not exceed several tens nm. In this paper, we demonstrate a novel measurement technique to expand the measurement wavelength range to several wavelength bands, namely, up to several hundreds nm.

2 Operation principle
Fig. 1 shows the schematic diagram of the proposed measurement system. It is arranged to form an optical closed-loop composed of the C- and the L-band F-P LDs, a wavelength-division multiplexer (WDM), an optical circulator and a fiber-under-test (FUT). Each F-P LD is pre-biased and the modulation currents generated by the signal generator are also applied to both LDs.

The F-P LDs emit multi-mode optical pulse train and the outputs are re-injected into the LDs after traveling the optical close-loop. We assume either of the F-P LDs emits three modes (λ₁, λ₂, λ₃) for the simplicity. The pulses of different modes generated by either of the LDs depart the LD at the same time and synchronize with the applied current. However, the chromatic dispersion of the FUT causes a round-trip time difference (ΔTᵦ) between the pulses as shown in Fig. 2(a). The time difference is given by ΔTᵦ=DLΔλc where D is the dispersion parameter of FUT, L is the length of FUT and Δλ is the wavelength difference between the modes.

If a specific mode pulse (λᵢ) arrives at the moment when the current applied to F-P LD is higher than the threshold current, a laser oscillation occurs only at the mode (λᵢ). Similarly, the laser oscillation may occur at one of the other modes (λⱼ) if the modulation frequency is changed so that the pulse of a specific mode (λᵢ) is re-injected to the F-P LD at the moment when the applied current is higher than the threshold current, as shown in Fig. 2(b). If the relative time deviation (ΔTᵦ) due to the modulation frequency change (ΔFᵦ) is given by ΔTᵦ= nLΔFᵦcF, where n is the refractive index of FUT, c is the velocity of light in vacuum and F is the initial modulation frequency. We can change the lasing mode from one to another if the time deviation due to the frequency change (ΔFᵦ) is equal to the round-trip time deviation (ΔTᵦ) between the modes. Thus, the dispersion parameter of the FUT is given by

\[ D = \frac{n \Delta F_{\text{mid}}}{cF \Delta \lambda} \]  \hspace{1cm} (1)

The laser oscillation may occur at the specific mode (λᵢ) again if the pulse arrival time of the mode satisfies the condition shown in Fig. 2(c). Assume that the self-seeding laser oscillation occurs at the λ₂ mode when the modulation frequency is F and the minimum frequency change required to induce the laser oscillation at the λ₂ mode again is ΔFₘᵦ. The round-trip time of the optical pulse is NL/c and should be equal to N/F to induce the laser oscillation at the λ₂ when the modulation frequency is F, where N is an integer. It also should be equal to (N+1)(F+ΔFₘᵦ) if the modulation frequency is changed to F+ΔFₘᵦ to induce the lasing at the same λ₂ mode again. From this, the length of FUT is given by

\[ L = \frac{c}{n \Delta F_{\text{mid}}} \]  \hspace{1cm} (2)

By using the Eq. (1) and the Eq. (2) we can calculate the dispersion parameter and the length of the FUT if measure the required frequency changes, the ΔFₘᵦ and the ΔFₘᵦ. We can use a direct lasing mode detection technique to measure them.[6]. Since a self-seeding laser oscillation occurs at a specific mode, the total output
power through the close-loop increases compared with the case without the laser oscillation, which caused an output power fluctuation when the modulation frequency is swept continuously. We could detect the lasing mode from the fluctuation by using a photo-detector and a signal processor as shown in Fig. 1. It is notable that the F-P LDs are coupled with WDM and operate independently and thus we can use several LDs with different gain spectra. This means that we can measure the dispersion parameters over multi-wavelength bands at the same time by using multiple F-P LDs with different gain peaks.

3 Experimental results
To confirm the performance of the proposed system, we measured the dispersion and the length of the standard single mode fiber (SMF) by using conventional C- and L-band F-P LDs. The mode-spacings ($\Delta \lambda$) of the LDs was about 0.8 nm and the center wavelengths of the LDs were 1552.5 nm and 1591.1 nm, respectively.

Fig. 3(a) shows the output of the photo-detector when we sweep the modulation frequency from 500 MHz to higher. It fluctuates with the modulation frequency and the self-seeding laser oscillations occur at the local peaks. In the C-band, the frequency change required to repeat a lasing at the same mode was 187.79 kHz and the calculated length of FUT by using the eq. (2) was 1.085 km. Also, the frequency change was 187.65 kHz and the calculated length was 1.086 km in the L-band. The relative measurement error compared with the length measured with a commercial OTDR (Anritsu, MW9060A) was less than 0.36 % in both cases.

We measured the frequency changes ($AF_{MD}$) for different wavelengths from the outputs of photo-detectors shown in Fig 3(a) and calculated the dispersion parameters using eq. (1). The measurement results are shown in Fig. 3(b) and we also show the dispersion measurement results by using a commercial instrument (PerkinElmer, FD440) to examine the accuracy of the proposed technique. Two results agree well and the relative measurement error was less than 2.8 % over C- and L-band wavelength ranges. We also measured the dispersion parameters and the lengths of optical fibers with different lengths. The results were very similar but the measurable wavelength range slightly decreased for longer fibers.

In summary, a simultaneous measurements technique of the dispersion parameter and the fiber length over multi-wavelength bands has been demonstrated by using the self-seeding laser oscillation of the F-P LDs and the direct lasing mode detection technique. The relative measurement errors compared with the commercial instruments were less than 0.36 % for length and 2.8 % for dispersion parameter measurements. It is notable that we can easily expand the multiple-band wavelength range by using multiple F-P LDs.

This work was partially supported by the grant no. D00184 from KRF and Korea Ministry of Education under the BK21 project.

References
3 K. Yoon, J. Song, H. D. Kim, OFC’04, ThP5, 2004
6 K. Yoon, J. Song, H. D. Kim, OFC’07 will be presented

![Fig. 1. The schematic diagram of the proposed measurement system](image1)

![Fig. 2. Time diagrams of applied current and optical pulse, (a) when the modulation frequency is $F_m$, (b) $F_m + AF_{MD}$ and (c) $F_m + AF_{LD}$](image2)

![Fig. 3. The outputs of the photo-detector for different modulation frequencies (a) and the measured dispersion parameters (b)](image3)
Dispersion and nonlinear coefficient measurements in optical fibres using soliton-effect compression


1: UMR FOTON, ENSSAT, 6 rue de Kerampont, 22300 Lannion, France; Thanh-Nam.Nguyen@enssat.fr
2: PERFOS, 11 rue de Broglie, 22300 Lannion, France; ntraynor@perfos.com
3: Université de Nantes, Nantes Atlantique Universités, IREENA, EA1770, Faculté des Sciences et des Techniques, 2 rue de la Houssinière - BP 9208, Nantes, F-44000 France, Dominique.Leduc@univ-nantes.fr

Abstract We present a novel and simple method to measure both the value of the second-order dispersion coefficient and the nonlinear coefficient in optical fibres. This method is based on the higher-order soliton-effect pulse compression phenomenon.

Introduction Nonlinear pulse propagation in optical fibres is influenced by both the group-velocity dispersion parameter $D$ and the nonlinear coefficient $\gamma$ [1]. The knowledge of both of these parameters is essential for a wide range of applications including optical transmission, nonlinear fibre optics or mode-locking fibre lasers. The recent development of highly-nonlinear holey fibres allows a broad range of values for these parameters to be covered ($D$, $\gamma$). Simple and efficient methods for the simultaneous measurement of both $D$ and $\gamma$ are therefore of great interest for the manufacturers or the users of these kinds of fibres. Some methods for the simultaneous measurement of $D$ and $\gamma$ are based on (1) four-wave mixing [2,3] or modulation instabilities [4,5], but are only valid for low-dispersion fibres (around the zero-dispersion wavelength). The method proposed in Ref. [6] is valid for any value of the dispersion but requires non-conventional features like a frequency resolved optical gating technique and a numerical minimization algorithm.

In this paper we report a novel and simple method for the simultaneous measurement of the dispersion parameter $D$ and the nonlinear coefficient $\gamma$ for all types of fibres in the anomalous dispersion regime. This method is based on higher-order soliton pulse compression effect and is an extension of the method proposed in Ref. [7].

Theory In the anomalous dispersion regime, the soliton effect occurs as a result of the interplay between the group-velocity dispersion (GVD) and self-phase modulation. N-th-order solitons follow a periodic evolution pattern along the fibre such that they are periodically compressed by a factor that depends on the soliton order $N$ given by [1]

$$N^2 = \frac{2\pi n^2 T_0^2}{\lambda_0^2 D}$$

(1)

where $c$ is the speed of light in vacuum, $T_0$ the pulse duration, $P_0$ the input peak power and $\lambda_0$ the wavelength. Previous work has shown that, for a given fibre length $L$, it is possible to find, by adjusting the peak power $P_0$, the lowest soliton order $N$ that leads to the maximum compression of the output pulse [7,8]. In this case, the shape of the compressed output pulse depends only on the value of the soliton order $N$ (depending itself on $D$ and $\gamma$) and the relation between $L$ and $N$ is expressed by an empirical equation as follows [9]

$$L \left( \frac{\lambda_0}{\lambda_0^2} \right)^2 = \frac{0.32}{N^3} + \frac{1.1}{N^2}$$

(2)

We propose a simple method, related to the shape of the autocorrelation trace of the compressed pulse, to extract the values of $D$ and $\gamma$. The first parameter we measure is the compression factor $F_a$ defined as the ratio between the full width at half maximum of the initial and compressed pulses (see Fig. 1). The second parameter is the ratio $R_a$ defined as the ratio between the main peak and the level of the secondary peaks of the compressed pulse (see Fig. 1). The last parameter is the peak power $P_0$ of the initial pulse. Similar to the work in Ref. [9], we have found approximate empirical relations to extract the values of $D$ from the measured values of $F_a$, $R_a$ with good accuracy,

$$L \left( \frac{\lambda_0}{\lambda_0^2} \right)^2 D = \frac{3.224}{F_a} + \frac{3.73}{F_a^2} + \frac{1.774}{F_a^3} - 0.007$$

(3)

$$L \left( \frac{\lambda_0}{\lambda_0^2} \right)^2 D = \frac{9.775}{R_a} + \frac{18.075}{R_a} - \frac{11.347}{R_a^2} + 2.438$$

(4)

The method we propose works as follows. For a given length $L$ of an anomalous dispersion fibre, by adjusting the power of a launched soliton pulse of duration $T_0$ at wavelength $\lambda_0$, we obtain the soliton order $N$ that leads to the maximum pulse compression. Then, we measure the compression factor $F_a$ (method A) and the ratio $R_a$ (method B). According to equation (3), the value of $F_a$ gives a value of dispersion which we shall call $D_A$ and, according to equation (4), the value of $R_a$ gives a value which we shall call $D_B$. In the ideal case (measurement without error) $D_A$ and $D_B$ are identical.

![Fig. 1. Theoretical autocorrelation traces of an input pulse and a compressed pulse and definition of $F_a$ and $R_a$.](image)

In fact, both values of $D$ obtained with method A and method B can differ because they are obtained using approximate relations. Moreover, in these relations no impact of cubic dispersion, attenuation and higher-order nonlinearity are included.

Calculations of $D_A$ and $D_B$ when taking into account the effect of third order dispersion ($\beta_3 = 0.1$ ps$^3$/km), the effect of fibre loss ($\alpha = 0.2$ dB/km), the effect of self steepening and stimulated Raman scattering ($\gamma_S = 3$ ps$^{-1}$/km) are carried out numerically. Fig. 2(a) shows the relative error between the theoretical values and the calculated...
values of dispersion over a wide range of dispersion (from 0.2 ps/km/nm to 100 ps/km/nm). We note that these effects have a detrimental influence on the accuracy of the methods for the lower values of dispersion (D<1 ps/km/nm). We also note that the errors of each method are of the opposite sign. This can be explained as follows. When an output pulse is more compressed than in the ideal case (the measurement of $F_2$ is over-estimated), the quality of the compressed pulse decreases [1] (the main peak is lower and the side-peaks are higher) and the measured value of $R_1$ is underestimated. Conversely, when an output pulse is less compressed than the ideal case, $F_2$ is under-estimated and $R_2$ over-estimated. By taking the average value of dispersion calculated from the two methods we have the final value of dispersion with an error less than 1% for $D>1$ ps/km/nm and less than 6% for $D>0.5$ ps/km/nm (see Fig. 2(b)).

Replacing the value of $D$ in relation (2) allows us to determine $N$. Introducing the value of $D$ and $N$ in relation (1) allows us to find $\gamma$. The accuracy in the calculation of $\gamma$ is on the same order than for $D$ (Fig. 2(b)).

From the autocorrelation traces we find $F_2 = 20.7$ and $R_2 = 3.9$. Therefore, the value of dispersion $D$ and nonlinear coefficient $\gamma$ are 112.5 ps/km/nm and 44 W km$^{-1}$ respectively. Before comparing these results with other conventional methods, we also performed the measurement on a 304 m-long SMF. The wavelength is set to $\lambda_0 = 1560$ nm and the pulse duration is $T_0 = 8.5$ ps. The procedure of measurement is the same as above and we obtain $P_0 = 13.5$ W, $F_3 = 26.2$, $R_3 = 3.8$. The dispersion is found to be $D = 17.2$ ps/km/nm and the nonlinear coefficient $\gamma = 1.4$ W km$^{-1}$.

The comparison between our results and the results obtained by conventional methods are summarized in Table 1. The reference method used to measure the dispersion is the low coherence interferometry method [10]. The reference method used to measure the nonlinear coefficient is based on the measurement of the effective area of the optical fibre, knowing the nonlinear refractive index $n_2$ of silica [1].

<table>
<thead>
<tr>
<th>Fibre</th>
<th>$D_{\text{eff}}$ (ps/km/nm)</th>
<th>$\gamma_{\text{eff}}$ (W km$^{-1}$)</th>
<th>$D$ (ps/km/nm)</th>
<th>$\gamma$ (W km$^{-1}$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>NHF</td>
<td>115</td>
<td>12.5</td>
<td>112.5</td>
<td>44.2</td>
</tr>
<tr>
<td>SMF</td>
<td>17.6</td>
<td>17.2</td>
<td>44</td>
<td>1.4</td>
</tr>
</tbody>
</table>

Table 1. Results of dispersion and nonlinear coefficient measurements of NHF at 1555 nm and SMF at 1560nm.

We find good agreement between our measured values and the values measured by other conventional methods, indicating that our method is valid and reliable.

**Conclusions**

We have proposed a novel and simple method for simultaneous measurement of both the dispersion and the nonlinear coefficient of optical fibres in the anomalous dispersion regime. This method is based on the soliton-compression effect in optical fibres and is reliable for dispersion values greater than 0.5 ps/km/nm. Experimental measurements were performed and have demonstrated the accuracy and reliability of the method.

**Acknowledgements**

This work is supported by the Conseil Regional de Bretagne and the European Union (Fonds Europeen de Developpement Regional).

**References**

Tunable Lasers for Optical Systems; Technologies and Commercialisation

Jens Buus, Gayton Photonics, 6 Baker Street, Gayton, Northants, NN7 3EZ, UK
tel +44 (0)1604 859253, email jbuus@btinternet.com

Abstract

We will give an overview of different tunable laser technologies and compare the different aspects of their performance. This will be related to the current state of commercialisation. Some present lines of development will be discussed.

Why tunable lasers?

The main argument for the use of tunable lasers is the reduced cost (and logistic complexity) of sparing. If tunable lasers become sufficiently cheap, they might even replace fixed wavelength lasers, at least for use in high data rate DWDM systems. In addition tunable lasers offer flexible provisioning and also support new network features based on the use of wavelength routing. One can use wavelength switching either as circuit switching or as packet switching. Due to the attractions of tunability there has been intense activity on research and development in the last 20 years, and tunable lasers was one of the hottest areas in optical components during the telecom bubble. The arguments for tunability are still valid, and we now see the emergence of a significant market, and a second wave of interest is building up.

Technologies and standards

We restrict the discussion below to lasers capable of covering a full band (at least 30nm), this means that thermally tuned DFB lasers and standard DBR lasers will not be included. Widely tunable lasers can broadly be subdivided into 4 groups of structures which each have specific advantages and drawbacks:

- DBR lasers with extended tuning range. These have two special design gratings, with either one at the front and one at the back, or with both in parallel at the back. The main advantages are tuning speed, integrability (with SOAs and/or modulators), and component size. However, wavelength control is a challenge.
- DFB (or DBR) arrays. A combiner function is required. If this is implemented monolithically the output power has to be boosted with an SOA and the chip becomes large. Use of an external combiner allows a high output power, but involves a moving part (MEMS mirror). In DFBs the wavelength tuning of each array element is thermal, and therefore simple but slow.
- DBR arrays offer fast (partially continuous) tuning, but again a combiner is required. Tunable VCSELS. These are small and relatively simple, but the power level is low. There is little current commercial activity.
- External cavity lasers (ECLs). A variety of versions exist, and generally they offer a high optical power, a wide tuning range, and a narrow linewidth. The main issues are: size, drive requirements for the tunable filter or mirror, and tuning speed (at least in the case of thermal tuning). More details on the different technologies can be found in [1].

In order to promote the introduction of tunable lasers, groups of manufacturers have entered into multi source agreements (MSAs). One example is the Integrable Tunable Laser Assembly (ITLA). This agreement specifies the physical size (74mm x 30.5mm x 10.5mm), the optical performance, and the control interface of a CW tunable source. ITLAs from
different manufacturers are completely interchangeable, and the actual laser structure and tuning mechanism is “invisible” to the user.

Companies (past and present) and market

Due to the large interest in tunable lasers during the “bubble” years, a number of companies were started with tunable lasers as the main or only product. With two exceptions these companies no longer exist: Agility (SG-DBR) now taken over by JDSU; Santur (DFB array) is still privately held.

Some companies have tunable lasers as a part of a wider portfolio of products, examples are: Intel (ECL), Bookham (DS-DBR) and Pirelli (ECL). A number of companies had tunable laser activities (mainly on DBRs) in the past, but seem to have stopped, these include: Agere (Lucent), Avanex (Alcatel), Multiplex and Agilent.

With the renewed interest after the burst, a couple of new specialist tunable laser companies have emerged during the last few years: Paxera (ECL) now taken over by Neophotonics and Syntune (MGY-DBR).

It is noticeable that developments in Japan have been much less turbulent with no start-ups, no take-overs and no company closures. It is also noticeable that the Japanese activities are biased towards DFB arrays (Fujitsu, Furukawa, NTT) and ECLs (Fujitsu, NEC). In addition there is work on DBR arrays, ring resonator structures and the TDA-DFB.

In a study published in 1999, a market of 30M$ was forecast for 2002, growing to almost 1G$ by 2005. In reality the development has been lagging this prediction by at least 3 years, but we are now finally seeing a strong commercial interest with 10s of thousands of units being shipped per year and a significant growth rate. Currently the largest suppliers are Santur, JDSU (Agility), and Intel.

Pluggable tunable?

With the emergence of small form factor pluggable transceiver modules, such as the XFP, the sparing argument for tunable lasers has been somewhat eroded since the cost of sparing has been significantly reduced. The challenge is therefore to develop a tunable and pluggable transceiver module for 10Gbit/s with full band coverage. This requires a small laser package (e.g. TOSA), and a low power consumption since the thermal management is a major issue. A step towards this goal is the recent development of a laser monolithically integrated with a MZ modulator.

DWDM vs CWDM

Current lasers for DWDM requires the use of a thermo electric cooler (TEC) in order to avoid thermal drift of the wavelength. CWDM on the other hand does not require cooling, but the resulting 20nm channel separation reduces the spectral efficiency significantly. An interesting approach is to use tuning to compensate for thermal drift. This would allow a channel spacing of a few nm without the added cost of the TEC. A couple recent developments in this area will be described in more detail.

Conclusion

The attractions of tunability have not disappeared with the burst of the bubble, but the commercialisation has been slowed down by a number of years, and most of the start-up companies from the late 90’s have not survived. The market now finally seems to have taken off with a significant number of units being sold and with a good growth rate.

New companies are emerging, new technologies are being explored, and there is significant interest in miniaturisation, improved performance, increased functionality and cost reduction.

References

**Abstract**

We achieved a mode-hop-free tuning range of more than 5 nm for phase controlled wavelength tunable DFB lasers with high coupling coefficient gratings by changing the initial phase between their two DFB sections.

**Introduction**

In future optical network systems and optical sensing systems, it will be necessary for wavelength tunable lasers to change wavelength quickly. Carrier variation in a semiconductor induces a large and fast change in refractive index. Many types of wavelength tunable lasers operated by current injection have been reported including short cavity distributed Bragg reflector (DBR) lasers [1], super structure grating (SSG) [2] and sampled grating (SG) [3] DBR lasers and tunable twin guide (TTG) lasers [4].

We have studied wavelength tunable DFB lasers with high coupling coefficient ($\kappa$) gratings that can be controlled easily because they have only a single electrode for tuning. Moreover, they can be integrated with other optical components such as semiconductor optical amplifiers (SOA) and multi-mode interferometer (MMI) couplers to form a tunable laser array [5].

We have reported a continuous tuning range of 4.6 nm for a tunable wavelength DFB laser with high $\kappa$ gratings of about 390 cm$^{-1}$ [6]. In this work, we investigate the relationship between the two gratings in our lasers. As a result, we achieved a continuous wavelength tuning range of over 5 nm.

**Experiment**

Figure 1 shows a schematic diagram of our wavelength tunable DFB laser, which consists of two DFB sections (high $\kappa$ gratings with gain) and one phase shift section. The lasing wavelength changes within a wide reflection band (stop band) according to the phase shift. The amount of phase shift can be changed smoothly and quickly by injecting current into the phase shift section. A grating with a $\kappa$ as high as 390 cm$^{-1}$ yields a stop band more than 10 nm wide.

---

**Effect of Initial Phase in Wavelength Tunable DFB Laser with High Coupling Coefficient Gratings**

N. Nunoya, Y. Shibata, H. Ishii, H. Okamoto, Y. Kawaguchi, Y. Kondo, and H. Oohashi

**NTT Photonics Laboratories, NTT Corporation,**

3-1, Morinosato Wakamiya, Atsugi-shi, Kanagawa, 243-0198, Japan

Tel: +81 46 240 3220, Fax: +81 46 240 4345, E-mail: nunoya@aecl.ntt.co.jp

**Fig. 1 Wavelength tunable DFB laser**

We fabricated wavelength tunable DFB lasers with four kinds of grating patterns providing different initial phase shift ($0, \lambda/8, \lambda/4, 3/8\lambda$), as shown in Fig. 2. The actual optical phases are determined by both the initial grating position and the optical length of the phase shift section.

**Fig. 2 Images of four grating patterns**
Figure 3 shows the dependence of the lasing wavelength on the voltage supplied to the phase shift section for tunable DFB lasers with different initial phase shifts. All lasers consist of two 40-μm long DFB sections and a 15-μm long phase shift section. The DFB current \( (I_a) \) was fixed at 50 mA, and the tuning voltage \( (V_b) \) ranged from 0 to 1.8 V. The lasing wavelength can be tuned from the longer wavelength side to the shorter wavelength side. The initial lasing wavelength at \( V_b = 0 \) V depends on the initial phase. The tuning range is the largest (5 nm) for the laser with an initial phase of 0, where the lasing wavelength moves continuously from 1550.3 to 1545.3 nm. Whereas the laser with an initial phase of \( \lambda/8 \) exhibits mode hopping from a shorter to a longer wavelength during tuning, the tuning ranges for lasers with initial phases of \( \lambda/4 \) and \( 3\lambda/8 \) are 3.9 and 4.6 nm, respectively. The initial lasing wavelength should be set on the shorter wavelength side as long as there is no mode hopping during tuning.

Figure 4 shows the tuning spectrum of the laser with an initial phase of 0, which has a tuning range of 5 nm. We achieved an output power of more than 1 mW and a side mode suppression ratio (SMSR) of more than 35 dB during tuning. Although the output power decreases, a wider tuning range of more than 5 nm can be obtained with a small DFB current.

**Conclusion**

We achieved a mode-hop-free tuning range of 5 nm for wavelength tunable DFB lasers with optimized initial phase shifted gratings for a \( \kappa \) value of 390 cm\(^{-1}\).

**Acknowledgments**

The authors thank the members of NTT Photonics Laboratories for fruitful discussions. The authors also thank the technical staff for their experimental help.

**References**

Short-cavity DBR Laser using an InP/InGaAsP Deep-ridge Waveguide with Vertical-groove Gratings

Toru Segawa (1), Boudewijn Docter (1, 2), Takaaki Kakitsuka (1), Shinji Matsuo (1), Tetsuyoshi Ishii (1), Yoshihiro Kawaguchi (1), Yasuhiro Kondo (1), Fouad Karouta (2), Meint K. Smit (2), and Hiroyuki Suzuki (1)

1 NTT Photonics Laboratories, NTT Corporation, 3-1 Morinosato Wakamiya, Atsugi, Kanagawa, 243-0198, Japan,
2 COBRA Research Institute, Technical University of Eindhoven, P.O.Box 127, 5600 MB Eindhoven, The Netherlands,
Tel: +81-46-240-3214, Fax: +81-46-240-3259, e-mail: segawa@acl.ntt.co.jp

Abstract
A compact distributed Bragg reflector (DBR) laser was fabricated using an InP/InGaAsP deep-ridge waveguide with vertical-groove gratings. We achieved stable single-mode laser operation with an active length of only 25-µm with a low threshold current.

1 Introduction
Large-scale photonic integrated circuits (PICs) are important for future optical networks, because they can provide low-cost, compact, and low-power consumption high-functionality devices through the monolithic integration of many photonic components on a single wafer. In large-scale PICs, short-cavity DBR lasers will be key components: the lasers could be employed as various functional light sources with a small structure, such as a mode-hop-free tunable laser [1] or a high-speed direct modulation laser because of its short-cavity length [2]. For application in large-scale PICs, these lasers should be compact, low power consumption, and easy to integrate.

In this paper, we present short-cavity DBR lasers using an InP/InGaAsP deep-ridge waveguide with vertical-groove gratings, where a corrugation is applied on both sides of the waveguide. This structure enables us to fabricate DBRs with a simple one-step-etching technique [3]. In this structure, the coupling strength is controlled by electron beam (EB) lithography. Various types of DBR lasers can therefore be integrated on a single chip. For example, unidirectional light output from DBR lasers, which is desirable in large-scale PICs, can therefore be integrated on a single chip. For example, unidirectional light output from DBR lasers, which is desirable in large-scale PICs, can therefore be achieved by using an asymmetric-κ DBR structure [4]. In addition, high coupling strengths can be obtained with vertical-groove gratings, allowing us to make the device compact. Furthermore, the gratings can easily be integrated with other components, such as AWGs, because all these components can be fabricated using the same process. A fabricated device with an active area of only 25 µm and a total device length of 200 µm shows stable laser operation. The threshold current is 13 mA and the side mode suppression ratio (SMSR) is more than 35 dB.

2 Device structure and fabrication
A schematic top view and a photograph of the fabricated device are shown in Fig. 1. The device consists of an active region and front/rear gratings, which are connected by a tapered waveguide. A deep-ridge waveguide structure was used in the grating and tapered waveguide, whereas a shallow-ridge waveguide was used in the active region. The active region consists of a multiple quantum well (MQW) active layer containing 10 QWs on an n-InP substrate. The grating and tapered waveguide regions (InGaAsP layer, λPL = 1.4Q) were butt-jointed to the active region using selective area growth. Then, a 1.5 µm-thick p-InP cladding layer and a p+-InGaAsP contact layer were grown over the entire surface. All layers were grown by metal organic vapor phase epitaxy (MOVPE).

The deep-ridge, containing the vertical-groove grating structures, was formed by chlorine-based inductively coupled plasma reactive ion etching (ICP-RIE). The waveguide pattern was defined by EB lithography and then transferred into a predeposited SiN layer (600 nm) by RIE, which served as a mask for ICP-RIE. This process provides a deep etch (4 µm) with steep and smooth sidewalls. After fabricating the deep-ridge structure, the shallow-ridge was formed by selective wet etching with diluted HCl, which stops at the top of the active layer. The shallow-ridge waveguide was made 3-µm wide to provide enough gain for the short cavity. The shallow-ridge waveguide structure was coated with a benzocyclobutene (BCB) and etched back for planarization. Then, Ni/Zn/Au electrodes were deposited on the contact layer by the lift-off process. After the metallization, the sample was cleaved and both facets were coated with antireflection films.

The deep-ridge waveguide with vertical-groove
Measurements showed that a cavity by using the high different. Modal overlap 17 We presented compact sh 0.55 35 We thus used 23- 0.75 25 L 30 0.75 e laser operation of DBR 0.55 200 µm. Despite a few kinks in the L-I curve for the devices show threshold currents of 14, 13 and 10 mA for active region length as shown in Fig. 3. The measured data includes the fiber-chip coupling losses of 4 dB. The temperature as shown in Fig. 3. The measured data of the smallest device became 200 µm. For the longer devices, the excess loss. The power consumption was on the order of 30 mW per device, which means that several hundred devices could easily be integrated on a single chip. These devices are promising components for future large scale PICs.

4 Conclusion
We presented compact short-cavity DBR lasers using an InP/InGaAsP deep-ridge waveguide with vertical-groove gratings, which makes the devices useful as building blocks in PICs. Devices with an active region of only 25 µm showed single-mode laser operation with a high SMSR. The power consumption was on the order of 30 mW per device, which means that several hundred devices could easily be integrated on a single chip. These devices are promising components for future large scale PICs.

References
The Impact of InAlGaAs Barriers on Material and Differential Gain of Quantum Wells on Low Indium Content InGaAs Ternary Substrates

T. Fujisawa, M. Arai, T. Yamanaka, Y. Kondo, and H. Yasaka
Photonics Laboratories, NTT Corporation, 3-1 Morinosato-Wakamiya, Atsugi, Kanagawa, 243-0198, Japan; fujisawa@aecl.ntt.co.jp

Abstract: The impact of InAlGaAs barriers on material and differential gain of quantum wells (QWs) on InGaAs substrates is theoretically investigated. Material gain of QWs with InAlGaAs barriers can be largely increased because of deeper $\Delta E_c$.

1. Introduction
1.3-μm lasers for data communications systems operating in a wide temperature range are strongly desired. Quantum Well (QW) lasers grown on InGaAs ternary substrates are promising candidates for high-temperature operation because of their large conduction band offset $\Delta E_c$ [1]. Lasers fabricated on InGaAs substrates with relatively large In content exhibit excellent temperature characteristics [2]. But the results were limited to substrate with In content of 0.2 to 0.3 because a highly strained well layer is required in order for low-In-content substrate to obtain 1.3-μm light. However, deeper $\Delta E_c$ can be available and the fabrication of substrates is easier for low-In-content substrates. Recently, lasers on low-In-content (In:0.1) InGaAs substrate have been realized [3]. In [3], an InGaAs was used for the barrier layer. By introducing an InAlGaAs barrier, gain and temperature characteristics can be improved further due to the large barrier height of InAlGaAs compared with that of InGaAs barrier. Furthermore, although the In content of the substrate is an important design parameter for QWs on InGaAs substrates, the influence of the In content of the substrate has not been discussed theoretically. Therefore, it is important to investigate the performance of QWs on ternary substrate for various materials and structural parameters.

In this paper, the material gain and differential gain of InGaAs/InAlGaAs QWs on InGaAs ternary substrates are theoretically investigated. It is shown that the material gain of InGaAs/InAlGaAs QWs becomes 30% larger than that of InGaAs/InGaAs QWs and the temperature dependency is largely improved.

2. QW Structures and Analysis Methods

The left side of Fig. 1 is a schematic of the QW structure we analyzed. The structure consists of three QWs with 120-nm SCH layers and InGaP cladding layers. The thicknesses of the well and barrier layers are 10 and 20 nm, respectively. The well layer is InGaAs and barrier and SCH layers are InGaAs or InAlGaAs. The strain of the well layer was adjusted to obtain 1.3-μm band-gap between the first conduction and valence sub-bands.

Band lineups were obtained using the tight-binding approach, taking into account of repulsion from a cation d-orbital [4], and a flat band was assumed because of the deep $\Delta E_c$. The valence band energy dispersion relation was calculated using the block diagonalized 6×6 Luttinger-Kohn Hamiltonian to include the effect of spin-orbit split-off bands [5] while parabolic bands were assumed for conduction bands. The band structure of the QWs was calculated for given barrier and SCH materials and substrate In content by solving the effective-mass equation discretized by finite-element method. Material parameters used in the simulation were extracted from [6]. Based on the calculated band structure, the quasi-Fermi level, material gain, and differential gain were obtained for given carrier density and temperature. Lorentzian broadening of energy spectrum was assumed in calculating material gain.

3. Results for Material Gain, Differential Gain, and Temperature Dependency

The right side of Fig. 1 shows the potential profiles of the QWs on In$_{0.15}$Ga$_{0.85}$As substrates. For the InAlGaAs barrier, the band-gap
wavelength is 0.8 μm (Al:0.26). Compressive strain of 2.35% is assumed in the well layer. By using InAlGaAs barriers, ΔEc almost doubles (410 meV) compared with that of InGaAs barriers (210 meV).

Fig. 2 shows the peak material gain as a function of carrier density. We can see that the material gain of InGaAs/InAlGaAs QWs is 30% larger than that of InGaAs/InGaAs QWs. This is because the deeper ΔEc prevents electrons from filling higher energy sub-bands. For high temperatures, the material gain of the QWs is reduced because the quasi-Fermi level becomes smaller due to the large density of states in higher energy states. When the temperature is increased, the degradation of material gain for the InGaAs/InAlGaAs QWs is smaller than that of the InGaAs/InGaAs QWs, indicating the superiority of InAlGaAs barriers for high-temperature operation. Fig. 3 shows the differential gain as a function of carrier density. At room and high temperature near the threshold, differential gain of the InGaAs/InAlGaAs QWs is 20% larger than that of the InGaAs/InGaAs QWs, showing the possibility of high-speed operation in the high temperature environment.

One of the interesting features of QWs on ternary substrates is the flexibility of the substrate composition. By changing the In content of the substrate, various characteristics of QWs can be largely tuned. Fig. 4 shows the carrier density for obtaining material gain of 1000 cm⁻¹ as a function of temperature for various In contents of the substrate. Compressive strains of 2.8 and 1.9% in the well layer were assumed for In0.1Ga0.9As and In0.2Ga0.8As substrates. Barrier band-gap wavelengths are assumed to be 0.76 (Al:0.26) and 0.84 μm (Al:0.25) for In0.1Ga0.9As and In0.2Ga0.8As substrates. The slope of carrier density to temperature of the InGaAs/InGaAs QWs is larger than that of the InGaAs/InAlGaAs QWs as discussed above. This feature is more evident for QWs on high In content substrate. For In0.2Ga0.8As substrate, the carrier density to obtain 1000 cm⁻¹ material gain is increased 25%, while the increase is 16% for In0.1Ga0.9As substrate. This is because, for low In content substrate, electrons are strongly confined in the well region even in the high-temperature environment due to the larger values of ΔEc.

4. Conclusion
We have theoretically clarified the material gain, the differential gain, and the temperature dependency of QWs on InGaAs ternary substrates can be greatly improved by using InAlGaAs barriers and low In content substrates. The material gain of InGaAs/InAlGaAs QWs is 30% larger than that of InGaAs/InGaAs QWs.

References
Analysis of Large Kink Mechanism in I-L Characteristics of Tunnel Injection Lasers

Hiroshi Nakajima¹, Tomoyuki Miyamoto¹, Takahiro Iwasaki¹, Yasutaka Higa¹, Fumio Koyama¹
¹ Microsystem Research Center, P&I Lab., Tokyo Institute of Technology
R2-39, 4259 Nagatsuta, Midori-ku, Yokohama 226-8503, Japan
Tel: 045-924-5077 ext 4459, Fax: 045-924-5977, Email: nakajima.h.ae@m.titech.ac.jp

Abstract
Large kink in I-L and lasing prohibition are interesting characteristics of tunneling-injection lasers. We suggested these characteristics are due to change of the carrier accumulation in the tunneling-well and ballistic injection to p-cladding layer, respectively.

1 Introduction
Semiconductor lasers with high-speed direct modulation capability and good temperature characteristics are strongly required for local- and metro-area networks. The modulation speed of conventional quantum well (QW) lasers is limited by a carrier capture time.¹ A novel tunnel injection structure was proposed to overcome this limit.²,³ It is possible to inject carriers directly into emission energy level of the QWs by a resonant tunneling structure. A high speed direct modulation is expected by high speed direct carrier injection.⁴ Good temperature characteristics and high quantum efficiency are also expected because hot carriers with an excess energy can be eliminated by the tunnel barriers.

Lasing characteristics of large kink in I-L and prohibition of lasing are also reported by lasers with tunneling structures.⁵ However, these novel lasing mechanisms has been not understood. Clarifying of these mechanisms is important for improvement of the lasing performance of tunneling injection lasers.

In this report, we discussed carrier dynamics of tunnel injection lasers to clarify the lasing mechanism of large kink in I-L and prohibition of lasing by theoretical analysis of the tunneling current in comparison with the measured characteristics.

2 Analysis and measurement
Figure 1 shows a schematic band diagram of MBE-grown tunneling injection lasers. The emission layer is Ga₀.₈In₀.₂As single QW with the thickness of 80 Å. Tunneling barrier layers are Al₀.₈Ga₀.₂As with the thickness of L_b and the tunneling well layer is Ga₀.₈In₀.₂As with the thickness of L_w. Stripe laser structures with no-facet coating and the cavity length of approximately 700μm were measured under room temperature pulsed condition.

Fig. 1 Schematic band diagram of grown and model tunneling injection lasers

Figure 3a) is experimentally measured I-L, I-V characteristics for L_b of 30, 20, 10 Å and L_w of 40 Å. Figure 3b) is theoretically calculated J-V characteristics. Detail of the fabrication and measurement of tunneling injection lasers are shown in reference 5). The theoretical analysis was carried out by calculation of the tunneling current from n-side SCH-layer to p-side one. The tunneling probability was calculated by the transfer matrix method for the biased potential structures. The dotted line of Fig. 3b) shows a voltage at which the
quantized energy level in the tunneling well becomes above the band edge of the p-SCH layer.

In Fig. 3a), there is kink in I-L characteristics for $L_b$ of 30 Å. It is thought that only a sample with $L_b$ = 30 Å reached to a voltage at which the quantized level in the tunneling well reaches the band edge of the p-SCH layer in the measured current density range. The quantized level of the injection well of the fabricated laser is located below the band edge of the n-SCH layer, and thus the carrier may accumulate in the tunneling well. The accumulated carrier flowed out to the p-SCH layer when the bias condition reached the kink. This may cause the potential structure change and the gain characteristics in the emission well are also modified, resulting in the kink in I-L characteristics.

Figure 4a) shows measured I-L, I-V characteristics for $L_b$ of 20Å, 10 Å and $L_w$ of 40Å. as shown in Fig. 4a), though sufficient current was obtained under a high bias for the analysis.

In conclusion, we discussed about lasing mechanism of tunneling injection lasers by comparing measured result with analyzed one. The tunnel injection laser shows various characteristics by tunneling structures. A large kink may be caused by the carrier accumulation in the tunneling well and prohibition of laser is due to carrier leakage to the p-cladding.

References
Photon Funneling from Photonic Crystal Nanolasers

Yong-Hee Lee, In-Kag Hwang, and Se-Heon Kim

Department of Physics, KAIST
373-1 Kusung-dong, Yusung-gu, Taejon, Korea 305-701
E-mail: yhlee@kaist.ac.kr

ABSTRACT

Efficient photon out-coupling from photonic crystal nanolasers are to be discussed. Specifically the vertical beaming scheme and the all-fiber coupling scheme are discussed in detail.

The ability to localize photons into photonic bandgap semiconductor microcavities having wavelength-scale volumes and high quality factors is expected to enable us to study the cavity quantum electrodynamics in semiconductor material systems. Several ultra-small, photonic crystal lasers have been recently reported. Once we identify the high-Q small-V resonator, the next issue to collect these valuable photons generated from this nano-cavity efficiently. In general, the far-field radiation pattern coming from the wavelength-scale resonator is diverging over a wide solid angle and the output beam is hard to be collected by using conventional optics. Here we are discussing two possibilities: vertical beaming and all-fiber coupling schemes.

For the efficient vertical beaming, the hexapole single-cell photonic crystal resonator was chosen and studied. The hexapole mode has a central node and suitable electrical pumping. The nondegeneracy of the mode is also advantageous to satisfy the condition of the single-modeness. The distribution of the electric field in the hexapole mode is delicately balanced. And this balance can be perturbed by simple structural modification. In fact, we break the symmetry along the horizontal axis by moving two nearest air holes slightly outward as indicated in Fig. 1(a). As one can witness in Fig. 1(b), the far field pattern does converge along the vertical direction. In essence, this can be thought of the loss engineering. Remember that the original hexapole mode has a Q factor in excess of 100,000. In other words, in the beginning the original hexapole mode is almost lossless. In fact, by adjusting the two nearest air holes, we are controlling the loss channel of output photons and the total Q of the cavity becomes smaller. When the vertical beaming condition is optimized, the Q factor is degraded down to ~10,000, still large enough to provide a large Purcell factor. It is interesting to find that, through this structural perturbation, the DC component of E_x-field is newly created and escape into the free space. Therefore, the far field radiation mostly x-polarized becomes available. It is confirmed that over 70% of the photons generated in the resonator is contained within the angular span less than 30 degrees from the vertical direction under the optimum condition. In other words, photons can be easily collected by using a commercial microscope objective lens. Moreover, the collection efficiency can be increased up to 84% with the help of bottom distributed feedback reflector.

In addition to the vertical beaming, we proposed and demonstrated the all-fiber coupling scheme. To this end, photonic crystal small-linear-resonator that has a high-Q value and small mode volume was investigated. The waveguide-type linear photonic crystal resonator was modified to introduce the better overlap of the resonant mode with the micro-fiber, in both k-space and real space. To achieve efficient coupling, the optical fiber was tapered to its extreme. The final diameter of the micro-fiber is ~1 micron where only one propagating mode is allowed. For longitudinal phase matching, a few small air holes are inserted inside the linear waveguide-type resonator. This simple modification brings two main advantages over the conventional linear photonic crystal resonator. The first merit is that the wave vector of the modified cavity could now have a reasonable overlap with those of the silica waveguide. In other words, two modes share a certain region of overlap in k-space to ensure efficient optical coupling. The second
point is that of better spatial overlap along y-direction. Even after the modification, the linear resonators still have Q-factors in excess of 10,000. The modal volume of the cavity is somewhat larger than that of the single-cell cavity. The Purcell factor can be easily made larger than 1,000.

In this configuration, pump light from a laser diode is to be injected through the tapered microfiber into the cavity by very local evanescent coupling. Once the resonant mode of interest is excited, photons will be collected into the curved microfiber, in both forward and backward directions. The photons coming out of the two ends into a single fiber can be collected straightforwardly by using a commercial fiber coupler. In this scheme where photons are coupled directly into the fiber without the help of conventional bulk optics, the photon coupling efficiency in excess of 70% is realized [7]. Since the photon generator unit contains only a miniature photonic crystal resonator integrated with a micro-fiber, the final unit can be very simple and mechanically robust. When optically pumped using a 980-nm InGaAs diode lasers, this all-fiber coupled modified photonic crystal resonator functions as a laser with small threshold pump power of 25 μW and generates continuous wave output optical power larger than 10 nW.

In summary, we demonstrated two efficient ways to collect photons from the wavelength-scale photonic crystal resonators and lasers.

REFERENCES


Fig. 1 Vertical beaming. (a) Modification of nearest air holes. (b) Far field pattern with various air hole size. (c) Q factor as a function of detuning parameter.
GaInNAs Distributed Feedback (DFB) Laser Diode with High Resistive Semiconductor Current-blocking Layer

Jun-ichi Hashimoto\textsuperscript{1,2}, Kenji Koyama\textsuperscript{1,2}, Takashi Ishizuka\textsuperscript{1,2}, Yukihiro Tsuji\textsuperscript{1}, Kousuke Fujii\textsuperscript{1}, Takashi Yamada\textsuperscript{1,2}, Chie Fukuda\textsuperscript{1}, Yutaka Onishi\textsuperscript{1} and Tsukuru Katsuyama\textsuperscript{1,2}

1) Transmission Device Laboratories, Sumitomo Electric Industries, LTD.  
2) OITDA

I, Taya-cho, Sakaeku, Yokohama, 244-8588, Japan  
Tel : +81-45-853-7308, Fax : +81-45-852-2913, E-mail : hashimoto-junichi@sei.co.jp

Abstract
GaInNAs-DFB laser with a high resistive semiconductor current-blocking layer was developed. CW oscillation could be possible up to 110 °C with SMSR > 42dB. 10-Gbps-modulation with a clear eye-opening could be possible between 25 °C and 80 °C.

1. Introduction
Distributed feedback (DFB) lasers with excellent single-mode lasing characteristics are now commonly used in optical networks. However, for the wider use of them, especially in access networks, uncooled operation is essential to reduce the cost and power-consumption drastically. DFB lasers using GaInNAs [1] seem to be most promising for this purpose because carriers (especially electrons) can be confined in the active layer strongly in this material system, leading to much better temperature characteristics than the conventional GaInAsP/InP DFB lasers.

Recently we developed a buried-ridge GaInNAs-DFB laser [2], in which a GaAs grating was formed just above a GaInNAs active region in a current injection region, and was overgrown by an upper cladding layer. In this structure, the center part of a propagating light with high-intensity can be coupled to the grating, so that large coupling coefficients can be obtained easily. Although this structure is common in conventional DFB lasers, we introduced it into the GaInNAs-DFB laser for the first time. Under CW condition, it could oscillate up to 120 °C with SMSR more than 40 dB.

However, this laser had a large parasitic capacitance due to the pn junction in the current-blocking region, and therefore it was not suitable for a high-speed operation. To address this issue, we newly introduced a high resistive semiconductor layer into the current-blocking region to reduce the parasitic capacitance significantly.

In this paper, we report on the first successful operation of a buried-ridge-stripe GaInNAs DFB laser with a high resistive semiconductor current-blocking layer. Good CW single-mode oscillations with SMSR > 42dB were obtained up to 110 °C, and I-L characteristics with good linearity were obtained up to 100 °C. In addition, 10-Gbps-modulation with a clear eye-opening could be possible between 25 °C and 80 °C.

2. Device Structure and Fabrication
A cross-sectional view of the laser is shown in Fig.1. The epitaxial growth was done by low-pressure OMOVPE.
Zn-GaInP cladding layer was successfully overgrown on the GaAs-grating without defects or anomalous growth at the regrowth interface of the GaAs grating region. These merits contribute to a good uniformity and a good reproducibility of the device fabrication.

3. Results and Discussions

The AR(<0.1%)/HR(>90%) coated laser chip having a 300 μm-long cavity and a 2 μm-wide stripe was bonded on a heatsink in a p-side-up configuration. Measurements were performed under CW condition. Fig.2 shows the I-L characteristics. The threshold current and the slope efficiency (SE) at 25 °C were 19.9 mA, and 0.28 W/A, respectively. Oscillation could be possible up to 110 °C, and I-L characteristics with good linearity and small SE change were obtained up to 100 °C. This temperature characteristic seems to be good, considering that no structural optimization was performed. Strong carrier confinement in the GaInNAs active region would contribute to this preferable result.

Fig. 3(a) shows lasing spectra at 2mW. The lasing wavelength of 1237.76 nm at 25 °C was nearly equal to the designed value (1237.7 nm), and it shifted towards the longer wavelength region with the rate of about 0.1 nm/K as shown in Fig. 3(b), which is almost equal to those of the conventional DFB lasers. As also shown in Fig. 3(b), high SMSR values above 42 dB were obtained in all temperatures between 25 °C and 110 °C. This result clearly shows that an excellent single-mode operation was maintained up to high temperature range.

In addition, we performed a direct-modulation experiment using 10-Gbps 2\(^{11}\)-1 NRZ signals. Fig. 4 shows eye-diagrams at 25 °C and 80 °C with extinction ratios (E.R.) around 5 dB. In both temperatures, clear eye-openings were obtained, although relatively large pattern jitters were observed at 80 °C, demonstrating that this laser can be operated at 10Gbps up to high temperature region.

The capacitance reduction due to the high resistive current-blocking layer contributed to this improvement in high-speed operation.

4. Conclusions

First successful operation of the buried-ridge-stripe GaInNAs-DFB laser with a high resistive semiconductor current-blocking layer were realized. Excellent CW single-mode oscillations with SMSR > 42dB were obtained up to 110 °C, and oscillation could be possible up to 110 °C. In addition, 10Gbps-modulations with clear eye-openings were obtained between 25 °C and 85 °C. These are the promising results for the uncooled use of the GaInNAs-DFB laser. Further improvements can be expected by optimizing the laser structure.

5. Acknowledgement

This work was performed under management of the OITDA supported by NEDO.

References
1.55-μm-Wavelength λ/4-Shifted DFB Lasers with High-Density InAsSb Quantum-Dot Active Layers

M. Matsuda¹, K. Kawaguchi¹², A. Uetake¹², H. Kuwatsuka¹², M. Ekawa¹, T. Yamamoto¹, M. Sugawara¹³ and Y. Arakawa¹⁴
¹Fujitsu Laboratories Ltd., ²Fujitsu Limited, 10-1 Morinosato-Wakamiya, Atsugi 243-0197, Japan, Tel: +81-46-250-8249, Fax: +81-46-248-5192, e-mail: matsuda.manabu@jp.fujitsu.com
²QD Laser Inc., 1-1-14-17 Kudan-Kita, Chiyoda-ku, Tokyo 102-0073, Japan
³RCAST, ⁴IIS, ⁵CINQIE, Univ. of Tokyo, Komaba, Meguro-ku, Tokyo 153-8505, Japan

Abstract
1.55-μm-wavelength λ/4-shifted DFB lasers have been demonstrated for the first time, using high-density InAsSb quantum-dot active layers. CW lasing operation up to 100°C with stable single mode oscillation was achieved.

1. Introduction
Quantum dots [1] with emission in the 1.55-μm wavelength range have attracted for optical telecom device applications. Recently, several DFB lasers using quantum dots and quantum dashes as active layers have been reported [2-4]. However, the characteristics are still needed to improve, since there are no DFB lasers with AR-coated facet. Therefore, research and development for quantum dot growth technology as well as investigation as a DFB laser in detail are required.

For improving 1.55-μm-wavelength quantum-dots (QDs), we have investigated InAsSb QDs on InP slightly-tilt 001) substrate grown by MOVPE. We have recently realized high-density quantum-dots of over 1 x 10¹¹ cm⁻² with photoluminescence (PL) spectrum at 1.55-μm wavelength range [5].

In this report, we demonstrate 1.55-μm-wavelength λ/4-shifted DFB lasers with both AR-coated facets for the first time, using the high-density InAsSb QDs as active layers. We achieved CW lasing operation up to 100°C with stable single mode oscillation. We also evaluated the dependence of relaxation oscillation frequency on wavelength detuning.

2. Device Structure and Fabrication
The active layers of the fabricated lasers consist of ten-period-stacked InAsSb-QD sandwiched by GaInAsP layers on a 2°-off n-type InP substrate grown by MOVPE. Figure 1 (a) shows an AFM image of the InAsSb QDs grown on GaInAsP layer. The density of the quantum-dots was as high as 1.1 x 10¹¹ cm⁻². The 1st-order diffraction grating with a λ/4-shift at the center of the laser cavity was located beneath the QD active layers. Figure 1 (b) shows a PL spectrum of the InAsSb QD active layer grown on the epilayers with gratings used for laser fabrication. The full-width half-maximum (FWHM) of the PL spectrum was 92 meV. We have observed little degradation of the PL intensity and FWHM although the QD active layers were grown on the grating structure. The laser had buried heterostructure waveguide with p-n current-blocking layers by using dry etching and MOVPE-regrowth technique.

![AFM image of InAsSb QDs](image1)

![PL spectrum of InAsSb QD active layer](image2)

Figure 1 - (a) 1 x 1 μm² AFM image of InAsSb QDs grown on GaInAsP epitaxial layer, and (b) PL spectrum at 295K of InAsSb QDs used for device fabrication.

Here, we fabricated 600-μm long DFB lasers with the coupling coefficient (k) of the grating of about 50 cm⁻¹. The kL value was about 3. To investigate the influence of wavelength detuning, we prepared ten different grating periods for ten different lasing wavelengths from 1500 nm to 1590 nm by 10 nm in one wafer using electron-beam lithography. Both facets of the lasers were coated with anti-reflection films.
3. Device Characteristics

Figure 2 shows the light-current characteristics of the fabricated DFB laser. CW lasing operation up to 100°C was achieved. The threshold currents were 16.7 and 43.2 mA at 25°C and 85°C, respectively. The slope efficiency at 25°C was 0.072 mW/mA. We observed stable single longitudinal mode operation up to 100°C in keeping with a side-mode suppression ratio more than 45 dB. Figure 3 shows the lasing spectra of this laser at 25°C and 85°C, with a drive current of 100 mA.

![Figure 2 - Light output power against drive current characteristics of a 600-μm-long λ/4-shifted DFB laser.](image)

![Figure 3 - Lasing spectra of the DFB laser at 25°C and 85°C with drive current of 100 mA.](image)

We also evaluated the dependence of the relaxation oscillation frequency on wavelength detuning by measuring relative intensity noise (RIN) spectra of these lasers. Here, the detuning was evaluated from the measured spectra at threshold current (Ith). Figure 4 shows the relaxation oscillation frequencies at the drive current of Ith + 50 mA and threshold currents at 25°C of ten lasers with different detuning. We found that both the relaxation oscillation frequency and the threshold current increased with decrease of wavelength detuning. This revealed that the detuning set negative without excess threshold current increase was one of the important ways to improve the characteristics of quantum-dot-based DFB lasers, as well as quantum-well-based ones.

![Figure 4 - Dependence of f_r and I_th on lasing wavelength detuning at 25°C.](image)

4. Conclusion

We have demonstrated 1.55-μm-wavelength λ/4-shifted DFB lasers with both AR-coated facets, using high-density (1.1 x 10^11 cm^-2) InAsSb quantum-dot active layers. We achieved CW lasing operation up to 100°C in keeping with stable single longitudinal mode oscillation. We have also evaluated the dependence of relaxation-oscillation frequency on wavelength detuning and have found that the detuning set negative is also important to improve the characteristics of QD DFB lasers.

Acknowledgment

This work was supported by Ministry of Internal Affairs and Communications as a part of the R&D of High-Performance Network Subsystems Using Nano-technology Project and the Special Coordination Funds for Promoting Science and Technology.

References

Fabrication and Optical Properties of Hybrid-type Pillar Microcavity

T. Yamaguchi (1, 2), T. Tawara (1), H. Gotoh (1), H. Kamada (1), H. Okamoto (3), H. Nakano (1), O. Mikami (2)

1. NTT Basic Research Laboratories, 3-1 Morinosato-Wakamiya, Atsugi, Kanagawa 243-0198, Japan, Tel: 046-240-4106, Fax: 046-270-2342, E-mail: t-yama@will.brl.ntt.co.jp
2. Tokai University, 3. NTT Photonics Laboratories

Abstract — We studied the fabrication and optical properties of a hybrid-type microcavity that consists of dielectric and semiconductor distributed Bragg reflectors. Moreover, we observed discrete emission lines from quantum dots coupled to the cavity mode of a pillar microcavity.

I. INTRODUCTION

A quantum dot (QD) in a pillar microcavity is expected to provide a highly efficient single photon emitter for quantum cryptography [1]. Single photons in the telecom band are required for optical fiber facility applications.

A semiconductor microcavity in the telecom band will have a total thickness of about 12 μm. This causes difficulties as regards crystal growth and processing. Furthermore, the optical property of a QD will deteriorate as a result of the higher temperature growth of a distributed Bragg reflector (DBR) on top of the cavity layer.

Our approach involves using a hybrid-type microcavity consisting of a dielectric top DBR and a semiconductor bottom DBR. This approach can reduce the total thickness to about 9 μm, because the refractive index difference of a dielectric DBR is high. A dielectric top DBR can also avoid any QD deterioration caused by crystal growth at a higher temperature.

In this paper, we describe the fabrication of a hybrid-type pillar microcavity and our observation of the optical properties of a small number of QDs coupled to the cavity mode of the microcavity.

II. FABRICATION

Our sample is a hybrid-type microcavity consisting of a SiO\textsubscript{2}/TiO\textsubscript{2} top DBR, InAs QDs, a GaAs \(\lambda\) cavity and a GaAs/AlGaAs bottom DBR in the 1.3 μm band [2]. The SiO\textsubscript{2}/TiO\textsubscript{2} DBR was etched by using reactive ion etching (RIE) with CF\textsubscript{4} gas. Semiconductor layers were etched by using inductively coupled plasma (ICP) with a Cl\textsubscript{2}-based reactive gas. The etching patterns were ranged in area from 0.5×0.5 to 2×2 μm\textsuperscript{2} with Ni masks. TiO\textsubscript{2} layers are generally etched isotropically by employing RIE on the SiO\textsubscript{2}/TiO\textsubscript{2} DBR with CF\textsubscript{4} because of the effect of radicals that react with the TiO\textsubscript{2} layers. Therefore, we reduced the radicals to suppress side etching. When etching semiconductor layers using ICP with a Cl\textsubscript{2}-based reactive gas, under-cut can easily occur. One solution involves adding Si to the plasma for vertical etching [3]. When the RIE and ICP etching conditions were balanced, we achieved the vertical etching of a hybrid-type microcavity with a high aspect ratio (Fig. 1).

III. OPTICAL PROPERTIES

We evaluated the optical properties using photoluminescence (PL) measurements. We measured the QD wafer, hybrid-type planar (before etching)
microcavity and pillar microcavity. Figure 2a) shows the PL spectrum of a QD wafer excited with a CW Ar+ laser (488 nm) at 4 K. The laser power was 400 nW. The PL peak is close to 1260 nm with a full width at half maximum (FWHM) of 25 nm. We designed the cavity mode to be 1260 nm. Reflection and PL spectra from the planar hybrid-type microcavity are shown in Fig. 2b). Cavity modes were observed close to 1260 nm with an FWHM of 2 nm. The difference between the wavelengths of the cavity mode and the PL peak from the planar cavity is caused by fluctuations in film thickness.

![Figure 2](image)

Figure 2  a) PL spectrum from a QD wafer. b) Reflection spectrum (upper curve) and PL spectrum (lower curve) from a planar hybrid-type microcavity.

Figure 3a) shows the PL spectrum from a pillar microcavity with a 2×2 μm² pattern. The laser power is 7 μW. The PL peak is 1240 nm with an FWHM of 2 nm. The PL peak position is shifted to a shorter wavelength because the effective refractive index of the cavity layer was reduced by light leaking into the air. The cavity mode wavelength is proportional to the effective refractive index. A comparison of Fig. 2b) and Fig. 3a) show that the FWHMs are almost the same. Therefore, we used the PL spectrum to evaluate the quality factor (Q-factor) of the cavity mode, which is 620. The theoretical value is 1300, which we calculated by the transfer matrix method at 1.3 μm. This difference is caused by the fluctuation in film thickness and reabsorption by uncoupled QDs.

The PL spectrum obtained from a 1×1 μm² pattern pillar microcavity is shown in Fig. 3b). A discrete spectrum was observed because of the reduction in the number of QDs as a result of the smaller area. Each sharp line corresponds to the emission from a single QD. Figure 3b) shows that the quality of the QDs is retained after the deposition of the SiO$_2$/TiO$_2$ DBR and the dry etching process. The PL peak position is near the cavity mode of 2×2 μm². This means the PL peak is coupling with the cavity mode of the pillar microcavity. This result reveals the possibility of coupling a single QD, and we can expect to realize a single photon emitter in the telecom band.

![Figure 3](image)

Figure 3 PL spectra from a pillar microcavity. The etching patterns are a) 2×2 μm² and b) 1×1 μm².

**Conclusion**

We studied the fabrication conditions and optical properties of a hybrid-type pillar microcavity for a single photon emitter in the 1.3 μm band.

Radical reduction is effective when etching a SiO$_2$/TiO$_2$ DBR by RIE with CF$_4$. It is effective to add Si to the plasma when etching a semiconductor layer by using ICP with a Cl$_2$-based reactive gas. By optimizing the RIE and ICP etching conditions, we were able to fabricate a vertical hybrid-type pillar microcavity with a high aspect ratio.

With the PL measurement of a pillar microcavity, since the etching pattern is smaller, we can observe discrete emission lines from QDs coupled with the cavity mode. The quality of the QDs is maintained after processing. If we increase the Q-factor by increasing the number of pairs of DBRs and design a cavity taking the effective refractive index into consideration, we can expect to realize a single photon emitter in the telecom band.

**References**

Differential gain in InGaAsN quantum well structures

M.S. Wartak
Senior Member IEEE
Department of Physics and Computer Science
Wilfrid Laurier University
Waterloo, Ontario N2L3C5, Canada
Email: mwartak@wlu.ca
Telephone: (519) 884-1970, ext.2436
Fax: (519) 746-0677

P. Weetman
Department of Physics and Computer Science
Wilfrid Laurier University
Waterloo, Ontario N2L3C5, Canada
Email: pweetman@eml.cc
Telephone: (519) 884-1970, ext.2685
Fax: (519) 746-0677

Abstract—Numerical studies of differential gain in InGaAsN quantum well systems are reported. Our approach is based on a $10 \times 10$ Hamiltonian solved self-consistently with the Poisson’s equation. It was found that the properties of that system can be effectively modified.

The differential gain is one of basic parameters which characterize operation of quantum well (QW) based semiconductor lasers [1]. It is defined as the derivative of the optical material gain with respect to the injected carrier density. It is linked with the modulation bandwidth of these devices. The differential gain is often determined from measurements of the relaxation oscillation frequency as a function of the emitted optical power under the approximation of small signal modulation [2], [3]. The differential gain is an intrinsic property of the material in the active layer and is independent of the laser structure.

In recent years we have witnessed tremendous progress in the research on a new class of materials based on nitride semiconductors [4], [5]. Dilute Nitride lasers based on GaAs are considered as replacements for InP based lasers in metropolitan and local area networks. It has been found that replacing a small amount of the group V element by nitrogen in a III-V material systems reduces the energy gap. This reduction significantly changes band structure and offers new possibilities of improving optoelectronic properties of devices based on those materials. For example, impressive improvements of in-plane lasers [6] and [7] as well a VCSELs [8] based on those materials have been reported.

Differential gain of quantum well nitride structures have been studied recently [9]. It was found that differential gain with respect to either current or carrier concentration is reduced in dilute-nitride devices. Differential gain among other factors was recently analyzed theoretically by Alexandropoulos and Adams [10] in the BAC model. It was found that an increase of Nitrogen content reduces the peak differential gain. We have extended their analysis to $10 \times 10$ Hamiltonian and also included electrostatic effects on the heterostructure potentials of electrons and holes.

Our approach is based on a $10 \times 10$ k-p Hamiltonian [11] which is a generalization of the $8 \times 8$ Lüttinger-Kohn Hamiltonian [12] that accounts for coupling between the conduction and hole bands. Substitution of nitrogen splits the conduction band to create the additional ‘nitrogen band’. This new system necessitates the introduction of an additional band in the description of the electron-hole bandstructure. For this purpose, a $10 \times 10$ Hamiltonian which accounts for nitrogen, conduction, light- and heavy-holes and spin bands is used here.

We have recently applied $10 \times 10$ Hamiltonian [13] to numerically analyze the effective masses in InGaAsN quantum-well structures with self-consistent effects. In that paper we were able to obtain detailed band structures for various well parameters such as nitrogen compositions. We use this information in the material gain computations.

The structure simulated is similar to that used previously [13] and consists of undoped $In_{0.36}Ga_{0.64}As_{1-y}N_y$ system. The Nitrogen composition $(y)$ changes in the range $0.00 < y < 0.05$. We have considered several values of well widths.

We performed the following steps of our computations:
1. produce gain spectrum at a given temperature
2. determine differential gain at peak value
3. increase temperature and repeat process
4. plot differential gain such determined vs temperature.

Below, we report on some representative results.

In Fig.1 we plotted differential gain vs energy for two values of barrier composition and three values of well widths. One can observe that the effect of Nitrogen content in the barrier is relatively small for all well widths considered. One can also observe that by varying well width, we can significantly change maximum value of differential gain. Its peak value stays roughly the same at a particular energy. There is however optimum value of well width for which differential gain reaches maximum. In Fig.2 we plotted differential gain as a function of Nitrogen composition for four values of well width. As was noticed with data shown in Fig.1, for a given Nitrogen composition, that the value of differential gain is the largest for 6 nm well width. For all values of...
well widths, the differential gain shows little dependence on Nitrogen content although its dependence for narrower wells is stronger (and resemblances linear for the values considered).

The temperature dependence of differential gain is shown in Fig. 3 for several values of well widths ranging from 6 nm to 9 nm. Nitrogen composition is 1.5% and carrier density is equal to \(4 \times 10^{18} \text{cm}^{-3}\). It is roughly independent of temperature, although some degradation is observed at larger temperatures.

In conclusion, we have analyzed differential gain for some range of parameters important in the design of semiconductor lasers based on single quantum well using InGaAsN material system. The effects of self-consistency were considered for all our computations since it is known that they can significantly affect properties of quantum well systems. This is due to an increase of the density of conduction electrons and simultaneously the reduction of the hole densities in the active region [14] which modifies the amplitude of the differential gain and the change in relative transition strengths which modifies the spectral dependence. These effects have a large practical importance in the design of an efficient laser as they can reduce the differential gain by significant amounts.

We would like to acknowledge the support from the Natural Science and Engineering Research Council of Canada (NSERC). This work was made possible by the facilities of the Shared Hierarchical Academic Research Computing Network (SHARCNET: www.sharcnet.ca).

REFERENCES

Abstract We have developed a compact and high speed InP Mach-Zehnder modulator by employing a novel n-i-n structure. We have also developed a compact wavelength tunable transmitter by integrating a wavelength tunable laser array and the n-i-n InP Mach-Zehnder modulator.

Introduction
The recent increase in data traffic on the Internet has led to the need for large capacity and highly functional photonic networks. High-speed optical modulators are key components for systems where high-speed optical signals must be transmitted through long-haul optical fibers. We devised a novel waveguide structure as an alternative way of fabricating a compact Mach-Zehnder (MZ) modulator with a low driving voltage [1,2]. Our modulator has an n-i-n isotype heterostructure designed to eliminate the electrical and optical signal losses caused by the p-type cladding layer. We obtained a half-wavelength voltage (\(V_\pi\)) of 2.2 V with a 3-mm-long phase-shifting region. And we confirmed 40-Gbit/s operation with a push-pull driving voltage of 1.3 V_{pp}.

Small size transponders, which include wavelength tunable lasers and small size external modulators, are indispensable to reducing the cost and size of ROADM systems [3,4]. We also report a compact 10 Gbit/s tunable laser module that we realized by the hybrid integration of a wavelength tunable DFB laser array (TLA) and an n-i-n MZ modulator [5]. It can operate at 80 channels in the full C-band wavelength region. The modulator characteristics have little wavelength dependence. As a result, fixed operation of the MZ modulator driving condition can be achieved even when the output wavelength is switched.

In this paper we summarize our recent activities on the InP based MZ modulator.

40 Gbit/s Mach-Zehnder modulator
A semiconductor Mach-Zehnder modulator consists of two n-i-n phase-shift waveguides. The n-i-n structure offers 50-Ω impedance matched signal lines and velocity matching for electrical and optical signals. These features enable us to realize high-speed and small size MZ modulators. The device chip size is 4.5x0.8 mm. The 6-dB bandwidth of electrical S\(_{21}\) is over 40 GHz and that of S\(_{11}\) is less than -15 dB at frequencies up to 50 GHz. The fabricated modulator chip was installed in a compact package as shown in Fig. 1. Two RF-input V-connectors for dual driving are placed on one side for ease of connection with the differential output modulator driver. The package is 21x17 mm in size and the footprint size is the same as that of conventional DFB laser modules. Figure 2 shows the eye diagram obtained for the operation of a non-return-to-zero (NRZ) 2\(^{31}\)-1 pseudo random bit sequence at 40 Gbit/s. A 1.3-V_{pp} differential driving voltage was supplied to the dual ports of the modulator.

Wavelength tunable transmitter
The full C-band tunable laser source that we employed is our wavelength tunable DFB laser array (TLA) [6]. The TLA is an attractive candidate for practical use because the main structure is based on the conventional DFB laser, which is widely used in existing optical network systems. It can provide one
wavelength from a wide wavelength tuning range when we select one DFB laser and control its temperature. The characteristics of these DFB lasers do not exhibit any abrupt change, such as mode hopping. This is an important feature for a working optical communication network system.

Figure 3 shows a photograph of the wavelength tunable transmitter module. It consists of a TLA, an n-i-n MZ modulator and a wavelength locker. The TLA and the MZ modulator are co-packaged in a compact module, which is 41 mm (L) x 13 mm (W) x 9 mm (H) in size. The TLA and MZ modulator are mounted on the same metal carrier and thermo-electric cooler (TEC). These two devices are coupled by using lenses via an isolator. The output from the MZ modulator is coupled into a fiber pigtail after passing through the wavelength locker mounted on another TEC. All of the lenses are welded with a YAG laser, which secures long-term stability.

The inset in Fig. 4 shows the 10-Gbit/s eye diagrams we obtained back-to-back and after transmission through 100 km of single-mode-fiber (SMF) in the 1530 to 1560 nm wavelength range. For this measurement, we fixed the driving and bias voltages of the MZ modulator at constant values even when the output wavelength was switched. The bias voltages were kept at -3.3 and -2.8 V for the two arms of the modulator. The driving voltages were fixed at 0.75 and 2.25 V for the push-pull operation. The modulation signal has a negative chirp under this driving condition. Clear eye opening can be obtained from 1530 to 1560 nm without changing the driving condition of the MZ modulator under a back-to-back condition and even after a 100-km SMF transmission.

The bit error rate (BER) performance measured back-to-back and after 100 km SMF transmission is also shown in Fig. 4. The power penalties after transmission through a 100-km SMF were less than 3 dB for all wavelength channels.

Figure 3. Photograph of a wavelength tunable laser module integrated with an n-i-n MZ modulator.

Figure 4. BER performance at various wavelengths. Insets are back-to-back and 100 km SMF transmission eye patterns for various output wavelength signals.

Conclusion
We developed a compact push-pull drive n-i-n MZ modulator, and confirmed 40-Gbit/s operation with a push-pull driving voltage of 1.3 Vpp. We also fabricated a compact wavelength tunable transmitter by the hybrid integration of a TLA and an n-i-n MZ modulator in a small package. The low wavelength and temperature dependent characteristics of the modulator enabled it to be operated with a fixed driving condition even when the output wavelength was switched over the wide wavelength range of the C-band. A 100-km SMF transmission was also demonstrated with the module, and a low power penalty of less than 3 dB was confirmed for the entire output wavelength range of 1530 to 1560 nm.

References
Multi-carrier Light Generator Using Phase Modulators and Chirped Fiber Bragg Grating

Takashi Yamamoto*, Tetsuro Komukai, Kazunori Suzuki, and Atsushi Takada

*NTT Access Network Service Systems Laboratories
1-7-1, Hanabatake, Tsukuba, Ibaraki, 305-0805 Japan
Tel. +81 29 868 6114, Fax +81 29 868 6440
Email: yamamoto.takashi@anl.ntt.co.jp

NTT Network Innovation Laboratories
1-1, Hikari-no-oka, Yokosuka, Kanagawa, 239-0847 Japan

Abstract — This paper describes a multi-carrier light generation technique that utilizes a CW light source, two phase modulators and a chirped fiber Bragg grating (FBG). By adopting a chirped FBG as a dispersion medium instead of a long normal dispersion fiber, we can increase the stability of the optical output spectrum and reduce the size of the multi-carrier light generator. We have built a prototype of the multi-carrier light generator with this configuration. The prototype apply a 25 GHz sinusoidal phase modulation with a modulation index of $\pi/4$, and a dispersion with $D = 98$ ps/nm to a CW light followed by a 25 GHz sinusoidal phase modulation with a modulation index of 9.7$\pi$. We have obtained 61 carrier-light with the power deviation of less than 8 dB.

A multi-carrier light generator, which emits many optical line spectra arranged with an equal frequency interval, is expected to play an important role to realize flexible and robust photonic networks. Several configurations have been proposed for a multi-carrier light generator; a phase modulator in an optical resonator [1], the combination of an optical pulse generator and a nonlinear fiber [2], and the combination of a CW light source, a phase modulator, and a Mach-Zehnder intensity modulator [3,4]. We have proposed a novel configuration for a multi-carrier light generator, where a CW light source, two phase modulators, and a dispersion medium are used [5]. A 5-km long normal dispersion fiber was used as the dispersion medium and was inserted between the two phase modulators. In this configuration, it is important to suppress the timing change between the two phase modulations to stabilize the whole optical spectrum of multi-carrier light. To avoid the timing change, it is necessary to shorten the fiber length between the two phase modulators and suppress the effect of temperature change. In this paper, we describe a multi-carrier light generator in which a chirped FBG is used as a dispersion medium instead of the long normal dispersion fiber. By adopting the chirped FBG, we can increase the stability of the optical output spectrum, and reduce the size of the multi-carrier light generator.

Fig. 1 shows a block diagram of our proposed multi-carrier light generator. A CW light source, a 1st phase modulator, a chirped FBG, and a 2nd phase modulator are connected serially. First, a sinusoidal phase modulation with a frequency of $f_m$ and a modulation index of $\pi/4$ is applied to a CW light. Then the chirped light is coupled into the chirped FBG whose group velocity dispersion (GVD) is set at $\pm 1/(4\pi f_m^2)$. The output of the chirped FBG is coupled into the 2nd phase modulator, and a sinusoidal phase modulation with a large modulation index is applied. The calculated optical spectrum at the 2nd modulator output for a modulation index of $10 \pi$ is also shown in Fig. 1. A multi-carrier light with an equal frequency interval of $f_m$ is generated. The number of the carrier light is nearly proportional to the modulation index of the 2nd phase modulation. The spectral flatness is preserved for any modulation index of the 2nd phase modulation.

Fig. 1 Block diagram and calculated output spectrum of the proposed multi-carrier light generator.
We have built a prototype of the multi-carrier light generator with this configuration. Fig. 2 shows the appearance of the prototype. The prototype contains phase modulators, a chirped FBG, an EDFA, and a 25/50 GHz interleaver. By injecting a CW light ($\lambda = 1585.8$ nm) and a 25 GHz sinusoidal RF signal into the prototype, multi-carrier light with 50 GHz interval is obtained.

Fig. 3 shows the delay characteristic of the chirped FBG used in the prototype. The delay linearly changes with the wavelength and the dispersion at the CW light wavelength is $D = 98$ ps/nm.

Fig. 4 shows the optical spectrum and the relative intensity noise (RIN) characteristics of each carrier light measured at the 2nd phase modulator output. We can see that 61 carrier-light with the power deviation of less than 8 dB was obtained. By adopting a chirped FBG in place of a long normal dispersion fiber, the fiber length between the two phase modulators was reduced to less than $10^7$ and the power stability of each carrier light was significantly improved. The RIN of each carrier light was less than -140 dB/Hz.

Acknowledgements
The authors thank Y. Hibino and K. Hagimoto for their constant encouragement. Part of this work was supported by the National Institute of Information and Communications Technology (NICT).

References
**LiTaO$_3$ Electro-Optic Polarization Modulator Utilizing Periodically Poled Structure**

Hiroshi Murata, Asuka Takahashi, Yasuyuki Okamura

*Osaka University, Graduate School of Engineering Science*
1-3 Machikaneyama, Toyonaka, Osaka 560-8531 Japan

Tel.: +81-6-6850-6306, Fax.: +81-6-6850-6341, E-mail: murata@ee.es.osaka-u.ac.jp

**Abstract**—New electro-optic polarization modulators utilizing periodically poled z-cut LiTaO$_3$ were demonstrated experimentally for the first time as far as the authors know. They lead to high-speed (>10GHz) polarization control devices operating in wide wavelength ranges.

**I. INTRODUCTION**

A high-speed optical polarization modulator/converter is an important device in many opto-/quantum electronic systems. Several electro-optic (EO) optical polarization modulators/converters have been proposed and implemented [1]-[3]. LiNbO$_3$ and LiTaO$_3$ have a large Pockels coefficient of $r_{42}$ applicable for polarization modulation/conversion in high frequency ranges. A key point for an efficient polarization control is the phase matching between two orthogonally polarized modes, and several specific comb-like electrode structures were adopted for the phase matching [1], [3]. However, these electrodes with a complicated structure are not suitable for the operation in microwave frequency ranges. Another approach using a Pockels coefficient of $r_{61}$ ($r_{22}$) is applicable with no comb-like electrode structure, but its Pockels coefficient value is about 1/10 of $r_{42}$.

In this report, a new optical polarization modulator using a periodically poled structure is proposed, and its operation is demonstrated experimentally. Utilizing the quasi-phase matching technique with a periodically poled structure, an efficient polarization control is possible by using standard coplanar electrodes with no comb-like structure. This modulator should lead to novel high-speed polarization control devices.

**II. EO POLARIZATION MODULATOR WITH A PERIODICALLY POLED STRUCTURE**

Figure 1 shows the structure of the EO polarization modulator we have proposed. It consists of a single-mode channel waveguide and coplanar electrodes fabricated on a z-cut LiTaO$_3$ crystal substrate. A light propagation direction is set to the x-axis of the LiTaO$_3$ crystal and a modulation electric field is applied along to the y-axis as shown in the Fig. 1 (b). With this configuration, coupling between TE and TM modes is induced through a Pockels coefficient of $r_{42}$. Compared with the z-axis propagating polarization modulator by using a Pockels coefficient of $r_{61}$ ($r_{22}$), a lower operational voltage is expected since $r_{42}$ is about 10 times larger than $r_{61}$. A periodically poled structure is designed for the quasi-phase matching between the TE and TM modes.

Figure 2 shows the calculated wavelength dependences of the x-propagating polarization modulators with periodic poling by using z-cut LiTaO$_3$ or z-cut LiNbO$_3$, where we assumed an operational wavelength of ~630nm and an electrode length of 10mm. LiNbO$_3$ has a large material birefringence ($|n_o - n_e| \approx 0.086$), therefore, the poling period required for the quasi-phase matching is $2L \approx 7\mu m$ for the wavelength of ~630nm. On the other hand, LiTaO$_3$ has a small birefringence ($|n_o - n_e| \approx 0.004$), and the poling period for the quasi-phase matching is $2L \sim 180\mu m$ for ~630nm. Therefore, the wavelength bandwidth in the LiTaO$_3$ device is over 20 times wider compared to LiNbO$_3$.

![Fig. 1. Basic structure of the EO polarization modulator with periodically poled structure](image1)

![Fig. 2. Calculated wavelength dependence of EO polarization modulators with periodic poling.](image2)
III. DEVICE FABRICATION AND EXPERIMENT

In order to construct the proposed polarization modulator, it is necessary to fabricate an optical waveguide supporting TE and TM modes with the periodically poled structure in a z-cut LiTaO$_3$ substrate. The APE method is applicable for waveguide fabrication in a periodically poled LiTaO$_3$ substrate [4], however, the APE waveguide supports a single polarization mode only. Ti diffused waveguides are widely used for LiNbO$_3$ devices, however, the high temperature diffusion process over the Curie point of LiTaO$_3$ (~600 degrees Centigrade) is necessary. Therefore, we adopted the Ni diffusion method with a relatively low diffusion temperature of ~580 degrees Centigrade [5].

Firstly, the periodically poled structure was fabricated into a z-cut LiTaO$_3$ crystal by using the pulse voltage applying method. The poling period, $2L$, was set to ~180$\mu$m for the phase matching between the TE and TM modes at ~630nm. Next, a ~20mm long z-propagating optical waveguide was fabricated by thermal diffusion (70h, 580 degrees Centigrade) of a 35nm thick Ni stripe into the surface of the periodically poled LiTaO$_3$ crystal. The Ni width of 5$\mu$m was determined so as to be a single-mode both for TE and TM modes around the operation wavelength of ~630nm. Finally, 10mm long Al coplanar electrodes were formed on the surface of the LiTaO$_3$. The spacing of the two electrodes was set at 10$\mu$m.

The polarization modulation operation of the fabricated device was tested. Figure 3 (a) shows the output intensity change from the device through an analyzer plate when a sinusoidal voltage of 400Hz was applied to the electrodes. The polarization change according to the applied voltage was clearly observed. The required voltage for the complete TE-TM conversion was about 35V for the device with the periodic polarization reversal of $2L$=188$\mu$m, which was several times larger than the calculated voltage. We believe that it would be reduced by using optimized design and fabrication conditions. Figure 3 (b) shows the measured TE-TM conversion efficiency for the lightwave wavelength. The wavelength bandwidth was in good agreement with the calculated value.

IV. CONCLUSIONS

The optical polarization modulator utilizing periodically poled LiTaO$_3$ was proposed and its basic operation was demonstrated. By adopting a travelling-wave electrode structure to the proposed device, a fast polarization modulator operated in quasi-millimeter-wave/millimeter-wave ranges can be obtained.

ACKNOWLEDGEMENT

This work was supported in part by Grants-in-Aid for Scientific Research from the Ministry of Education, Science, Sports and Culture, Japan.

REFERENCES

An 80-GHz carrier-suppressed optical pulse generator using two cascaded phase modulators followed by a delay-interferometer

Guo-Wei Lu, Tetsuya Miyazaki
National Institute of Information and Communications Technology (NICT)
4-1-1, Nukui-Kitamachi, Koganei, Tokyo 184-8795, Japan, Tel: +81-42-327-5439; Fax: +81-42-327-7035; Email: gwlu@nict.go.jp

Abstract
We propose and experimentally demonstrate a cost-effective 80-GHz carrier-suppressed optical pulse generator. The proposed optical pulse generator uses two cascaded time-interleaved phase modulators driven by only 20-GHz clock signals, and a delay-interferometer following the modulators.

1 Introduction
In high-speed optical transmission systems, return-to-zero (RZ) and carrier-suppressed return-to-zero (CSRZ) optical pulses are widely employed for both on-off keying (OOK) and differential phase-shift keying (DPSK) systems. Optical pulses are usually generated by a half-rate clock driven Mach-Zehnder modulator [1, 2] or a half-rate clock driven phase modulator followed by delay interferometer [3, 4]. Compared with the actively mode-locked laser, these approaches are cost-effective and flexible in configuration. However, for the clock-driven modulation based schemes, the achievable repetition rate of pulse source is limited by the available electronic devices.

In this paper, we propose and experimentally demonstrate a cost-effective CSRZ optical pulse generator using two cascaded phase modulators followed by a delay-interferometer. The cascaded phase modulators are driven by a clock with only quarter of the pulse repetition rate. Instead of using a 40-GHz clock driven phase modulator, a cost-effective 80-GHz CSRZ pulse generator is successfully demonstrated by driving the phase modulators with clock signals with repetition rate of 20 GHz, which is only quarter of the pulse repetition rate.

2 Operation Principle
As an example, the operation principle is illustrated in Fig. 1 and Fig. 2 for an 80-GHz CSRZ pulse generator. The proposed CSRZ pulse generator consists of a CW laser source and two cascaded phase modulators followed by a Mach-Zehnder delay-interferometer (MZDI). In [3], an optical CSRZ and chirp-free pulse has been demonstrated using a half-rate clock driven phase modulator followed by a MZDI. In our proposed pulse generator scheme, the repetition rate of optical pulse source can be further doubled by phase interleaving technology [5]. As shown in Fig. 1, the phase of the optical light from laser is modulated by the two cascade phase modulators driven by RF clock signals with a modulation depth of \( \pi \) and modulation frequency of \( f = \frac{1}{4T} \), where \( T \) is the desired pulse period. Two cascaded phase modulations are offset in time with relative delay of \( \Delta T = T \). After the cascaded modulators with a relative time delay between them, these two phase modulations are interleaved in time domain, thus resulting in the doubled phase modulation speed to the half of the pulse repetition rate. For example, for a CSRZ optical pulse with repetition rate of 80 GHz, \( T \) is equal to 12.5 ps and \( f = 20 \) GHz. As shown in Fig. 2, a 40-GHz phase modulation is successfully achieved using two serially-cascaded phase modulators driven by only 20-GHz clock signals with relative 12.5-ps time delay, which is equivalent to a 40-GHz clock driven phase modulation. However, as the two phase modulators are driven by 20-GHz clock signals only, the implementation cost is effectively reduced. After the 40-GHz-clock phase modulation introduced by the phase interleaving of two cascaded modulations, the periodical phase modulation is converted into intensity modulation of the light at the destructive output of a MZDI with FSR \( \geq 1/T \). According to the analysis in [3], a CSRZ pulse train is successfully generated at the destructive port of MZDI. The pulse-width of generated CSRZ pulse train depends on FSR of MZDI and the rate of the phase modulation.

![Fig.1. Schematic diagram of the proposed 80-GHz CSRZ pulse generator, PM: phase modulator, MZDI: Mach-Zehnder delay interferometer.](image-url)
modulators. Each of the phase modulator is driven only by a 20-GHz RF clock signal. However, in this case, there is a high response requirement at the raising and falling edges for the cascaded phase modulators.

![Diagram of phase modulations](a) first and (b) second phase modulators, as well as (c) the resulted phase modulation by the cascaded phase modulators.

### 3 Experiment and Results

![Measured intensities](a) 40-GHz CSRZ optical pulses using one phase modulator followed by MZDI and (b) 80-GHz CSRZ optical pulses using proposed two cascaded modulators followed by MZDI (10ps/div).

![Measured spectrum](Optical spectrum of generated 80-GHz CSRZ optical pulse.

An experiment for an 80-GHz CSRZ pulse generator is carried out to verify the proposed scheme. The experiment setup is similar to what is illustrated in Fig.1. Light from a CW laser is phase modulated by two cascaded phase modulators driven by 20-GHz clock signals with modulation depth of \( \pi \). The 3-dB bandwidth of the employed phase modulators is only 20 GHz. A 12.5-ps relative time delay between the two cascaded phase modulations is introduced by tuning the RF delay line. Thus, after the phase interleaving, the light is phase-modulated with the repetition rate of 40 GHz. Through the MZDI with FSR of 164 GHz, a CSRZ pulse train with repetition rate of 80 GHz is generated when destructive interference is present at the output of MZDI.

The generated optical pulse at the destructive port of MZDI is measured using a high-speed optical sampling oscilloscope. The measured intensity of the generated 80-GHz CSRZ pulse train is shown in Fig. 3 (b). The slightly un-uniform pulse shape is attributed to the performance difference of employed phase modulators. For comparison, Fig. 3 also illustrates a generated 40-GHz CSRZ optical pulse using only one 20-GHz clock driven phase modulator followed by a MZDI.

An experiment for an 80-GHz CSRZ pulse generator is carried out to verify the proposed scheme. The experiment setup is similar to what is illustrated in Fig.1. Light from a CW laser is phase modulated by two cascaded phase modulators driven by 20-GHz clock signals with modulation depth of \( \pi \). The 3-dB bandwidth of the employed phase modulators is only 20 GHz. A 12.5-ps relative time delay between the two cascaded phase modulations is introduced by tuning the RF delay line. Thus, after the phase interleaving, the light is phase-modulated with the repetition rate of 40 GHz. Through the MZDI with FSR of 164 GHz, a CSRZ pulse train with repetition rate of 80 GHz is generated when destructive interference is present at the output of MZDI.

The generated optical pulse at the destructive port of MZDI is measured using a high-speed optical sampling oscilloscope. The measured intensity of the generated 80-GHz CSRZ pulse train is shown in Fig. 3 (b). The slightly un-uniform pulse shape is attributed to the performance difference of employed phase modulators. For comparison, Fig. 3 also illustrates a generated 40-GHz CSRZ optical pulse using only one 20-GHz clock driven phase modulator followed by a MZDI. Measured by an autocorrelation, the pulse-width of the generated 80-GHz optical pulse is around 7.2 ps. The corresponding optical spectrum of the generated 80-GHz optical pulses is measured and shown in Fig. 4. The power of carrier tone is suppressed by more than 25 dB, compared with that of clock tones. Some undesired clock tones are observed in Fig. 4, which are mainly due to the finite extinction ratio of the MZDI in the experiment. Better performance is expected by using fast response phase modulators and a MZDI with high extinction ratio.

### 4 Conclusion

In this paper, we proposed and experimentally demonstrated a cost-effective 80-GHz CSRZ optical pulse generator. Rather than a costly 40-GHz-clock driven phase modulator, two cascaded time-interleaved phase modulators, which are driven by 20-GHz clock, are employed to introduce 40-GHz phase modulation. The proposed scheme is scalable to high repetition rate optical pulse source using two or more cascaded phase modulators driven by low-speed RF clock signals.

### Reference


Optical Frequency Comb Generation Using Semiconductor Optical Modulators
Fumiaki Ono, Azusa Komoro, Hiroyuki Tsuda, Hiroaki Takeuchi†, and Takashi Kurokawa††
Keio University, 3-14-1 Hiyoshi, Kouhoku-ku, Yokohama, 223-8522 Japan
Phone: +81-45-563-1151 Facsimile: +81-45-566-1529 E-mail: fumiaki@isd.elec.keio.ac.jp
†NTT Photonics Laboratory, 3-1 Morinosato-Wakamiya, Atsugi, Kanagawa, Japan
††Tokyo University of Agriculture and Technology, 2-24-16 Naka-machi, Koganei-shi, Tokyo, Japan

Abstract
We propose the optical frequency comb generation using semiconductor optical modulators. The optical frequency comb spectra were simulated using the modulation characteristics of the fabricated multiple quantum well (MQW) optical modulators.

1. Introduction
An optical frequency comb is suitable for multi-wavelength light source, an optical frequency measurement and so on. It can be generated in many ways, one of which uses LiNbO3 (LN) modulators [1], [2]. However, the size of the LN modulator is large and cannot be integrated with other functional components. In this paper, we have proposed the optical frequency comb generator using semiconductor modulators, and simulated its spectra with the measured parameters of the multiple quantum well (MQW) semiconductor optical modulators.

2. Optical frequency comb generator configuration

\[ V_{\text{fe}} (= V_a \sin(\omega t)) \]
\[ V_{\text{fb}} (= V_b \sin(\omega t)) + V_{\text{dc}} \]
\[ \alpha \rightarrow \delta \rightarrow \alpha' \rightarrow \delta' \]

Fig. 1 Schematic of the optical frequency comb generator with a phase modulator and an intensity modulator.

Figure 1 shows the configuration of the optical frequency comb generator using the phase modulator and the intensity modulator. The phase modulator driven at large amplitude sinusoidal signal generates a lot of sidebands, and the intensity modulator flattens the sidebands. The optical field at output is expressed as follows:

\[ E_{\text{out}} = A \left\{ \sum_{k=-\infty}^{\infty} J_k(\delta) + B \times \sum_{k=-\infty}^{\infty} J_k(\delta+\delta') \exp(\Phi_{\text{eb}}) \right\} \] (1)

\[ A = \exp \left[ - \left( \alpha(\lambda) + \Delta \alpha(V_{\text{fe}}, \lambda) L_{\text{fe}} + \Delta \alpha'(\lambda) L_{\text{fe}} \right) \right] \] (2)

\[ B = \exp \left[ - \left( \alpha_{\text{dc}}(\lambda) + \Delta \alpha'(V_{\text{fe}}, \lambda) \right) L_{\text{fe}} \right] \] (3)

\[ \delta = n(V_{\text{fe}}, \lambda) \frac{2\pi L_{\text{fe}}}{\lambda_0}, \] (4)

\[ \delta' = n'(V_{\text{fb}}, \lambda) \frac{2\pi L_{\text{fe}}}{\lambda_0}, \] (5)

\[ \Phi_{\text{eb}} = n'(V_{\text{fe}}, \lambda) \frac{2\pi L_{\text{fe}}}{\lambda_0}, \] (6)

where \( J_k \) denotes \( k \)-th order Bessel function, \( \lambda_0 \) is the wavelength of the input light, and \( L_{\text{fe}} \) and \( L_{\text{fe}} \) are the lengths of the phase modulator and the intensity modulator. \( \delta \) and \( \delta' \) are the phase modulation degree of phase modulator, and that of the intensity modulator. \( \Delta \alpha \) and \( \Delta \alpha' \) are the absorption change when the voltage is applied. \( V_{\text{fe}} \) and \( V_{\text{fe}} \) denote amplitudes of the applied sinusoidal signals. \( V_{\text{dc}} \) and \( \Phi_{\text{eb}} \) are the bias voltage to the intensity modulator and phase difference by caused by the bias, respectively. \( \alpha_{\text{dc}} \) is the change in the absorption coefficient when the bias is applied. \( n \) and \( n' \) are the change in the refractive index of the phase modulator and that of the intensity modulator, respectively.

3. MQW optical modulator
Two types of modulator were fabricated, one was a phase modulator and the other was a Mach-Zehnder intensity modulator. Figure 2 shows the cross sectional view of the MQW optical modulator and the schematic of the phase modulator. The MQW had 10 cycles of 110-Å thick InGaAlAs layer and 50-Å thick InAlAs layer. The band gap wavelength was 1470 nm. The width of the ridge waveguide was 3μm and the length of the electrode was 1 mm.

Fig. 2 Structure of the MQW optical modulator.

We measured the excess loss of the phase modulator for various applied voltage as a function of wavelength. When there was no applied voltage, the loss of the phase modulator including the coupling loss was 19.7 dB. The absorption coefficient change calculated from the experimental data was shown in Fig. 3. We also
measured the loss characteristics of the intensity modulator. Figure 4 shows the phase modulation indexes calculated from the experimental results.

![Graph of absorption coefficient change for various applied voltages as a function of wavelength.](image)

Fig. 3 Absorption coefficient change for various applied voltages as a function of wavelength.

![Graph of phase modulation index for various wavelengths as a function of applied voltage.](image)

Fig. 4 Phase modulation index for various wavelengths as a function of applied voltage.

The maximum phase modulation indexes for each wavelength varied from 2.26 to 2.28.

4. Optical frequency comb spectra
We calculated the power spectrum of the optical frequency comb assuming that the length of electrode was 3 mm, the modulation frequency was 10 GHz, applied voltage was 2.3 V and the absorption coefficient derived from the measurement. The calculated comb spectra for the 1560-nm input light, and the 1580-nm input light are shown in Fig. 5(a) and 5(b), respectively. The basis of the power was set to that of the center frequency.

![Graph of power spectra of the optical frequency comb.](image)

Fig. 5 Power spectra of the optical frequency comb. (a) 1560-nm input light, and (b) 1580-nm input light.

The flatness of the comb spectrum for the input light with a wavelength of 1580 nm is better than that for the input light with a wavelength of 1560 nm due to the larger absorption coefficient change. Consequently, the optical frequency comb generation, using the MQW optical modulators, with 11 flattened modes and a frequency spacing of 10 GHz, may be possible.

5. Conclusion
The flatness of the comb spectrum for the input light with a wavelength of 1580 nm is better than that for the input light with a wavelength of 1560 nm due to the larger absorption coefficient change. Consequently, the optical frequency comb generation of 11 flattened modes and a frequency spacing of 10 GHz using MQW optical modulators may be possible.

Reference

Acknowledgement
We are grateful to Mr. Y. Oote of NTT Electronics Corporation for the fabrication of modulators.
Ultra fast optical frequency sweep with an integrated lithium niobate modulator

Tetsuya Kawanishi, Takahide Sakamoto, Akito Chiba and Masayuki Izutsu
National Institute of Information and Communications Technology
4-2-1 Nukui-kitamachi, Koganei, Tokyo 184-8745, Japan
Tel +81 42 327 7490, Fax +81 42 327 7938, e-mail: kawanish@nict.go.jp

Abstract—Ultra fast and precise optical frequency sweep for fine spectral measurement was demonstrated by using single-sideband modulation. The sweep time and range were 0.1 \( \mu s \) and 6.4 GHz, where the sweep repetition rate was 10 MHz.

I. INTRODUCTION

Optical frequency sweep plays an important role in precise measurements of optical properties of materials or components. Many of materials or components with very fine structures in wavelength domain have resonant structures or are sensitive to mechanical vibrations or thermal fluctuations. When the sweep speed is lower than temperature or mechanical fluctuations, we should use some stabilization techniques such as feedback control. Thus, the frequency sweep speed is very important to obtain feasible optical frequency sweepers. In conventional techniques, tunable lasers are often used for generation of optical frequency sweep signals, but the relation between the output optical frequency and the electric signal for tuning the laser can not be described by a simple function, where the sweep repetition rate was smaller than 100 kHz. In addition, we should also handle mode hopping of lasers and frequency aberration due to fluctuation in temperature or injection current. On the other hand, recently, we reported optical frequency sweep technique by using an optical single-sideband (SSB) modulator consisting of two Mach-Zehnder (MZ) modulators [1], [2]. The frequency of the output lightwave depends on rf-signal frequency and dc-bias voltage fed to the modulator, which can be electronically controlled. Thus, by sweeping the rf-signal frequency, we can obtain an optical frequency sweeper whose output frequency can be swept agilely. The frequency shift from optical input is precisely equal to the rf-signal frequency, so that we can achieve precise measurement of optical properties in wavelength domain. The frequency sweep range is limited by the modulation bandwidth of the SSB modulator. Recently, we developed high-speed versatile modulator which can generate optical SSB signals, where the optical frequency tuning range was 80 GHz [3]. Thus, we can expect that an optical frequency sweeper with wide tuning range and precise control can be achieved by using a combination of a tunable laser source for coarse tuning and an SSB modulator for fine tuning. In this paper, we investigate rapid optical frequency sweep technique using high-speed optical modulation. In order to achieve very fast optical frequency sweep without losing stability, we investigate a novel optical frequency sweep signal generation system which consists of a wideband fast sweep signal generator and an optical SSB modulator. Fast optical frequency sweep was demonstrated, where the sweep range and time were, respectively, 6.4 GHz and 0.1 \( \mu s \), where the sweep repetition rate was 10 MHz.

II. OPTICAL FREQUENCY SWEEP USING SINGLE-SIDEBAND MODULATION

The SSB modulator consists of a pair of MZ structures as shown in Fig. 1 (a) [3]. When we apply a pair of rf-signals, which are of the same frequency \( f_m \) and have 90° phase difference, to the electrodes RF\(_A\) and RF\(_B\), frequency shifted lightwave, which corresponds to upper sideband (USB) or lower sideband (LSB) can be generated at the output port of the modulator, where the output optical frequency is \( f_0 + f_m \) or \( f_0 - f_m \), where \( f_0 \) is the input optical frequency. Thus, we can control the output frequency by changing the rf-frequency \( f_m \). Fig. 1 (b) shows our experimental setup for optical frequency sweep. We used a wideband fast sweep signal generator consisting of an arbitrary waveform generator (Tektronix AWG710B) having a fast digital-to-analog converter, and of a wideband frequency multiplier which can generate the 32th order harmonic component. Electric linear frequency sweep signal whose frequency range was 300-500 MHz was generated by the arbitrary waveform generator, where the sweep time was 500 ns, 250 ns, 200 ns or 100 ns. In order to enhance the frequency sweep range and speed, the sweep signal was fed to the frequency multiplier. The multiplier output signal whose sweep range was from 9.6 GHz to 16.0 GHz was fed to an SSB modulator via a microwave amplifier and a 90° hybrid coupler, where the microwave power at each modulator input port RF\(_A,B\) was 23 dBm. A tunable laser source was used for coarse wavelength control. The output optical signal from the modulator was fed to a device under test (DUT). Transmitted optical power from the DUT was measured by using a photodetector. We also measured the transmitted power without DUT for calibration. The ratio between the optical powers with and without DUT should be equal to the transmittance of the DUT. The bias voltages applied on the modulator was adjusted for USB generation. Fig. 1 (c) shows an optical spectrum at the output port of the modulator, where the sweep time of the optical frequency sweeper was 100 ns.
III. FINE MEASUREMENT OF NARROW OPTICAL BANDPASS FILTERS

A narrow passband in the reflection band of a dual-section FBG was measured by using the fast optical frequency sweep technique with a repetition rate of 10 MHz. Fine spectral measurement of a 200 MHz narrow passband was demonstrated with 100 ns sweep time, where the sweep time was 10 MHz. This technique can be used for time-resolved frequency response measurement, with a very simple setup.

REFERENCES

50-nm Wavelength Tunable Ultra-Flat Comb Generation Using Single-Stage Mach-Zehnder Modulator

Takahide Sakamoto, Tetsuya Kawanishi and Masayuki Izutsu
National Institute of Information and Communications Technology (NICT)
4-2-1 Nukui-Kitamachi, Koganei-shi, Tokyo 184-8795, Japan
Tel: +81-42-327-6944, Fax: +81-42-327-5328, Email: tsaka@nict.go.jp

Abstract
50-nm widely wavelength tunable ultra-flat comb generation was demonstrated, where a highly stable electro-optic comb generator consisting of a standard dual-driven Mach-Zehnder modulator was used. 10-GHz spaced comb with 21 components was generated within 5-dB bandwidth.

1 Introduction
Recently, electro-optic (EO) modulation using LiNbO$_3$ waveguide modulator are increasingly attractive for optical frequency comb generation [1],[2],[3] due to an improvement of modulation bandwidth and a decrease in the driving voltage of the modulator [4],[5]. Introducing a high-voltage radio-frequency (RF) signal into the modulator, higher-order frequency components of the driving signal are promptly generated, which can be used as the frequency comb signal. Unfortunately, however, it is difficult to generate a frequency comb with good spectral flatness since each frequency component obeys Bessel function of different order. To solve this problem, proceeding reports proposed to utilize two optical modulators, where an optical phase modulator and a Mach-Zehnder modulator (MZM) should be cascaded in tandem [3].

Instead of the conventional technique, we found the optimal operating condition for flatly generating an optical frequency comb by using only a set of conventional MZM [6]. Therein, a continuous-wave (CW) light is EO modulated using an MZM dual-driven with in-phase signals but have slightly different amplitudes. The generated comb is spectrally flattened if the MZM is driven under the condition of $\Delta I + \Delta \theta = \pi/2$.

Fig. 1. Experimental setup;
TLD: tunable laser diode, PC: polarization controller, ATT: RF attenuator
where $\Delta A$ and $\Delta \theta$ denote amplitude difference of the driving signals and the bias condition of the MZM, respectively.

In this paper, we report on widely wavelength tunable ultra-flat comb generation using a conventional LiNbO$_3$ MZM. 10-GHz spaced ultra-flat optical frequency comb generation over ~200-GHz bandwidth was successfully demonstrated. Spectral ripple was less than 1.1 dB for 11 number of frequency comb components; 21 components were generated within a 5-dB bandwidth. The operation was almost wavelength-shift free: 50-nm wavelength tunable operation was demonstrated.

2. Widely wavelength-tunable ultra-flat comb generation

Fig. 1 shows the experimental setup for ultra-flat optical frequency comb generation using an MZM. The optical frequency comb generator consisted of a wavelength tunable external cavity semiconductor laser diode (TLD) and a LiNbO$_3$ dual-drive MZM having half-wave voltage of 5.4 V. A CW light was generated from the TLD with the intensity of 5.8 dBm and its
wavelength tuning range was 1524 nm-1574 nm. The CW light was introduced into the modulator through a polarization controller to maximize modulation efficiency. The MZM was dual-driven with sinusoidal signals with different amplitudes (RF-a, RF-b). The RF sinusoidal signals at a frequency of 10 GHz were generated from a synthesizer, amplified with a microwave booster, divided half with a hybrid coupler, and then fed to each modulation electrode of the modulator. The intensity of RF-a injected into the electrode was attenuated a little by giving loss to the feeder line connected with the electrode. The input intensities of RF-a and RF-b were 35.9 dBm and 36.4 dBm, respectively. The phase difference between RF-a and RF-b was aligned to be zero by using a mechanically tunable delay line, which was placed in the feeder cable for RF-a. The spectra obtained from the frequency comb generator were measured with an optical spectrum analyzer.

Fig. 2 shows optical spectra obtained from the comb generator, where modulator was dual driven under the optimal condition of $\Delta f + \Delta \theta = \pi/2$. The frequency spacing and 5-dB spectral width of the generated comb were 10 GHz and 210 GHz, respectively. The optical conversion loss from the input CW light to the average power of the generated comb ($-\eta$) was about 18 dB. The measured spectrum had good flatness under the operating condition. The number of frequency components for $<1.1$-dB spectral ripple was 11. 21 components of the comb were flatly generated within the 5-dB bandwidth. In addition to the demonstration of ultra-flat comb generation, Fig. 2 demonstrates widely wavelength tunable operation of the comb generator. The optical spectra were measured when the centre wavelength was tuned in the range between 1524 nm and 1574 nm, (the spectra were plotted with the wavelength interval of 10 nm.) This confirms that ultra flat frequency combs were generated over the whole wavelength tuning range of the input CW laser source.

3 Summary
In this paper, we have demonstrated widely wavelength tunable ultra-flat optical frequency comb generation using a conventional MZM. Widenly wavelength tunable operation over 50 nm was also demonstrated.

Acknowledgements
The authors would like to express their appreciation to Dr. M. Tsuchiya, T. Miyazaki for fruitful discussions.

References
Fabrication and Applications of Deep Gratings Buried in Silica Glasses

Kenji Kintaka¹, Junji Nishii¹, Tatsuhiko Nakazawa², and Kazuaki Oya²
1: Photonics Research Institute, National Institute of Advanced Industrial Science and Technology (AIST),
1-8-31 Midorigaoka, Ikeda, Osaka 563-8577, Japan
Phone: +81-72-751-8527, Fax: +81-72-751-4027, E-mail: kintaka.kenji@aist.go.jp
2: Nippon Sheet Glass Co., Ltd.

Abstract: Our recent work is reviewed on fabrication of binary gratings with deep grooves and high spatial-frequency buried in silica glasses and on applications of such gratings to a wavelength demultiplexer and one-dimensional photonic crystals.

1. Introduction
Binary gratings with deep grooves and high spatial-frequency are very attractive elements in integrated optics because of their high diffraction efficiencies and birefringence properties. Polarization beam splitters, wave plates, and polarization-independent transmission diffraction devices can be realized by use of specially designed deep gratings. The microstructures of such deep gratings are, however, highly fragile and difficult to handle. Recently, we have demonstrated that deep gratings could be buried in silica plate by overcladding without decrease of diffraction efficiencies and polarization dependencies [1]. This paper reviews our recent work in fabrication and applications of the buried gratings.

2. Fabrication of buried gratings [1]
Surface-relief gratings were fabricated in a SiO₂ substrate by electron beam (EB) lithography and inductive coupled-plasma reactive ion etching (ICP-RIE). An 0.4-µm-thick EB resist (ZEP-520, Nippon Zeon) was spin-coated on a SiO₂ substrate, and grading pattern with 1.5-µm period was patterned by EB direct writing and developing. A 300-nm-thick Ni film was deposited by evaporation followed by liftoff. The Ni mask pattern was transferred to the SiO₂ substrate by ICP-RIE (RIE-200iP, Samco) with C₆F₅ gas. The residual Ni mask was removed by a FeCl₃ solution. Figure 1(a) shows a scanning electron microscope (SEM) photograph of the cross-sectional structure of the fabricated grating. The groove depth, the ratio of groove width to period, and the side slope were 2.8 µm, 0.4, and 88°, respectively. Small cracks near the top of grating teeth were due to cleaving of the sample for SEM observation.

The SiO₂ cladding layer was deposited on the surface-relief grating by plasma-enhanced chemical vapor deposition (PECVD) using Si(OCH₃)₄ as precursor. The substrate temperature, pressure, and rate in the deposition were 400 °C, 53 Pa, and 100 nm/min, respectively. Figure 1(b) shows a SEM photograph of the cross-sectional structure of the grating after the overcladding. Although the grating grooves had triangular shape at upper part, the groove widths of middle and lower parts were almost the same as those of the original grating.

Figure 2 shows 1st-order transmission diffraction efficiency of the gratings before and after the overcladding measured in the wavelength range from 1.5 to 1.58 µm with incident angle of 30°. Open and solid circles show diffraction efficiency of the original grating for TE and TM-polarized light, respectively.

Open and solid triangles show diffraction efficiency of the buried grating for TE and TM-polarized light, respectively. Diffraction efficiencies higher than 80% were obtained with both gratings for TE and TM-polarized light, which were comparable with theoretically predicted values. Therefore, the deep grating can be successfully buried in SiO₂ by overcladding.

A compact four-channel demultiplexer using a high-spatial-frequency transmission grating buried in a silica-based waveguide have been proposed and investigated. A schematic view of the proposed demultiplexer with 20-nm wavelength spacing at 1550-nm band is illustrated in Fig. 3. The demultiplexer consists of a buried high-spatial-frequency transmission grating, a pair of parabolic mirrors, slab waveguide, and channel waveguides. The input light expands horizontally from the end of input channel waveguide to the slab waveguide, depending on the numerical aperture of the waveguide. The expanded light is reflected and collimated toward the buried grating by total internal reflection of the first parabolic mirror with an interface between the vertically etched waveguide and the air gap. The collimated light is diffracted by the buried grating. The 1st-order diffraction efficiency of the buried grating is more than 90 % across 200-nm bandwidth centred at 1550 nm. The diffracted light is reflected and focused onto the ends of the output channel waveguides by the
second parabolic mirror.

A 10-µm-thick SiO₂ undercladding and 5-µm-thick Ge-SiO₂ core layers were sequentially deposited on a SiO₂ substrate by PECVD. Channel waveguides of 5-µm width and a grating with 1.5-µm period, 5-µm groove depth, and 3.4-µm interaction length were patterned by EB lithography and ICP-RIE. A 10-µm-thick SiO₂ overcladding layer was deposited by PECVD. The parabolic mirrors were fabricated by photolithography and ICP-RIE. The device was annealed at 800°C in air. The device size was 5.1 mm × 9.2 mm.

The device characteristics were measured by use of a wavelength-tunable laser diode, optical spectrum analyser, and single-mode fibers. The single-mode fibers were butt-coupled to input and output waveguides. Figure 4 shows transmission spectrum of the fabricated device. The average insertion loss was 13 dB, which includes coupling losses. The average crosstalk noise level was –16 dB for the adjacent channels. The polarization-dependent loss (PDL) was about 0.6 dB. The insertion loss and PDL will be reduced by optimization of the fabrication condition.

4. Triangular one-dimensional photonic crystals [3, 4]

Novel wavelength-division devices using triangular one-dimensional photonic crystals (1D-PhCs) buried in glass substrates have been proposed and investigated for realization of large angular dispersion. Figure 5 illustrates a schematic view of the proposed device consisting of Ta₂O₅ core and SiO₂ cladding [4]. The input and output surfaces were tilted at an angle of 45° from the grating vector, so that the incident beam could be coupled with the first photonic band in the second Brillouin zone. The grating period and incident angle were designed to be 500 nm and 10°, respectively.

A 3-µm-thick Ta₂O₅ core layer was deposited on a SiO₂ substrate by RF ion plating (NSG Techno-Research). The grating with 3-µm-deep grooves was fabricated by two-beam interference exposure and ICP-RIE with CHF₃ gas. A 10-µm-thick SiO₂ over-cladding layer was deposited by PECVD.

TM mode was end-coupled in the slab waveguide through a selfoc microlens and a rod lens. Figure 6 shows the wavelength dependencies of diffraction efficiency and the change of output angle, which indicates the angular difference from that at 1405 nm in wavelength. The output angle change was 4.8° with the incident wavelength change of 1% in wavelength range from 1400 to 1500 nm, which was six times as large as that of conventional gratings. A maximum diffraction efficiency of 25% was obtained at around 1440-nm wavelength. Higher diffraction efficiency will be obtained by improving of the fabrication accuracy.

5. Conclusions

We have fabricated deep gratings with high spatial-frequency buried in silica-based glasses, and applied these gratings to the wavelength demultiplexer and the triangular 1D-PhCs.

Acknowledgement

This work was conducted in the Nanotechnology Glass Project as part of the Nanotechnology Materials Program supported by the New Energy and Industrial Technology Development Organization (NEDO).

References

Realization of 1-D Diffraction Grating on the End Surface of Hollow Optical Fiber

S. Lee¹, J. Kim², Y. Jung², W. Shin³, I. Sohn³, J. Lee³, D. Ko³ and K. Oh¹
1. Institute of physics and applied physics, Yonsei University 134 Shinchon-dong, Seodaemun-gu, Seoul, 120-749, Republic of Korea, Tel: 82-2-2123-7657, Fax: 82-2-365-7657, Email: koh@yonsei.ac.kr
2. Department of Information and Communications, GIST, Buk-gu, Gwangju, 500-712, Republic of Korea,
3. Advanced Photonics Research Institute, GIST, Buk-gu, Gwangju 500-712, Republic of Korea

Abstract
We fabricated 1-D diffraction grating on the end surface of cleaved hollow optical fibers with a direct inscribing method using femtosecond laser pulses. The measured near and far fields were modulated by diffraction grating.

1 Introduction
Diffraction pattern can be observed in nature and also be obtained by an artificial repetitive array such as diffraction grating.[1] Propagation properties of light can be effectively controlled using diffractive optical elements as well as refractive elements such as lenses. Combination of diffractive optics and guided wave optics can provide versatile control of electromagnetic wave, and in recent years, especially, there have been various efforts to inscribe an index modulation pattern on the end surface of optical fiber.[2]

One of the methods is the surface relief grating using azo-polymer film, where the actual mass of layer is modulated.[3] Another approach is the direct inscribing microstructures on fiber surface using femtosecond laser pulses. Formation of these photorefractive index modulation structures can provide effective diffraction grating even on very tiny region with ultra high accuracy.[4,5]

In this study, we inscribed 1-D diffraction grating on the surface of cleaved hollow optical fiber (HOF) using a femtosecond laser. An unique ring-shaped beam of HOF with modulated grating experimentally demonstrated a potential for optical beam shaping in the visible range, which was not possible in conventional solid core single mode fiber (SMF).

2 Experiment
The experimental setup for diffraction pattern fabrication is schematically described in Fig. 1. Femtosecond laser pulses of 184 fs duration and 450μW power were generated with Ti-Sapphire laser with 785.5nm wavelength for inscription of 1-D groove arrays on the HOF surface. Focused beam with help of ×50 objective lens which had N.A 0.42 inscribed 1-D diffraction pattern as it moved horizontally. The period of the grating was 2μm and the thickness of etched line was less than 1μm.

Fig.1. The schematic diagram of the experimental setup for fabrication of diffraction grating.

The actual cross-section images of the HOF before and after inscription are shown in Fig. 2. The dark line region indicates etched line by focusing femtosecond laser and the dark circle in the center is air-hole of the HOF which diameter was 10μm.
With the grating on the surface of HOF, the near field diffraction beam patterns shown in Fig.3 and 4 were measured by a charge-coupled device (CCD) with a He-Ne laser source of 635nm wavelength and a white light source.

The figure 3-(a)/(b) shows the ring-shaped near field images of HOF without / with surface grating modulation, respectively. The effects of the grating were clearly shown in Fig. 3(b) with dark lines modulated on a ring-shaped beam.

The figure 4 shows the far-field image of modulated HOF with white light and He-Ne laser sources. The zero’th and the first order are observed in both figures. It is interesting to note that the diffraction patterns strongly depended on the wavelength and the blue light was conspicuously diffracted in a Gaussian type as shown in Fig 4(a). To investigate the monochromatic diffraction pattern, we used 635nm He-Ne laser. As shown in Fig. 4(b), the pattern denotes somewhat complexity mainly caused by two factors, grating and ring-shaped HOF. The grating affected the separation of $0^{th}$ and $1^{st}$ order lobes and the ring-shaped HOF resulted in the shape of a lobe.

Not only the surface grating but also a unique beam shape of specialty optical fiber gives a new freedom of diffraction patterns design. The authors are in the process of investigating various core-shape of optical fiber with 1-D and 2-D surface modulation.

In conclusion, we fabricated one-dimensional surface grating on the end surface of HOF and verified the zeroth and the first order diffraction pattern. Furthermore, the composite effect of grating and ring-shape core generated more complicated pattern compared to the case of Single Mode Fiber (SMF).

This work was supported in part by the KOSEF (Program Nos. R01-2006-000-11277-0, R15-2004-024-00000-0), and the Science and Technological Cooperation program between and from MOST.

3 References

Polarization Insensitive and Low-loss Tunable 3D Hollow Waveguide for Tunable Photonic Devices

Mukesh Kumar and Fumio Koyama

Abstract - The modeling and experiment of a new 3D hollow waveguide is presented for polarization insensitive tunable photonic devices. Low polarization dependence and low propagation loss can be expected at the same time by optimizing the shape of the 3D structure.

Key Words: hollow waveguide, tunable photonic devices, photonic integrated circuits.

1. Introduction

As the data traffic and transmission speeds in optical communications are increasing, the photonics technology has shown great progress and promise for future flexible photonic networks. Various tunable optical devices will play a key role in these photonic networks. We have proposed and demonstrated tunable hollow optical waveguide (HWG) with either dielectric multilayers or semiconductor multilayers, as well as a number of tunable optical devices based on the hollow waveguide, exhibiting unique features such as temperature-insensitivity, wide tunability and polarization-insensitivity [1]. The shape of hollow waveguides can be changed by using a MEMS element which leads to the tunability of propagation characteristics. In practical applications, 3D hollow waveguide is necessary for the purpose of compactness of photonic integrated circuits. We presented the modeling and experiments on 3D hollow waveguides with either etched groove or lateral periodic structure [2, 3]. In addition, there have been reports on integrated antiresonant reflecting optical waveguides (ARROW) with 3D hollow cores and a microelectromechanical optical switch inside a hollow waveguide [4, 5]. The investigation of low loss 3D hollow waveguide with polarization insensitive operation is necessary for the realization of widely tunable devices based on these waveguides. Though we have presented tuning characteristics and loss of 3D hollow waveguides but the loss of the 3D structure was considerably large [3].

In this paper, a low loss, polarization insensitive and simply fabricable 3D hollow waveguide (HWG), which we call a 3D nanostep hollow waveguide, is proposed. The modeling on its propagation loss, including the polarization dependence, spot size and tunability is presented. In addition, the fabrication of a nanostep 3D hollow waveguide is presented.

2. Structure of Nanostep Hollow Waveguide

Fig 1. Schematic of 3D Nanostep Hollow Waveguide.

A 3D hollow waveguide has potential of a wider tuning range than that of a slab hollow waveguide [6]. The proposed 3D hollow waveguide is shown in Fig 1 in which light can be confined in three dimensions by the step in the top three layers of a bottom DBR. Both the top and bottom DBR layers consist of Si/SiO2 for high refractive index contrast. We optimized the parameters h (step height in bottom DBR), W (width of air-core) and t (thickness of the remaining third layer, Si, in the bottom DBR) for lateral confinement, low loss and less polarization dependence. In assumed calculation models, 6 pair Si/SiO2 in the top DBR and 7 pair Si/SiO2 in the bottom DBR are assumed with thickness of quarter wavelength where the operating wavelength is chosen to be 1.55 µm. The width of the air-core is kept at 20 µm. In order to confirm the optical confinement in 3D step hollow waveguide we calculated intensity distribution with different step heights by Film-Mode Matching Method as shown in Fig 2. We note strong confinement of light in our new 3D hollow waveguide.

Fig 2. Calculated intensity distribution in 3D nanostep hollow waveguide with W = 20 µm, D = 8 µm and t = 40 nm.

3. Propagation Characteristics

The calculated propagation loss versus the thickness of a remaining layer of silicon in the bottom DBR is shown in Fig 3. The calculated loss is lowest for the case of the quarter wavelength thick layer. For comparison, the
propagation loss of a slab hollow waveguide is also shown. The propagation loss of a 3D step hollow waveguide is much less than 0.1 dB/cm with \( h = 200 \) nm and it is around 0.1 dB/cm with \( h = 376 \) nm. So the calculated loss is in an acceptable level even at a narrower air core where its tunability will be larger [6].

![Propagation Loss vs Thickness of Remaining Layer](image)

**Fig. 3.** Propagation loss with “t”, at \( D = 4 \) \( \mu \)m and \( W = 20 \) \( \mu \)m.

![Air Core Thickness Dependence and Calculated Tunability](image)

(a) Air core thickness dependence of spot size and (b) calculated tunability of 3D nanostep hollow waveguide.

The relation between lateral spot size and air core thickness is shown in Fig 4 (a). We note that the spot size is widely tuned with air core thickness and can be matched with that of single mode fibers. To investigate the tunability of the 3D hollow waveguide the change in propagation constant as a function of air-core thickness is calculated which is shown in Fig 4 (b). We note that tunability of 3D hollow waveguide is (around 2.4 % by changing the air core from 9 \( \mu \)m to 5 \( \mu \)m) wider than that in slab hollow waveguide. The tunability will be higher for narrower air cores [6].

![Polarization Dependence](image)

**Fig. 5.** Polarization dependence of 3D nanostep hollow waveguide.

In a slab hollow waveguide, the polarization dependence is normally large due to strong polarization dependence of reflectivities of multilayer mirrors, while in case of 3D hollow waveguide, it is lower as shown in Fig 5. By decreasing the first layer thickness of the multilayer mirror, the birefringence \( B \), defined by \( B = (\beta_{TE} - \beta_{TM}) / \beta_{TE} \), where \( \beta \) is propagation constant, goes down. By optimizing the shape of 3D hollow waveguide polarization insensitivity can be realized.

By making the step in three layers of bottom DBR and leaving the third layer silicon 40 nm thick, we found polarization-insensitive condition. The birefringence of this new polarization insensitive 3D hollow waveguide is less than 4 \times 10^{-5} at air core thickness of 9 \( \mu \)m which is small enough to use in practical devices.

### 4. Fabrication

The 3D nanostep hollow waveguide was fabricated by wet etching (around 80nm silicon) the 6-pair Si/SiO2 DBR, followed by re-deposition of one pair SiO2/Si and then lifting off this newly deposited pair resulting in the required nanostep. The measured near field pattern of fabricated structure is shown in Fig 6. The measured lateral spot size is 55 \( \mu \)m at 20 \( \mu \)m thick air core and the calculated lateral spot size is 47 \( \mu \)m. The calculated and measured spot sizes are almost in agreement.

![Near Field Pattern](image)

**Fig. 6.** Measured near field pattern.

### 5. Conclusion

Our new 3D hollow waveguide (HWG) gives us simple fabrication, low loss, polarization insensitivity and wider tunability. This 3D waveguide also shows confinement of light in three dimensions and less polarization dependence with the optimized structure than a slab waveguide. Its spot size allows its coupling with a single mode fiber. The result will be useful for realizing temperature-insensitive, widely tunable and polarization-insensitive photonic devices based on the 3D hollow waveguide for future flexible photonic networks.

### References

Box-like filter response of quadruple series coupled microring resonator by coupling efficiency control

Masahiko Hisada, Yuta Goebuchi, Tomoyuki Kato*, Yasuo Kokubun
Yokohama National University, Graduate school of Engineering, *presently with Tokyo. Inst. Tech.
79-5 Tokiwadai, Hodogaya-ku, Yokohama, Japan 240-8501
Phone: +81-45-339-4237, Fax:+81-45-338-1157, Email: ykokubun@ynu.ac.jp

Abstract
To realize box-like spectrum response, we designed and fabricated a hitless wavelength selective switch using Thermo-Optic (TO) tuning of quadruple series coupled microring resonator. Both Butterworth and Chebyshev filters were successfully demonstrated by the control of coupling efficiency.

1 Introduction
We proposed and demonstrated a hitless wavelength selective switch using TO-tuning of double series coupled microring resonator [1]. This switch can serve as the building block of the wavelength selective switch matrix. However, to apply this switch element to large scale switch matrix, more box-like spectrum than that of double series coupled microring was required [1], [2]. This is because the bandwidth is reduced by cascading filter elements and only the box-like can be free from the spectrum narrowing due to cascading.

Therefore in this study, we proposed and demonstrated a box-like filter response (Butterworth and Chebyshev filters) of quadruple series coupled microring resonator by the control of coupling efficiency [3]-[5].

2 Principle and design
The Butterworth filter has no ripple in the pass band and gradual roll-off, while the Chebyshev filter has equi-ripples and sharp roll-off. The Chebyshev filter is more suitable for the box-like filter than the Butterworth filter, if the ripple is allowably small.

In the quadruple series coupled ring resonator, the Butterworth condition is given by $\kappa_1=\kappa_0/(2-\kappa_0)$, and $\kappa_2=\kappa_1/\kappa_0$ [6]. In our device, we designed as $\kappa_0=0.5$ to reduce the loss in the pass band. In this case, the theoretical spectrum response was calculated as shown in Fig. 1. The shape factor (the ratio of -1dB bandwidth to -10dB bandwidth) can be improved to 0.59, which is greater than that of double series coupling (0.41).

On the other hand, the Chebyshev condition depends on the allowable ripple. In our device, the allowable ripple was assumed to be 0.3dB and we designed as $\kappa_0=0.5$, $\kappa_1=0.100$, and $\kappa_2=0.0678$. In this case, the theoretical spectrum was calculated as shown in Fig. 2. The theoretical shape factor was calculated to be 0.74.

Fig. 1: Theoretical spectrum of Butterworth filter.

Fig. 2: Theoretical spectrum of Chebyshev filter.
3 Device fabrication

The structure of hitless wavelength selective switch is shown in Fig. 3. The core and the upper cladding materials were sputter deposited 17mol%Ta₂O₅-SiO₂ (n=1.660 @λ=1550nm) and SiO₂ (n=1.445 @λ=1550nm), respectively. The busline waveguide and microring resonator were laterally coupled. The resonator was shaped like a racetrack, and the coupling efficiency was controlled by the overlap length of straight waveguide portion and the distance between cores. The core height and width were 1.3μm and 1.3μm, respectively. The round trip length of racetrack resonator was 611μm, which corresponds to the FSR of 2.22 nm.

Fig. 3: Perspective view of wavelength selective switch.

4 Experiment

In the initial stage, the resonant wavelengths of individual microrings were slightly different due to the fabrication error. By controlling the electric current to each Cr thin film heater above individual ring resonators separately, we realized the Butterworth and Chebyshev filter responses as shown in Figs. 4 and 5, respectively.

In the device with coupling efficiencies κ₀=0.5, κ₁=0.0452, and κ₂=0.0215, the Butterworth spectrum response was successfully obtained as shown in Fig. 4. Other two peaks around λ=1549.44nm are slightly mixed orthogonal polarizations. The shape factor was 0.699, the FWHM bandwidth was 0.092nm, and the ripple depth was -0.44dB. The discrepancy between theoretical and experimented values is attributed to the error of coupling efficiency resulting from the fabrication error.

On the other hand, in the device in which the coupling efficiencies were designed as κ₀=0.5, κ₁=0.100, and κ₂=0.0678, the Chebyshev filter response was obtained as shown in Fig. 5. Other three peaks around λ=1551.1nm are the slightly mixed orthogonal polarizations. The shape factor was 0.852, the FWHM bandwidth was 0.201nm, and the ripple depth was -2.13dB.

The shape factor in the double series coupled microring resonator was 0.41, and we successfully improved the shape factor using the quadruple series coupled microrings.

Fig. 4: Measured drop port spectrum response of Butterworth filter.

Fig. 5: Measured drop port spectrum response of Chebyshev filter.

5 References

Characterization of interlayer directional coupler in layered silica-based planar lightwave circuit

Kenya Suzuki, Mikitaka Itoh, Takashi Saida, Shinsuke Matsui and Akio Sugita

kenya@aecl.ntt.co.jp

NTT Photonics Laboratories, NTT Corporation.

3-1 Morinosato Wakamiya, Atsugi, Kanagawa 243-0198, Japan.

Abstract—We characterize interlayer directional couplers in a layered silica-based PLC, and demonstrate a core width tuning method to reduce imperfect coupling and polarization dependence. We achieved almost 100% coupling between layers without polarization dependence.

I. INTRODUCTION

Silica-based planar lightwave circuits (PLCs) play an important role in optical communications systems, because they offer excellent optical characteristics, cost effectiveness and high reliability. Several approaches have been proposed for the large-scale integration of circuit components, including high contrast waveguides [1] and waveguide stacking on a single wafer [2-4]. The latter is attractive because it is capable of offering a novel function by using a 3-dimensional circuit topology with interlayer directional couplers. In this paper, we report the characteristics of interlayer directional couplers in a monolithically layered silica-based PLC. The couplers initially exhibited imperfect coupling characteristics and a polarization dependent coupling ratio due to the stress induced index mismatch between the layers and polarizations. To eliminate these undesirable characteristics, we developed a core width tuning method that matches the internal stresses of waveguides between layers.

II. INTERLAYER DIRECTIONAL COUPLER

Figure 1 shows the schematic configuration and a microscope image of a cross-section of our interlayer directional coupler. The upper and lower waveguides, which are first separated at the input region to avoid optical coupling, converge in the horizontal direction and remain close together for a certain length $L$. They then diverge with sufficient separation for us to disregard the optical coupling.

The interlayer directional coupler is fabricated as follows. First, the lower layer of a layered PLC is fabricated by the conventional PLC fabrication technique using flame hydrolysis deposition (FHD) and reactive ion etching (RIE). Then, we polish the middle cladding layer to flatten it and adjust the gap between the upper and lower cores. Finally, we repeat the same procedure to fabricate the upper layer.

The coupling ratio $C.R.$ of a [2x2] directional coupler is expressed as [5]

$$C.R. = F \sin^2 \left( \frac{\pi}{2} \left[ q (L - L_0) \right] \right)$$

$$F = \frac{\kappa}{q} \frac{\beta_1 - \beta_2}{2}$$

Here $\beta_1$, $\beta_2$, $\kappa$, $L$ and $L_0$ show the propagation constants of the upper and lower waveguides, the coupling coefficient between the waveguides, the length of the straight waveguides, and the equivalent length of the curved waveguide, which contributes to optical coupling, respectively. It is clear that if the propagation constants of the two waveguides are different, the coupler exhibits imperfect coupling characteristics (i.e. $F < 100\%$). In other words, unless the two waveguides have the same propagation constant, the coupling ratio is less than unity for any straight length.

Here, it should be noted that the FHD process requires a high temperature, and the glass composition of each layer of the PLC must be different to avoid deformation of the lower layer core. In other words, the upper layer must be composed of a softer glass than the lower layer. This causes stress mismatches between the waveguide layers and between the polarizations, resulting in imperfect coupling and polarization dependence caused by the photo-elastic effect. Therefore, we investigated a method to tuning the core width to match the internal stress and refractive indices of the waveguides.
III. EXPERIMENT

Figure 3 shows typical characteristics of the fabricated interlayer directional couplers. We used 0.75 % refractive index contrast waveguides, and fixed the waveguide gap and core heights at 3 and 6 μm, respectively. Figures 3(a) and (b) show the coupling characteristics of couplers with a core width of 5.5 μm for both layers, and with a 6.0-μm upper core and a 5.5-μm lower core, respectively. Owing to the internal stress mismatch, the coupler with the same core width for both layers exhibits polarization dependence and imperfect coupling as shown in Fig. 3(a), while the coupler with a wider core for the upper layer exhibits little polarization dependence and almost perfect coupling characteristics.

Figure 4 summarizes the measured maximum coupling ratio for various combinations of core widths. It can be seen that better coupling is obtained when the width of the upper layer is wider than that of the lower layer. This agrees with reports stating that softer cladding provides a small birefringence [6] and a wider core width leads to a larger birefringence [7]. That is, with the same core width, the birefringence of the upper layer is smaller than that of the lower layer. An increase in the width of the upper core increases its birefringence. The condition seen in Fig. 3(b) is the best.

If we accept that there was a small error in the fabrication, the prediction in the previous section also agrees with the experiment. As predicted, for the same core width, the TE mode shows perfect characteristics while the TM mode does not. As the upper width increases, the effective index difference between layers for the TM mode approaches zero, and the optimum width for balancing both modes is between 5.5 and 6 μm.

IV. SUMMARY

We investigated interlayer directional couplers in a layered silica-based planar lightwave circuit. They exhibited inherent polarization dependence and imperfect coupling due to the stress mismatch induced by the photo-elastic effect. We demonstrated a core width tuning method as a countermeasure, and achieved almost 100 % coupling without polarization dependence. A layered silica-based PLC is a promising candidate for the greater integration of optical functional devices.

REFERENCES

MEMS Devices and Technologies for Photonic Network
Hiroshi Toshiyoshi
Institute of Industrial Science, the University of Tokyo
Room Ee-306, 4 – 6 – 1 Komaba, Meguro-ku, Tokyo 153-8505, Japan
Phone: +81-(0)3-5452-6276, Fax: +81-(0)3-5452-6648
E-mail: hiro@iis.u-tokyo.ac.jp

Abstract
We present recent research activities on microelectromechanical systems (MEMS) and devices in the field of fiber optic applications such as variable attenuators, tunable wavelength filters, and modulators integrated with silicon light wave circuits.

Introduction
Microelectromechanical systems have found a wide range of application in the field of micro-optics thanks to the larger optical effect generated by the mechanical motions than most solid-state solution could provide. For instance, angle change of mirror can make a large offset of reflected light, which could be used in optical crossconnects and variable attenuators. Micromechanical motion in the wavelength scale can also be used for wavelength filtering when it is combined with tunable interferometers. Figure 1 illustrates typical optomechanical modulations that could be combined with MEMS, which implies a plenty of potential of MEMS approach in micro optics and photonics. Amongst the recent research activities in this field, we give three points of view: (1) MEMS approach towards small components rather than large systems, (2) hybrid assembly technique for better optomechanical performance, and (3) a challenge towards the silicon photonics.

Towards Small Components
Large scale optical crossconnect was the main target of the optical MEMS devices in late 1990s but it has been shifted towards smaller components for their large number of sales shipping. For instance, we have developed a VOA using an electrostatically controllable tilt mirror [1], as shown in Figure 2 (a) and (b). Starting material was a bonded silicon wafer (SOI, silicon on insulator) processed by high aspect-ratio silicon dry etching. A micro mirror of 0.6 mm in diameter and 0.03 mm in thickness was electrostatically controlled with dc voltages of lower than 5 V. Reliability has been the most critical pass of MEMS devices for commercialization; the VOA has cleared its low-voltage operation (5 V for 50 dB attenuation or higher), electrostatic drift issue, temperature dependent fluctuation, and mechanical shock tolerance [2]. Compared to large scale optical crossconnects used in the trunk line of optical fibers, small components such as VOA are used in quantity everywhere in the fiber systems.

Figure 2 Bulk-micromachined electrostatic mirror for fiber optic variable attenuator: (a) Electrostatic mechanism, (b) SEM image of the device on a stem, and (c) VOAs used in optical fiber systems [1, 2].

Hybrid Assembly with MEMS
MEMS is an enabling technology for integrating an optical component with its motion control actuators by batch processing. However, optical components in MEMS devices sometimes suffer from poor optical performance such as temperature-sensitive mirror flatness and process-dependent surface smoothness, when compared with mirrors and lenses of macro scale. To overcome this problem, we have developed an optical platform for mounting an optical component (glass cube...
of high-reflection and anti-reflection coating) to compose a wavelength-tunable Fabry-Perot interferometer [3], as shown in Figure 3. Bonding of mirror cubes is currently done manually under microscope observation, but it could be replaced with automatic assembly using die bonders. Special care was paid for the blue material to fix the micro optics on the silicon platform. DC voltage around 70 V was used to control the gap between the mirror cubes and to tune the peak wavelength of transmission in a range from 1479 nm to 1579 nm with bandwidth of 2.6 nm to 2.9 nm, as shown in Figure 3.

Figure 3 Fabry-Perot interferometer with a pair of HR-mirror cubes mounted on a bulk micromachined translation stage for wavelength filter application [3].

Integration with Silicon Photonics
Lightwave traveling in a high-refractive index material (silicon) can be modulated by mechanically inducing disturbance in the evanescent field of the waveguide. For instance, micro disk resonators combined with micromechanical motion have been reported by Lee et al. for integrated wavelength-selective switch application. In our group, we have developed a process for making a silicon surface micromachined structure on a silicon photonic waveguide such as photonic crystal (PhC) waveguide and silicon nanowire waveguide. Figure 4 shows the schematic illustration and an SEM image of the optical modulator made of polysilicon (0.38 micron thick) deposited on a pre-fabricated photonic PhC waveguide. Electrostatic force of applied voltage brought the cantilever in the close vicinity (100 nm or less) over the PhC waveguide to induce modulation in both intensity and phase. Thanks to the high refractive index of silicon, the effective length of modulator was only 5 microns long. Further reduction of device size is under development by using a single layer of silicon rather than using additional deposited layer of polysilicon.

Figure 4 Surface micromachined polysilicon structure for optical modulator integrated with silicon PhC waveguide [5].

Conclusions
There still remain various types of optical MEMS principles unexplored, and their effects become more significant when device is made to be as small as the waveguide of interest, which is now possible by the fast-evolving micro and nano fabrication technologies.

Acknowledgements
The author would like to thank the co-workers in his research group. The MEMS VOA is by the joint work with Keiji Isamoto et al with Santec Co. The tunable Fabry-Perot filter was developed by Toshiro Yamanoi et al with Koshin-Kogaku Co. The MEMS PhC modulate was co-developed with Prof. K. Isamoto et al with Santec Co. The tunable Fabry-Perot filter was developed by Toshiro Yamanoi et al with Koshin-Kogaku Co. The MEMS PhC modulate was co-developed with Prof. Santoshi Iwamoto and Prof. Yasuhiko Arakawa with RCAST, the Univ. of Tokyo, and also by Dr. Akio Higo.

References
Nanoscale couplers for large scale photonic integrated circuitry

Duncan L. MacFarlane
Department of Electrical Engineering
The University of Texas at Dallas
Richardson, Texas 75083
972-883-2165 (w) 972-883-2710 (f)
dlm@utdallas.edu

Marc P. Christensen and Gary A. Evans
Department of Electrical Engineering,
Southern Methodist University
Dallas, Texas 75275

Abstract—A compact, efficient nanophotonic coupler, and its application to programmable photonic integrated circuitry is presented. The coupler operates on the partial coupling of the evanescent wave of a total internal reflection into a continuing waveguide.

I. INTRODUCTION

Important inter-related frontiers for photonic integrated circuitry include the increased level of integration, higher yield and smaller footprint building blocks. A common task is the splitting of one optical signal, in one waveguide, into two components in two waveguides, and this must be done with high efficiency and minimal back-reflection. Currently, the conventional approach to this problem is the adiabatically tapered “Y” coupler [1-3]. Using this configuration it is common to achieve very high efficiencies with virtually no back-reflection in splitters that are several hundreds of microns in length. Resonant structures may also be included in Y-type couplers to reduce the size, however not necessarily the group delay of the device [4]. Parallel waveguide based directional couplers have also been realized in a variety of material systems [5-7], and in these structures gratings may be included to provide wavelength selectivity and shorter length scales [8]. Recently, compact “air trench” tapers have been demonstrated for low index contrast systems with efficiencies in the high 90% range that occupy approximately 30x30 um areas [9]. In Reference 9, the authors appropriately point out the problem of loss in photonic crystal couplers. Careful sidewall control and tapers contribute to low Fresnel back-reflections and loss in Reference 9.

The authors have been developing an active 1.5 micron wavelength photonic true time delay device in the InGaAsP/InP material system that requires a right angle split of the mode [10] at a T intersection of two ridge waveguides as shown in Fig. 1. In this letter the authors discuss a compact coupler design based on frustrated total internal reflection to achieve an equal split. In contrast to the coupler shown in Fig. 1, the trench is etched deeply to completely overlap the waveguide mode, but its width is reduced to frustrate the total internal reflection and allow coupling to the continuing waveguide. In this configuration arbitrary splitting ratios can be achieved through careful control of the trench width. Efficiencies in excess of 95% are predicted in a 2 micron footprint.

The coupler operates on the partial coupling of the evanescent wave of a total internal reflection into a continuing waveguide, so that half the signal from the evanescent field is coupled into the continuing waveguide and therefore only the remaining half is reflected into the side waveguide. For material systems with low index of refractions an air gap of reasonable dimensions (>200 nm) can be used. Importantly, for higher index III-V semiconductor waveguides with high refractive index, the gap can be filled with a dielectric using, for example, plasma enhanced vapor deposition (PECVD) [11,12] thereby increasing the required gap width for 3 dB coupling.

II. DESCRIPTION

The coupler was designed using finite difference time domain software installed on a Linux based platform. The complex optical scattering problem was solved in two dimensions, and design variables included different material systems, coupler dimensions and coupler materials. Reported here, however, is one geometry, a 45 degree cut aligned to couple signal into the branch of the “T”. For most material systems used in integrated photonics, such as III-V semiconductors, lithium niobate, SiO2, silicon and polymers, the cut is totally internally...
reflecting. Therefore, in the FDTD runs, the width of the trench is scanned numerically to find the desired coupling percentage, which for our application is 50%. Figure 2 shows a contour plot of the waveguide “T”, a 105 nm air gap, and the resulting electric fields in an InGaAsP/InP multiple quantum well epitaxial waveguide structure. In these TM calculations, the ridge waveguide width was 2.5 microns. The diagonal polygon is the air trench, and the

Figure 2. Finite Difference Time Domain calculations for a 3 dB coupler in an AlGaAsP/InP ridge waveguide using an air gap of 105 nm.

other thin polygons are power monitors for the launch, transmission and reflection of the incident wave. Improved manufacturability and less sensitivity to wavelength and gap tolerances is attained by depositing a dielectric to fill the etched gap. For instance, in the InGaAsP/InP case above, the use of Al2O3 (n=1.75) extends the gap dimension to 137 nm.

The role of frustrated internal reflection in the operation of the coupler discussed herein can be described through an analytic derivation based on a plane wave approximation to the mode. For our configuration the barrier index and incident angle (45-degrees) should be chosen to result in total internal reflection had the thickness of the barrier been large. Figure 3 is a family of curves of the resulting 3dB coupling solution. It plots the necessary trench width as a function of the waveguide index of refraction for a variety of fill materials: air (n=1), PMMA (1.48), SU8 (1.57), sapphire (n=1.75), and zirconium (n=2.1). The ranges of the plotted lines are limited by the domain over which this analysis holds, i.e., total internally reflecting waveguide – trench interfaces. Note that in the range of polymer and glass waveguide indices, near 1.5, air trenches provide trench dimensions that can be readily fabricated. With waveguide indices typical of semiconductor waveguides (n>3), the air trench approach results in great challenges for high aspect ratio etching. The fabrication difficulty can be reduced if the gap is filled with a higher index material.

III. CONCLUSION

A novel 1x2 photonic coupler has been presented. The coupler operates on the partial coupling of the evanescent wave from a total internal reflection interface into a continuing waveguide, and was analyzed both by finite

difference time domain calculations and by the closed form expression for frustrated total internal reflection. These analyses are in good agreement.

REFERENCES

Beam steering and coupling in tunable hollow waveguide with narrow air core

Akihiro Imamura and Fumio Koyama

Microsystem Research Center, P&I Laboratory, Tokyo Institute of Technology, 4529 Nagatsuta, Midori-ku, Yokohama 226-8503, Japan
Tel: +81-45-924-5077, Fax: +81-45-924-5977, Email: koyama@pi.titech.ac.jp

Abstract
We present the new function of tunable hollow waveguides with a variable air core. A possibility of wide-angle beam steering from the hollow waveguide and efficient tilt-coupling in narrow air core hollow waveguides is suggested.

1. Introduction
Tunable hollow waveguides [1] based on Bragg reflector waveguides [2] exhibit various unique features such as wide tunability of propagation constants, temperature insensitivity, and slow light propagation in narrow air cores. We have demonstrated various device applications of tunable hollow waveguides [3-6]. The small air core hollow waveguide gives us giant tuning characteristics [5]. Propagation characteristics including slow light in sub-wavelength narrow air cores are still open for discussions. Also, the coupling issue for such narrow air cores is an important issue.

In this paper, we discuss a possibility of wide-angle beam steering from tunable hollow waveguides and efficient coupling in narrow air core hollow waveguides.

2. Radiation and beam steering
Figure 1 shows the model of hollow waveguides for calculating the radiation from the hollow waveguide. The top mirror is terminated and thus the radiation takes place in an upper side. We used a full-vectorial Maxwell’s solver (FIMMwave, provided by Photon Design Company), which is based on a film-mode-matching method [7], for calculating field distributions. Figure 2 shows the calculated field distribution of the radiation from hollow waveguides with an air core of (a) 3 μm, (b) 1.5 μm and (c) 1 μm, respectively. It is noted that we are able to realize wide beam steering of over 40 degrees with small displacement of air cores. Figure 3 shows the radiation angle versus the air core thickness of hollow waveguides. The result shows a possibility of continuous and wide angle beam steering, which results from the wide tunability of propagation constants.

Figure 4 shows the calculated insertion loss as a function of air core thicknesses, indicating an increase of loss for small air cores, which can be understood by the reflection at the edge of the terminated top mirror as shown in Fig. 5. If the core thickness is closed to a cut-off condition, the guided light in the hollow waveguide is perfectly reflected at the edge as shown in Fig. 5.

3. Tilt-coupling into narrow air-core waveguide
Figure 4 shows a possibility of low loss coupling into a narrow air core by tilting an input beam as shown in Fig. 6. The incident angle and spot size of an input beam are 40 degree and 2 μm, respectively. Figure 6 shows the calculated field distribution. The insertion loss is 2.5 dB, while the structure is not optimized yet.

4. Conclusion
We present wide-angle beam steering from the tunable hollow waveguide and efficient tilt-coupling in a narrow air core. Wide tunability in sub-wavelength air cores enables us to realize new functions of tunable hollow waveguides.

References
Fig. 1 Model of hollow optical waveguides for beam steering.

Fig. 2 Calculated field distribution of radiation from hollow waveguide with an air core of (a) 3 µm, (b) 1.5 µm and (c) 1 µm.

Fig. 3 Calculated radiation angle from the hollow waveguide versus air core thickness.

Fig. 4 Calculated insertion loss as a function of air core thickness.

Fig. 5 Perfect reflection at the edge of hollow waveguide with 0.78 µm air core.

Fig. 6 Schematic and calculated field distribution of hollow waveguide with tilt-coupling scheme.
Micro-optic MZI filter composed of a dual-fiber collimator and a reflective beam-splitting plate

Jong Hoon Lee¹, Chang-Hee Lee¹, Sung Jo Park², Hyung Woo Kwon², Jae-Won Song², and Hyun Doek Kim²

¹Department of Electrical Engineering and Computer Science, Korea Advanced Institute of Science and Technology 373-1 Kusong-Dong, Yusong-Gu, Daejeon 305-701, South Korea, Tel: +82-42-869-3463, Email: chhl@ee.kaist.ac.kr
²School of Electrical Engineering and Computer Science, Kyungpook National University, 1370 Sankyuk-dong, Puk-gu, Daegu, 702-701 South Korea, Tel: +82-53-950-7578, Email: hyundkim@ee.knu.ac.kr

Abstract

We demonstrate a micro-optic MZI filter composed of an optical collimator with dual fiber-pigtails and a reflective optical plate with a periodic mirror pattern. The insertion loss and the PDL of the implemented filter were less than 1.52 dB and 0.12 dB, respectively.

1 Introduction

Mach-Zehnder interferometric (MZI) filters are widely used in fiber-optic communication systems since they have wide pass-bands and linear phase characteristics. Especially, micro-optic MZI filter shows excellent optical performances such as a low insertion loss, a very low polarization dependent loss (PDL) and offers a high design flexibility. However, the micro-optic MZI filter requires a strict alignment process, which increases manufacturing difficulties. In this paper, we propose and demonstrate a novel micro-optic MZI filter structure to alleviate the optical alignment requirements of the conventional micro-optic MZI filter.

2 Configuration and principles

The proposed MZI filter is composed of an optical collimator with dual fiber-pigtails and a beam-splitting optical plate as shown in Fig. 1 (a). The pigtail-fibers of the collimator are used as an input port and an output port of the MZI filter, respectively. The optical plate has mirrors on its both facets. The rear facet of the optical plate has a uniform mirror pattern while its front facet has a periodic mirror pattern, a stripe mirror pattern. We assume that the width of the optical plate, the length of the stripe pattern in y-direction, is much longer than the beam diameters of the collimated beam when the beam propagates in z-direction. The centers of the input and the output fibers of the collimator are aligned as shown in Fig. 1 (b) so that the position deviation of the plate in y-direction does not affect on the characteristics of the filter.

The optical beams entering through the input fiber are expanded by the collimator and reflected by the mirror on either facets of the optical plate. Since the mirror area and the open areas are interleaved on the front facet, some beams are reflected by the front facet mirror while the other beams by the rear facet mirror after passing through the open areas of the front facet. The optical collimator and the optical plate are placed so that the reflected beams are coupled into the output fiber through the collimator. The optical plate splits the beams and the beams reflected by the rear facet of the optical plate propagate relatively longer optical paths before they arrive at the collimator. Namely, the optical plate induces the optical path length difference between the beams reflected by the front facet and the beam reflected by the rear facet. The Mach-Zehnder interference occurs between the beams with different phases during they are collimated into the output fiber, which leads to a wavelength-dependent spectral response of the filter.

3 Fabrication and experiment

We fabricated the optical plate by implementing mirrors on both facets of a glass substrate. The pyrex glass with a 500 µm thickness is used as the substrate. We deposited Aluminum (Al) on the front facet of the Pyrex glass substrate through an E-beam evaporation process to form the optical mirror pattern. We used etching process to implement a periodic stripe mirror pattern on the front facet of the optical plate. The thickness of the deposited Al mirror region on the front facet of the glass substrate was about 2000 Å. The period of the stripe mirror pattern was 100 µm and the width of the etched region was about 50 µm, which means the width of the Al mirror pattern on the front facet was also 50 µm.

Fig. 2 shows the microscopic picture of the front facet of the fabricated optical plate. The white regions are the Al-deposited mirror areas while the dark regions are the etched open areas. The stripe mirror pattern was well defined and has clear open areas. The measured reflectivity of the mirror pattern of the front facet was higher than 85% over the wavelength range from 1520 nm to 1580 nm. The uniform mirror of the rear facet of the plated was implemented by evaporating Al thermally.
on the rear side of the glass substrate and the thickness of the deposited Al region was about 8000 Å. We used a conventional optical collimator with a 500 µm beam diameter and about 2 degree of crossing angle between in-out beams. We aligned the fabricated optical plate and the optical collimator as shown in Fig. 1 (a). The working distance between the collimator and the optical plate was about 2 mm.

Fig. 3 shows the measured spectral response of the implemented MZI filter using a tunable laser and power meter. The free spectral range (FSR) of the filter was about 1.6 nm (about 200 GHz). The insertion loss of the implemented MZI filter was less than 1.52 dB and the extinction was greater than 24 dB over 1510 nm to 1640 nm. We also measured the spectral responses of the implemented micro-optic MZI filter by changing the relative position of the optical plate from 0 µm to 50 µm in x-direction to ensure the constant spectral responses irrespective to the position deviation in lateral direction. The collimator and the plate were aligned as shown in Fig. 1 (b) in the initial state (x = 0 µm). The extinction ratio variation due to the change of the relative position of the optical plate was less than 0.2 dB and the measured insertion loss variation at the peak wavelength was less than 0.05 dB. This means that the location error due to a misalignment between the optical plate and the collimator does not affect on the characteristics of the MZI filter. Thus the only parameter one should care in the alignment process to implement MZI filter is the tilt angle of the optical plate, which simplifies the manufacturing process of the optical filter.

Fig. 4 shows the spectral response and the PDL of the implemented MZI filter measured with a tunable laser and a PDL meter in the wavelength range from 1545 nm to 1555 nm. The measured PDL of the filter was less than 0.12 dB within the 3 dB passband of the filter.

4 Summary
In this paper, a novel micro-optic MZI filter has been proposed and demonstrated to overcome the alignment and the fabrication difficulties of the micro-optic filter. It is composed of a collimator with dual pigtail-fibers and a reflective optical plate with a periodic mirror pattern on the front facet. We implemented the MZI filter with a 1.6 nm FSR. The maximum insertion loss and the minimum extinction ratio of the filter were 1.52 dB and 24 dB, respectively. We also show that the lateral alignment error does not affect the characteristics of the filter.

This work was partially supported by Korea Ministry of Science and Technology under the National Research Laboratory and Brain Korea 21 Project.

References
Finite-difference time-domain calculation of radiation dynamics in advanced photonic crystal devices

M. J. Steel1,2, D. P. Fussell3, M. M. Dignam3, C. M. de Sterke2, R. C. McPhedran2, F. Borda4, C. Seassal4 and A. Rahmani4

1 RSoft Design Group, Inc., 65 O’Connor Street, Chippendale, NSW 2008, Australia
2 Centre for Ultrahigh-bandwidth Devices for Optical Systems (CUDOS), School of Physics, University of Sydney, NSW 2006, Australia
3 Dept. of Physics, Queen’s University, Kingston, Ontario, Canada K7K 3N6
4 Institut des Nanotechnologies de Lyon, UMR 5270 Ecole Centrale de Lyon, 36, Avenue Guy de Collongue 69134 Ecully, France
5 Centre for Ultrahigh-bandwidth Devices for Optical Systems (CUDOS), Department of Physics, Macquarie University, NSW 2109, Australia

Abstract

Finite-difference time-domain (FDTD) simulation is central to studying the radiation dynamics of complex open nano-photonic systems. I review some important techniques and discuss applications in laser cavities and LDOS characterization.

Introduction

The enormous interest focused on Photonic Crystal (PC) cavities of late is in large part attributable to the way in which they enable precise control of light, especially strong manipulation of radiation properties [1]. The enormous Q factors possible in PC cavities [2] will enable low threshold lasers while at modest Qs they permit lasers with very high modulation rates [3]. Similarly, manipulation of the density of states near band edges enhances the spontaneous emission rate allowing high-efficiency LEDs. These are all intrinsically open systems where leakage of radiation to the environment is central part of the physics. While mode expansion techniques can be applied to some problems, especially in 2D, and significant physics can be revealed through semi-analytic approaches such as elegant Green function techniques [4], to fully account for complex geometries and coupling between waveguides, cavities and the environments, large-scale FDTD simulation remains the dominant modeling tool.

FDTD simulation has a number of advantages—it is easy to incorporate material dispersion, arbitrary shaped features, and imperfections in design. The algorithm can exploit structural symmetries, and parallelizes naturally. Since the whole electromagnetic field is known, essentially any derived quantity can be found such as mode volumes, coupling coefficients, and direction-dependent leakage rates. Finally FDTD simulation is accessible: it is conceptually simple to use, easily applied to experiments, and is available in numerous commercial and free implementations. Against these positives must be balanced the requirement of very large computational resources and the reduced physical insight that comes from an over-reliance on purely numerical techniques.

In this paper we discuss two applications, slow-light mode PC laser cavities and efficient calculation of the Local Density of States (LDOS) in PC cavities.

Trapping slow-light in photonic crystal cavities

Slow light modes of a photonic crystal near a band-edge are attractive for enhanced nonlinearity and low-threshold laser emission because they exhibit a very high density of states. However, as modes of the perfect crystal they are extended states, not suitable for laser emission. We designed a range of cavities in which a central crystal is contained in a larger crystal of different hole radius (Fig. 1a).

This allows the confinement of a slow-light mode to the inner cavity. Similar structures were considered in Ref. [5]. We performed a large series of FDTD simulations to optimize the Q factor with respect to the size of the inner cavity, its hole radius and the number of “adaptation layers” smoothing the border between the cavities. An important part of these simulations was the exploitation of the Filter Diagonalization Method (FDM) [6] which allows the resonant frequency and decay rate to be found from a single relatively short calculation. This is much faster and more reliable than traditional FFT-based approaches or Padé-interpolation methods. As shown in Fig. 1b, the slow-light cavities produce cavity resonances with very high Q factors, but quite small modal volumes. By measuring separate in-plane and out-of-plane Qs, we find that the optimum Q is driven to the right (higher inner hole radius) by decreasing in-plane losses as the mode moves deeper into the band gap of the outer crystal, and to the left by increasing vertical losses as the mode becomes more spatially confined and more spectral components enter the light cone.

FDTD calculation of LDOS

For many radiation calculations, the primary quantity of interest is the local density of states (LDOS)
FIG. 1: (a) Examples of $V^2_j$ cavity. Only the top-right quadrant is shown. (b) Quality factor $Q$ (solid lines) and frequency (dashed lines) of the confined slow-light mode versus the filling factor of the core for $V^2_j$ cavities with $j=0, 1, 2$.

$L(r, \omega)$ since spontaneous emission enhancement is directly proportional to $L(\omega, r)$. The LDOS itself is given by the electromagnetic dyadic Green function as $L = -2/(\pi c^2)G(\omega; r, r)$. As first pointed out in Ref. [7], time-domain method such as FDTD are well-suited for calculating Green functions since the response of the medium to a source localized in time and space is easily found. We recently formulated this approach [8] to give a general expression for the Green function in terms of the field excited by any localized source of area $A$ as

$$G_k(\omega, r) = \frac{i}{2\omega n A^{1/2}} \int_0^\infty d\tilde{t} \tilde{x}_k \cdot \mathbf{E}(r, t) \exp(-i\omega t). \quad (1)$$

While this provides a description for finding the spontaneous emission rate throughout a photonic crystal, it remains a daunting calculation since a separate simulation must be performed for launch position at which the emission rate is required.

In fact, for coupling between confined cavity states, the problem can be greatly reduced by projecting onto a finite subspace of modes. Consider two neighboring cavities in a 2D photonic crystal. Interference between the states gives rise to a suppression of spontaneous emission at points between the two cavities (see Fig. 2). We have shown that the physics of such an open system can be accurately modeled using just two FDTD propagations in which the launch fields are not localized point excitations, but the bound modes of a corresponding closed system [8]. Such an approach can in principle be extended to any number of coupled well-confined PC resonances.

By direct calculation of modal properties and extraction of fundamental properties such as the LDOS and Green function, FDTD is liable to remain a key tool in nanophotonic modeling for a considerable period.

FIG. 2: Spontaneous emission ratio $R_{se}$ as a function of frequency $\omega$ and position $x$ along the $x$ axis of the microcavity.

Acknowledgments

This work was supported by the Ministère délégué à la Recherche et aux Nouvelles Technologies (Programs “ACI Jeune Chercheur” and “ACN Nanooqub”), by the Australian Research Council Centre of Excellence and Linkage International Fellowship schemes and by the Merit Allocation Scheme on the National Facility of the Australian Partnership for Advanced Computing. CU-DOS is an ARC Centre of Excellence.

Optimal Design and Analysis of Ultra Compact 1.3/1.55μm Demultiplexer

Dae-Seo Park, Jae-Hyun Kim, Beom Hoan O, Se Geun Park, El-Hang Lee and Seung Gol Lee
Optics and Photonics Elite Research Academy (OPERA), Inha University
253 Yonghyun-Dong, Nam-Gu, Incheon, 402-751, South Korea
Tel: +82-32-860-7433, Fax: +82-32-873-8970, E-mail: sglee@inha.ac.kr

Abstract
Ultra compact 1.3/1.55μm demultiplexer was newly proposed based on the self-imaging phenomenon and the bandgap property of photonic crystal. The characteristics of the proposed device were analyzed and optimally designed by using the FDTD method.

1. Introduction
From a communication capacity point of view, the electron based conventional communication system has shown the physical limitation due to a rapid growth of the internet and multimedia. Therefore the solution should be found. One of the answers is optical communication system which has an extremely wide communication capacity. And one of the key components in the optical communication system is a wavelength demultiplexer. Especially, 1.3/1.55μm wavelength demultiplexer plays a very important role in the passive optical network systems. Therefore many researchers have tried to enhance the performance of a demultiplexer.

Multimode interference (MMI) based device is a good candidate for the wavelength demultiplexer. Since the length of the device using an interference phenomenon is determined to be a common beat length for the multiple wavelengths, however, the device is a quit long. Though some researchers used the Bragg gratings to resolve the size problem [1], it has very wavelength-sensitive characteristics by the Bragg grating’s properties. In 2006, Chung et al. suggested the photonic crystal (PhC) instead of the Bragg gratings to enhance the isolation ratio [2].

Almost conventional dielectric waveguide devices can be realized with PhC devices having a very small size. In 2004, Kim et al. demonstrated that self-imaging phenomenon also was valid in the PhC waveguide as well as in the dielectric waveguide [3].

In this paper, we proposed an ultra compact 1.3/1.55μm demultiplexer based on the self-imaging phenomenon in the PhC waveguide and the photonic band gap (PBG) property of the PhC device. And the characteristics of the proposed device were analyzed and optimally designed by using the finite-difference time-domain (FDTD) method.

2. Design and Evaluation
The PhC is a kind of dichroic mirror. So the PhC device can transmit or reflect the light according to the wavelength of an incident light because of the PBG [4]. And a self-imaging phenomenon in the MMI device is the property that the profile of an incident wave is reproduced with a single or multiple images periodically along the propagation direction in the MMI device [5]. By using these two properties, an ultra compact demultiplexer can be realized.

In this study, 2D PhC structure consisting of a square lattice of dielectric rods in the air was used to get a wide band gap. The refractive index of the dielectric rods is set to be 3.4 and the radius(r) is 0.18a, where a is the lattice constant of the crystal. In this structure, the band gap opens for the frequency ranges of 0.303-0.445 (a/λ) for the E-polarization (electric field parallel to the rods) where λ is the wavelength in free space.

Schematic diagram of the proposed 1.3/1.55μm demultiplexer is depicted in Fig. 1. In this structure, five consecutive rows were removed to form a multi-mode PhC waveguide. And as an access waveguide, one-line-defect PhC waveguide was introduced. This
access PhC waveguide supports single-mode. We choose an operating frequency of 0.35 \((a/\lambda)\) and 0.417 \((a/\lambda)\), these are corresponded to the wavelength of 1.55\(\mu\)m and 1.3\(\mu\)m by setting the lattice constant as 542.5\(\text{nm}\).

\(L_{m, 1.3\mu m}\) and \(L_{m, 1.55\mu m}\) is the length of the first mirrored image for the wavelength of 1.3\(\mu\)m and 1.55\(\mu\)m, respectively. The size defects of the rods \((r = 0.33a)\) were formed in the position of \(L_{m, 1.3\mu m}/2\) within the multi-mode PhC waveguide. In this structure, the band gap opens for the frequency ranges of 0.392-0.479 \((a/\lambda)\). Therefore the wavelength of 1.3\(\mu\)m could be reflected from the defects and the wavelength of 1.55\(\mu\)m could be transmitted through the defects.

The characteristics of the proposed device were analyzed and optimized by using the FDTD method. Fig. 2 shows the steady-state electric field distribution for the frequency of 0.417 \((a/\lambda)\) and 0.35 \((a/\lambda)\), respectively. As we expected before, the wavelength of 1.3\(\mu\)m reflected and the wavelength of 1.55\(\mu\)m transmitted through the defect as shown in Fig. 2. The transmittance of the device is about 98.8\% for 1.3\(\mu\)m and 87.7\% for 1.55\(\mu\)m, respectively. The extinction ratio was defined as the ratio of the power at the desired output port to the power at an undesired port for a specific wavelength. In the optimally designed structure, the extinction ratios for 1.3\(\mu\)m and 1.55\(\mu\)m were less than about -21.7 dB and -16.3 dB, respectively.

3. Conclusion

Optimally designed 1.3/1.55\(\mu\)m demultiplexer has the size of 10.3 x 7.6\(\mu\)m\(^2\), it is believed to be the smallest one in the state of the art to our knowledge. The transmittance of the device is about 98.8\% for 1.3\(\mu\)m and 87.7\% for 1.55\(\mu\)m, respectively. And the extinction ratio of the proposed device is less than about -21.7 dB for 1.3\(\mu\)m and -16.3 dB for 1.55\(\mu\)m, respectively. Now we are trying to enhance the extinction ratio for 1.55\(\mu\)m.

Acknowledgement

This work was supported by grant No. R11-2003-022 from the Engineering Research Center of the Korea Science and Engineering Foundation.
Design and fabrication of optical multichannel filter with changeable channel spacing through phase sampling

Li Xia, Ping Shum, and Cheng Tee Hiang
(Network Technology Research Centre, Nanyang Technological University, Singapore 639798
Tel: +65-67904527, Fax: +65-67926894, Email: xiali@ntu.edu.sg)

Abstract
An optical multichannel filter with changeable channel spacing is designed in the chirped fiber Bragg grating (CFBG) through phase sampling technique, and we achieve the wavelength spacing changes from 0.65 nm to 1.15 nm among 23 channels.

1 Introduction
Nowadays, optical fiber passive devices are more attractive in various areas, due to their immunity to electromagnetic interference, high sensitivity, compactness and simplicity of fabrication. Among these, optical multichannel filter is particularly useful in several situations, such as wavelength division multiplexed system and network, multiwavelength fiber lasers, optical sensors, etc. Among the approaches, fiber Bragg grating (FBG) has received much attention to generate a number of equally spaced reflective channels through imposing a periodic modulation sampling on amplitude or phase parameter [1, 2]. These equally spaced channels come from the constant sampling period. In fact, the applications of unequally spaced filtering, different microwave frequency generating by optical interference, channel spacing tuning, demand on the flexible channel spacing optical devices, especially different channel spacing integrated in one device. Therefore, the technique needs to be enhanced to achieve the special requirements. In this work, we demonstrate an optical multichannel filter which can achieve different channel spacing filtering in one FBG. The filter is designed through introducing different phase shift at different CFBG position according to different sampling period. The phase sampling technique requires the lower index modulation, and makes the approach effective and flexible.

2 Filter design and principle
Traditionally, a multichannel filter based on chirped sampled FBG and phase shift has equal channel spacing [3]. The sampling period $sp$ is constant along the FBG length and the phase shift $k\phi$ is introduced before the $k$th sample. In order to obtain changeable channel spacing, in our approach, different sampling period $sp_k$ is applied. And corresponding phase shift $\Phi_k$ is calculated through Talbot effect existed in FBG [4]:

$$\frac{2\pi g_1}{\Lambda_0}sp_k^2 + \phi_k = \frac{2\pi}{T}$$

Here, $cg_1$ is the chirp coefficient of grating, $\Lambda_0$ is the central grating pitch, and $T$ is a positive integer. When $T$ is an infinite integer, the corresponding phase shift $\Phi_k$ can be expressed below:

$$\phi_k = -\frac{2\pi g_1}{\Lambda_0}sp_k^2$$

The scheme of FBG inscribing with phase sampling technique is illustrated in Fig. 1. The whole grating length is divided into multiple sub-gratings with different length, phase shift, and constant index modulation.

![Fig. 1 The scheme of phase sampling.](image)

3 Simulation and experimental results
In order to verify our design principle, here, a transfer matrix approach is adopted to calculate the corresponding reflection spectrum [5]. The beginning sampling period
sp₀=0.6916 mm and phase shift Φ₀=0, leads to the initial channel spacing of around 1.2 nm. The next sampling period spₖ=sp₀+k×Δsp, (k=1, 2…) with sampling period increment amount Δsp=2 μm. Other parameters are chosen as grating central pitch Λ₀=529.02 nm, chirp coefficient cg₁=1.125 nm/cm, coupling coefficient κ=500 m⁻¹. The simulation result is plotted in Fig. 2.

Based on our simulation condition, a changeable channel spacing filter with 23-channel is fabricated. In the experiment, the phase mask scanning method is adopted, meanwhile, the PZT (the precision <10 nm) is used to introduce the different phase shift amount. The effective refractive index of used photosensitive fiber n=1.4565. The reflection spectrum is plotted in Fig. 3, showing a changeable channel spacing performance across a bandwidth of nearly 18.5 nm. The smallest wavelength spacing, which lies in the shorter wavelength area, is around 0.65 nm. And the largest wavelength spacing is around 1.15 nm at longer wavelength area. The side mode suppression ratio at each channel is beyond 20 dB. It is noted that the sampling period increment amount Δsp should be carefully selected. If Δsp is too large, the reflectivity of each channel, as well as the side mode suppression ratio, will decrease dramatically. When Δsp=0, the channel spacing is constant among the whole bandwidth.

Fig. 4 The channel spacing distribution.

4 Conclusions

In summary, an optical multichannel filter with changeable channel spacing is proposed and designed. The channel spacing can vary from 0.65 nm to 1.15 nm among around 18.5 nm bandwidth along the grating length of 6 cm. The initial spacing and last spacing can be easily controlled by carefully selecting the starting sampling period and its increment amount. The introduced phase sampling technique makes it easy to fabricate various channel spacing using one chirped phase mask. This kind of mulíchannel filter is potentially useful for arbitrary wavelength spacing filtering, multiple microwave frequency generating, etc.

References

Data Vortex switch network with shared cylindrical traffic control

Qimin Yang
Harvey Mudd College, Engineering Department, Claremont, CA 91711, qimin_yang@hmc.edu

Abstract
Data Vortex networks that operate with shared traffic control at each cylinder is shown to improve routing performance in both throughput and latency, and such control mechanism is especially beneficial at high load operations.

Introduction
Packet switched optical interconnection networks are key subsystems in high capacity communication systems. Due to limited optical processing and buffering capabilities, the main challenge in packet switching is to deal with traffic contention. Alternatively, Data Vortex networks implement a single-packet-routing rule at each node through a traffic control mechanism combined with deflection routing as well as the virtual buffering topology.

Fig. 1. Routing nodes and links within Data Vortex of A=5, H=4. (Note: the control links are not shown for clarity).

Data Vortex routing nodes are arranged in concentric cylinders with A, H and \( C = \log_2 H + 1 \) designating the number of nodes along angle, height and cylinder respectively, as shown in Fig. 1. The links along the same cylinder route a packet back and forth between two height groups, while the links to a neighbor cylinder maintain the same height. During the synchronous operation, each node routes a single arrival packet to either the same or the next inner cylinder based on the packet’s target and the inner node’s traffic. A traffic control signal properly permits or blocks the packet to the inner cylinder so that the single-packet-routing rule is satisfied. Packets blocked are deflected to stay on the current cylinder for virtual buffering purpose with slight delay penalty. More detailed description of the Data Vortex network and physical layer testbed can be found in [1][2][3][4].

Shared traffic control mechanism
Although it facilitates Data Vortex’s optical implementation, the deflection mechanism introduces traffic backpressure at high load conditions which compromises the throughput and latency performance. The much better performance is achieved only if the sufficient routing redundancy, i.e. small enough \( A_n/A \) is provided, where \( A_n \) is the number of I/O or injection angles. Therefore the system may require high cost for the given number of I/O ports.

To improve the routing efficiency, we propose to share cylindrical traffic

Fig. 2 (a) Traffic controller in regular Data Vortex
Fig. 2(b) Data Vortex with shared traffic controller
information within the control mechanism. As shown in Fig.2 (a), in regular Data Vortex nodes of the same height and same cylinder make the routing decisions individually. Since these nodes are connected to the same height in the next cylinder, sharing traffic information among these nodes allows for reuse of the nodes when their own inputs become idle. As an example, Fig.2 (b) shows that by adding a 1x1 switch and two couplers, in the case of deflection in regular Data Vortex, we may instead route the packet to the neighbor angle node at the same height and same cylinder if that node has idle input traffic. In system implementation, nodes of same height and same cylinder can be localized in one unit, so very slight delay can be introduced between different coupling points to switch the path at proper time based on the inputs as well as the controls from the inner cylinder.

It is possible to further increase the routing efficiency if nodes of non neighboring angles share traffic information and routing paths. However this requires more advanced components such as fast reconfigurable MxM switches in order to set up alternative paths without significant delay within the shared controllers.

**Performance evaluation**

To verify the feasibility of the new control mechanism and the potential improvement in the routing efficiency, we simulate the new Data Vortex routing for various traffic and network conditions. We only allow for a packet to be redirected to one neighbor angle node with shared traffic control mechanism because of its simplicity and feasibility in implementation. We particularly focus on cases when the routing redundancy is rather limited and when the deflection induced backpressure effect is more significant. This is typically when the Data Vortex advantages get diminished compared with other interconnection network topologies.

In Fig.3 (a) and Fig.3 (b), the performance of a network A=5 and H=512 with and without shared control mechanism are compared. The throughput represented by the successful injection ratio at the input ports has been improved dramatically, especially at high load conditions. For example when load ≥ 0.5 in A_in=5 case, the improvement is roughly 45%-50%. The latency by the average number of hops is also improved with the shared traffic control and an average of 5-6 hops reduction can be achieved. The performance improvement in A_in=3 case has similar trends, and such improvement is also shown to be similar for different network sizes.

**Conclusion**

Data Vortex with a new traffic control mechanism is proposed. By sharing the traffic conditions among the nodes of the same height and same cylinder, routing performance can be improved greatly while the complexity of the routing node is kept reasonably simple.

**Reference:**

On the Challenges of Transporting RF Signals over an FSO Channel

Pham Tien Dat†, Alam Shah‡, Kamugisha Kazaura‡, Kazunori Omae†, Toshiji Suzuki†, Harukazu Watanabe†, Mitsuji Matsumoto†, Takeshi Higashino†, Yuji Aburakawa†, Katsutoshi Tsukamoto†, Shozo Komaki‡, Takuya Nakamura‡, Kazuhiko Wakamori†, Koichi Takahashi‡

† GITS/GITI, Waseda University, 1011 Okuboyama, Nishitomida, Honjo-shi, 367-0035, Japan
‡ Division of Electrical, Electronic and Information Engineering, Graduate School of Engineering, Osaka University, 2-1 Yamada-oka, Suita-shi, 565-0871, Japan

Abstract

In free-space optical (FSO) communication links, atmospheric turbulence induced intensity fluctuations can significantly impair the link performance. In this paper, we explore the challenges on utilizing FSO links to transport modulated radio frequency (RF) analog signals.

1. Introduction

Free Space Optical (FSO) communication technology has been successfully utilized to offer high-speed digital transmission for a variety of applications. FSO can provide a cost effective alternative to fiber optic systems in last mile applications, enterprise connectivity, metro network extension, fiber backup. They have many advantages over fiber optic systems which include easier and faster deployment, and in some cases, more economical. This is important, especially in the urban areas where it is difficult to install fiber, or in presence geographical barriers like roads or rivers. It also has many advantages compared to other wireless technologies, including much higher data rates and no licensing requirement or frequency allocation. Beside the above-mentioned applications, FSO links can be attractive means for RF signal transmission. FSO can play a similar role to an optical fiber in the layer 1 of the Radio over Fiber (RoF) network to transport radio signal in some applications, such as CATV, cellular radio and WLAN. However, up to now, not enough research has been devoted on this. One of our project objectives is to establish an experimental RoFSO link with a distance up to 1 km. This link will be used to examine transmission performance for some kinds of RF signals, including CDMA, terrestrial digital TV and WLAN signals over an extended period and measure system parameters to verify the suitability of FSO technology for RF signals transmission [1].

This paper introduces the RoFSO concept and report some measured result of refractive index structure constant parameters, which characterize FSO scintillation channels.

2. RoFSO systems

Optical fiber has traditionally been used for transmission of both digital and analog signals. The transmission of RF intensity-modulated signals over optical fibers has been widely investigated and applied. By using RoF, the capacity of optical network can be combined with the flexibility and mobility of wireless access network. Since RoF involves analog modulation, and detection of light, it is fundamentally an analog transmission system. Therefore, signal impairments such as noise and distortion, which are important in analogue communication systems, are important in RoF and RoFSO systems as well. These impairments tend to limit the Noise Figure (NF) and Dynamic Range (DR) of the RoFSO links. Although RoF and RoFSO transmission system itself is analog, the radio system being distributed need not be analog as well, but it may be digital (e.g. WLAN, UMTS), using comprehensive multi-level signal modulation formats such as QAM, or Orthogonal Frequency Division Multiplexing (OFDM).

In [2] the authors report experiment work on transmitting multiple RF signals over a single FSO link using WDM technology. The link distance is 3 m which is sufficient to characterize the end-to-end communication channel while omitting the atmospheric losses. The transmission response and dynamic range measurements results demonstrate suitability of FSO links for transmission of RF signals.

Similar work is reported in [3], in which the authors analyzed the most important parameter in the analog optical systems, i.e. spurious free dynamic range (SFDR) and thereafter set up a full duplex optical link with a distance of 500 m to transfer cellular signals using 810 nm and 1550 nm wavelengths. The performance analysis showed that FSO links can be satisfying for both CDMA and GSM cellular techniques.

It should be borne in mind that in FSO links, atmospheric turbulence has a significant impact on the optical beam propagating through the atmosphere over long distances. Research has focused on mitigating these atmospheric induced effects which include beam broadening, scintillation and angle-of-arrival fluctuations to improve the performance and reliability of FSO links.

3. RoFSO system concept

Figure 1 depicts the concept of multiple RF signals transported over an FSO link using DWDM [1]. In our experimental development, we want to realize an FSO link which has an equivalent capacity for heterogeneous wireless services as fiber
in RoF networks. Direct optical amplification and emission of RoF signal into free space will be employed and at the receiver the signal will be focused directly to the SMF core using receiver optics. This technology has successfully been demonstrated in our previous work reported in [4]. Since RoF is essentially an analog transmission, higher stability and reliability will be required for RoFSO link. In the development, therefore, the accuracy of the optical beam tracking mechanism should be improved as well as improvement in the atmospheric induced turbulence mitigation techniques so as to suppress the errors occurring during periods of strong atmospheric turbulence occurring mostly at midday.

4. RoFSO system evaluation consideration

The performance of RoFSO links can be evaluated in the same generally accepted criteria typically applied to conventional RF links. Standard performance criteria for conventional RF links include RF loss, frequency response. In RoFSO links, additional noise generated by the laser source and the photodiode can degrade the carrier-to-noise ratio CNR, and nonlinearity of the modulation device can reduce the SFDR.

The important RF performance criteria for RoFSO include RF gain and frequency response, additional noise and CNR and SFDR.

Some of the critical parameters that should be measured to evaluate the system performance include:

a) Optical Power
b) Transmission Response: provide the relative loss, or gain, in a communications link with respect to the signal frequency. Any signal attenuation due to the communications link will manifest itself in the transmission response measurement.

c) Reflection Response: provides an evaluation of the amount of reflected power relative to the incident power versus frequency at the insertion input of the transmitter.
d) Group Delay: the group delay is a measure of the total delay a signal experiences when traversing a communications link, which thus gives rise to a phase shift in the signal.

e) CNR Measurements: CNR is an important measure that quantifies the performance of communication channels relatively to the existing noise. CNR plays an important role in determining the minimum average signal power that can provide error-free transmission through a communications channel.

In our experiment work, the above-mentioned RoFSO system performance parameters will be measured and correlated with the atmospheric induced effects. The refractive index structure constant parameter, $C_n^2$, is one of the critical parameters along the propagation path in characterizing the effects of atmospheric turbulence. From previous FSO experimental measurement data, the variation of the midday maximum value of $C_n^2$ can change by a factor of 2.3 for winter and summer seasons. This is significant because the RoFSO experiment will be conducted at same location. Figure 2 shows measured values of $C_n^2$ in our deployment environment with occurrence of higher values of $C_n^2$ ($\geq 10^{-13}$ m$^{-2/3}$) around midday indicating strong atmospheric turbulence thus increase in the degree of scintillation.

Considering that different wireless services have different data-rate, modulation formats and required sensitivity, it is necessary to investigate the system deterioration as a result of variation of $C_n^2$ for each wireless service. From numerical results reported in [1], considering WLAN signal formats, a BER of $10^{-3}$ can be achieved for a scintillation variance of 0.35 for 1 km RoFSO link when the received power is $-10$ dBm.

The objective of our work is by using a similar concept for the previously developed experimental FSO system capable of transmitting WDM digital signals [4], an RoFSO system capable of transporting more than 4 different aforementioned wireless services will be conducted. With the 1 km RoFSO link, the experimental results and performance analysis will reflect on actual deployment scenario, including the atmospheric induced effects like scintillation. This will provide an opportunity to derive more accurate performance of RoFSO compared to previously reported work in [2] and [3], which were performed in the laboratory environment with/or short link distances, and therefore omitting the effects of atmosphere in performance analysis of RoFSO systems.

![Figure 2 Variation of $C_n^2$ values for one month period](image)

5. Conclusions

We have outlined the design concept and important parameters that should be considered in the evaluation of RoFSO system performance. Studies on the performance of the actual implementation of 1 km RoFSO link under various atmospheric and weather conditions will be helpful in characterizing the end-to-end communication channel performance and suitability of transporting heterogeneous wireless services using FSO links.

Acknowledgement

This work is supported by a grant from the National Institute of Information and Communication Technology (NICT) of Japan.

References


Proposal for Simple Bit Timing Synchronization
Technique of Mapping Optical Burst Signals
for Optical Burst Switching

Ayako Iwaki, Motoharu Matsuura, Naoto Kishi, and Tetsuya Miki
Department of Information and Communication Engineering, University of Electro-Communications,
1-5-1 Chofugaoka, Chofu-shi, Tokyo 182-8585, JAPAN
Tel: +81-42-443-5218, Fax: +81-42-490-7096, Email: iwaki@ice.uec.ac.jp

Abstract
A simple bit timing synchronization technique for optical burst switching was proposed by monitoring the timing lag between optical burst signals.

1 Introduction
In future photonic networks, optical burst switching (OBS) is one of the attractive technologies due to fine bandwidth granularity and high feasibility of hardware architecture. To send control information for bandwidth allocation in OBS, two schemes have already been proposed; one is to use a channel for the control signal differing from data channel, and the other is to transmit on a channel same as data channel. Although the latter has a better channel availability for data transmission, it requires a bit timing synchronization between the control and the data signals, since both signals must be received by an optical receiver.

On the other hand, in the case of optical time division multiplexing (OTDM) systems, a number of synchronization techniques have already been proposed [1], so far. However, it is difficult to realize precise bit timing synchronization. To solve this problem, we proposed a novel bit timing synchronization technique between optical burst data.

2 Mapping burst data in optical frame
In dynamic path photonic networks [2][3], the optical burst data are mapped into an optical frame and dropped at a destination node at header information of the optical frame.

Fig. 1 shows the configuration of the optical burst data mapping using an optical switch. An optical burst data from a source node is mapped into an optical frame using a 2 × 1 optical switch. The optical burst data is injected into an input port of the optical switch, while the optical frame is injected into the other port. At the same time, an electrical control signal is injected into the optical switch from the source node. The electrical control signal is employed to change passed input port of the optical switch.

3 Bit timing synchronization
Fig. 2 shows configuration of the proposed bit timing synchronization system. The bit timing synchronization system consists of a timing lag meter and a tunable fiber delay line (FDL). The optical frame and burst data are launched into the lag meter, and the timing lag is detected. Using the feedback circuit based on the FDL, the timing lag can be controlled automatically.

The schematic view of the lag meter with dual-stage AND gates is depicted in Fig. 3(a). At the AND gate #1 of the timing lag meter, the part of continuous bit pattern of “010101...” in the burst signal is drawn by using the window signal as shown in Fig. 3(b). The clipped signal from the optical frame consist of a part of the padding of header, which has a continuous bit
pattern of “010101...”. The signal from the source node also has same bit pattern. Next, the frame signal and the output signal at the AND gate #1 are injected into the AND gate #2 in order to measure the overlap part of the mark bits between the frame and burst signals with each same bit pattern. Thus, the timing lag can be evaluated by the output average power at the output of the dual-stage AND gate, because the power is determined by the overlapped area of mark bits. Since high speed receiver is not needed in this scheme, we will realize a cost effective bit timing synchronization.

4 Experimental verification of the timing lag meter

To verify the availability of the proposed bit timing synchronization technique using AND gates, we demonstrated experimentally a single stage AND gate scheme by means of four-wave mixing (FWM) [4] in a semiconductor optical amplifier (SOA). Fig. 4 shows the experimental setup of the AND gate. The input lights were coupled by a 3-dB coupler and modulated by a 10 Gb/s data with the bit pattern of “010101...” using a LiNbO$_3$ modulator (LNM). The modulated lights were demultiplexed by an array waveguide grating (AWG). One of them was passed through a FDL. Erbium Doped Fiber Amplifiers (EDFAs) were employed to compensate the insertion loss of 3-dB coupler, LNM and AWG. After that, the signals were coupled by 3-dB coupler and injected into a SOA. In the SOA, the overlapped light component between two lights was generated by FWM, and only passed through the AWG. At the output of the AWG, the average output power was monitored by an optical power meter. In this experiment, the wavelength of input lights were 1555.98 nm and 1556.79 nm, and their input powers into the SOA were 8 dBm and 5 dBm, respectively.

Fig. 4 Experimental setup of the AND gate #2

Fig. 5 shows the measured output average power with changing the time delay of the FDL. It can be clearly seen that the output power was changed periodically, and the period corresponded to the 2 bit time of the 10 Gb/s data, ‘0’ and ‘1’.

This indicates that the timing lag can be detected by the output average power. Therefore, the timing lag will be controlled automatically, if the time delay of the FDL is adjusted by monitoring the change of the output power.

5 Conclusion

We proposed a bit timing synchronization system for optical burst switching by measuring the timing lag between two optical signals. We also verified the operation using a prototype of the timing lag meter.

References

Millimeter-wave true-time delay measurement in WDM-based optically controlled array antenna

Wataru Ohuchi\(^1\), Wataru Chujo\(^2\), Yoshiyuki Fujino\(^2\), Yahei Koyama\(^1\)

\(^1\) Ibaraki University, 4-12-1 Nakanarusawa, Hitachi, Ibaraki 316-8511 Japan
\(^2\) National Institute of Information and Communications Technology, 3-4 Hikarino-oka, Yokosuka 239-0847 Japan

Phone: +81-46-847-5057, Fax: +81-46-847-5059, Email: chujo@nict.go.jp

Abstract Microwave/millimeter-wave, true-time delay characteristics of WDM-based optically controlled array antenna is experimentally verified. True-time delay characteristics with phase deviation of less than 3 deg. are obtained using 4-element WDM-based optically controlled array antenna in the frequency range from 20 to 40 GHz.

1 Introduction

Optically controlled beam forming of microwave/millimeter-wave phased array antenna is superior in wide/multi-bandwidth operation to microwave/millimeter-wave phase shifters and digital beam forming. Especially optical dispersion technique \([1]–[4]\) has inherently true-time delay characteristics suitable for wide/multi-bandwidth operation, although temperature instability of the delay in the optical fiber is one of the technical issues. However, WDM-based optically controlled array antenna consists of WDM optical sources and an optical dispersion fiber does not cause microwave/millimeter-wave phase instability due to temperature because only the difference in chromatic dispersion at each wavelength within the same SMF is used to control the phase of microwave/millimeter-wave signals. In this letter, we measured the millimeter-wave true-time delay in order to clarify the phase accuracy of the WDM-based optically controlled array antenna.

2 Experimental setup

Figure 1(a) shows schematic diagram of the proposed optically controlled array antenna for transmission. Thick line shows optical component and thin line shows microwave/millimeter-wave one. In the transmitting array antenna, chromatic dispersion of a conventional single mode fiber is used to control microwave/millimeter-wave phase to be fed to each antenna element for beam steering. First, optical carrier \(\lambda_1 – \lambda_n\) with different delay is detected by photodiode after DEMUX. Since obtained microwave/millimeter-wave phase at each antenna element is different, antenna beam is able to be steered to arbitrary direction by adjusting the optical wavelengths and SMF length. Delay time difference \(r_{opt}\) between adjacent wavelengths can be expressed as

\[
r_{opt} = DL\lambda_{opt}
\]

where \(D\) is wavelength dispersion of the SMF, \(L\) is the SMF length and \(\lambda_{opt}\) is wavelength spacing. Microwave/millimeter-wave phase difference \(\Delta \phi\) between adjacent antenna element can be expressed as

\[
\Delta \phi = 2\pi DL\lambda_{opt}f_{RF}.
\]

where \(f_{RF}\) is microwave/millimeter–wave frequency.

Since only one SMF acts as microwave/millimeter–wave phase shifters, number of phase shifters can be reduced when optical carrier of a different wavelength is assigned to each antenna element. In addition, since wavelength dispersion characteristics of the SMF is invariant and each wavelength

![Schematic diagram and (b) experimental setup of WDM-based optically controlled array antenna.](image-url)
can be adjusted to control the phase of corresponding microwave/millimeter–wave signal. WDM-based control is less affected by surrounding environmental condition including temperature and expected as one of ways to be able to attain very stable microwave/millimeter–wave phase shifters. Figure 1(b) shows experimental setup of WDM-based optically controlled array antenna. We carried out phase control experiments using 4-element array antenna for transmission in the frequency range from 20 to 40 GHz. Four DFB lasers with wavelengths of 1559.79, 1558.98, 1558.17 and 1557.36 nm are multiplexed and modulated by 20 - 40 GHz signals. Length of SMF for optical delay is changed in 20, 30 and 40 GHz, respectively. (b) Figure 2 shows obtained 20 - 40 GHz phase characteristics of 4-element WDM-based optically controlled array antenna for transmission. SMF length is 100, 70 and 50 m at $f = 20$, 30 and 40 GHz, respectively in Fig. 2(a) and 130, 100 and 100 m at 20, 30 and 40 GHz, respectively in Fig. 2(b). Microwave/millimeter-wave phases relative to the phase corresponding to $\lambda_\text{d} = 1559.79$ nm are shown. Since both experimental and calculated phases are coincided very well, standard deviation of the obtained phase at each microwave/millimeter-wave frequency is shown in Table 1 to know the phase variation correctly. Although obtained phase deviation is slightly increased as microwave/millimeter-wave frequency is increased, phase deviation of less than 3 deg. at 40 GHz is realized.

3 Experimental results

Figure 2 shows obtained 20 - 40 GHz phase characteristics of 4-element WDM-based optically controlled array antenna for transmission. SMF length $L$ is 100, 70 and 50 m at $f = 20$, 30 and 40 GHz, respectively. Microwave/millimeter-wave phases relative to the phase corresponding to $\lambda_\text{d} = 1559.79$ nm are shown. Since both experimental and calculated phases are coincided very well, standard deviation of the obtained phase at each microwave/millimeter-wave frequency is shown in Table 1 to know the phase variation correctly. Although obtained phase deviation is slightly increased as microwave/millimeter-wave frequency is increased, phase deviation of less than 3 deg. at 40 GHz is realized.

Fig. 2 Measured phase characteristics of 4-element optically controlled array antenna. (a) SMF length $L = 100$, 70 and 50 m at $f = 20$, 30 and 40 GHz, respectively. (b) $L = 200$, 130 and 100 m at $f = 20$, 30 and 40 GHz, respectively.

Table 1 Standard deviation of microwave/millimeter-wave phase at each frequency and wavelength.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>SMF Length (m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
</tr>
<tr>
<td>40</td>
<td>40</td>
</tr>
</tbody>
</table>

Relative phase [deg.]

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>SMF Length (m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
</tr>
<tr>
<td>40</td>
<td>40</td>
</tr>
</tbody>
</table>

Relative amplitude [dB]

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>SMF Length (m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
</tr>
<tr>
<td>40</td>
<td>40</td>
</tr>
</tbody>
</table>

4 Conclusion

20-40 GHz phase accuracy of 4-element WDM-based, true-time delay optically controlled array antenna was investigated. WDM-based microwave/millimeter-wave phase shifters composed of four DFB lasers and one SMF is experimentally verified. Measured 20 – 40 GHz phases are coincided well with calculated values and phase deviation of less than 3 deg. are realized independent of microwave/millimeter-wave frequencies.

References

Two Novel Configurations and an Adaptive Round-Time MAC for Light-trail IP-Centric Applications

Nguyen Cac Tran¹, Ealle Kim¹, Kwangjoon Kim² and Jinwoo Park¹
¹Department of Electrical, Korea University, Rep. of Korea
²Electronics and Telecommunications Research Institute, Rep. of Korea
tncac@korea.ac.kr

Abstract
To enhance light-trail robustness, we proposed two novel light-trail configurations: Light-ring and light-bus and a light-trail medium access protocol to efficiently support the highly dynamic traffic patterns of while maintain the fair access.

1. Introduction
Light-trail is promising solution for sub-wavelength rate service in optical domain that make WDM optical network more flexible as well as cost-effective. A light-trail is similar to light path in that it requires the establishment of a circuit between the source and destination. The key difference is that intermediate nodes are able to access into the circuit for transmitting and receiving data in a time multiplexed manner.

To implementation light-trail technology in current metro optical network, this paper present two new light-trail configurations and the way it is seamlessly combined with reconfigurable add drop multiplexer (ROADM). Because the light-trail efficiency depends on a light-trail medium access control protocol (MAC), sect. 3 presents a novel light-trail MAC method in detail and those analyses. Finally, Sect. 4 concludes the paper.

2. Light-Trail Configurations
Light-trail has inherent characteristic that is light signal propagating in the downstream direction from head node to tail node so that conventional CSMA/CD MAC is impractical because of unfairness problem[3]. In most cases, the only possible choice is scheduling method that need to exchange in-band signaling messages among light-trail node with very low delay. To solve this problem, this paper proposes light-trail ring and bidirectional light-trail configuration called light-bus, respectively in Fig.2. b) and c).

The ability of in-band signaling induces the great advantage to design a MAC protocol based on scheduling to adapt traffic patterns and maximum wavelength utilization. Additionally, almost current metro optical networks are based on ring or multiple ring topologies and the Reconfigurable Optical Add Drop Multiplexer (ROADM) is being deployed that consistent with light-ring concept. We proposed an approach to implement light-ring in Fig. 3 that consists of a ROADM with light-trail extension module. The user interface can be arbitrary however with the dominance of Ethernet technology and the emerging of metro Ethernet carriers this interface should be Ethernet interface. Therefore, it promises to be a practical and cost-effective approach to implement light-trail. In case of light-bus, conventional CSMA/CD can be used similar to Ethernet LAN.
3. Adaptive Round-Time Light-Trail MAC

Adaptive Round-Time (ART) MAC is described as follow, \( T_{\text{max}} \) is defined as maximum period for each node to access light-trail in one round. A particular node finishes transmission either \( T_{\text{max}} \) is reached or its buffer is empty. Its actual light-trail access period is added to Accumulative Access Period Variable (AAPV) that is sent to next node via a message called Downstream Control (DSC) message. After sending DSC message, current node waits for next round turn. The following node receives DSC and begins its transmission. When DSC message is received, last-transmitting node, right previous tail node, performs a time-compensated algorithm to limiting switching frequency in case of low load and diminishes propagation delay of End of Round (EoR) message. Last-transmitting node sends a signaling message EoR to head node to for starting a new round. Head node begins new round in two cases, first case, its round time counter is exceeded maximum round time (time out). Second case, EoR is received before time out and head nodes begin transmitting appropriate to actual_round_time read from EoR. Note that in case of general light-trail, EoR is send via out-of-band signaling and in case light-ring EoR can be send as in-band signaling message.

Following equations are derived to analyze ART-MAC performance

\[
T_c = \frac{(N-1)T_r + T_r}{1 - \rho} \quad (1)
\]

\[
\rho_{\text{max}} = \frac{T_{\text{max}} - (N-1)T_r}{T_{\text{max}} + T_r} \quad (2)
\]

\[
\rho_{\text{r, max}} = \frac{T_r}{T_{\text{max}} + (N-1)T_r} \quad (3)
\]

Where: \( N \) - number of nodes;

\( T_r \) - switching time between nodes;

\( T_r \) - mean round time;

\( \rho \) - total load normalized to theoretical maximum load;

\( T_c \) - compensated time (\( T_c = 0 \) when \( T_r > T_{\text{max}} + (N-1)T_r \)).

From (1) we observe that round time \( T_r \) auto-enlarges or auto-shrinks when total load \( \rho \) increases or decreases, respectively. (2) implies that to increase maximum load \( \rho_{\text{max}} \) we can increase \( T_{\text{max}} \) or try to reduce \( T_r \). (3) calculates maximum total load \( \rho_{\text{r, max}} \) in that traffic focus only one node, and other node buffers are empty. It implies that ART-MAC promptly adapts with high traffic focus on one node.

An extensive simulation uses a 5-node light-trail in bit rate 2.5 Gbps and traffic density for node 1 to 4 are divided in the ratio 4:3:2:1. Based on number of node is 5 and maximum access delay bound 10 ms, \( T_{\text{max}} \) equal to 2.5 ms and the switching time \( T_r = 300\mu s \). The numerical results of existing light-trail MAC called Light-Trail Fair Access (LT-FA) [2] was collected for comparison.

Observation from Fig. 3 shows that average delays achieved by ART-MAC are much lower than LT-FA MAC about 44% in various loads. Especially, node 2 using LT-FA is overflow in 0.8 and 0.83 total load while node 2 still works well with ART. This consistent with the claim in [1] that LT-FA performs well until loads about 0.7. The average delay of ART more stable in both aspects of various loads and among nodes compares with LT-FA.

4. Conclusion and Future Work

We have seen that light-ring and light-bus, two novel light-trail configurations, and the ART-MAC for practical light-trail implementation. The analyses above have showed that ART-MAC outperforms the existing method in both heavy traffic and traffic variation while maintain the fairness. Further work will conduct a lighting test-bed using ART-MAC.

References


Network Performance of the Multi-Protocol Optical Switch Routing Algorithm using Multimedia Packet Traffic

Bernard HL Lee1,2, Zailani Omar1, Sunarti Mohd Shukor1, Romli Mohamad1, Kaharudin Dimyati2
1TM Research & Development, UPM-MTDC, Lebuh Sillikon, 43400 Serdang, Selangor, Malaysia, 2University Malaya, Faculty of Engineering, 50603 Kuala Lumpur, Malaysia.
Tel: +60-3-89413239, Fax: +60-3-89413239, Email: bernard@tmrnd.com.my

Abstract
In this paper, authors compared the performance of the MPOS [1] routing algorithm against conventional packet schemes. The MPOS routing segregates the different types of multimedia traffic into dedicated classes/priorities and assigns a light path for each of them. Improved traffic performance has been observed by the MPOS technique at high traffic load.

1 Introduction
All-optical networks using Wavelength Division Multiplexing (WDM) and wavelength routing offers an excellent bandwidth capacity for real-time multimedia traffic such as IPTV, Voice over IP (VoIP) and also the conventional Internet Data. Furthermore, with the current efforts to automate and expedite dynamic wavelength and bandwidth provisioning over the optical layer, trends are leading towards more intelligent optical networks with more efficient routing algorithms. The concept of a single routing algorithm that suits all types of traffic may not apply to multimedia applications which require a dedicated routing algorithm that suits its individual needs.

2 Theory and Concept
Alternatively to applying a single routing algorithm to cater for all types of multimedia traffic, the MPOS routing techniques applies a ‘divide and route’ approach. Sets or clusters of wavelengths and optical switching planes are assigned to cater for each of the three main traffic types (i.e. Video, Voice and Data). The MPOS routing technique is similar to the conventional Priority/Class Routing algorithms [3] which priorities the incoming traffic and store them in individual queues depending on the packet’s priority. The switch later will retrieve the packet from the queue and route them through the switching plane and out towards the output port. When contention occurs, the packets with the highest priority will be routed first while the rest will be queued awaiting available slots. During high traffic load where the switch is unable to handle all the packets, packets with the lowest priority will be dropped to ease congestion. Although such priority routing scheme increases the chances of higher priority packets to be delivered, low priority packets suffers from high packet loss. Thus an ‘unfair’ scenario occurs.

Figure 1: MPOS Routing Algorithm

Figure 2: (a) Conventional and (b) MPOS Priority Routing

In the MPOS routing scheme, incoming packets are also prioritized and put into their individual queues according to their priorities, similar to the conventional method. But instead of having each switching plane routing packets from all priority classes, each of them will only route packets from
a single priority class. Therefore the MPOS routing technique avoids contention among priority classes. Contention will only occurs among packets with the same priority thus reducing high packet loss for a specific traffic type.

3 Simulation Results

The objective of this simulation setup (Figure 3) is to investigate the performance differentiation between the MPOS method versus the conventional by use of scheduling at the router interface. Two scenarios are simulated using OPNET, which are the conventional method and the MPOS method in a 6 node mesh network. The multimedia traffics used in this case study are video conferencing, voice over IP (VoIP) and data traffic (web browsing, FTP, e-mail and database). For both conventional and MPOS methods, web browsing, FTP, e-mail and database uses exponential inter-arrival time and each is set with constant transaction/packet/file size. Traffics are set to best efforts except for video streaming traffic. From the simulated results, the average packet jitter and the average packet delay variation are collected for both scenarios. In packet jitter, if two consecutive packets leave the source node with time stamps t1 & t2 and are played back at the destination node at time t3 & t4, then:

\[ \text{Jitter} = (t_4 - t_3) - (t_2 - t_1) \]

Negative jitter indicates that the time difference between the packets at the destination node was less than that at the source node. On the other hand, packet delay variation is the variance among end to end delays for packets received by each node. End to end delay for a voice packet is measured from the time it is created to the time it is received. As the network utilization increases from 0% to 100% (occurs at 2m) packet jitter occurs in the conventional approach but not in the MPOS. Packet delay also increases with network utilization for the conventional approach but not the MPOS. Therefore, both the Packet Jitter and Packet Delay Variation results show that the MPOS approach has an improved performance over the conventional method when the network utilization reaches 100% (approximately at 2m). This is due to the dedicated link for each type of traffic thus avoiding contention between different classes.

4 Conclusion

In this paper we have shown a novel routing method which has an improved performance over the conventional IP routing with multimedia traffic. At high network utilization, the MPOS routing method was able to maintain the end-to-end delay variation and also avoided packet jitter.

5 References

Employment of Optical Phase Conjugators in Transparent DWDM Single-Ring Network

Nuttha Chuenprasertsuk\textsuperscript{1} and Pasu Kaewplung\textsuperscript{2}
Department of Electrical Engineering, Faculty of Engineering, Chulalongkorn University, Bangkok, Thailand.
E-mail: \textsuperscript{1}jugjunG\textsuperscript{1}@Gmail.com, \textsuperscript{2}Pasu.K@chula.ac.th

Abstract - The employment of optical phase conjugator (OPC) in DWDM single-ring network is studied. We propose an algorithm that can ensure the minimum number of OPC, as well as the most suitable position in the network.

I. Introduction
The SONET/SDH-based ring topology has been widely employed as metro-area networks (MANs) owing to its provision of reliability, superior protection and restoration schemes. However, the exponential increase in transmission capacity nowadays demands for the migration from the legacy time-division multiplexed (TDM)-based technology to the transparent dense-wavelength-division multiplexing (DWDM) scheme over the incumbent SONET/SDH ring networks. When an optical signal remains in optical domain during the transmission in a network that has relatively long links, signal power attenuation, dispersion, and the Kerr effect become main sources of signal distortion. For compensating the fiber attenuation, the optimum amplifier placement for the transparent ring network has been already studied [1]. However, an attractive method for overcoming the dispersion and the Kerr effect in the ring network still has never been reported.

For dispersion compensation, the optical phase conjugation technique is very interesting since the Kerr effect is by-product reduced by performing optical phase conjugation [2], [3]. Nevertheless, up to now, the optical phase conjugation has been focused only on long-haul point-to-point applications, and has never been extended its performance to any scales of optical networks [4], [5].

In this paper, we propose, for the first time in our knowledge, a practical algorithm for the optimum placement of optical phase conjugators (OPCs) in DWDM single-ring network in order to overcome the dispersion. By applying our propose algorithm to a sample network consisted of 6 nodes with total length of 637 km, 12 OPCs are necessary for all communication traffic to be able to transmit in the network within a dispersion limit.

II. Algorithm for OPC Placement
The network sample for demonstrating our algorithm is shown in Fig. 1. The network is consisted of 6 nodes, 5 links, and the communication in the network can be bidirectional. Our algorithm can be explained in 3 main steps.

A. Find all possible traffics
The total number of traffic, which is the shortest path, between any 2 nodes is \(N^2-(N-1)\) where \(N\) is number of nodes in network. From the sample network, we show all possible traffics in term of distances from sources to destinations in Table 1.

B. Find placement ranges of OPC
In the dispersion compensation mechanism using the optical phase conjugation, the accumulated dispersion along the length of the fiber before the OPC is cancelled by the amount of the accumulated dispersion along the fiber length after the signal is phase-conjugated by the OPC. Thus, among the different dispersion values resulted from the wavelengths used in the network, we must choose the maximum dispersion value for determining the locations of OPC since the channel wavelength that exhibits the maximum dispersion will always reach the maximum dispersion tolerance \(D_{\text{max}}\) at the distance shorter than other channels. Next, for each traffic obtained from step A, we explore the possible range of the location of OPCs with the constraint that both the accumulated dispersion values entering the OPCs and the destinations must be less than \(D_{\text{max}}\), and at the same time the possible location range of OPCs must yield the minimum number of OPC. This procedure will be repeated for obtaining the location ranges of OPCs for all possible traffics in the network.

For our sample network, we assign 5 wavelengths centered at 1550 nm with spacing of 0.8 nm. The standard single-mode fiber (SMF, G.652) with the fiber dispersion \(D_s\) equals to 16.5 ps/km/nm, and the dispersion slope \(D_s^\prime\) of 0.05 ps/ km/nm\(^2\), both at 1550 nm, is used for signal transmission. As a result, the maximum channel...
dispersion can be calculated by using $D_z$ and $D_{z'}$, and is 16.58 ps/km/nm. With $D_{max} = \pm 1600$ ps/nm, Table 2 shows the possible range of location for installing the OPCs, in term of distance from source to destination, correspondingly to each possible traffic shown in Table 1.

<table>
<thead>
<tr>
<th>Traffic</th>
<th>OPC Location Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>1→2</td>
<td>41-96</td>
</tr>
<tr>
<td>1→3</td>
<td>78-96</td>
</tr>
<tr>
<td>1→4</td>
<td>61-96, 183-288</td>
</tr>
<tr>
<td>1→5</td>
<td>66-96</td>
</tr>
<tr>
<td>1→6</td>
<td>no need OPC</td>
</tr>
<tr>
<td>4→1</td>
<td>61-96, 183-288</td>
</tr>
<tr>
<td>4→2</td>
<td>62-96</td>
</tr>
<tr>
<td>4→3</td>
<td>no need OPC</td>
</tr>
<tr>
<td>4→5</td>
<td>45-96</td>
</tr>
<tr>
<td>4→6</td>
<td>79-96</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Traffic</th>
<th>OPC Location Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>2→1</td>
<td>41-96</td>
</tr>
<tr>
<td>2→3</td>
<td>38-96</td>
</tr>
<tr>
<td>2→4</td>
<td>62-96</td>
</tr>
<tr>
<td>2→5</td>
<td>64-96, 192-288</td>
</tr>
<tr>
<td>2→6</td>
<td>73-96</td>
</tr>
<tr>
<td>5→1</td>
<td>no need OPC</td>
</tr>
<tr>
<td>5→2</td>
<td>64-96, 192-288</td>
</tr>
<tr>
<td>5→3</td>
<td>69-96</td>
</tr>
<tr>
<td>5→4</td>
<td>45-96</td>
</tr>
<tr>
<td>5→6</td>
<td>34-96</td>
</tr>
<tr>
<td>6→1</td>
<td>no need OPC</td>
</tr>
<tr>
<td>6→2</td>
<td>73-96</td>
</tr>
<tr>
<td>6→3</td>
<td>62-96, 186-288</td>
</tr>
<tr>
<td>6→4</td>
<td>79-96</td>
</tr>
<tr>
<td>6→5</td>
<td>34-96</td>
</tr>
</tbody>
</table>

Figure 2 shows the map contains all location ranges of OPC for all traffics in (a) clockwise and (b) counter-clockwise directions. Since the installed OPCs have to support all traffics in the network, the placement location of OPCs in a link between two nodes must be within the intersected product of the location ranges given by all traffics existed in that link. For example, on the link between Node #1 and #2 in the clockwise direction (Fig. 2(a)), we have the location range of OPCs equal to 41-96 km from the traffic from Node #1 to #2, the range of 78-96 km from Node #1 to #3, the range of 61-96 km from Node #4 to #1, and the range of 192-288 km from Node #5 to #2. Therefore, the intersection among these ranges results in the range of 78-91 km. This range of distance is the location range suitable for installing the OPC in the link between Node #1 and #2. By performing this procedure, we finally have 12 OPCs for the sample network. (6 OPCs on clockwise and 6 OPCs on counter-clockwise).

C. Choose the OPC positions, which yields the minimum accumulated dispersion at nodes

In the ring topology, signal may be dropped at arbitrary nodes. Hence, within the possible range for placing the OPC, we should choose the OPC position, which yields the minimum accumulated dispersion at node. The optimum position to achieve the lowest accumulated dispersion at node is always the minimum boarder of the OPC location range. For example, the optimum position is 78 km for the link between Node #1 and #2.
Optimum Placement of Dispersion Compensating Unit for Transparent DWDM Ring Network

Ismaeane Mohara \(^1\) and Pasu Kaewplung \(^2\)
Department of Electrical Engineering, Faculty of Engineering, Chulalongkorn University, Bangkok, Thailand
Tel: +66-2-218-6907, Fax: +66-2-218-6912, E-mail: 1yamala29@hotmail.com, 2Pasu.K@chula.ac.th

Abstract

This paper first time introduces an optimal algorithm for placing the dispersion compensating unit (DCU) in DWDM ring network. Our proposed algorithm can be applied for both non-slope-compensated and slope-compensated DCUs.

1. Introduction

SONET/SDH-based optical ring network has a widespread of adoption as metro-area networks (MANs) due to its reliability and outstanding restoration scheme \([1]\). The exponential growth in data traffic leads to the requirement of upgrading the incumbent SONET/SDH ring to support transparent DWDM technologies, where the multi-wavelength signal remains in optical domain during transmission in the network \([1]\). For such a network that involves with long links, the fiber attenuation and the fiber dispersion become the serious problems that limit the transmission performance. The optimal amplifier placement method for the long-haul DWDM ring network has been already developed \([2]\), while no any attempt has been made on the dispersion compensation although the dispersion compensation is also a serious issue that has to be taken into account in practical design of the transparent DWDM ring network.

In this paper, we propose, for the first time in our knowledge, an algorithm for optimal placement of DCUs in the transparent DWDM ring network. Our algorithm can support both non-slope-compensated (NS) and slope-compensated (SC) DCUs, and can ensure the minimum number of DCUs. We assign a part of the optical pan-European network (OPEN) with the total length of 1,882 km as a 4-ring-intersected sample network, as shown in Fig. 1. By applying our algorithm, we show that for the channel spacing range of 0.2 nm - 1.0 nm, the required total number of DCUs for 9-channel signal is 44 for both NS- and SC-DCUs.

2. Optimum DCU placement algorithm

Our algorithm consists of 4 steps as follows:

**Step1: Communication light paths between any two nodes.**

Let the communication between any two nodes can be bidirectional, in this step, the possible light paths between any two nodes are generated, and finally only the shortest paths are selected for signal transmission.

**Step 2: Generate the constraints.**

First, we assign a group of wavelengths which will be used in the network. The number of wavelength can be larger, smaller, or equivalent to the number of node. Then, the following constraints are generated.

(A) Path Constraints

\[
D_{acXZi} + (D_{\lambda} \times L_{XY}) + (D_{COMP} \times N_{XY}) = D_{acYXi}
\]  

(1)

According to Eq. (1), for a path started from node \(X\) to node \(Y\) with the length of \(L_{XY}(X \neq Y)\), the accumulated dispersion \((D_{acXZi})\) from node \(Z\) of signal at wavelength \(\lambda\) will increase with the amount of \(D_{\lambda} \times L_{XY}\), where \(D_{\lambda}\) is the dispersion at \(\lambda\) of the transmission fiber of the path. At the same time, the accumulated dispersion will be compensated by the amount of \(D_{COMP} \times N_{XY}\) where \(D_{COMP}\) is the dispersion of DCU at \(\lambda\) and \(N_{XY}\) is the number of DCU on path \(XY\). Finally, we obtain the total...
accumulated dispersion at the output of node Y equals
to $D_{acYZ}$.  

(B) Maximum dispersion constraints.
For each wavelength, it is required that the
accumulated dispersion at any point in the network
should not exceed the maximum acceptable
accumulated dispersion $D_{\text{max}}$ [4]. Therefore, we have
the constraint,

$$-D_{\text{max}} \leq D_{acYZ} \leq D_{\text{max}}$$  \hspace{1cm} (2)

(C) Integrality constraints
For each path $XY$, $N_{xy}$ must be an integer.

(D) Objective function

$$\text{Minimize}(N)$$  \hspace{1cm} (3)

Where $N$ is total number of DCU in the network

Step 3: Solve the constraint equations.

Since our problem is a type of mixed-integer linear
programming (MILP), this optimization problem can
be solved by the soft wares such as X-Press,MP and
C-plex. As a result from solving the problem, the
number of DCU in each link and the accumulated
dispersion at every node is obtained.

Step 4: Place the DCUs.
The DCUs are placed in the network at a position
where at least one wavelength exhibits the
accumulated dispersion that reaches $D_{\text{max}}$.

3. Optimal DCUs placement in sample network

For demonstrating our algorithm in really existing
network, we use part of OPEN as a sample network as
shown in Fig. 1. Our sample network is assumed to
operate as ring topology and is consisted of 4-
intersected rings with 10 nodes, 13 links, and has the
total length of 1,882 km. For transmission fiber, the
standard single-mode fiber (SMF, G.652), which
exhibits the dispersion ($D$) of 16.5 ps/nm/km and
the dispersion slope ($D_s$) of 0.05 ps/nm$^2$/km at
1,550 nm [5]. We employ 2 types of DCUs: the NS-
DCU that has $D$ of (~82)ps/nm/km with $D_s$ of 0.25
ps/nm$^2$/km both at 1,550 nm, and the length of the
NS-DCU can perfectly compensate for the
accumulated dispersion of the 100-km-transmitted
length of G.652 fiber, and the SC-DCUs that has same
characteristics as the NS-DCU except the reverse sign
of the dispersion slope [5]. Next, we assign the group
of 9 wavelengths with the center wavelength located at
1,550.12 nm. Following the procedures described in
section 2, we obtain the minimum number of DCU and
the appropriate location of DCUs in the network. By
varying the channel spacing ($\Delta \lambda$) from 0.2 nm to 1.4
nm, we compare the total number of both types of
DCUs that requires in the sample network in Table 1.
From Table 1, the total number of DCUs equals to 44
for both types of DCUs although $\Delta \lambda$ increases from
0.2 nm to 1.0 nm. These results indicate that we can
use the NS-DCU to obtain the same network
performance as using the SC-DCU. This significantly
helps reducing the cost of the network because the NS-
DCU is usually less expensive than the SC-DCU. For
$\Delta \lambda = 1.2-1.4$ nm, the number of NS-DCU necessary
for the network becomes greater than that of SC-DCU
due to large over- and under-compensation.

4. Conclusion

In this paper, an optimal DCU placement algorithm
that gives the minimum number of DCU for
transparent DWDM ring network was presented. By
applying the algorithm to the sample network based on
part of OPEN, we obtained equivalent number of
DCUs for both SC-DCU and NS-DCU for $\Delta \lambda = 0.2$
mm up to 1.0 nm.

Acknowledgement

This work is supported by the cooperation project
between the Department of Electrical Engineering and
private sector for research and development.

<table>
<thead>
<tr>
<th>Type Of DCU</th>
<th>$\Delta \lambda$ (nm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>NS-DCU</td>
<td>0.2</td>
</tr>
<tr>
<td></td>
<td>0.4</td>
</tr>
<tr>
<td></td>
<td>0.8</td>
</tr>
<tr>
<td></td>
<td>1.0</td>
</tr>
<tr>
<td></td>
<td>1.2</td>
</tr>
<tr>
<td></td>
<td>1.4</td>
</tr>
<tr>
<td>SC-DCU</td>
<td>44</td>
</tr>
<tr>
<td></td>
<td>44</td>
</tr>
<tr>
<td></td>
<td>44</td>
</tr>
<tr>
<td></td>
<td>44</td>
</tr>
<tr>
<td></td>
<td>46</td>
</tr>
<tr>
<td></td>
<td>46</td>
</tr>
</tbody>
</table>

Table 1: Comparison of number of DCU between NS-DCU and SC-DCU

Reference

“Transport Metropolitan Optical Networking: Evolving
Trends in the Architecture Design and Computer Modeling,”
IEEE Journal of Light wave Technology 22 (Nov 2004):
2653-2670
methods for metropolitan WDM ring networks,” IEEE
Topological Design of Metro WDM Optical Network Using
Computer Simulation,”IEEEJ. On selected area in
communication.,20(Jan 2002), No.1...
[5] Data. Sheet of PowerForm® DCM ® Modules for SMF,
Reducing of Kerr Effect in Long-haul Broadcast-and-Select Transparent DWDM Optical Network by Static Wavelength Assignment

Preeda Jarupoom¹ and Pasu Kaewplung²
Department of Electrical Engineering, Faculty of Engineering, Chulalongkorn University, Bangkok, THAILAND.
E-mail: ¹cvl_eng84@hotmail.com, ²Pasu.K@chula.ac.th

1. Introduction
Broadcast-and-select (B&S) DWDM optical network is very attractive for the network scales of metropolitan-area network (MAN) or long-haul network because it can provide the full transparency of optical signal transmission from transmitter to a receiver with relatively low system cost. For a network that has relatively long links, signal power attenuation, dispersion, and the Kerr effect become main sources of signal distortion. The optimal amplifier placement for MAN or long-haul B&S network has been studied [2], while the optimal placement of dispersion compensation unit (DCUs) was proposed by us [3]. Through the Kerr effect, the cross-phase modulation (XPM) and the four-wave mixing (FWM) are induced and result in the nonlinear crosstalk among transmission channels. Thus, the suppression of the Kerr effect becomes very necessary and has to be seriously considered in practical design.

In this paper, we propose a static wavelength assignment (WA) algorithm that can provide the globally maximum channel spacing for signal transmission in the B&S network for a given set of wavelengths. Then, the XPM and FWM consequently reduce because of large group velocity mismatch among channels.

2. Kerr effect reduction by wavelength assignment
In our work, each network station has its own wavelength that assigned from our algorithm. A group of stations is connected to a passive star couple (PSCs) in such a way that the network passes the modified feasibility, [2], [4]

$$P_{\text{max}} - 10\log(DG_{\text{max}} - 1) - 10\log(\lambda_{j/\text{max}}) - \sigma_y \geq P_{\text{sen}},$$

where $$P_{\text{max}}$$ [dBm] is the maximum total output power of amplifier and transmitter, $$P_{\text{sen}}$$ [dBm] the minimum single channel signal power that is detectable for amplifiers and receivers, $$DG_{\text{max}}$$ the maximum degree of PSC $$i$$, $$\lambda_{j/\text{max}}$$ the maximum number of wavelengths in link $$j$$ entering the PSC $$i$$, and $$\sigma_y$$ the total additional losses such as the insertion loss, the splicing loss, and the margin considered for signal distortion. We should identify the link $$j$$ and the PSC $$i$$ for the test in such a way that $$(DG_{\text{max}} - 1)\lambda_{j/\text{max}}$$ is maximum among all such pairs in the network. When a set of wavelengths that will be used in the network is given in such a way that the number of wavelength is identical to the number of station, the solution for optimal WA for reducing the Kerr effect is to assign wavelengths to stations in such a way that the summation of channel spacing between two neighbor channels in each link becomes maximum, as shown in the following Eq. (2) and (3).

$$SCS_i = \sum_{j=1}^{n} \left| \lambda_{j} - \lambda_{j+i} \right|,$$

$$\text{Maximize } SCS = \sum_{i=1}^{L} SCS_i,$$

where $$SCS_i$$ is channel spacing between two neighbor channels in link $$i$$, $$\lambda_{j}$$ is wavelength $$j$$ in link $$i$$, and $$L$$ is the total number of links in the network. $$SCS$$ relates to the group-velocity mismatch between the two pulses [1]. Therefore, our objective equation (3) gives the maximum $$SCS$$ of the network for reducing the Kerr effect. In the optimization, the objective equation was solved by numerical simulation to find WA formats that give the maximum $$SCS$$.

3. Simulation
In order to validate our proposed WA in reducing the Kerr effect, we perform numerical simulations of DWDM signal transmission in the B&S network with and without our WA. The network shown in Fig. 1 is used for example. The sample network shown in Fig. 1 consists of 6 stations (6 wavelengths) and 4 PSCs. The wavelengths used for the network range from 1549.12 to 1550.12 following the 25-GHz ITU grid. For 6 wavelengths, we have as much as 720 possible WA patterns to assign to each station of the network. By using our WA algorithm, we obtain 220 WA patterns that have the maximum $$SCS$$ equals to 10 for this network. Some WA patterns with their corresponding $$SCS$$ are shown in Table 1.

From Table 1, the data set No. 1 is a representative of WA patterns that gives the maximum value of $$SCS = 10$$, while the data set No. 720 is a representative of such WA patterns without using our WA ($$SCS = 8.4$$). We use these two patterns for the numerical simulation of multi-wavelength signal transmission. The standard single-
mode fiber (SMF: G.652) with the dispersion of 16.5 ps/nm/km and its slope of 0.05 ps/nm^2/km both at 1550 nm is used for signal transmission. The nonlinear refractive index \(n_2\) is \(2.6 \times 10^{-20} \text{ m}^2/\text{W}\). The optical signal is composed of pseudo-random 1024 bits with the data rate of 10 Gbps.

4. Conclusion
In conclusion, we have proposed an optimized WA for long-haul B&S DWDM network in order to reduce the inter-channel crosstalk induced by the Kerr effect. By using our WA algorithm, the numerical results have shown the significant improvement in transmission performance comparing to the case without the optimized WA.

Acknowledgement
This work is supported by the cooperation project between the Department of Electrical Engineering and private sector for research and development.

References

Fig.1: Sample network.

Table 1: SCS values in some WA patterns.

<table>
<thead>
<tr>
<th>Vz, Data Set</th>
<th>Station#1</th>
<th>Station#2</th>
<th>Station#3</th>
<th>Station#4</th>
<th>Station#5</th>
<th>SCS</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1946.12</td>
<td>1946.12</td>
<td>1548.32</td>
<td>1548.32</td>
<td>1550.12</td>
<td>10</td>
</tr>
<tr>
<td>37</td>
<td>1950.12</td>
<td>1949.92</td>
<td>1549.72</td>
<td>1549.72</td>
<td>1546.52</td>
<td>5</td>
</tr>
<tr>
<td>25</td>
<td>1955.12</td>
<td>1954.92</td>
<td>1549.52</td>
<td>1546.32</td>
<td>1546.12</td>
<td>3.4</td>
</tr>
</tbody>
</table>

Fig.2: Numerical \(Q\) factor of the signal from 5 other stations entering station#1; (a) using our WA, (b) without using our WA.

4. Conclusion
In conclusion, we have proposed an optimized WA for long-haul B&S DWDM network in order to reduce the inter-channel crosstalk induced by the Kerr effect. By using our WA algorithm, the numerical results have shown the significant improvement in transmission performance comparing to the case without the optimized WA.

Acknowledgement
This work is supported by the cooperation project between the Department of Electrical Engineering and private sector for research and development.

References

721
Pseudo-Random Sequences for Modeling of Quaternary Modulation Formats

D. van den Borne¹, E. Gottwald², G. D. Khoe³, H. de Waardt¹
(¹ COBRA Institute, Eindhoven University of Technology, The Netherlands, e-mail: D.v.d.Borne@tue.nl)
(² Siemens Networks GmbH & Co. KG, Munich, Germany, e-mail: Erich.Gottwald@siemens.com)

Abstract
We discuss the influence of pseudo-random quaternary sequences on DQPSK transmission performance and present an efficient method to generate such sequences.

1 Introduction
Multi-level modulation formats are considered for next-generation optical transmission systems. In particular differential quadrature phase shift keying (DQPSK) is attractive as it encodes 2 bits/symbol and as such offers a combination of high spectral efficiency and large chromatic and polarization-mode dispersion tolerance [1, 2]. To assess the linear and nonlinear transmission properties of such modulation formats, inter-symbol interference as occurs along the transmission line should be properly modeled [3]. Binary modulation formats are routinely modeled using pseudo-random binary sequences (PRBS), but for quaternary modulation formats no such standard exists. Hence, to generate a quaternary sequence typically two PRBS are multiplexed with a cyclic shift for de-correlation. However, such a sequence does not necessarily include all possible combinations of symbols up to a given length, which can result in inaccurate modeling of system penalties. This can be avoided by using pseudo-random quaternary sequences (PRQS).

In this paper we discuss the influence of pseudo-random sequences on 42.8-Gbit/s NRZ-DQPSK transmission. In addition, we present an efficient method for PRQS generation by multiplexing two PRBS with a suitably chosen cyclic shift.

2 Pseudo-random quaternary sequences
Assume a sequence to have length $k^n$, where $k$ is the alphabet size and $n$ an integer value. A pseudo random sequence can then be defined as a sequence that contains all possible combinations of symbols up to length $n$ (subsequences) exactly once. Such sequences are known as de Bruijn sequences and are considered in the remainder of this paper. Note that de Bruijn sequences are equal to standard PRBS with length $k^n-1$ with the exception that PRBS lack the subsequence with $n$ consecutive zeros. It is evident that for $k \geq 4$ the required sequence length to consider all possible subsequences of a given length $n$ grows very rapidly. For example, consider subsequences of length $n = 7$. For binary sequences this results in a sequence length of $2^7 = 128$ symbols, whereas for quaternary sequences $4^7 = 16384$ symbols are required. Hence, the subsequence length $n$ for $k \geq 4$ is restricted by simulation complexity. This underlines the need for a properly chosen PRQS to correctly model system penalties.

A simple method to generate PRQS is to multiplex two PRBS ($u$ and $v$) with a cyclic shift for de-correlation. A figure of merit for the de-correlation in the resulting sequence can be obtained by computing,

$$\text{correlation} = \sum \left| \Pr(r) - \frac{1}{k^n} \right|^2$$

where $\Pr(r)$ is the probability of each possible subsequence $r$ for a pseudo-random sequence with alphabet size $k$ and subsequence length $n$. Fig. 1 shows the correlation as function of the cyclic shift between two PRBS of length $2^6$, normalized with respect to the correlation value obtained when there is no cyclic shift between $u$ and $v$. As evident from Fig. 1 the relative correlation is zero when $v$ is cyclically shifted over $+/3$ symbols. It can be shown that for a $2^6$ PRBS and a cyclic shift of $+/n/2$ symbols (for $n$ is even) always a PRQS with length $4^{2n}$ is obtained. Shifting $v$ over $-n/2$ symbols is equal to shifting $v$ over $+n/2$ symbols and inverting it.
Fig. 1: Correlation as a function of the cyclic shift.

Fig. 2: BER for 42.8-Gbit/s NRZ-DQPSK over 1140 km for different pattern length and cyclic shift, dotted lines denote in-phase and quadrature tributaries.

3 Simulation results

We now simulated the influence of different 4-level sequences on 42.8-Gbit/s NRZ-DQPSK transmission. Single channel transmission over 12×95-km (1140 km) of standard single mode fiber (SSMF) is simulated with a pre-compensation of 680-ps/nm, 85-ps/nm/span in-line under-compensation and zero residual dispersion. The launch power is 4 dBm per channel into the SSMF and -1 dBm into the DCF. At the receiver the OSNR is set to 18.3 dB (0.1 nm res.) which results in a back-to-back bit-error ratio (BER) of 10^-9. The BER is computed using a Karhunen-Loève series expansion [4].

Fig. 2 shows clearly that the performance is strongly dependent on the cyclic shift between u and v. Short PRBS result on average in a lower BER, thereby incorrectly modeling the transmission penalties. For longer PRBS (1024 symbols) the variation in BER as a function of the cyclic shift is reduced, but differences of more than an order of magnitude are still apparent. This shows that even for long sequences an arbitrary cyclic shift can potentially result in inaccurate modeling of transmission penalties.

Fig. 3: Worst-case, median and best-case BERs for all cyclic shifts and the BER obtained with PRQS.

Fig. 3 shows the worst-case, median and best-case obtained BER for all possible cyclic shifts between u and v larger than two (to ensure all symbol transitions exist). For a sequence length longer than 256 symbols the median value is relatively constant, indicating this is sufficient to properly determine transmission penalties. It furthermore shows that the BER obtained with the proposed PRQS is close to the median value over all possible cyclic shifts which indicates that these sequences are suitable for modeling of DQPSK transmission impairments.

The proposed method can also be used to generate multi-level sequences for modulation formats with >2 bits/symbol. This can be helpful to explore modulation formats such as polarization-multiplexed DQPSK (with 4 bits/symbol) [5], which are likely to be used in future optical transmission systems.

4 References

1. G. Charlet, et. al., in proc. ECOC’05, PD paper Th4.1.3.
2. D. van den Borne, et. al., in proc. OFC’06, paper OFD2.
5. D. van den Borne, et. al., in proc. OFC’06, paper PDP34.
Chromatic Dispersion Compensation by Monitoring Bit-Pattern-Dependent Intersymbol Interference in NRZ Format

Yuji Hayami and Katsushi Iwashita
Kochi University of Technology
185, Miyanokuchi, Tosayamada-cho, Kami-shi, Kochi, 782-8502, Japan
Phone: +81-887-57-2117, Fax: +81-887-57-2120
E-mail: 070323h@ugs.kochi-tech.ac.jp

Abstract
A novel chromatic dispersion compensation technique by monitoring bit-pattern-dependent intersymbol interference is proposed. This method is based on pulse compression/broadening by self-phase modulation. The feasibility of this technique is verified by simulation.

1. Introduction
Fiber chromatic dispersion is a limiting factor for high bit-rate and long-distance optical transmission systems. The degradation resulting from chromatic dispersion can be improved by using a dispersion-compensating fiber (DCF) or optical compensation devices, such as fiber Bragg gratings or planar lightwave circuits. However, since fiber chromatic dispersion varies with temperature, in-service dispersion compensation requires more than 40 Gbps, since the dispersion tolerance is reduced. Many chromatic dispersion compensation methods have been proposed, including eye diagram monitoring[1,2] and optical frequency modulation[3,4].

In this paper, we propose a novel chromatic dispersion compensation method for monitoring bit-pattern-dependent intersymbol interference (ISI) in NRZ format. The proposed method is verified by simulation.

2. Principle of dispersion compensation
We propose a chromatic dispersion compensation method by monitoring the bit-pattern-dependent intersymbol interference in NRZ format. When self-phase modulation (SPM) occurs, the optical pulses are compressed in the anomalous dispersion region, as shown in Fig-1; by contrast, the optical pulses are broadened in the normal dispersion region. These phenomena result from optical power changes around the rise/fall time regions because the frequencies of these regions are varied by SPM. Thus, the height of narrow pulses is changed more than that of wide pulses. Consequently, the ISI of narrow pulses in the normal dispersion region is larger than that in the anomalous region. By contrast, the height of wide pulses does not vary. Therefore, by comparing the ISI between a narrow pulse and a wide pulse, we can determine the sign of the chromatic dispersion. The pulse width is determined by the data in NRZ format. For example, narrow pulses are produced by the [010] bit stream, while wide pulses are produced by the [111] bit stream. The chromatic dispersion region can be determined by comparing these two-bit patterns.

In the chromatic dispersion compensation system, two specific patterns are monitored at the optical receiver. The ISI of these two patterns are then compared. The dispersion compensator is controlled using the compared ISI signal.

3. Simulation Result
The proposed method was verified by simulation. We applied 40 Gbps NRZ format with a $2^{11}$ pseudo-random bit sequence (PRBS). A Mach-Zehnder intensity modulator with a push-pull configuration was used to produce a chirp-free optical signal. The modulated optical signal was transmitted over an 80-km single-mode fiber. The ISI was measured at the optical receiver. The ISI is defined as the temporal amplitude divided by the average distance between 1 and 0. The ISI was measured at the $10^{-7}$ bit error rate by controlling the received optical power. The ISI was affected by the 2 bits before and after the decision point. We measured the ISI with 5-bit pattern included in the $2^{11}$ PRBS bit stream.
The simulated results of the ISI in a non-repeater transmission system are shown in Fig-2. While we have simulated all the 5-bit patterns, just two remarkable results are shown in Fig-2. Eye diagrams at a chromatic dispersion of (-0.6 ps/km/nm) and (0.6 ps/km/nm) are shown in Fig-3. The ISI of the pattern [01110] does not depend on chromatic dispersion. However, the ISI of the pattern [10101] depends on chromatic dispersion. Therefore, by monitoring the ISI of these two patterns, we can determine the sign of chromatic dispersion. Then, using this sign, the chromatic dispersion compensator can be controlled.

\[ \text{Fig-2} \quad \text{Intersymbol interference for non-repeater transmission system. The fiber launch power is +13 dBm.} \]

\[ \text{Fig-3} \quad \text{Eye diagrams for (a) chromatic dispersion -0.6 ps/km/nm and (b) chromatic dispersion +0.6 ps/km/nm} \]

The simulated results of the in-line amplifier transmission system are shown in Fig-4. There are three in-line optical amplifiers and each single-mode fiber is 80 km in length, as shown in Fig-4(a). The optical launch power of each transmission fibers is the same. Figure 4(b) shows the difference in the ISI between the two patterns at each optical launch power. The ISI increases in the negative dispersion region, where the optical launch power is less than 4 dBm, because the influence of SPM is reduced. At powers greater than 10 dBm, the available dispersion region is restricted by SPM degradation. These results are similar to those for non-repeater transmission. Therefore, we can compensate the chromatic dispersion by monitoring the ISI.

\[ \text{Fig-4} \quad \text{The difference in the ISI for in-line amplifier transmission system. (a) the system configuration, where Tx is transmitter, Rx is transmitter, (b) the difference of ISI} \]

4. Conclusion
We have proposed a novel chromatic dispersion compensation technique by monitoring the bit-pattern-dependent ISI in NRZ format. It is shown that the sign of chromatic dispersion can be determined by measuring the ISI of two specific patterns. This signal can be utilized as the dispersion compensator control signal.

Reference
Proposal of Control Mechanism for Delay Sensitive Traffic in Simplified WFQ Scheduling

Sachiko Taniguchi, Ryusuke Kawate, Tetsuya Yokotani, Kuniaki Motoshima, *Shinichi Yoshihara
Information technology R&D center, Mitsubishi Electric Corporation
(5-1-1, Ofuna, Kamakura, Kanagawa, 247-8501, Japan, Tel: +81-467-41-2872, Fax: +81-467-41-2419)
(E-mail: Taniguchi.Sachiko@cj., Kawate.Ryusuke@db., Yokotani.Tetsuya@eb., Motoshima.Kuniaki@cs., MitsubishiElectric.co.jp)
*NTT Access Network Service Systems Laboratories, NTT Corporation
(6-1, Nakase, Mihama, Chiba, 261-0023, Japan, Tel: +81-43-211-3256, Fax: +81-43-211-8282)
(E-mail: yosihara@ansl.ntt.co.jp)

Abstract
This paper proposes the new control mechanism for delay sensitive traffic based on the simplified WFQ Scheduling. This mechanism provides low delay transmission and guarantee of bandwidth on multiplexing a huge number of traffic flows.

1 Introduction
WFQ (Weighted Fair Queuing) scheduling controls bandwidth among multiplexed traffic flows according to packet length on these traffic flows and pre-assigned weight. In general, since WFQ should investigate and schedule a lot of traffic flows simultaneously, it is difficult that WFQ supports a number of traffic flows by limitation of hardware circuit. On the other hand, recently, a huge number of traffic flows have been multiplexed one physical interface by VLAN family architecture, e.g., IEEE802.1Q, IEEE802.1ad, and IEEE802.1ah. For this purpose, we have proposed the simplified WFQ scheduling based on two approximation technologies to provide these multiplexing [1]. However, WFQ intends to bandwidth control among flows, and does not pay attention to provide delay control. This paper proposes the new mechanisms by enhancement of the simplified WFQ to support low delay transmission for delay sensitive traffic class, such as transfer of voice and video applications.

2 Problems to provide low transit day in WFQ
The simplified WFQ multiplexes a huge number of traffic flows on a physical interface, and provides fair share among these flows with high accuracy. However, this mechanism does not mention the delay control. In a real system, support of delay sensitive services, such as VoIP and streaming video, are required. In such a case, switch should precede data for these services as high priority class to reduce transit delay. On the other hand, bandwidth is fairly allocated per user aggregating all of services. Therefore, it is required that priority control for delay sensitive services and bandwidth allocation for every user go together.

Typically, for this purpose, queues are configured by combination of WFQ and SP (Strict Priority). The two configuration types are typifies as follows.
Type 1: Several queues are grouped and controlled by SP among these queues. These groups are controlled by WFQ. In this type, although bandwidths among groups are fairly controlled, transit delay is increased in proportion to the number of groups. Therefore, the transit delay cannot be guaranteed.
Type 2: Dedicated queues for delay sensitive services are isolated [2]. Queues for non delay sensitive services isolated from dedicated queues are controlled by WFQ. These dedicated queues are multiplexed with others by SP. Although this type provides low delay transmission for delay sensitive services, bandwidth cannot be shared between dedicated and other queues. In short, even in no traffic of delay sensitive services, non delay sensitive services cannot utilize this room.
3 Proposed mechanism for delay sensitive services

It is mentioned that both types have some problems in the previous section. This section proposes the new mechanism to solve these problems. In proposed mechanism, the type 2 queue configuration can be adopted. As in Fig.1, the Coordination function is defined between control modules for delay services and non delay sensitive services.

![Fig. 1 Configuration of proposed mechanisms](image)

In simplified WFQ described in [1], continuance time is quantized by slots. Therefore, virtual output time is mapped into the quantum, and is en-queued in the queue for this quantum as shown in Fig.2. For cooperating scheduling between delay sensitive and non delay sensitive services, virtual output time of the head packet in the queue for non sensitive delay is calculated. After that, this time is quantized and is stored in the queue for this quantum. When the packet for delay sensitive services is transmitted, virtual output time for non delay sensitive packet should be sifted.

The virtual output time of the head packet of User #i in non delay service queue represented by \( F_i(1) \), which is the key point in these operations, is specified as follows.

**1) Non empty case of the queue for User #i**

When \( F_i(0) \) represents the virtual output time of the previous packet, \( F_i(1) \) is described in Eq. (1).

\[
F_i(1) = F_i(0) + (L_i(1) + L_{DS}) \times W_i
\]

In Eq. (1), \( L_i(1) \), \( L_{DS} \), and \( W_i \) represent the packet length of User #i, the total length of packets for delay sensitive services and weight for User #i, respectively.

**2) Empty case of the queue for User #i**

If the packet arrives at \( t \) in the queue for User #i in the non delay sensitive service, the virtual output time of this packet is described in Eq. (2).

\[
F_i(1) = t + L_i(1) \times W_i
\]

![Fig.2 Virtual output time by simplified WFQ](image)

4 Conclusion

This paper has proposed the mechanism to provide fair share of bandwidth and low delay transmission for delay sensitive services based on the simplified WFQ. This proposal intends to expand advantages of the simplified WFQ which has been proposed in the previous work.

This work has been in part supported by "Lambda Access Technology" in a project of National Institute of Information and Communications Technology in Japan.

References


Complexity Reduction of MSPE by an Iterative Algorithm for Optical D(Q)PSK

K. Piyawanno, F.N. Hauske, M. Kuschnerr, B. Lankl
University of the Federal Armed Forces, Munich, Institute for Communications Engineering, D-85777 Neubiberg
phone: +49-89-60043925, fax: +49-89-60043921, kittipong.piyawanno@unibw.de

Abstract Differential detection combined with electrical signal processing applying MSPE can nearly reach the performance of coherent detection. We present an iterative algorithm that replaces the recursive MSPE reducing complexity without performance loss.

Introduction
Higher order modulation formats like quaternary phase-shift keying (QPSK) inherit a high spectral efficient, increased tolerance to chromatic dispersion and polarization-mode dispersion [2,3]. Coherent (homodyne) and differential (self-homodyne) detection are the two candidates to detect PSK signals. Coherent detection provides the best performance, but requires complex high speed controlling in the electrical domain. Differential phase shift keying (DPSK) shifts complexity into the optical domain and avoids electrical control circuits enabling high speed transmission, but typically suffers a 3dB penalty compared to coherent detection [1]. MSPE improves differential detection performance close to coherent detection with minor signal processing.

Multi-symbol phase estimation (MSPE), as described in [4] and [5] applies recursive signal processing with the feedback of the detected signal. Thus, the combination of MSPE with digital equalization like an MLSE is difficult. Furthermore, the recursion limits the processing rates by the time required for detection, logic operation and feedback. The replacement of the recursive MSPE by an iterative algorithm overcomes this drawback and reduces implementation complexity. Furthermore, this structure can be combined with iterative equalization as described in [6].

Principle of the Recursive MSPE
The received QPSK signal is differentially detected by an add-and-delay interferometer (ADI), which compares the phase of two consecutive symbols. Noisy distortions of each symbol contribute to the total phase error, which explains the sensitivity penalty compared to coherent detection. After the balanced photo diodes, we receive the complex electrical signal

\[ u(n) = e^{j\theta(n)} y(n) y(n-1) \]

where \( y(n) \) is the normalized optical field of the \( n \)th bit before demodulation. Now, MSPE recursively generates an improved decision variable on basis of a noise estimation from previous detected symbols[4][5].

Fig. 1 shows the proposed MSPE receiver for DQPSK. With the aid of the feedback signal and the received signal \( u(n) \), a decision flip-flop (D-FF) detects the digital sequence \( c(n) \) and \( c_0(n) \). The logical operation unit (LOU) recovers the phase shift variable \( \Delta \Phi(n) \) from \( c(n) \) and \( c_0(n) \) expressed by

\[ e^{-j\Delta \Phi(n)} = c(n) e^{j\theta(n)} y(n) y(n-1) + j \cdot c_0(n) e^{j\theta(n)} y(n) y(n-1) \]

Now, \( \Delta \Phi(n) \) is used for the noise estimation of the previous symbols. The improved decision variable is approximated by

\[ \bar{x}(n) = u(n) + w \cdot u(n-1) \]

where \( l(n) \) is the normalized intensity of the received signal detected by the intensity arithmetic (IA). The weight factor \( w \in [0,1] \) adjusts the memory length of MSPE. The value of \( w \) should be chosen inversely proportional to the noise power in the system [4,5].

\[ x(n) = u(n) + w u(n-1) \cdot \frac{1}{l(n-1)} e^{-j\Delta \Phi(n-1)} e^{-j\Delta \Phi(n)} \]

Iterative Algorithm for MSPE
To derive the iterative algorithm, we rewrite the recursion of Eq. (3) in summand form. For large values of \( N \) the last term containing \( w^n x(N) \) from the right hand side of Eq. (3) can be omitted

\[ x(n) = u(n) + \sum_{k=1}^{N} \sum_{i=0}^{N-i} \frac{u(n-i)}{l(n-i)} e^{-j\Delta \Phi(n-i)} e^{-j\Delta \Phi(n)} \]

In the next step, we apply a window length \( k \), which limits the memory length. Furthermore, we substitute the recursive phase shift variable \( \Delta \Phi(n-i) \) by \( \Delta \Phi_i(n-i) \), which is obtained from a previous decision \( c_i(n-i) \).

\[ x(n) = u(n) + \sum_{n=1}^{N} \sum_{i=0}^{N-i} \frac{u(n-i)}{l(n-i)} e^{-j\Delta \Phi_i(n-i)} e^{-j\Delta \Phi_i(n)} \]

Eq. (5) requires a first decision without MSPE to obtain the initial \( \Delta \Phi_i(n) \), following iterations can employ the decisions from each previous iteration. An erroneous decision \( c_i(n) \) will possibly lead to a burst error in the
following iteration \( p+1 \) within the range of \( n-i < 2k \), which refers to twice the window length. Thus by comparison of \( c_i(n) \) and \( c_{i+1}(n) \) we apply a save-interval, which allows only to flip the first bit out of a burst of flips.

Fig. 2 shows the block diagram of the iterative MSPE with three cascades. The structure clearly avoids any feedback.

**Simulation Results**

Back-to-back Monte-Carlo simulations were carried out to compare the performance of the recursive MSPE to the iterative MSPE. We fed 20 Gb/s DQPSK (PRBS 2\(^{15}-1\)) at a low launch power loaded with additive Gaussian noise into the receiver, which applied a Gaussian optical bandpass filter (2\(^{nd}\)order, 12.5 GHz) and an electrical low-pass Bessel filter (20\(^{th}\) order, 12.5 GHz) after balanced detection. BERs were evaluated at various OSNR (0.1nm) values. The weighting factor of the MSPE was set to \( w=0.9 \).

Fig. 3 shows the sensitivity improvement of the iterative MSPE with respect to the number of cascades. No further improvement is visible after the 5\(^{th}\) cascade, which determines the number of required cascades to reach the maximum performance of the iterative structure.

In Fig. 4, the BER performance as a function of the window length \( k \) is evaluated. The sensitivity increases with the window length and converges to the sensitivity of the recursive MSPE for large values of \( k \). For \( k > 12 \) only a minor improvement can be made at high OSNR at the cost of a higher complexity. In our simulations we regarded a window length of \( k=12 \) as sufficient.

Fig. 5 compares the BER vs. OSNR of the recursive MSPE with the iterative MSPE. Increasing the number of cascades, the iterative MSPE enhance its performance nearly reaching the recursive MSPE after the 5\(^{th}\) cascade with a sensitivity improvement of 1.7 dB at BER=10\(^{-3}\) compared to conventional DQPSK. Further cascades only lead to a minor improvement with a remaining penalty of 0.1 dB compared to the recursive MSPE.

**Conclusion**

We have presented an iterative structure for MSPE in optical DQPSK receivers. Controlling burst errors with a save interval, employing a limited window length and a low number of cascades, we nearly reach the performance of the recursive MSPE with a remaining penalty of 0.1 dB at BER=10\(^{-3}\). Avoiding any feedback, we reduce the complexity of the electronic signal processing, which enables implementation at high clock rates.

**References**

[1] Griffin et al., OFC’03, paper FP6
[2] A. H. Gnauck et al., ECOC’04, pdl paper Th4.4.1
[4] D. van den Borne et al., IEE ’05, paper 12
[5] X. Liu et al., ECOC’06, paper We2.6.6.
[6] Hauske et al., OFC’07, paper OMG2
Heuristic Approximation of Transient Gain Dynamics due to Network Reconfiguration

Stephan Pachnicke\textsuperscript{1}, Erich Gottwald\textsuperscript{2}, Peter M. Krummrich\textsuperscript{2}, and Edgar Voges\textsuperscript{1}

\textsuperscript{1}University of Dortmund, High Frequency Institute, 44227 Dortmund, Germany,
Tel: +49-231-755-6675, Fax: +49-231-755-4631, E-mail: stephan.pachnicke@uni-dortmund.de
\textsuperscript{2}Siemens Networks, Fixed Networks, 81379 Munich, Germany.

Abstract
Transient gain dynamics due to changes in WDM channel count have been investigated. Based on extensive numeric simulations a worst-case approximation for the maximal power overshoot in electronic feedback gain controlled EDFA chains is developed.

1 Introduction
In reconfigurable wavelength-division-multiplexed (WDM) transparent optical transmission systems the addition and dropping of channels - without countermeasures - leads to transient events with power over- and undershoots with respect to the steady-state. In future dynamic transparent automatically switched optical networks switches can initiate adds and drops with rise and fall times in the order of milliseconds and total power changes of more than 10 dB. Beyond that faults such as fiber cuts or component failures may lead to a sudden loss of all express channels. In both cases the remaining traffic should be maintained error-free. In the past several control strategies have been proposed to stabilize the EDFA gain at a given operating point so that the output powers of the surviving channels remain constant. In this paper we assume an electronic pump-power adjustment based on a proportional-integral (PI) controller [1]. The reason for choosing a simple PI controller without D-part lies in the fact that it is very insensitive to power fluctuations induced by noise. In extensive numerical simulations we analyzed the transient behavior of EDFAs under different operating conditions and channel load changes. Based on these results a heuristic formula is derived to approximate the maximal power fluctuation during the transient event. Such an approximation formula is of great importance for the system designer to assess the system response due to dynamic reconfiguration without the need for time-consuming simulations.

2 Simulation methodology
An 80 channel WDM system with 50 GHz channel spacing, which is equivalent to a channel load covering the entire C-band (Fig. 1), has been examined. At \( t = 0.25 \) ms the channel load is changed. In our simulations we investigated different channel loading conditions. The falltime has been chosen to be either 160 \( \mu \)s or 1 ms, which are realistic values for component failures or switching events in reconfigurable optical cross connects, respectively.

![Fig. 1: System setup. A total number of 80 channels has been launched with 50 GHz channel spacing. The power transients due to a change in channel loading after up to 20 spans are analyzed. In our simulations we used the average inversion level model introduced in [2]. The signal channels are represented by their time-averaged power levels. Amplified spontaneous emission (ASE) noise is included by the spontaneous emission factor \( n_{sp} \) in the rate equation, which gives an approximate solution for the ASE influence in EDFAs under dynamic conditions [3]. To facilitate the simulations we assumed the gain spectrum in the relevant wavelength range to be flat – corresponding to ideal gain flattening filters.](image-url)
electronic control circuit works as follows. The control deviation $e$ is calculated from the difference between the total input power $P_{\text{in, tot}}$ multiplied by the desired gain $G$ and the total output power $P_{\text{out, tot}}$. Both powers are obtained from a 5% tap coupler inserted before and after the EDFA. In our simulations we modeled the transmission fiber as a lumped loss element. We deliberately neglected the tilt induced by stimulated Raman scattering (SRS), which has been investigated in e.g. [4]. Furthermore, the effect of spectral hole burning (SHB) has not been considered. In all simulations the same controller parameters have been selected ($k_C = 60$, $\tau_I = 4.5$ ms), which have been determined by extensive simulations. In the following only the results for the dropping of channels are depicted because usually adding channels can be controlled, whereas dropping may be uncontrolled in the case of failures.

3 Heuristic approximation

In the following figures the maximal overshoot calculated for the different setups with activated gain control is depicted. The first graph shows the results for a falltime of 160 $\mu$s, the second graph for a falltime of 1 ms.

It can be seen that the maximal power overshoot increases almost linearly (in dB scale) with the number of cascaded EDFAs.

$$\text{Overshoot}_{\text{db}} = (C - 0.01 \cdot \Delta G) \cdot (\Delta ch - 0.9) \cdot N^{0.03} \quad (1)$$

Based on these findings a heuristic formula (1) has been developed, which gives a worst-case approximation for an arbitrary number $N$ of cascaded EDFAs, a gain difference $\Delta G$ (in dB) with respect to the reference gain of 26 dB and different power dropping ratios $\Delta ch$. The simulated overshoot showed very low dependence on the EDFA output power, and (1) yields a good approximation for saturated amplifiers with gain control. As can bee seen from Fig. 2 the heuristic equation is in good agreement with the simulated values in all cases. In (1) variable $C$ has a value of -0.11 for a falltime of 1 ms and -0.15 for a falltime of 160 $\mu$s. As an example for a 6 dB power drop, a gain value of 29 dB and a falltime of 1 ms the parameters are $\Delta ch = 0.25$, $\Delta G = 3$ and $C = -0.11$ and eq. (1) yields an overshoot of 1.47 dB.

4 Conclusion

We have presented a study of power transients in transparent optical networks stemming from reconfiguration and changes in WDM channel count. The investigated electrical PI feedback loop can effectively stabilize the EDFA gain level. However, remaining power over- and undershoots will accumulate along the transmission line. The developed heuristic formula allows a fast approximation of the remaining power fluctuations based on basic parameters such as gain, number of cascaded EDFAs and number of switched channels. The approximation formula shows good agreement with the results from numerical simulations and enables the system designer to estimate the expected transient behavior.

References

Simple fiber-type wavefront-splitting interferometer using hollow optical fiber

Y. Jung¹, J. K. Kim¹, B. H. Lee¹, and K. Oh²
(¹Department of Information and Communications, GIST 1 Oryong-dong, Buk-gu, Gwangju, 500-712, Republic of Korea)
(²Institute of Physics and Applied Physics, Yonsei University, Seoul, 120-749, Republic of Korea)
Tel: 82-2-2123-7657, Fax: 82-2-365-7657, Email: koh@yonsei.ac.kr

Abstract
We have developed a compact and stable fiber optic wavefront splitting interferometer using hollow optical fiber (HOF) spliced between standard single mode fibers. The use of a HOF at grazing incidence is to provide two optical beams with interference.

1 Introduction
Varieties of smart fiber-optic sensing elements have been proposed for fiber-optic sensors to cope with ever increasing varieties of physical and bio-chemical measurands. In the field of the interferometric sensors, many efforts have been devoted and promising results have also been achieved with Michelson, Mach-Zehnder, Sagnac, Fabry-Perot, and multimode interference (MMI) techniques [1].

In this paper, we propose and demonstrate a novel sensing structure using HOF segment spliced between standard single mode fibers (SMFs). Similar to Lloyd’s mirror, an optical phase difference occurs between the directly proceeding wavefront and the reflecting wavefront at the fiber-air interface, which modulates the intensity at the detector for the interference.

2 Experiment
The proposed fiber interferometer is schematically illustrated in Fig. 1. The device consists of a short HOF segment with the diameter of 8/125μm (hole/outer) and the length of 100μm [2] spliced between standard SMFs to form the center air region. We used micro-translation stage that provides the desired cleave position on the HOF. The SMF was spliced to the HOF with particular control of arc-conditions keeping the hole open. First of all, we investigated the reflection and transmission spectra due to the similar configuration of all-fiber Fabry-Perot (FP) interferometer [3]. It is noticeable that FSR of transmission is significantly different each other from the reflection as shown in Fig. 2. Reflection fringes were well explained from the equation for FSR of FP cavity, \( \frac{\lambda^2}{2L} \), and \( L \) is estimated to be 98.7μm. However, the transmission spectrum shows more big value of FSR with high visibility. This means that there is other interference with smaller optical path difference than FP cavity.

Fig. 1. Configuration of the proposed interferometer.

Fig. 2. Reflection and transmission spectra of interferometer
Most of the incident light of SMF1 is transmitted to SMF2 and undergoes significant beam divergence through the air gap. However, hollow optical fiber that serves as a Lloyd’s mirror, reflects a portion of the diverging wavefront emitting from SMF1. Note that silica-glass internal wall has good reflectivity at grazing incident angle (85~89°). In order to investigate how an input fundamental LP$_{01}$ mode in SMF changes to an air-core mode in HOF was analyzed using a beam propagation method (BPM). The contour displays are similar to the concept of re-imaging in MMI effects associated with multimode waveguides as widely used in planar waveguides as shown in Fig. 3(a). The transmitted power becomes highly sensitive to the spectral position and the optical path length of the air cavity. In a short length of HOF, diverging Gaussian beam from SMF1 experienced multiple reflections, when the air hole was decreased from 50 to 8.2μm. The guiding mechanism of hollow optical fiber is different from conventional glass-fiber optics because the refractive index of the core is lower than that of the cladding. There is no total reflection at the boundary between the core and the cladding; thus, light transmits in the core as a leaky mode with a transmission loss.

The relationship between transmittance and air gap from the experimental results is shown in Fig. 4. Longer HOF segment gives narrower spacing between two adjacent maxima, and that the spacing increases along wavelength. From the Fig.2 and 4, the proposed transmission-type interferometer shows low transmission loss and strong fringes with high visibility than previous reflection-type FP interferometers. Intensity of light is sensitive to optical path length which is applicable to various measurands such as temperature, tension, refractive index in air cavity.

We have demonstrated the innovative fiber-type wavefront splitting interferometer with SMF-HOF-SMF structure. The ultra compact and stable sensing structure with high visibility will overcome previous limitations of all-fiber PF interferometers and endow a new degree of freedom to design next generation fiber optic sensor applications.

This work was supported in part by the KOSEF (Program Nos. R01-2006-000-11277-0, R15-2004-024-00000-0), and the Science and Technological Cooperation program between and from MOST.

3 References
Common-Path Interferometer for Characterization of Fiber Bragg Gratings

Lih-Gen Sheu and Hui-Mei Fang
Department of Electro-Optical Engineering, Vanung University
1 Van Nung Road, Chung-Li, Tao-Yuan 320, Taiwan, R.O.C.

Abstract -- A simple measurement for characterization of fiber gratings by the common-path interferometer is proposed. The complex reflection-coefficient spectrum is obtained by processing the interference spectrum. The results excellently match the data by optical network analyzer.

I. Introduction

Measurement of the group time delay is an essential technique for developing a new fiber Bragg grating (FBG) device. Although the modulation phase-shift method [1] is usually applied to measure the chromatic dispersion of a fiber grating device due to its accuracy and repeatability, it requires the complicated setup and expensive modules such as high-speed optical modulator and detectors. To simplify the dispersion measurement setup for fiber grating devices, the techniques based on the Michelson interferometers with Fourier transform spectroscopy have been proposed [2]. However, these measurement systems must be carefully isolated for good stability, because the interferometer is very sensitive to the environmental perturbations between the two discrete optical paths. In this paper, we propose a stable and accurate measurement technique for characterization of FBGs by using a common-path interferometer. This method just requires a broadband source and an optical spectrum analyzer (OSA) to scan the interference pattern in wavelength domain. Moreover, the common-path interferometer configuration can much improve the stability of the measurement system. We will demonstrate the accuracy and repeatability of group time delay measurement for the test FBGs. The results agree very well with the data measured by using the optical network analyzer (Advantest Q7760).

II. Principle of Measurement

The complex reflection coefficient measurement setup for a fiber grating consists of a Fabry-Perot interferometer which first and second reflectors are the end of the input fiber and the tested FBG respectively, as shown in Fig.1. The incident light is accurately coupled to the tested FBG by using a mechanical splice without matching oil. The ASE source and OSA have been used to obtain a wide spectral range in a single fringe scan. The reflectivity of the fiber end is so small (≈ 4%) that the Fabry-Perot interferometer can be approximated as a Fizeau interferometer. The interference spectrum of the Fizeau interferometer measured by the OSA can be expressed as

\[
I_{\text{os}}(\lambda) = I_{\text{ref}} + I_{\text{FBG}} + 2\sqrt{I_{\text{ref}}I_{\text{FBG}}} \cos \left( \frac{4\pi}{\lambda} n_{\text{eff}}L + \phi(\lambda) \right)
\] (1)

where \(I_{\text{FBG}}\) and \(I_{\text{ref}}\) are the intensity reflection spectra of the FBG and the reference end respectively, \(n_{\text{eff}}\) is the effective index of the core mode, and \(L\) is the effective distance between the two reflectors. The phase term \(\phi(\lambda)\) is the wavelength-dependent phase delay of the test FBG. According to (1), the Fourier transformation of the measured interference spectrum consists of the -1, 0, and +1 order harmonic components. If the optical path difference \(2n_{\text{eff}}L\) is large enough to avoid the overlap of the 0 and +1 order harmonic components, then the +1 order component can be conveniently calculated from the inverse Fourier transformation of the Fourier-transformed spectral data with the band-pass filtering to select only the +1 order component. Finally, we evaluate the group time delay of the tested device by differentiating the phase \(\phi(\lambda)\) with respect to the angular frequency. The above-mentioned data processing procedure is summarized in Fig.2. Likewise, the intensity reflection spectrum \(I_{\text{FBG}}(\lambda)\) of the test FBG can be also calculated by applying the proposed data processing with the band-pass filtering to select only the 0 order component and deducting the reflection contribution \(I_{\text{ref}}(\lambda)\) of the reference end.

III. Results and Discussions

In contrast to the common-path interferometer shown in Fig.1, Skaar proposed a similar Fabry-Perot structure in which the first and second reflectors are the tested FBG and its fiber end, respectively [3]. To compare the performance difference between the two common-path interference schemes and to verify the accuracy and repeatability of our proposed measurement, we have repeatedly measured the group delay of a uniform-period FBG for five times by using the two common-path interferometers and the above-mentioned data processing technique. We performed the Skaar’s measurement by coupling the broadband light from the other side of the fiber pigtail. Fig.3(a) shows the measured interference spectra \(I_{\text{int}}\) of the FBG filter, where the thin solid line is the measured result by our common-path interferometer (END-FBG) and the thick solid line is the data measured by the inverse common-path scheme (FBG-END). Fig.3(b) and (c) show the measured intensity reflection spectra and group delays of the FBG filter, where the solid lines are the results calculated by using the interference spectra shown in Fig.3(a) and the dash line is the data measured by the optical network analyzer. We notice that the group delay and reflection spectrum obtained by our proposed method are in good agreement.

---

References


---

734
with the measured results of the optical network analyzer. The uncertainty of this measurement is better than 0.5ps in the 10dB bandwidth of the tested FBG. For the inverse common-path configuration case, the larger error of the group time delay is observed in the wavelength range with high reflectance, although the reflection spectrum very match the data measured by Q7760. The reason is that only little amount of the light in the stop-band can transmit through the FBG and reach the second reflector (the fiber end), such that the interference between the two reflected lights is very weak. To verify the inference, we have also measured the group delay of another lower reflective FBG by using the FBG-END configuration with the bare end and higher reflective end coated a metal film. Fig.4(a) shows the measured interference spectra, where the thin and thick solid lines are respectively the measured results by using the coated end and bare end. Fig.4(b) shows the group delays of the test FBG, where the solid lines are the results calculated by using the interference spectra shown in Fig.4(a) and the dash line is the data measured by Q7760. One can see that the measured result with the stronger interference is more accurate than the data by using the bare end. Therefore, we believe our proposed common-path interferometer is better than the Skaar’s configuration for FBGs with high reflectivity.

IV. Conclusion

In this paper, we have proposed a simple and accurate measurement scheme for characterization of the fiber grating devices by using the common-path interferometric technique. In this measurement, a broadband source and an OSA are used to scan the interference spectrum, and then the group time delay and intensity reflection spectrum can be obtained from the Fourier transformation of the interference pattern. We have repeatedly measured the group delay of a test FBG filter. The experimental results excellently match the data obtained by the optical network analyzer. We also compare the measurement performance with the inverse configuration proposed by Skaar to show that better accuracy can be obtained with our scheme. Because of its accuracy, repeatability, as well as simplicity in the experimental setup, we believe this technique has the potential to become a very powerful method for measuring the complex reflection spectrum during the manufacture of fiber grating devices.

References

Less Polarization Coupling and High Extinction Ratio of A Near-Elliptic Cladding Polarization-Maintaining Photonic Crystal Fibre

Lin Wang (1), Fengping Yan (2), Shuisheng Jian (3)

1 : Institute of lightwave technology, Beijing Jiaotong University, Beijing, 100044, China, phone:+86-10-51684015 email:linda_wl@163.com
2 : Institute of lightwave technology, phone:+86-10-51688584, email:fpyan@center.njtu.edu.cn
3 : Institute of lightwave technology, phone:+86-10-51683615, email:ssjian@center.njtu.edu.cn

Abstract The new near-elliptic cladding PM-PCF is proposed. Less polarization coupling and higher extinction ratio (>30dB) for a wide wavelength range are obtained and it is bending-independent and fairly insensitive to structural irregularity for practical use.

Introduction Nowadays, Polarization-maintaining photonic crystal fibres (PM-PCF) were successfully developed by introducing highly asymmetric fibre structures [1-3]. However, in real situation, maintaining the structure throughout the fibre fabrication processes is nontrivial, which results in the inevitable additional polarization coupling and extinction ratio deterioration even more susceptible than the conventional fibres.

In this paper, the new near-elliptic cladding PM-PCF with five different hole diameters is proposed. Since the refractive index decreases gradually from x-axis to y-axis, less polarization coupling and higher extinction ratio can be obtained compared to the conventional cross-sectional structures with sharp corners or sloping walls. Moreover, the new PM-PCF is proved to be fairly insensitive to structural irregularity and bending-independent which is useful from the practical point of view.

Fibre Configuration
The near-elliptic cladding consists of eighteen air-holes and six silica-filled holes along x-axis with five different hole diameters (d1~d5) as shown in Fig.1, where A is the hole pitch. According to coupled-mode theory, good polarization maintaining can be obtained by small power-coupling coefficient h and the relationship among h, birefringence (B) and polarization extinction ratio (PER) can be expressed as[4]

\[ h = \left( \frac{k_0^2}{4} \right) \left( \Gamma(L_B) \right) \]  
(1)

\[ \text{PER} = 10 \times \log \left[ \tanh \left( hz \right) \right] \]  
(2)

Therefore, higher extinction ratio can be obtained by higher birefringence in the same structures. Both of the two parameters of near-elliptic cladding PM-PCF (fibre#1) with six rings and fibre#2 (the same as fibre#1 except for uniform hole diameters) are calculated by full vector beam propagation method (FV-BPM) [5] in Fig.2 (a)and (b), respectively.

As it shown in Fig.2. Since the refractive index decreases gradually from x-axis to y-axis, less polarization coupling and higher extinction ratio can be obtained in near-elliptic cladding PM-PCF even if the birefringence is
lower compared to the cross-sectional structures with sharp corners or sloping walls such as fibre\#2.

**Polarization Characteristics**

Recently, the configuration of fibre\#3 (shown in Fig.3) of which the core is surrounded by eight enlarged and other normal-sized air-holes was proposed to show good performance because of its high birefringence, low confinement loss, low bending loss and low polarization crosstalk for a large range\(^6\).

Compared to near-elliptic cladding PM-PCF (fibre\#1) in the condition of uniform birefringence (1.965×10\(^{-3}\) at 1.55\(\mu\)m) by adjusting \(d_1\) and \(d_2\), the extinction ratio of the two fibres are shown in Fig.4. As it expected, less polarization coupling in near-elliptic cladding PM-PCF results in high extinction ratio (>30dB) for a wide wavelength range and 7~8 dB higher extinction ratio than fibre\#3.

Moreover, the refractive index which decreases gradually from x-axis to y-axis can also reduce polarization coupling induced by nonorthogonal polarization axis because of structural irregularity such as changed size or position of air-holes. Two of air-holes in each fibre are moved to vary the \(x\)-polarization axis in the same degree (\(\Delta \theta=2^\circ\)) and the results of percentage extinction ratio drop are shown in Fig.5. It is proved that the extinction ratio of near-elliptic cladding PM-PCF is fairly insensitive to structural irregularity (percentage extinction ratio drop<3% when \(\Delta \theta=2^\circ\)).

**Conclusions**

The new near-elliptic cladding PM-PCF with five different hole diameters is proposed. Since the refractive index decreases gradually from x-axis to y-axis, less polarization coupling and higher extinction ratio can be obtained compared to the conventional cross-sectional structures with sharp corners or sloping walls. Moreover, the new PM-PCF is proved to be fairly insensitive to structural irregularity and bending-independent which is useful for polarization sensitive optical devices and sensors.

**References**

7. N.Guan et al, ECOC2004 paper Mo4.3.4
Tunable Fiber Laser Using Broadband Fiber Mirror
Integrated Tunable Fiber Bragg Gratings

Dept. of Electronic Engineering, National Taiwan University of Sci. & Tech., Taipei 106, Taiwan , R.O.C.
Cheng-Ling Lee
Department of Electro-Optical Engineering National United University, Miaoli 360, Taiwan, R.O.C.

Abstract
A linear-cavity fiber laser using broadband fiber mirror and tunable fiber Bragg gratings is experimentally demonstrated. The power variation and the tunable range of laser are +/- 1.0 dB and 29.0 nm, respectively, in the conventional band.

1. Introduction
In recent years, laser sources have various applications in sensing, instrument test, optical signal process, optical communications, and in photon analog-to-digital converters (ADCs) [1]. Moreover, the tunable lasers are flexible for usage in wavelength routing, wavelength protection and optical measurement. Several methods have been proposed to investigate tunable lasers, for example, by using photonic crystal mirrors [2] and a multiple quantum-well waveguide. Among them, erbium-doped fiber ring laser (EDFRL) is a potential candidate because of the feature of low temperature sensitivity of wavelength. To select different wavelength under different operations has been achieved by using different optical filter devices like Mach–Zehnder filters or in-fiber comb filters. However, the tunable range is narrow [3] or the line width is wide due to large filter bandwidth. Another approach is the dual-wavelength tunable fiber laser. Lasing occurred between two different wavelengths alternatively by using a cavity is consisted of two overlapping cavities with a common gain medium [4]. Fiber Bragg grating (FBG) pair lasers are popular. However, it has the risk of reflected wavelength misalignment between the FBG pair. The paper addresses a novel tunable laser design. Our results show that the widely tunable ability could be realized by appropriate switching, one 1x2 optical switch (OSW), pair and the tunable FBGs (TFBGs) in compressed or strained condition.

2. Operation Mechanism
For the investigation of tunable FBG, we embed the FBG between the outer lamination. The composite with TFBG embedded inside is attached on a 3-point tuning device by using instant adhesive glue. By strained or compressed tuning the precise screw of 3-point bending device, we can apply both directions in the transverse displacement for increasing the tunable range up to ±10 nm easily. No other complicated and bulky components are needed to perform the tunable function. A 1x2 OSW pair is used to choose an appropriate FBG between 2 band operations. The tunable range of FBG1 and FBG2 is about 15 nm (not the record sample in lab) for each. Without tuning, the FBG1 has central reflective wavelength at 1540.5 nm and the FBG2, 1552.7 nm. We demonstrate the FBG1 to be tuned from 1535- to 1545 nm, while FBG2 to be tuned from 1547- to 1561 nm. Fine tuning resolution as precise as 0.1 nm for FBG is possible. The decided tuning wavelength, either in the shorter C band (1530-1545 nm) or the longer C band (1546-1560 nm), is dependent on the switching status of OSW. This switching speed is 10 ms for this OSW. The transmission (or reflection) spectrum is not degraded when the FBG are tuned to the boundary points (1533-, 1546, and 1562 nm). We replace a conventional FBG with the fiber mirror and add another TFBG to construct the laser cavity.

Fig. 1 Proposed wavelength-tunable fiber laser based on broadband fiber mirror and tunable FBGs.

Fig. 1 shows proposed scheme of the wavelength tunable laser. This laser cavity includes 3 meters of erbium-doped fiber (EDF), a wavelength division multiplexer (WDM) coupler and a 1480 nm pumping laser. The Broadband fiber mirror (BFM) is a broadband wavelength reflector. It is broad band with high reflectivity up to 97% from 1450- to 1600 nm. It has the benefit to act as a broadband mirror to reduce the possible reflective wavelength misalignment as the case when FBG pairs are used as cavity ends. At the right hand side, there is one tunable FBG (TFBG) with 50% reflectivity. Because the laser signal comes out from the transmission port of FBG, the best reflection of TFBG is calculated to be 50%. A laser cavity is formed because the lasing signal will resonate between the fiber mirror.
and TFBG. Then the cavity starts lasing by passing 50% of lasing power through the TFBG for each round-trip. The lasing wavelength is determined by the reflected wavelength of TFBG. The tunable range could be doubled by appropriate switching the 1x2 OSW pair and compress or strain the TFBGs.

3. Experimental Results and Discussion

Fig. 2 shows the lasing power versus pump power using 3 meters EDF and 100 mW pumping power. The pumping efficiency \( PE \) is defined as:

\[
PE = \frac{\lambda_2}{\lambda_3} \times \frac{P_s}{P_p} \times 100\% \quad (1)
\]

where \( P_p \) is pumping power @1480 nm and \( P_s \) is lasing power @1545.3 nm. \( PE \) with a high value of 31.26% is obtained. It is a linear curve with low threshold current. Fig. 3 (a) shows superposed optical spectra of the BFM-integrated-TFBG tunable fiber laser. The overlapping spectra of tunable laser could cover the longer C band (1546-1562 nm). To further increase the tuning range for the C band, two TFBGs and a 1×2 optical switch (OSW) pair are used instead. Its overlapping spectra are shown in Fig. 3 (b). The average lasing power is 12.8 dBm for each channel and power variation is less than ±1.0 dB over the whole tuning range. The side mode suppression ration (SMSR) of laser is ranging from 57.8 to 62 dB and narrow line width is less than 0.05 nm, limited by resolution of the OSA. No polarization mode competition effect is observed. A versatile and cost-effective laser source should have tunable ability to allow choosing which wavelength is needed, or to scan over a range of bandwidth. Our proposal TFBG lasers may fulfill such requirements. In theory, the proposed method could apply to the whole C+L band by parallel connection of several TFBGs with a 1xN OSW pair. Of course, TFBGs with appropriate originally central reflective wavelengths are necessary.

4. Conclusion

In summary, we propose and demonstrate C-band linear-cavity tunable fiber laser based on BFM and TFBGs. Brandband band tunable range could be obtained using a 1x2 OSW pair. The laser linewidth, power variation and tuning range are 0.05 nm, ±1.0 dB and 29.0 nm, respectively, in this demonstration. High SMSR of better than 57.8 dB is achieved. This proposed laser may find vast applications in optical communication or optical measurement.

5. Acknowledgement:

The work was supported by National Science Council, Taiwan, under grant numbers of NSC 95-2622-E011-013-CC3 and NSC95-2219-E-011-007.

References

Gain-Clamped Fiber Amplifier with Reflective Configuration

Hsi-Shan Huang¹, Shih Hsu², and Tsair-Chun Liang¹

¹Graduate Institute of Electro-Optical Engineering, National Kaohsiung First University of Science and Technology, Kaohsiung, Taiwan 811, R.O.C.
Tel: 886-7-6011000 ext. 2712, Fax: 886-7-6011096, E-mail: tcliang@ccms.nkfust.edu.tw

² Department of Electronic Engineering, Cheng Shiu University, Kaohsiung, Taiwan, R.O.C.
Tel: 886-7-7310606 ext. 3212, Fax: 886-7-7331758, E-mail: hsushih@csu.edu.tw

Abstract

A new configuration of gain-clamped for L-band fiber amplifier is proposed and demonstrated. A laser cavity was formed by a fiber grating and a fiber mirror containing an erbium-doped fiber. When we used 1615-nm center wavelength with 1-nm bandwidth of fiber grating (FBG) to clamp the gain can offer the minimum gain variation and moderate low noise figure.

I. Introduction

Owing to the operating wavelengths of the L-band EDFAs are very far from the peak emission band of Er³+ ion, the amplifiers are relatively inefficient. In order to improve gain efficiency, the EDFAs by double-pass scheme have been suggested and experimentally verified [1-3]. Gain clamping (GC) of EDFAs have received widely attention in dense wavelength division multiplexing (DWDM) networks with optical add-drop multiplexers (OADMs) and optical cross-connects (OXC's) for stabilizing the channel-gain constant of the surviving channels in the presence of dynamic input power variation or add-drop of optical channels [4,5]. Though the double-pass technique began to be employed in L-band system, the configuration selections of two-stage reflective-type and add gain clamped EDFA for achieving high and constant power characteristics have not yet been reported.

In this paper, we theoretically investigate the 1480-nm-pumping EDFA configuration by using two-stage reflective scheme to amplify 32 digital channel signals operated in the L-band. And by using different FBG’s center wavelengths and bandwidths to clamp the signal gain.

II. EDFA Configuration and Modeling

The configuration of the proposed gain-clamped EDFA is depicted in Fig. 1. In this scheme, the circulator is used to route the amplified signal power output and the fiber mirror is used to reflect pump and signal lights back to the EDF with a reflectance ratio of 100%. Owing to the signal be reflected by fiber mirror into the EDF gain medium to be amplified twice before it leaving the output port. The signal can get better small signal gain than the conventional EDFAs do. The 1480/1580-nm wavelength selective coupler (WSC) is employed to act as the pump/signal optical combiner. The FBG is used as the gain clamping of signal; it can depend on the FBG’s center wavelength and bandwidth to control the gain.

The simulation tool used in this work is named OptiAmplifier about optical fiber amplifier and Laser Design software (Optiwave Corporation, Version 4.0). The EDFA model is based on the model by Giles and Desurvire. The erbium-doped fiber with the erbium ion density of 1.4e+25/ m³ was used. Its core radius is assumed to be 1.8 μm, and numerical aperture is assumed to be 0.23. The insertion loss of WSC coupler is assumed to be the same with 0.5 dB at both 1480 and 1580 nm bands. The insertion loss and isolation of optical isolator and circulator are assumed to be 0.5 dB, 50 dB and 0.8 dB, 50 dB, respectively. The 1480-nm pump output power of pump laser diode is assumed to be 135 mW. A 32-channel signal in the 1574.54 ~ 1600.60 nm wavelength range with a channel spacing of 0.8 nm are considered. And the input power level of each digital channel is set to be –23 dBm.

Fig. 1 Scheme diagram of gain-clamped EDFA

III. Results and Discussion

For investigating the optimum gain and noise figure characteristics of two-stage EDFA configuration as shown in Fig. 1, the first step is to find the optimum EDF1 and EDF2 length in the non-clamped mode. Fig. 2 shows the optimum output spectrum without the FBG of non-clamped L-band EDFA, which design is based on 32-channels, each with an input power level of –23 dBm, from the figure, the 32-channel’s average gain is 24.2 dB, and the gain variation ΔG (= Gmax – Gmin) is about 0.6 dB. Compared with one-stage and single pass L-band EDFA scheme, the signal gain of...
double pass is greater than the gain of single pass about 5 dB.

![Fig.2 Optical spectra of a 32-channel WDM signal with an input signal power of –23 dBm/ch without the FBG.](image)

Then we add the FBG in the scheme for clamping the gain, which is shown in fig. 1. For obtaining the better gain clamping result, we alter the FBG’s center wavelength and bandwidth to control the EDFA gain. From the simulation, we find that the broader bandwidth of the FBG, the smaller signal gain in the L-band EDFA. From the simulation, we find that the FBG bandwidth from 1 nm increasing to 10 nm, the average signal gain from 24.8 dB dropped to 8.1 dB, and the gain variation from 0.5 increasing to 6.9 dB. In this work, we choose 1 nm bandwidth for FBG, which can get the better signal gain, the smaller noise figure and the minimum gain variation. Additionally, the position of the FBG center wavelength can also affect the gain of EDFA, the nearer of the 32 DWDM channels, the stronger of the gain clamping effect in the L-band, we choose 1615-nm center wavelength in the system, which can get fairly signal gain, the smaller noise figure and the minimum gain variation.

Figure 3(a) shows the non-clamped optical gain and noise figure characteristics of the multiple channel based EDFA with various channel numbers. For a normal operation of EDFA, the number of input channels launched into the amplifier is 32. The smaller channel number of surviving channels of 24, 16, 8 and 4 also simulates the different cases of EDFA in the presence of add/drop operation. In the non-clamped case, we find that the optical spectral gain variation from 32 channels dropping to 4 channels is 8.9 dB, and the corresponding maximum noise figure is about 4.5 dB. When an optimum FBG is inserted, the gain clamping effect is observed. Figure 3(b) shows the clamped optical gain and noise figure characteristics of the multiple channel based EDFA with various channel numbers. We find that the optical spectral gain variation from 32 channels dropping to 4 channels is 2.9 dB, and the corresponding maximum noise figure is about 4.8 dB. Compared with non-clamped EDFA, the gain variation of clamped EDFA has a less than non-clamped EDFA of 6 dB, and the noise figure of clamped EDFA does only degrade than non-clamped EDFA of 0.3 dB, so, the clamped EDFA is better than the non-clamped EDFA.

![Fig.3 The calculated optical gain and noise figure characteristics of the multiple channel based EDFA with various channel numbers. (a) Non-clamped EDFA (b) Clamped EDFA.](image)

IV. Conclusions
A gain-clamped EDFA with improved output gain, which uses double-pass technique, was proposed. Compared with single pass scheme, we find that the average gain of double pass scheme without FBG is better than the single pass gain about of 5 dB. When an FBG was inserted, the output gain was clamped. We find that when using FBG of 1615-nm center wavelength and 1-nm bandwidth can offer a minimum gain variation of 2.9 dB and moderate low noise figure of 4.8 dB. The investigative result provides the L-band gain-clamped configuration selection for multi-wavelength DWDM lightwave systems.

References
Dispersion Flattened Decagonal Photonic Crystal Fiber

S. M. Abdur Razzak\textsuperscript{a}, Yoshinori Namihira\textsuperscript{b}, Feroza Begum, Shubi Kaijage, Nguyen H. Hai, and Nianyu Zou

Graduate School of Engineering and Science, University of the Ryukyus,
1 Senbaru, Nishihara, Okinawa 903-0213, Japan.
E-mail: \textsuperscript{a}k058470@eve.u-ryukyu.ac.jp, \textsuperscript{b}namihira@eee.u-ryukyu.ac.jp

Abstract
Ultra-flattened dispersion of 0 \pm 0.80 ps/(nm-km) within a 370 nm band and low confinement loss 0.1 dB/km is obtained from a decagonal photonic crystal fiber. This fiber has a modest number of design parameter.

Key words: decagonal photonic crystal fiber (PCF), chromatic dispersion, effective area \((A_{eff})\), confinement loss

1. Introduction
In photonic crystal fibers (PCFs) \cite{1}, control of chromatic dispersion is crucial for practical applications to optical fiber communication systems, dispersion compensation and nonlinear optics \cite{2}.

Various index guiding hexagonal PCFs (H-PCFs) with remarkable dispersion and leakage properties \cite{3} have been reported to date including PCFs with two defected rings, PCF designs with uniform optimized air-holes, PCFs with several different air-hole diameters, and PCFs with a defected air-hole in the core \cite{4}. In almost all instances, either many rings layer, or many design parameters, or submicron adjustments are required; efforts continue to locate a simple PCF structure. As a part of the ongoing effort to locate a novel PCF structure, we propose a decagonal PCF (D-PCF) having a modest number of design parameters.

In this paper, it is numerically shown that a three-ring D-PCF can operate a single mode in the entire telecommunication windows with ultra-flattened chromatic dispersion and very low confinement losses. D-PCF with such novel properties as wideband single-mode operation with ultra-flattened chromatic dispersion and low confinement losses may pave the way for different potential applications in optical communication systems and nonlinear optics.

2. D-PCF structure
Fig.1 shows air-hole geometry of the D-PCF. For the purpose of simplicity only two rings are shown. Pitch, \(\Lambda\) is the spacing between air-hole centers on the adjacent rings and the spacing between air-hole centers on the same ring is \(\Lambda_1\). \(\Lambda_1\) is related to \(\Lambda\) by the relation \(\Lambda_1 \approx 0.618\Lambda\). Therefore, the maximum diameter of an air hole may have a value of 0.3075\(\Lambda\). The D-PCF is constructed by repeating a unit isosceles triangular lattice with vertex angle of 36° around the silica core. The core diameter is 2\(a\), where ‘a’ equals \(\Lambda-d/2\). The air-holes with diameter \(d\) are located at each corner of the isosceles triangle resulting lower refractive index due to higher air-filling fraction (AFF) around the core compared to an H-PCF. Using the definition of AFF \cite{5} it can easily be shown that the air-hole radius of a D-PCF can be around 18% smaller than that of the H-PCF for a same AFF and pitch values. Fig.2 is the proposed D-PCF structure with three rings of which the first ring is defected (hereinafter known as defected D-PCF). The air-holes on the first ring have a diameter \(\text{d}_1\) and \(\text{d}\) is the air-hole diameter on the other rings. The core diameter is 2\(a\), where ‘a’ equals \(\Lambda-d/2\).

3. Simulation results
The finite difference method (FDM) \cite{6} with anisotropic perfectly matched boundary layers (PML) is used to calculate the chromatic dispersion and confinement loss.

Fig.1 Geometry of a D-PCF with isosceles triangular lattices

Fig. 2 Geometry of the defected D-PCF with one defected ring
Once the modal effective indices $n_{\text{eff}}$ is obtained by solving an eigenvalue problem drawn from Maxwell equations using the FDM, the chromatic dispersion parameter $D(\lambda)$, confinement parameter $L_c$ and effective area $A_{\text{eff}}$ can be given by the equations [4].

Fig. 3 shows chromatic dispersion properties of the proposed defected D-PCF shown in Fig. 2. Optimizing the parameters ultra-flattened chromatic dispersion of 0 ± 0.80 ps/(nm·km) is obtained (solid line) for three-rings, $d_1 = 0.60$ µm, $d = 1.15$ µm and $\Lambda = 2.35$ µm. Fig. 4 shows the confinement loss and effective area of the same fiber for the said optimum parameters. It is seen that the confinement loss is as low as a 0.1 dB/km within the flat-dispersion wavelength band. The effective area corresponding to $\lambda = 1.55$ µm is about 10.0 µm$^2$. Fig. 5 shows the fundamental mode field width of the D-PCF at $\lambda = 1.55$ µm. The parameters are set at $d_1 = 0.60$ µm, $d = 1.15$ µm and $\Lambda = 2.35$ µm. It is confirmed that the fiber can operates as a single mode fiber as the confinement loss for second order modes at shorter wavelength is found to be more than a 200 dB/m.

Therefore, it can be concluded that the proposed defected D-PCF may be a suitable candidate for the optical communication systems because of their ultra-flattened dispersion and very low confinement losses.

4. Conclusion

Decagonal PCF for ultra-flattened chromatic dispersion has been reported for single mode operation within a 370 nm band with a low confinement loss 0.1dB/km. It was observed that there are two apparent advantages of this PCF. First, it can assume higher refractive index difference between the cladding and the core thereby reducing the confinement loss, and second, design parameters are reasonably a modest number.

References


Fig. 3 Wavelength dependence of chromatic dispersion properties of the defected D-PCF with four-rings, $d_1 = 0.60$ µm, $d = 1.15$ µm and $\Lambda = 2.35$ µm.

![Figure 3](image3.png)

Fig. 4 Confinement loss (solid line) and effective area (dash-dot line) of the defected D-PCF with four rings, $d_1 = 0.60$ µm, $d = 1.15$ µm and $\Lambda = 2.35$ µm.

![Figure 4](image4.png)

Fig. 5 Fundamental mode field distribution of the defected D-PCF at 1.55 µm wavelength with four rings, $d_1 = 0.59$ µm, $d = 1.15$ µm and $\Lambda = 2.35$ µm.

![Figure 5](image5.png)
Nearly chirp-free and pedestal-free pulse compression

Qian Li, P. K. A. Wai, K. Nakkeeran and K. Senthilnathan
1Photons Research Center and Department of Electronic and Information Engineering, The Hong Kong Polytechnic University, Hong Kong.
2Department of Engineering, Fraser Noble Building, King's college, University of Aberdeen, Aberdeen AB24 3UE, UK.

Abstract: We proposed a compact pulse compression scheme, which consists of a linear grating and a nonlinear grating, to effectively compress both hyperbolic secant and Gaussian shaped pulse profiles.

1. Introduction

Recently, we reported that pedestal-free compression of optical pulse is possible using a nonlinear fiber Bragg grating (NFBG) with exponentially decreasing dispersion profile. The input pulse however must have a hyperbolic secant pulse shape and a quadratic chirp [1]. In general, it is not easy to produce the precise pulse shape and chirp [2]. In this paper, we study the compression of initially chirp-free pulse using a linear chirped fiber Bragg grating (CFBG) to produce the required chirp profile for the compression in a NFBG with exponentially decreasing dispersion. In particular, we assume that the input pulse is a chirp-free Gaussian pulse or hyperbolic secant pulse. Our simulation results show that the pedestal generated from an input Gaussian pulse is much small than that of an input hyperbolic secant pulse showing that the compression by the NFBG is more sensitive to the chirp profile than the pulse shape. We also found that the initial Gaussian profile evolves into a hyperbolic secant profile after the compression in the NFBG.

2. Self-similar pulse compression

Self-similar pulse compression near photonic bandgap of gratings has been described in [1], where the pulse width \( T(z) = T_0 \exp(-2\alpha_{20}\beta_{20}z) \) and the normalized chirp \( C(z) = \alpha_{20}T_0^2 \exp(-2\alpha_{20}\beta_{20}z). \) As the normalized chirp decreases, the time-bandwidth product tends to 0.315 (transform-limited hyperbolic secant pulse). Fig. 1 shows the variation of the time-bandwidth product from 0.761 to 0.327.

Fig. 1. Evolution of time-bandwidth product.

We then launched a chirp-free Gaussian pulse (\( \exp(-T^2/T_0^2/2) \), \( T_0 = 10 \) ps) into linear grating. It is well known that the pulse after the linear CFBG is still Gaussian pulse with quadratic chirp. The chirp parameter \( \alpha_{20} \) is chosen to be negative, \( \alpha_{20} = -\beta_z z / (T_0^2 + \beta_z^2 z^2) < 0 \) (1)

Obviously, \( \left| \alpha_{20} \right| = 1/2 \left( T_0^2/\beta_z z + \beta_z z \right) \leq 1/4T_0^2 \) (2)

We choose a 4 cm long linear grating with constant dispersion \( \beta_z = 25 \) ps/cm here for demonstration.

3. Pre-chirp

First, we launched a chirp free hyperbolic secant pulse \( \left( \text{sech}(T/T_0) \right), T_0 = 10 \) ps) into the linear CFBG which is 4 cm long with normal dispersion \( \beta_z = 10 \) ps/cm.

Normal dispersive medium is used here to introduce negative \( \alpha_{20} \) required for the nonlinear pulse compression later. Fig. 2 shows the pulse profile before and after the linear CFBG in linear and logarithmic scale. From Fig. 2(d), the chirped pulse retains its hyperbolic secant pulse profile. Fig. 3 shows a polynomial fit of the phase of the pulse after linear grating CFBG. The coefficients of the \( t^4, t^2 \) and \( t^0 \) terms are 0.0111, -0.18555 and 0.168 respectively where \( t = T/T_0 \).

Fig. 2. Pulse profile of the initial ((a) and (b)) and chirped pulse ((c) & (d)) in both linear ((a) and (c)) and logarithmic scale (b) and (d).

Fig. 3. The phase profile after the linear CFBG.
\( \alpha_{20} \) is set to \(-0.0018555 \) THz\(^{2}\). We set \( T_{0} \sqrt{2}/|\beta_{20}|=1/\gamma_{G}/P_{0} \) and choose \( \beta_{20}=-40 \) ps\(^{2}\)/cm for illustration. Fig. 4 shows the pulse profile after the NFBG in both linear and logarithmic scales. The fitted hyperbolic profile (dashed lines) is of the same peak power and FWHM as final compressed pulse. We found that the pedestal energy of the final compressed pulse is 5.44\% [3]. Polynomial fitting of the phase of the pulse after nonlinear grating gives the coefficients of the \( \tilde{r}^{2} \) and \( \tilde{r}^{3} \) terms to be 0.0041, 0.0033 and 0.5661 respectively. Compared to the results in Fig. 3, the coefficients of \( \tilde{r}^{4} \) are greatly reduced, i.e. the final compressed pulse is close to chirp-free. Fig. 5 shows the evolution of the FWHM of an initial hyperbolic secant pulse in the linear CFBG and the NFBG. The final FWHM compression factor (compared to initial chirp-free pulse) is 7.23.

Next we launched the chirped Gaussian pulse into the NFBG. Following [4] we choose the parameters

\[
T^{2}(4 \text{ cm})/\sqrt{2} |\beta_{20}| = 1/\gamma_{G} / P(4 \text{ cm})
\]

where \( T(4 \text{ cm}) \) and \( P(4 \text{ cm}) \) are the pulse width and peak power of the pulse after linear CFBG, and \( \beta_{20} \) is the initial dispersion value of NFBG. In our example, \( \beta_{20}=-25 \) ps\(^{2}\)/cm, \( T(4 \text{ cm}) = 10\sqrt{2} \) ps, and the NFBG is 16 cm long. Fig. 6 shows the pulse profile after the linear CFBG and after the CFBG in linear and logarithmic scales. From Fig. 6 (d), the main portion of the compressed pulse is almost the same as the fitted hyperbolic pulse. We found that the pedestal energy of the final compressed pulse is only 0.0935\%. Polynomial fitting for the phase of the final compressed pulse gives the coefficients of the \( \tilde{r}^{4} \) and \( \tilde{r}^{3} \) terms to be \(-0.0007, -0.0487, -0.9355 \), respectively indicating that the pulse is close to transform-limited. Fig. 7 shows the evolution of the FWHM in the linear CFBG and the NFBG. The final FWHM compression factor (compared to the FWHM of initial chirp-free Gaussian pulse) is 6.3.

**5. Conclusions**

We have numerically demonstrated nearly chirp-free and pedestal-free optical pulse compression using a linearly chirped fiber Bragg grating and a nonlinear fiber Bragg grating with exponentially decreasing dispersion. A compact pulse compression schemes using fiber Bragg gratings is feasible.

**6. References**


Optic Fiber Sensor Based on Etching Long Period Gratings of Photonic Crystal Fibers

Wen-Fung Liu¹, Sheng-Zeng Peng¹, Hsin-Wen Peng¹, Ming-Yue Fu², Lung Ai³, Hao-Jan Sheng⁴ and Chuen-Lin Tien¹

¹Department of Electrical Engineering, Feng-Chia University, Taichung, Taiwan 407, R.O.C.
²Department of Avionics Engineering, Air Force Academy, Kaohsiung, Kaohsiung, 820, Taiwan, R.O.C.
³Department of Electrical Engineering, Chung-Cheng Institute of Technology, National Defense University, Taichung, Taiwan 335, R.O.C.
⁴The Graduate Institute of Electrical and Communications Engineering, Ph.D. Program, Feng-Chia University, Taichung, Taiwan 407, R.O.C.

Abstract: A new fiber sensor based on an etched long period grating in photonic crystal fibers is used for sensing temperature and various refractive index solutions.

1. Introduction

The device applications based on photonic crystal fibers (PCF) include silica/polymer hybrid tunable fiber waveguides, fiber couplers, grating based tunable filters, etc [1-4]. A long period grating (LPG) fabricated in a PCF has potential applications in a fiber sensing head. For a conventional LPG, in the transmission spectrum there exist several different mode loss peaks at different resonant wavelengths, which should satisfy the phase matching condition. In this study, the temperature effect of the etched an LPG in a PCF is investigated by using etching fiber technology to reduce the diameter of fiber cladding for improving the temperature sensitivity. Moreover, when the surrounding solution index of the etched LPG in a PCF is increased, the resonant wavelength shifts toward the shorter wavelength side. Thus, from the experimental results, this technique maybe provide a novel sensing head for discriminating different chemical solutions or distinguishing various constituents in a chemical solution by combining the thin-film coating technique on the surface of the etched LPG.

2. Basic principles

According to the reference of [2], the cladding effective index \( n_{\text{cl}}^{\text{eff}} \) of an LPG in a PCF can be simplified as

\[
 n_{\text{cl}}^{\text{eff}} = \left( \frac{\pi}{2 \sqrt{3}} (d / \Lambda_{\text{LPG}})^2 (n_{\text{co}}^2 - n_a^2) \right)^{1/2}
\]

(1)

where \( d \) is the average hole diameter, \( n_{\text{co}} \) is the core index and \( n_a \) is the refractive index of inside holes. By the combination of equation (1) and grating phase-matching condition, the resonant wavelength \( \lambda \) of an LPG in a PCF can written as

\[
 \lambda = (n_{\text{co}} - n_{\text{cl}}^{\text{eff}}) \Lambda_{\text{LPG}} = n_{\text{LPG}(\Omega)} \lambda_{\text{LPG}}
\]

(2)

where \( n_{\text{LPG}(\Omega)} = (n_{\text{co}} - n_{\text{cl}}^{\text{eff}}) \).

The resonant wavelength shift \( \Delta \lambda \) induced by the temperature perturbation \( \Delta T \) can be expressed as

\[
 \frac{\Delta \lambda}{\lambda} = \frac{\alpha + \zeta}{\lambda} \Delta T
\]

(3)

where \( \alpha = \frac{1}{\lambda} \frac{\Delta \lambda}{\Delta T} \) is the fiber thermal-expansion coefficient and \( \zeta = \frac{1}{n_{\text{LPG}(\Omega)}} \frac{\Delta n_{\text{LPG}(\Omega)}}{\Delta T} \) is the fiber thermal-optic coefficient. Refer to Eq. (3), while a part of the cladding layer is removed by means of etching fiber technique to keep the original symmetrical structure, the fiber thermal-optic coefficient will be increased due to the \( n_{\text{LPG}(\Omega)} \) to be decreased and thus to cause the temperature sensitivity to be further improved. Besides, from Eq. (1), when the index of surrounding air-holes solution of LPG is changed, it will result in a variation of the cladding effective index \( n_{\text{cl}}^{\text{eff}} \). Thus, from Eq. (3), the resonant wavelength should be shifted owing to the index change of surrounding solution. This phenomenon can be verified by the following experimental results.

3. Experimental set-up and results

In this experiment, the LPG is fabricated by utilizing the electric arc of a fiber splicer to heat the PCF and to cause the fiber index change. The PCF used is the holey fiber with the type of ESM_12_1 from Blaze Photonics Company. By controlling discharging time of the fiber splicer (Fujikura FSM-40S-B) and adjusting the translation-stage to periodically move the PCF to be heated, the fabrication of an LPG in a PCF can be achieved by the point-by-point writing technique.

In order to investigate the temperature effect for an etched LPG in a PCF, a non-etching LPG is firstly measured for observing the resonant
wavelength shift by using an OSA. The experimental results show the wavelength shift of the LPG loss peak from 1552.25 nm to 1553.8 nm with the total wavelength shift of 1.550 nm in the temperature range from 200°C to 300°C. The shift direction is toward the shorter wavelength side while the temperature is reduced. The temperature sensitivity of about 9.1 pm/°C is obtained with a nice linear property as shown in Fig. 1. An LPG in a PCF is etched down from 125 to 63 µm by using HF solution. By using this etched LPG in temperature measurements, the experimental results are shown in Fig. 2. The wavelength shift is from 1500.45 to 1504.45 nm in the temperature range from 200°C to 300°C with the temperature sensitivity of 23.5 pm/°C. Comparing the temperature sensitivity of the etched LPG with that of the non-etched LPG, it is improved to be 2.6 times. Thus, from experimental results we can predict that the temperature sensitivity of the etched LPG in PCFs should be further enhanced by reducing the cladding diameter.

For sensing different index solutions, the etched LPG of 63 µm in a PCF is inserted in four different index oils of 1.333, 1.358, 1.365, and 1.47. The experimental results are shown in Fig. 3. From these curves, we can clearly observe that the resonant wavelength is shifted toward shorter wavelength side when the oil index is increased. This phenomenon can be explained from Eq. (2). The resonant wavelength shift is proportional to the increment of cladding effective index surrounding the cladding air-holes of PCF as shown in Fig. 4. Thus, from above experimental results, we see that this new sensing head has the ability to discriminate different index solutions in the index resolution of 0.001 by using an OSA with the resolution of 0.01 nm. In the future, the sensitivity in discriminating index will be further improved by designing different types of gratings and combining the thin-film materials to be coated on the surface of the etched LPG.

4. Conclusions
We have experimentally demonstrated that the temperature sensitivity of an LPG in a PCF is improved to be 2.6 times while the cladding diameter is etched down from 125 to 63 µm. Additionally, the etched LPG of 63 µm in a PCF is used for discriminating different index oils with the sensitivity of 12.4 nm per unit index.

Reference

Fig. 1 The curve for the resonant wavelength shift of non-etched LPG versus temperature variations.

Fig. 2 The curve for the resonant wavelength shift versus temperature variations for etching the fiber to be 63 µm in diameter.

Fig. 3 The transmission spectra of an etched LPG of 63 micro-meters for measuring four different index oils.

Fig. 4 The wavelength shift of grating loss-dip versus different index oils.
RADIATION RESONANCE OF TRANSVERSE MAGNETIC WAVE IN DIELECTRIC MICRO-SPERES: AN ASYMPTOTIC SOLUTION

Anisur Rahman and Sunil Kumar
Thermal Optics Laboratory, Department of Mechanical Engineering, Polytechnic University
6 MetroTech Center, Brooklyn, NY 11201, USA
Email: mrahma07@utopia.poly.edu, mdanisur@gmail.com; Tel: (1)718 260 3012; Fax: (1)718 260 3532

ABSTRACT

We describe radiation resonances of Transverse Magnetic (TM) wave in dielectric micro-spheres. An Asymptotic solution is presented for TM wave and resonance condition is characterized. This study could potentially be used for designing optical sensors.

INTRODUCTION

The dielectric micro-sphere resonator has earned amazing research interest in recent years due to its unique characteristics for potential applications primarily in biomedical sensing, optical communications, multiplexing, spectroscopy, quantum mechanics, and others [1-4]. Over the years, study of morphology dependent resonances (MDR), also known as whispering gallery modes (WGM), has now been well established both theoretically and experimentally [5-10]. However, theoretical approaches published in the literatures are quite complicated. WGM and MDR are presented in the literatures [1, 5-7] based on classical quantum mechanics which are difficult to follow and do not give simple usable results or are not sufficient enough to explain the MDR peaks. The present study is the first to our knowledge of a complete asymptotic solution of TM wave in dielectric micro-spheres for large size parameter which is simpler and mathematically robust than existing approaches presented in the literatures [1, 5-7].

THEORY

When a dielectric micro-sphere and optical fiber coupled, Electro-Magnetic (EM) wave enters to micro-sphere from fiber and travels inner surface of the sphere due to total internal reflection (TIR). We have developed a robust asymptotic approach to explain MDR peaks for TM wave in dielectric micro-spheres. The details of mathematical expansion have not been included here due to space limitation, however, similar asymptotic expansions for TE wave have already been presented in our previous publication [13].

When $\vec{H}$ tangential to sphere surface (TM mode), electro-magnetic fields inside the sphere can be expressed as [11],

$$\vec{E}_{in} = -i \sum_{n=1}^{\infty} E_n d_n \vec{N}^{(1)}_n$$

$$\vec{H}_{in} = -\frac{k}{\omega \mu} \sum_{n=1}^{\infty} E_n d_n \vec{M}^{(1)}_n$$

where $\vec{E}$ & $\vec{H}$ are electric & magnetic field vectors, respectively. The scattered electro-magnetic fields from the micro-sphere can be expressed as [11]

$$\vec{E}_{out} = i \sum_{n=1}^{\infty} E_n a_n \vec{N}^{(3)}_n$$

$$\vec{H}_{out} = \frac{k_{out}}{\omega \mu} \sum_{n=1}^{\infty} E_n a_n \vec{M}^{(3)}_n$$

The essential boundary conditions for both electric and magnetic fields can be expressed as

$$\vec{E}_{in} \bigg|_{\tan\ g} = \vec{E}_{out} \bigg|_{\tan\ g} \quad \text{at} \ r=R$$

$$\vec{H}_{in} \bigg|_{\tan\ g} = \vec{H}_{out} \bigg|_{\tan\ g} \quad \text{at} \ r=R$$

Applying equation (1) and (2) into equation (3), the electric and magnetic field boundary conditions, and expanding the vector spherical harmonics in component form [11] yields,

$$a_n \frac{1}{k_{out}} \frac{d}{d\rho} [\vec{N}^{(3)}_n(\rho)]|_{\rho=R} + d_n \frac{1}{k_n} \frac{d}{d\rho} [\vec{M}^{(3)}_n(\rho)]|_{\rho=R}=0 \quad (4)$$

$$a_n \frac{k_{out}}{\mu_2} j_n(k_{out}R) + d_n \frac{\mu_1}{\mu_2} j_n(k_nR) = 0 \quad (5)$$

where $\mu_1$ & $\mu_2$ are permeability of micro-sphere & outside medium (air), respectively. Introducing size parameter:

$$x = \frac{2\pi R m_2}{\lambda_0}$$

$$k_{out} = \frac{2\pi m_2}{\lambda_0}, \quad \text{and relative refractive index:} \quad m = \frac{m_1}{m_2},$$

where $\lambda_0$ is the wavelength of light at free space, $m_1$ & $m_2$ are represented as refractive indices of micro-sphere & outside medium (air), respectively. So from equations (4) & (5), there will be two equations for $a_n$ & $d_n$ and after solving those equations we will have resonance condition of TM wave for micro-spheres.

REFERENCES

[1] Johnson, B., 1994, Morphol...

[2] Serpengüzel, A. et al., 1997, Enhanced coupling to...


[8] Johnson, B., 1994, Morphol...

[9] Serpengüzel, A. et al., 1997, Enhanced coupling to...

[10] Tapalian, H., et al., 2002, ...


Equation (6) is the natural resonance condition for TM modes of dielectric micro-spheres. \( j_s \) and \( h_0^1 \) are spherical Bessel functions of first & third kind, respectively. Assuming \( \mu_1 = \mu_2 \) for continuity. Equation (6) can be further simplified as follows [12]:

\[
\frac{d}{d\rho} \left[ \frac{j_s(\rho)}{h_0^1(\rho)} \right] = \frac{1}{m^2 \mu_2} \frac{d}{d\rho} \left[ j_s(\rho) \right] \quad (6)
\]

After applying asymptotic expansions of spherical Bessel functions [12] and performing rigorous mathematical operation, a simplified expression for TM mode can be achieved as,

\[
x \frac{d}{d\rho} \left[ h_0^{(1)}(\rho) \right] = \frac{m}{m^2} \frac{d}{d\rho} \left[ j_s(\rho) \right] \quad (7)
\]

After applying asymptotic expansions of spherical Bessel functions [12] and performing rigorous mathematical operation, a simplified expression for TM mode can be achieved as,

\[
x = \frac{n\pi}{2} + \frac{n+1}{mx} - i \ln \left( \frac{m+1}{m-1} \right) \quad (8)
\]

Considering the real parts of equation (8) and further simplification will yield a linear expression for radiation resonance condition of TM mode in dielectric microspheres.

\[
x = \frac{\pi}{2m} \left[ n + \frac{4(n+1)}{n\pi^2} \right] \quad (9)
\]

where, \( n \) is an integer. The second term of equation (9), \( \frac{4(n+1)}{n\pi^2} \), could be ignored as it is very smaller compare to first term, \( n \), that yields more simplified form of resonance condition for TM mode and comparable to resonance condition in dielectric micro-sphere for TE mode [13].

\[
x = \frac{\pi}{2m} n \quad (10)
\]

**VALIDATION OF THE THEORY**

According to the theoretical model presented in previous section, size parameter for each MDR peak wavelength will be the product of a constant \( \pi / 2m \) and an integer and for consecutive MDR peaks the integers will be consecutive numbers. The size parameter calculated with experimental MDR wavelength is very close to the size parameter calculated by equation (10). The error between experimental result and calculated results are in the negligible range (~10^{-3} %) which supports the validation of developed theory.

**DISCUSSION & CONCLUSION**

Electromagnetic solutions based on quantum physics from the literatures [1, 5-7] are compared to the new approach developed here, and it is found that the present approach is very accurate for large size parameters. Comparison between experimental results and calculated results based on developed theoretical approach yields negligible error which supports the validation of the presented approach. The present theoretical and experimental approaches can potentially be used in applications of Micro-Electro-Mechanical Systems (MEMS), proteomics, Surface Plasmon Resonance (SPR), biomedical and nano research.

**REFERENCES**

Fabrication and characterization of GaP two dimensional photonic crystals for terahertz-wave generation

K. Saito¹, T. Tanabe¹, Y. Oyama¹, K. Suto², T. Kimura², J. Nishizawa²

¹. Department of Materials Science, Tohoku University, Aoba-yama 6-6-11-1021 Sendai 980-8579, Japan, ². Semiconductor Research Institute, 519-1176 Aoba, Aramaki, Aoba-ku, Sendai 980-0845, Japan

phone: +81 795 7330, fax: +81 795 7329, email: a6td5303@stu.material.tohoku.ac.jp

Abstract- We have fabricated GaP crystal based two-dimensional photonic crystals (2D-PCs) for THz wave applications. We demonstrated THz wave generation from the fabricated 2D-PCs. The enhancement of THz power was observed at around 1.1 THz.

I. Introduction

One of the methods for generating coherent terahertz (THz)-wave is based on an excitation of phonon-polariton in crystals [1-5] using the difference frequency mixing (DFM) between two infrared lasers. The THz wave can be achieved by the excitation of transverse optical phonon. GaP is attractive material for THz wave generation because of its transparency in both near-infrared and THz region, and high nonlinear optical coefficient. We have reported THz-wave generation with high output power (peak power of 1.5W) from bulk GaP crystals under non-collinear phase matching condition. We also have confirmed the enhancement of THz wave conversion efficiency in GaP waveguides with the size of terahertz wavelength due to the confinement of THz wave into the waveguide structure [6].

We focus on photonic crystals (PC) in THz region, which is one of the structures that enable to confine THz waves effectively. PCs are artificial periodic structures that allow for photonic band gaps (PBG) and device designing for many applications. In these structures, electromagnetic waves are confined via Bragg reflection and decreased their group velocity. These properties make it possible to enhance the nonlinear optical effect such as THz wave generation.

In this presentation, we have developed GaP PCs in THz region. We have also measured transmission characteristics and demonstrated THz wave generation in GaP PCs.

II. Experimental

The PC was fabricated using reactive ion etching in parallel plate discharge plasma. Figure 1 shows the fabricated GaP photonic crystal slab waveguide. Air holes were etched in a semi-insulating 200 µm thick GaP wafer to a depth of 75 µm. Air holes were arranged in a triangular lattice with a lattice constant a of 200 µm. The radius of air holes r was 80 µm. The sample was cut into rectangle with 10 mm wide in <110> direction x 5 mm long in <110> direction.

![Fig. 1. Structure of the fabricated GaP PC slab waveguide](image-url)

The schematic of the experimental setup used for the THz wave generation is shown in Fig. 2. The 1064-nm fundamental beam of a Q-switched Nd:YAG laser was used as the signal light for the DFM, and the 355 nm third harmonic beam was used to pump a β-BaB₂O₄ (BBO)-based optical parametric oscillator (OPO). The OPO was tuned in the 1059-1063 nm range. THz-wave generation was carried out under collinear phase matching condition. The generated THz-wave was detected by a liquid helium cooled Si bolometer. We used two types of waveguides: the PC patterned and non-patterned one, respectively.

III. Results

Figure 3 shows the THz power generated from the GaP slab waveguides. Dashed line is the THz power from non-patterned GaP substrate. THz wave generated in the frequency range from 0.3 to 1.3 THz. In the case of the GaP slab waveguide structure, THz waves generate under the modal phase matching condition. THz output characteristics depend on the propagation modes of THz wave in the waveguide. For the PC patterned GaP waveguide (solid line),
increment of THz power was appeared at around 1.1 THz.

Figure 4 shows the in-plane transmission characteristics of PC patterned waveguide. The propagation of THz wave was parallel to $\Gamma M$ direction. The polarization of the THz wave was TE polarization. To obtain the transmission spectrum, we measured that of the non-patterned GaP substrate as reference spectrum.

Several low transmission regions were appeared in this spectrum. One region from 0.4 to 0.6 THz is corresponding to the photonic band gap. Other region at around 0.9 THz and 1.2 THz are corresponding to the photonic bands in which the electromagnetic wave cannot excite due to the existence of their anti-symmetric modes. In particular, THz transmission increased at round 1.1 THz. The enhancement of THz wave generation is possible due to the higher transmission features at around 1.1 THz. Some photonic band structure with low group velocity exists in this frequency region. These photonic bands affect the THz conversion efficiency.

IV. Conclusion

In conclusion, we have fabricated GaP crystal based two-dimensional photonic crystals (PCs) for THz wave applications. We demonstrated THz wave generation from the fabricated GaP PC waveguide.

References

High bandwidth semiconductor gain material for photonic active devices with a stacked quantum dots structure grown by strain-compensation technique

Kouichi Akahane, Naokatsu Yamamoto, Shin-ichiro Gozu, Akio Ueta, Hideyuki Sotobayashi and Masahiro Tsuchiya

National Institute of Information and Communications Technology (NICT)
4-2-1 Nukui-Kitamachi, Koganei, Tokyo, 184-8795, Japan

Phone/Fax: +81-42-327-6804/+81-42-327-6941 E-mail: akahane@nict.go.jp

We have developed a new scheme to fabricate QDs in order to expand the potential bandwidths of QD active regions. The full-width at half maximum of photoluminescence of QDs is as high as 240 nm.

Quite a few attentions have been paid on photonic active devices with optical gain of high bandwidths. Representative in such technical areas are broadband semiconductor laser amplifiers (SOAs) [1,2] and wavelength tunable semiconductor lasers. In order to extend performances of those devices, it is indispensable to develop new semiconductor materials of higher bandwidths and more desired features.

Quantum dots (QD) structures based on III-V compound semiconductors are one of the most promising materials for this purpose [1]. In addition to its inherent broadband nature, incorporation of QD structures into active devices are expected to lead to their unique and attractive properties in device performances, which have been partially confirmed experimentally [3-5]. However, there are rooms to improve in the present methods to enlarge a variety in the QD sizes, which are directly related to spectral widths of photo-emission bands. This is because the variety is restricted by delicate balance between the surface energy and strain energy during epitaxial growths in the S-K mode.

To overcome this restriction and expand the potential bandwidths of QD active regions, we have developed a new scheme, in which a deposition condition of QD can be varied during a molecular beam epitaxy (MBE) growth. Similarly to the introduction of varied thicknesses in multiple quantum well structures, which Kuroda et al. have already applied to SOA devices [2], we have intentionally changed QD size distributions in a layer-by-layer manner through a delicate control of an amount of QD material supply and its resultant generation of strain [6]: one of the most important issues in the scheme is to control strains of QD layers and make those compensated while any degradation of crystal quality is prevented. In the experiments, the strain and composition of InGaAlAs spacer layers for InAs QDs have been accurately controlled so that their lattice constant is slightly smaller than that of the substrate. This condition prevents accumulation of strain, and consequently generation of defects or dislocations, during the stack of QDs. We have also modulated the strain given by spacer layers so as to vary the acceptable amount of InAs supply and, eventually, the QD size distribution.

MBE samples were prepared on InP(311)B substrates. The substrates were heated up to 500°C, and then a 150-nm-thick lattice-matched In_{0.52}Al_{0.48}As buffer layer was grown at a growth rate of 0.5 ML/s. Finally, h-ML InAs QDs and d-nm-thick In_{1-x}Ga_{x}Al_{1}As spacer layers were grown alternatively, producing a stacked structure. The QD and spacer layers satisfy the strain compensation conditions so that the total strain.
energy of a QD layer/spacer layer pair is zero, which prevented degradation of QD quality [6]. The five QD layer/spacer layer pairs (h-ML/d-nm), starting from the buffer layer, were thus prepared and their thicknesses are 4.5/45, 4.0/40, 3.5/35, 3.0/30 and 2.5/25 ML/nm, respectively.

Figure 1 shows an atomic force microscopy (AFM) surface image of the stacked InAs QDs (topmost: 2.5 ML). Self-assembled QDs were formed even though the deposition thickness was less than that in the previous work [6], which indicates the modulated strains of the spacer layers. The average lateral size in the [01-1] direction and the height of the QDs were estimated to be 58.4 nm and 3.41 nm, respectively. The density of the QDs in the topmost layer was 3.4 x 10^{10}/cm^{2}.

The emission spectrum of this particular sample was characterized by a standard photoluminescence (PL) measurement system. As shown in Fig. 2, intense emission from the ground state around 1.55 μm appeared at room temperature. We confirmed by using a sample grown separately that no emission occurred from the excited QD states. The full-width at half maximum is as high as 240 nm, which is about 2.5 times higher than that for the sample used for broadband SOA by T. Akiyama et al [5], as long as the line-widths of luminescence from the ground states are concerned. This result indicates that our new QD growth method is a promising way to expand the gain bandwidth. Intense photoluminescence at room temperature could indicate suppression of defects or dislocations generation. In the next step, feasibility of high bandwidth devices with this QD material employed should be examined.

References

Fig. 1 AFM image of modulated stack of InAs QDs.

Fig. 2 PL spectrum of modulated stack of InAs QDs.
Blue-Light Emission from Si Ion Implanted Fused-Silica Substrates Generated by High Temperature Annealing

Kenta Miura, Takeshi Tanemura, and Osamu Hanaizumi
Department of Electronic Engineering, Graduate School of Engineering, Gunma University,
1-5-1 Tenjin-cho, Kiryu 376-8515, Japan
Phone&FAX: +81-277-30-1793, E-mail: miura@el.gunma-u.ac.jp
Shun-ya Yamamoto, Katsuyoshi Takano, Masaki Sugimoto, and Masahito Yoshikawa
Takasaki Advanced Radiation Research Institute, Japan Atomic Energy Agency,
1233 Watanuki-machi, Takasaki 370-1292, Japan

Abstract: We observed blue-light emission from fused-silica substrates containing Si nanocrystals formed by using Si ion implantation and annealing. Photoluminescence peak wavelengths are around 400nm, and the peak intensities can be remarkable after annealing above 1150°C.

1. Introduction

Various works on silicon (Si) based luminescent materials utilizing the quantum confinement effect [1], such as Si nanocrystal (Si-nc) [2-4], have been reported. Typical fabrication methods of Si-ncs are co-sputtering of Si and SiO\textsubscript{2} [2], laser ablation [3], and Si ion implantation into SiO\textsubscript{2} plates or thermal oxidized Si wafers [4]. Blue- and white-light emission were observed from Si:SiO\textsubscript{2} co-sputtered thin films without annealing [5,6].

Pavesi et al. reported that their samples including Si-ncs formed by using Si ion implantation exhibited photoluminescence (PL) and a large optical gain of 100cm\textsuperscript{-1} in a wavelength range from red to infrared after annealing at 1100°C [4]. The optical gain is nearly equal to that of III-V semiconductors. Therefore, it has recently been anticipated that light emission devices composed of Si-nc-containing materials could be realized.

In this paper, we fabricated Si-nc including fused-silica substrates by using the Si ion implantation method and annealing, and evaluated their PL properties. As a result, we found a blue-light emission band in addition to the longer wavelength band reported by Pavesi et al. We observed PL peaks around a wavelength of 400nm from our samples after annealing and succeeded in increasing the blue PL peak intensity to over four times higher than the longer wavelength peak by annealing at 1200°C.

2. Fabrication and Evaluation

Si ions were implanted into a fused-silica substrate (10mm x 10mm x 1mm\textsuperscript{3}) at room temperature in the Takasaki Ion Accelerators for Advanced Radiation Application (TIARA) of Japan Atomic Energy Agency. The implantation energy was 80keV, and the implantation amount was 1 x 10\textsuperscript{17} ions/cm\textsuperscript{2} [4]. The Si implanted substrate was cut into four pieces (0.5mm x 0.5mm x 1mm\textsuperscript{3}) using a diamond-wire saw and they were annealed in air for 25min at 1100, 1150, 1200, and 1250°C by using a siliconit furnace.

![Annealing Temperature vs. Wavelength Plot](image-url)

**Fig. 1** Measured photoluminescence spectra.

PL spectra were measured at room temperature with excitation using a He-Cd laser (Kimmon, IK3251R-F, \(\lambda=325nm\)). A monochromator (Nikon, P250), a photomultiplier (Hamamatsu, R2658), and a lock-in amplifier (NF, LI-572B) were used in our measurements.
Figure 1 presents the PL spectra of the four samples generated at room temperature. Blue PL spectra having peaks around a wavelength of 400nm were observed from all the samples. The peak wavelengths of the blue PL spectra of the samples are almost the same in spite of the varying annealing temperature. In our experiments, the blue PL peak had a maximum intensity after annealing at 1200°C, and the intensity was 4.2 times higher than that of the peak of the longer wavelength band of the sample annealed at 1100°C.

Liao et al. reported that a blue PL having a peak wavelength of 470nm was observed from Si ion implanted SiO$_2$ films and stated that the blue-light emission may have originated from an oxygen-defect level in SiO$_2$ [7]. However, the emission origin seems to differ from that of our samples owing to the difference of the PL peak wavelengths.

In addition to a blue PL peak, a PL peak around a wavelength of 800nm in a wavelength range from red to infrared was observed for the sample annealed at 1100°C. In our experiments, the Si ion implant conditions were almost the same as those reported in ref. [4] and the annealing temperature of 1100°C is common to ref. [4], therefore it seems that Si-ncs are also formed in this sample. There may be a common origin of blue-light emission in these samples and Si:SiO$_2$ co-sputtered films [5]. The SiO$_x$ layer located at the interfacial region of Si nanoclusters and surrounding SiO$_2$ medium may contribute to such emission.

3. Summary

We fabricated fused-silica substrates that emit blue light by using Si ion implantation and high-temperature annealing. Their blue PL peaks were located at around a wavelength of 400nm. We found that the peak intensity can be over four times greater than that of the longer wavelength peak reported by Pavesi et al. [4] after annealing at 1200°C. The blue-light emission seems to originate from both Si-ncs and interface layers between Si-ncs and SiO$_2$ media produced in fused-silica substrates. Blue light emission materials are expected to be useful as light sources for optical pick-up systems. Therefore, we are trying to optimize the conditions of Si ion implantation and annealing to improve the emission intensity and evaluate optical gains of the Si implanted samples.

Acknowledgements

This work was supported by a Joint Research Program, “Development of silicon-based functional materials.” Part of this work was supported by a Grant-in-Aid for Scientific Research (B) of the Japan Society for the Promotion of Science; a Grant-in-Aid for Scientific Research on Priority Areas and a Grant-in-Aid for Young Scientists (B) of the Ministry of Education, Culture, Sports, Science and Technology; the CASIO Science Promotion Foundation; the Yazaki Memorial Foundation for Science and Technology; the Iwatani Naoji Foundation; and the Murata Science Foundation.

References


Eu\textsuperscript{3+} doped polymer waveguide amplifiers

K.C. Tsang and E.Y.B. Pun

Department of Electronic Engineering, City University of Hong Kong
Tat Chee Avenue, Kowloon, Hong Kong
Phone: (852)2788 8609, Fax: (852)2788 7791, E-Mail: eeeybpun@cityu.edu.hk

Abstract

Eu\textsuperscript{3+} doped SU8 polymer materials have been characterized optically. The doped polymers are suitable for planar waveguide amplifiers operating in the visible wavelength, and channel waveguides have been fabricated using standard photolithographic process.

1. Introduction

Er\textsuperscript{3+}-doped planar glass waveguide amplifier operating at 1550nm wavelength was first demonstrated in 1992 [1], and recently Er\textsuperscript{3+}-doped polymer amplifier has also been developed as a suitable alternative [2]. There is an increased interest in exploiting polymer optical fiber low-loss visible wavelength window [3], and in optical amplifiers that operates in the visible spectrum [4,5]. Organic dyes can be used as amplifier’s dopants [6]; however, they may suffer from photobleaching, unstable emission problems, and triplet losses. In contrast, lanthanide dopants show steady fluorescence and longer emission problems, and triplet losses. In contrast, lanthanide dopants show steady fluorescence and longer emission lifetime.

In this paper, EuTTFA complexes with TTFA as asymmetric ligands encapsulating Eu\textsuperscript{3+} ions have been demonstrated as suitable dopants in SU8 polymer. The emission wavelength resulting from the \( \text{^5D}_0 \) to \( \text{^5F}_2 \) transition is close to the low loss window of polymer fibers, and optical channel waveguides have been fabricated using a one-step UV exposure process.

2. Experiment

Eu\textsuperscript{3+} doped polymers were prepared by adding EuTTFA organic complexes into SU8 polymer directly, and the complexes dissolve readily at room temperature. Polymers doped with 2 to 4 wt% Eu\textsuperscript{3+} complexes were prepared and spin-coated on quartz substrates and their material absorption and photoluminescence properties were studied. The UV-VIS absorption measurements were carried out using a Perkin Elmer Instruments UV-visible-near-IR \( \lambda_{10} \) spectro-photometer, and the photoluminescence (PL) spectra were recorded using an Acton Research Corporation Spectra Pro 500i.

Eu\textsuperscript{3+} doped polymer waveguide amplifiers 2\,\mu m high and 50\,\mu m wide were fabricated using standard UV photolithographic process, and thermally oxidized silicon wafers with \( \sim 2.5\mu m \) thick oxide layer were used as substrates. To measure the amplifier performances, the waveguide was top pumped by a pulsed 337nm UV laser, and the pump light was focused into a 1cm long line. The signal light at 612nm wavelength, after passing through a monochromator, was detected by a photodetector and measured using an oscilloscope.

3. Results and discussion

The absorption spectra of the undoped and Eu\textsuperscript{3+} doped SU8 samples are shown in Fig. 1. The absorption peaks of different Eu\textsuperscript{3+} doped SU8 concentrations are near the 350nm wavelength, and are due to the ground state to excited states absorption of the organic ligand (TTFA). The excitation of singlet state \( S_1 \) from the ground singlet state \( S_0 \) occurs in the TTFA ligand after absorption of energy. The TTFA ligands sensitize the absorption of 350nm UV light, and the singlet state \( S_1 \) undergoes non-radiative intersystem crossing process to the triplet states \( T_1 \).

![Fig. 1 Absorption spectra of undoped and Eu\textsuperscript{3+} doped SU8 films](image)

The excitation of the ligand band of energy \( T_1 \) couples with the \( ^3D \) energy levels. The \( ^3D \) upper level populates the lower levels non-radiatively to the lowest level \( ^1D_0 \) and the energy is transferred by emission from the \( ^1D_0 \) level to the \( ^7F_\ell \) levels of the Eu\textsuperscript{3+} ions where \( \ell=0-6 \). The PL spectra of undoped and Eu\textsuperscript{3+} doped SU8 films are shown in Fig. 2. There are seven emission peaks at 580, 593, 612, 652, 701, 750 and 806 nm, corresponding to transitions from \( ^1D_0 \) to the \( ^7F_\ell \) (\( \ell=0, 1, 2, 3, 4 \), and 6 respectively). The most intense emission line appears at 612 nm, and is due to the \( ^1D_0 \) to \( ^7F_2 \) transition.
This strong emission from $^5\text{D}_0$ to $^7\text{F}_2$ of Eu$^{3+}$ ion is a result of the highly polarized surrounding ligand fields.

The spectroscopic properties of Eu$^{3+}$ ions can be described quantitatively by Judd-Ofelt theory [7]. The intensity ratios between the $^5\text{D}_0$ to $^7\text{F}_{2,4,6}$ and the $^5\text{D}_0$ to $^7\text{F}_1$ transitions can be used to determine the parameters $\Omega_2$, $\Omega_4$, and $\Omega_6$. The radiative transition probability is proportional to the Judd-Ofelt parameters, and the intensity ratios between the transitions can be used to determine the parameters $\Omega_2$, $\Omega_4$, and $\Omega_6$. The spectra of samples A and B overlap with each other, and the fluorescence peak at 612nm saturates when the Eu$^{3+}$ concentration is greater than 3wt%, as shown in Fig. 2. This is the minimum concentration that provides high emission intensity when the sample is excited by UV laser at 337nm with 300µJ pulsed energy. The Judd-Ofelt parameters of the 3 wt% Eu$^{3+}$ doped SU8 films calculated from the fluorescence spectrum are $\Omega_2=11.4x10^{-20}\text{cm}^2$, $\Omega_4=2.9x10^{-20}\text{cm}^2$, and $\Omega_6=1.8x10^{-20}\text{cm}^2$. The $\Omega_2$ value is close to that reported in [5], which is $\sim18x10^{-20}\text{cm}^2$.

The measured lifetime $\tau_{\text{meas}}$ is estimated to be $\sim0.44$ms, as shown in Fig. 3. The emission cross section is calculated to be $\sim2.4x10^{-21}\text{cm}^2$, and is similar to that of erbium in silica optical fiber ($5x10^{-21}\text{cm}^2$ one). The radiative lifetime $\tau_{\text{rad}}$ of the $^5\text{D}_0$ level is calculated to be 1.85ms; hence the luminescence quantum efficiency $\eta$, defined as $\eta_{\text{meas}}/\eta_{\text{rad}}$, is $\sim24\%$.

The channel waveguide amplifiers doped with 3 wt% Eu$^{3+}$ complexes exhibit a 612nm wavelength signal response after being top pumped by focusing 1cm long UV pulsed laser. Although the absorption spectra of the Eu$^{3+}$ doped samples have an absorption peak at 350nm wavelength; due to limitation a pulsed 337nm UV laser pump source is used instead. It can be expected that more efficient excitation and higher gain can be obtained using a 350nm pump source. The gain measurement is currently in progress.

4. Conclusion
The material and the photoluminescence properties of Eu$^{3+}$ doped SU8 polymer have been characterized. The emission from the $^5\text{D}_0$ to $^7\text{F}_2$ transition is at 612nm wavelength, and the Judd-Ofelt parameters are $\Omega_2=11.4x10^{-20}\text{cm}^2$, $\Omega_4=2.9x10^{-20}\text{cm}^2$, and $\Omega_6=1.8x10^{-20}\text{cm}^2$. The measured lifetime is $\sim0.44$ms, and the radiative lifetime of the $^5\text{D}_0$ level is $\sim1.85$ms. The luminescence quantum efficiency is $\sim24\%$, and the emission cross section is $\sim2.4x10^{-21}\text{cm}^2$ which is similar to the value reported for erbium in silica optical fiber. Eu$^{3+}$ doped polymer channel waveguides have been fabricated using one step UV photolithographic process. The results indicate that the Eu$^{3+}$ doped SU8 polymer material is suitable for optical amplifiers and laser devices.

Acknowledgments
This work was supported by a grant from the Research Grants Council of the Hong Kong Special Administration Region, China [Project No. CityU 1102/06].

References
Optoelectronic Switch with S-Shaped Negative Differential Resistance

Jung-Hui Tsai¹, Der-Feng Guo², Ming-Yue Fu²
¹ Department of Electronic Engineering
National Kaohsiung Normal University, 116 Ho-ping 1st Road, Kaohsiung 802, TAIWAN
TEL: 886-7-7172930 Ext.7216; Email: jhtsai@nknucc.nknu.edu.tw
² Department of Electronic Engineering
Air Force Academy, P.O. Box 14-49 Kangshan, Kaohsiung County 820, TAIWAN
TEL: 886-7-6264185; FAX: 886-7-6264185; Email: g6f6guo@ms5.hinet.net

A great deal of research interest has been arisen in pnpn double-heterostructure optoelectronic switches. The switching characteristics in electrically or optically controlled npn bulk-barrier structures have also been reported. The npn bulk-barrier structures have great potentials as optoelectronic switches because of their high gains and high speed operations. In this report, an AlGaAs/GaAs/InAlGaP pn-p bulk-barrier optoelectronic switch (BBOS) with an AlGaAs/δ(n⁺)/GaAs/InAlGaP collector structure is proposed. The application of the collector structure causes an increasing switching voltage Vₘ and a decreasing switching current Iₛ, with illumination, which are different from the switching characteristics of the reported npn double-heterostructure or npn bulk-barrier optoelectronic devices. In the BBOS structure, a lightly-doped thin GaAs base layer is employed to create a bulk-barrier profile in the energy band diagram. This base layer is depleted of free charges and minimizes the problems of charge storage. After the onset of avalanche multiplication in the reverse-biased collector region, an S-shaped negative differential-resistance (NDR) phenomenon is observed in the current-voltage (I-V) characteristics. Because the GaAs well in the collector structure confines photogenerated electrons, the switching of the BBOS is influenced by the presence of incident light. Moreover, the BBOS presents obvious NDR characteristics at high environment temperature up to 160 °C.

The device structure consists of a 3000Å n⁻-GaAs (3×10¹⁷ cm⁻²) buffer layer, a 3000Å n⁻-ln₃(Al₈₃Ga₁₇)₅ P (5×10¹⁷ cm⁻²) layer, a 100Å i-GaAs layer, a δ(n⁺) (1×10¹⁷ cm⁻²) sheet, a 100Å n⁻-Al₈₃Ga₁₇As (5×10¹⁷ cm⁻²) layer, a 600Å p-GaAs (5×10¹⁷ cm⁻²) base layer, a 20Å i-GaAs layer, a 1000Å n-Al₈₃Ga₁₇As (5×10¹⁷ cm⁻²) emitter layer and a 3000Å n⁺-GaAs (5×10¹⁷ cm⁻²) cap layer.

Figure 1 shows the band diagram of the BBOS with incident photons under a positive collector-emitter (C-E) voltage Vₑ. Figures 2 shows the I-V characteristics of the BBOS under dark (solid lines) and illumination (dotted lines) conditions at 27°C. S-shaped NDR performances are observed under both dark and illumination conditions. Compared with the dark S-shaped NDR characteristics, the illuminated S-shaped NDR characteristics present a higher switching voltage Vₛ and a lower switching current Iₛ. These characteristics variation with illumination is quite different from other reported results. This is attributed to that the photogenerated electrons confined in well B elevate the potential spike at the collector region under illumination condition, as shown in Fig.1. In addition to the switching characteristics controllable by the input-light, optical switching can be observed by biasing the device to just below turn-off and then introducing light to turn the device off, or by biasing it to just before turn-on and removing the light source to turn the device on. This optical switching of the proposed BBOS is also different from that of other optoelectronic switches. Figure 3 shows the dark I-V characteristics at 160°C. Compared with the dark characteristics at 27°C, as shown by solid lines in Fig.2, Fig.3 also presents obvious NDR characteristics.

This work was supported by the National Science Council of the Republic of China under Contracts NSC 95-2221-E-013-004 and NSC 95-2221-E-017-013
Figure 1. Band diagrams of the BBOS under a positive bias voltage $V_{ce}$ with incident photons.

Figure 2. I-V characteristics of the BBOS under dark (solid lines) and illumination (dotted lines) conditions at 27°C.

Figure 3. Dark I-V characteristics of the BBOS at 160°C.
Synthesis and Discussion of 3-hexylthiophene related rigid copolymers for Optical Devices

Pei-Chen Huang, Yi-Jang Lee, May-Ying Chang, Wen-Yao Huang, Yu-Kai Han

a Department of Chemical and Material Engineering, National University of Applied Science, Kaohsiung, Taiwan R.O.C.
b Institute of Electro-Optical Engineering, National Sun Yat-Sen University, Kaohsiung, Taiwan R.O.C.

Tel: 886-7-3814526 x 5107, Fax: 886-7-3830674 E-mail: ykhan@cc.kuas.edu.tw

Abstract
Two conjugated copolymers, poly{(9,9-dihexylfluorene)-random-(3-hexylthiophene)} (PTFR) and poly{(9,9-dihexylfluorene)-alt-(3-hexylthiophene)} (PFTA) were obtained by using Kumada coupling based on 3-hexylthiophene moiety unit. The fluorescence (PL) and ultraviolet (UV) optical properties of these polymers were compared with poly(9,9-dihexylfluorene) and poly(3-hexylthiophene), and the fluorescence light emitting range cover from 400 to 550 nm which can be used for polymer white light emitting device and organic solar cell.

Keywords: conjugated polymer, organic solar cells, Kumada coupling.

1. Introduction
Since the first report of polymer light-emitting diodes (PLEDs) in 1990, the development of conjugated polymers are huge. Especially, polythiophene and its derivates (PTs), which are special conductive and optoelectronic materials[1-3]. Polyfluorenes are attractive class of polymer semiconducting material for electroluminescent, which synthesis easy and color tunable. To combine these two materials via nickel-catalyzed Kumada cross-coupling method, a random and a alternative copolymers base on 3-hexylthiophene and 9,9-dihexyl-2,7-dibromofluorene repeating units were synthesized.

2. Experimental details
The detail synthesis steps and chemical structures of the copolymers are shown in Fig. 1 and Fig. 2. All monomer have purified by crystallization, reduce pressure distillation, or reaction under dry nitrogen atmospheres. Polymers were purified via flash column chromatography to remove nickel catalysis.

3. Results and Discussion
Figure 3 shows copolymers PTFR and PFTA structure identify by using proton nuclear magnetic resonance spectra (NMR), which characters of poly(3-alkylthiophene) are around 2.8 ppm and 6.9 ppm[5]. If these two peaks are broaden with multiple peaks, that means conformation of 3-alkylthiophene is in random type (Fig. 3a.). If the peaks are indented, the polymer structure is in regular conformation. Fig.3b shows two peaks at the position of 6.9 ppm and 7.05 ppm respectively were resulted from different regular arrangement, head to tail form (H-T) and tail to tail form(T-T).

Figure 4 and Figure 5 are UV-Vis absorption spectra and fluorescence spectra of PFTA. Figure 4 indicates PTFA and PTFR have similar energy gap from UV-Vis absorbance. The energy gape was approximately between 2.7 ~ 2.8 eV in solid film. PTFA and PTFR both have similar maximum absorption around 375 nm in UV-Vis absorbance in solution and solid film. The width of these absorption peaks across ~ 100 nm shown that these two materials have stable energy gap.

In figure 5, PTFA PL spectra had red shift about ~60nm from solution to solid film. PTFR only slightly shifted about 10 nm - 20 nm, which indicates that PTFA have regular conformation and results in better structural aggregation in solid film as than PTFR.

4. Conclusions
The 3-hexylthiophene related rigid copolymers were obtained. The PL and UV spectra show that they cover wide range of visible light area and can be used in white light emitting diode and organic solar cell. The PFTA has regular structure which is easier resulting in aggregation form while comparing to PTFR random form.

References
Fig. 1. Synthesis of PTFR

Fig. 2. Synthesis of PFTA

Fig. 3. $^1$H NMR spectra of (a)PTFA and (b) PTFR

Table 1. Number average ($M_n$) and weight average ($M_w$) of PTFR and PTFA. Determine by GPC, RI detector.

<table>
<thead>
<tr>
<th>Polymer</th>
<th>$M_n$</th>
<th>$M_w$</th>
<th>$M_w/M_n$</th>
</tr>
</thead>
<tbody>
<tr>
<td>PTFR</td>
<td>2900</td>
<td>9200</td>
<td>3.17</td>
</tr>
<tr>
<td>PTFA</td>
<td>2600</td>
<td>7000</td>
<td>2.69</td>
</tr>
</tbody>
</table>
Apodization Structure for Chirped Quasi-Phase-Matched Wavelength Converter

H. Okayama, H. Ono, Y. Okabe and Y. Ogawa
R&D Group, Oki Electric Ind. Co., Ltd. and with
550-1 Higashiassakawa, Hachioji, Tokyo 193-8550, Japan
Phone: +81-426-62-6767 Fax: +81-426-62-6770
E-mail: okayama575@oki.com

Abstract
A quasi-phase-matching (QPM) wavelength converter with its coupling strength changed along the propagation length is proposed to improve the wavelength response. Sampling QPM structure can be used to attain a device with fabrication easiness.

1. Introduction
The quasi-phase-matching (QPM) structure which is implemented by periodically poling the domain of the ferroelectric LiNbO$_3$ crystal (PPLN) has been gaining much attention [1]. The device can be used in many applications including optical fiber communication, optical signal processing and unconventional light sources. Using sinusoidal sign reversed nonlinear coefficient, high wavelength conversion efficiency is achieved by compensating for the phase mismatch between the fundamental and wavelength converted light.

The spectral band width of the uniform period QPM tends to be very narrow. A chirped period QPM structure [1] (Fig. 1) has been used to widen the spectral bandwidth. However, the huge ripple of the conversion efficiency poses many problems such as distortion of the converted light pulse. The ripple can be eliminated by gradually increasing the coupling efficiency from the edge toward the center of the device, which is called the apodization structure [2]. The apodization has been achieved by changing the duty ration of the QPM structure [3]. However, it is very difficult to control the QPM width during the domain inversion process.

In this report we describe a design and new apodization scheme leading to an improved performance and easy fabrication.

2. Window function
For realizing the apodization, one of the most basic window functions used is the raised-cosine type slope placed at both ends of the QPM region. However, an ear tends to appear at the edge of the wavelength conversion band in the wavelength response [2]. In a wavelength filter device such as a fiber Bragg grating, the diffraction or coupling efficiency is so high (~100%) that the spectral shape in the wavelength filter pass band seldom becomes a problem. However, for the QPM device the low efficiency puts the problem forward. Inspired by the results of wavelet synthesis, we have found that to overcome the problem a slight coupling strength slope is required also at the middle of the QPM region. One of the window functions known to have this property is the tanh($z$) type [4]. However, in this design the parameter values those generate useful shape are limited. Instead...
we tested a window function given by Eq. 1. The function is the extended version of the raised-cosine type where 2L is the QPM region length and z is the propagation distance.

\[
F(z/L) = 1 - C_1 |z/L|^M \quad 0 < |z/L| \leq r \quad (1a)
F(z/L) = C_2 + C_3 \cos[\pi f(z)] \quad r < |z/L| \leq 1 \quad (1b)
F(z/L) = (|z/L|-r+C_4)/(1-r+C_4) \quad (1c)
\]

The continuous conditions of the value, slope and curvature at |z/L|=r and F(1)=0 makes only two parameters r and C_4 independent similar to a tanh(z) type. The structure becomes conventional raised-cosine window function when C_4=0. An example of the calculated wavelength response is shown in Fig. 2 with N=3000 reversed domains, r=0.8 and C_4=0.105. The Δk is the wavenumber detuning and Λ is the QPM period. The ripple in the conversion wavelength band is eliminated for optimized r and C_4 values.

3. Apodization by QPM sampling

It has been known that eliminating some gratings in the Bragg grating can alter the diffraction strength [5]. The same principle can be adapted to the QPM structure. When there are N polarization reversed domains in the QPM structure, we divide the structure into N^{1/2} sections. In one section some reversed domains are omitted according to the required coupling strength at the section. The reversed domains can be eliminated from one side in each section as shown in Fig. 3. Thus the coupling strength is divided into N^{1/2} levels and the window function is approximated by a multiple-step function. The response shown in Fig. 2 is for the sampled structure.

A random sampling of the reversed domain can also be used. In this case, we don’t need to divide the device into sections. However, the wavelength response of the device tends to be somewhat noisy compared to the device structure using the sectioning scheme.

The sampling of the reversed domain generates ghost peaks in the wavelength response due to sampling noise. The distance between the main conversion efficiency peak and unwanted ghost peak generated by sampling becomes wider for shorter section length.

A preliminary experimental wavelength response obtained from the fabricated 4cm-long PPLN ridge waveguide device is shown in Fig. 4. In the chirped QPM device without apodization, the power of the second harmonic light fluctuates largely producing many peaks, whereas we obtained much uniform second harmonic light power using sampled QPM structure.

4. Conclusion

A quasi-phase-matching (QPM) wavelength converter with its coupling strength changed along the propagation length (apodization) has been proposed to improve the wavelength response. Slight coupling strength slope at the middle of the QPM region together with the apodization at the QPM region edge generates a box like wavelength response with minimum ripple. Sampled QPM structure can be used to realize the apodization of the coupling strength. This structure is fabrication friendly since it does not require precise control of the fabrication conditions.

References
3. T. Umeki et al., Extended abstracts of 2006 spring JSAP meeting, paper 30p-ZX-12.

Fig. 3 QPM sampling for apodization

Fig. 4 Measured wavelength response
Theoretical Investigation of Light-Guiding Structures of Surface Plasmon Resonance Waveguide Sensors

Jun Shibayama, Shota Takagi, Tomohide Yamazaki, Junji Yamauchi, and Hisamatsu Nakano
Faculty of Engineering, Hosei University
3-7-2 Kajino-cho Koganei Tokyo 184-8584 Japan

tel: +81-42-387-6233, fax: +81-42-387-6381, e-mail: shiba@k.hosei.ac.jp

Abstract: Two waveguide structures are numerically investigated for improved performance of the SPR waveguide sensors: one is the waveguide supporting higher-order modes and the other is the waveguide with an embedded core.

1 Introduction

Surface plasmon resonance (SPR) waveguide sensors have been studied for the integration into optical circuits. Note that theoretical studies of the SPR waveguide sensors were limited to two-dimensional (2-D) models [1]-[4]. Very recently, practical three-dimensional (3-D) models of the SPR waveguide sensors have been analyzed, in which the effect of the metal width on the sensing characteristics has been revealed [5].

The purpose of this article is to investigate two waveguide structures for the SPR sensor with strong absorption using the 3-D semi-vectorial beam-propagation method (BPM): one is the waveguide supporting higher-order modes in the horizontal direction and the other is the waveguide with an embedded core. It is shown that these two waveguide structures are effective in increasing the absorption at the resonance wavelength, leading to improved measurement precision.

2 Discussion

Fig. 1 illustrates the configuration of the SPR waveguide sensor to be studied. The absorbing layer is inserted between the metal and analyte layers for the sensor operation around \( \lambda = 0.6 \) \( \mu \)m, where these layers are centered on the waveguide \( ( t_m = 0.045 \mu m \) and \( t_{ad} = 0.02 \mu m ) \). The sensing length with the metal is 200 \( \mu m \). The refractive indices are chosen to be \( n_{co} = n_{ad} = 1.47 \) and \( n_{sub} = 1.46 \). The metal is selected to be Au \( ( n_m = 0.131 - j3.654 \) at 0.6328 \( \mu m \) and the dispersion property of its refractive index is taken into account using the Drude model dielectric function [5]. We excite the input waveguide with the field of the TM mode. Varying the operating wavelength, we evaluate the output power from the waveguide, which depends on the refractive index of an analyte \( n_a \). Water is used as the analyte, which is sufficiently thick (1 \( \mu m \) to yield a converged solution.

First, we investigate the dependence of the core width \( d_w \) on the sensing characteristics, which cannot be treated in the 2-D model. The metal strip is wide enough to cover the core region. The core thickness is deliberately chosen to be \( d = 2 \mu m \) throughout this article so that the single-mode operation in the \( y \) direction may be maintained. This is based on the fact that the thick core supporting higher-order modes in the \( y \) direction gives rise to a broadening of the absorption dip in the wavelength response, when compared with the single-mode case [6].

Fig. 2 shows the output power of the multimode case \( ( d_w = 12 \mu m \) as a function of wavelength for \( t_b = 0 \) (semi-embedded core) and \( n_a = 1.332 \). For reference, the results of the single-mode case \( ( d_w = 2 \mu m \) [5]) and the 2-D case \( ( d_w = w = \infty ) \) are also included. The inset shows the enlarged view around the absorption peaks. It is seen that the output power for the multimode case is reduced when compared with the single-mode case. This is because the wide metal strip increases the absorption of light. It is interesting to note that the dips in the wavelength response are not broadened even for the higher-order modes. The slightly less absorption of the higher-order modes compared with the fundamental mode stems from the extension of the higher-order mode field. In Fig. 2, an appreciable difference can be found between the 3-D and 2-D results. Therefore, the 3-D analysis is indispensable for the accurate modeling of the practical model of the waveguide sensor.

Although the use of the multimode waveguide with the wide core has been shown effective for the strong absorption, the amount of the analyte increases in order for the analyte to cover the wide metal region. To achieve strong absorption maintaining a single-mode waveguide with a narrow metal region [5], we next investigate the characteristics of the waveguide sensor with an embedded core. Note that the effect of the embedded core is similar to that of a buffer layer placed between the core and the metal layer [2]-[4]. Theoretical investigations on this subject were, however, restricted to 2-D models. Although the em-
bedded core was used for the ion-exchanged graded-index waveguide in [7], its effect was not explicitly presented. Here, we treat the 3-D structure with the step-index single-mode waveguide for \( d_w = 2 \mu m \) and \( w = 3 \mu m \). The effect of the embedded core is illustrated in Fig. 3, in which the inset shows the enlarged results around the absorption peaks. It can be seen that the output power for \( t_b = 0.3 \mu m \) is reduced by more than 4 dB, compared with the semi-embedded case \((t_b = 0)\). This reduction of the power results from the strong coupling between the waveguide and surface plasmon polariton modes.

3 Conclusions

We have analyzed the practical 3-D structures of the SPR waveguide sensors using the BPM. It is shown that the waveguide supporting higher-order modes in the transverse direction and the waveguide with the embedded core are effective in reducing the output power at the resonance wavelength. For each structure, more than a 4 dB reduction can be achieved. The reduction leads to a sharp dip in the wavelength response, which is expected to contribute to the improved measurement precision.

Acknowledgment

The authors would like to thank Dr. M. Fujimura of Osaka University for sending the reference [7] and giving a chance to investigate the effect of the embedded core of the SPR waveguide sensor.

References

1-input 3-output Optical Interleave Filter with Group-Delay Dispersion Equalizer

Shafiul AZAM, Takashi YASUI, and Kaname JINGUJI

Signal Processing laboratory, Interdisciplinary Faculty of Science and Engineering, Shimane University, Japan

Abstract: The authors report, a 1-input 3-output optical interleave filter with group delay dispersion equalizer. Synthesis algorithm is derived to obtain all unknown circuit parameters. Finally, it is confirmed that developed circuit satisfies all desired filter characteristics.

1. Introduction:
In optical communication, both interleave filter and group-delay dispersion equalizer play an important role in wide division multiplexing (WDM) signal processing. Several techniques have been proposed to implement interleave filter. Recently, 1×3 flat pass-band lattice-form interleave filter is presented using 3×3 triangle structure directional coupler [1]. Group-delay dispersion equalizer is an optical frequency filter designed to equalize group-delay dispersion of long-distance optical fibers at a wavelength of 1.55 μm. Many ideas have been reported to compensate these basic circuit blocks. Degree of freedom confirms that external phase shifter in this circuit is justified.

2. Circuit configuration:
The 1×3 optical interleave filter consists of N stages optical delay-lines, 2×(N+1) tunable directional couplers, 2×(N+1) phase shifters and an external phase shifter (ϕc). This circuit consists of several basic blocks and each block has five elements shown in Fig. 1. Total filter characteristic is expressed as the multiple products of all these basic circuit blocks. Degree of freedom confirms that external phase shifter in this circuit is justified.

Output transfer function of a three-port lattice-form optical interleave filter can be written as

\[ H(z) = \sum_{k=0}^{N} a_k z^{-k}, \]

\[ F(z) = \sum_{k=0}^{N} b_k z^{-k} \]

\[ G(z) = \sum_{k=0}^{N} c_k z^{-k} \]

respectively.

Where \(a_k, b_k, c_k\) are the complex expansion coefficients with \(k = 0 - N\).

Fig. 1 Circuit configuration of 1-input 3-output lattice-form optical interleave.

3. Synthesis Algorithm:
In this section, an algorithm is introduced for synthesizing three-port optical interleave filter with group delay dispersion equalizer. The whole synthesis process can be presented in three steps as below:

Step 1: Calculate the delay time difference \(Δτ\) from the desired periodic frequency \(f_0\).

\[ Δτ = \frac{1}{f_0}. \] (1)

Step 2: In this step unknown complex expansion coefficients \(a_k, b_k, c_k\) are obtained from desired transfer function \(H(z), F(z), G(z)\) using least square approximation method. Restriction conditions from Eq. (2) are applied on objective function to keep total transmittance always 100%.

\[ H(z)\tilde{H}(z) + F(z)\tilde{F}(z) + G(z)\tilde{G}(z) = 1. \] (2)

Step 3: The key task of this step is to calculate the transfer function of each block and finally is to drive a set of recursion equations to calculate all coupling coefficient angles (\(θ_{ka}, θ_{kb}\)) of the directional couplers and phase values (\(ϕ_{ka}, ϕ_{kb}\)) of the phase shifters.

If the input of this circuit is specified by \([1, 0, 0]^T\), then, the input-output relation can be written as,

\[ S(z) = \begin{bmatrix} H(z) \\ F(z) \\ G(z) \end{bmatrix} = \begin{bmatrix} 0 \\ \prod_{k=0}^{N} S_k(z) \end{bmatrix} \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix}. \] (3)

Applying factorization, \(N\)-th block can be separated from \(S(z)\) and hence \(ϕ_{nb}, θ_{nb}, ϕ_{Na}\) and \(θ_{Na}\) can be obtained. Consequently, all the circuit parameters can be found successively from the following recursion equations in the order: \(n = N, N-1, N-2, \ldots, 2, 1, 0\):

\[ ϕ_{nb} = \arg \left[ \begin{bmatrix} jc[n]_n \\ jθ[n]_n \end{bmatrix} \right], \quad θ_{nb} = \tan^{-1}(\frac{j c[n]_n e^{-jϕ_{nb}}}{b[n]_n}) \]

\[ ϕ_{na} = \arg \left[ \begin{bmatrix} j a[n]_n \cos θ_{nb} e^{jϕ_{nb}} - c[n]_n \sin θ_{nb} \\ a[n]_n \end{bmatrix} \right] \]

\[ θ_{na} = \tan^{-1}(\frac{j a[n]_n \cos θ_{nb} e^{jϕ_{nb}} - c[n]_n \sin θ_{nb}}{a[n]_n}) \] (4)
4. Design Example:
The design of a three-port optical interleaver is to provide three channels of passband transmission that are shifted by $2\pi/3$ in phase but with an identical passband shape. Then the phase shift requirement can be described as

$$2\sum_{k=0}^{N} a_k e^{-j2\pi 2k\omega} \approx 0,$$  \hspace{1cm} (5)

where, $D_a(z), D_f(z)$ and $D_g(z)$ are the approximate desired responses. In this paper, objective function of $H(z), F(z)$ and $G(z)$ are considered as follows,

$$\xi_H = \left| \frac{1}{6} + e \right| D_H(\omega) - \frac{1}{2} \sum_{k=0}^{N} a_k e^{-j2\pi 2k\omega} \left| \omega \right|$$ (a) Delay slope: 377 ps/nm

$$\xi_F = \left| \frac{1}{6} + e \right| D_F(\omega) - \frac{1}{2} \sum_{k=0}^{N} b_k e^{-j2\pi 2k\omega} \left| \omega \right|$$ (b) Delay slope: 0 ps/nm

$$\xi_G = \left| \frac{1}{6} + e \right| D_G(\omega) - \frac{1}{2} \sum_{k=0}^{N} c_k e^{-j2\pi 2k\omega} \left| \omega \right|$$ (c) Delay slope: -514 ps/nm

with $\omega = \frac{0.2 \pi}{2\pi}$ and the transition band ranges are set with $\epsilon = 0.2$. We can therefore, optimize the above objective function as below,

$$\text{Minimize } J = \xi_H - \xi_F - \xi_G$$ subject to restriction conditions of Eq. (2).

Fig. 2 Synthesized power frequency response of 1×3 optical interleaver filter.

In this design example, the number of expansion coefficients is set at 30. Delay time is set to 0.01 ns, which corresponds to the free spectral range of 100 GHz. Fig. 2 shows the synthesized power frequency response of an 1×3 optical interleaver. The transmittance at the passband is -0.0014 dB which is near to 0 dB and the transmittance at the stop band was less than -26 dB. The passband of the filter are nonoverlapping. Fig. 3(a-c) shows the synthesized group delay response in delay slope 377 ps/nm, 0 ps/nm and -514 ps/nm respectively. Although there is a slight discrepancy (6 ps/nm, 3 ps and 8 ps/nm approximately) from desired group delay, the result agrees well with the required filter characteristics.

5. Conclusion:
We present a circuit configuration and synthesis algorithm for three-port optical interleaver with group delay dispersion equalizer in this paper. Proposed synthesis algorithm has been tested by example and work properly. It is expected to be employed in FDM and WDM optical communications.

Reference:
Effect of Defect Pillars in T-Junction Photonic Crystal 1310/1550 nm Demultiplexer

Azliza J.M. Adnan1, Suhbudin Shaari2, Imran A. Tengku2, Nur Hanani Zainal Abedin2
1TM Research & Development (M) Sdn. Bhd.
2Institute of Micro-Engineering and Nanoelectronics
Universiti Kebangsaan Malaysia, 43600 UKM Bangi, MALAYSIA
suhbudin@vlsi.eng.ukm
Tel:603-89450742 Fax:603-89450646 email: azliza@tmdrd.com.my

Abstract: The splitting characteristics between 1310 nm and 1550 nm wavelengths in T-junction photonic crystal is studied numerically. The higher number of defect pillar gives a high transmission and low reflectance of the split optical signals.

I. INTRODUCTION

Photonic crystals (PhC) have inspired a lot of research recently due to the ability to control the propagation of light in more drastic way compared to conventional planar waveguide. The periodic variation of the permittivity in photonic crystals can resort in photonic band gap, i.e. a range of frequencies with no allowed electromagnetic mode. By creating a defect in these structures, the periodicity and consequently the completeness of the band gap are broken and the propagation of light can be localized in the defect region. Its frequency is corresponding to the defect frequency inside the gap.

II. 1310/1550 nm PhC DEMULTIPLEXING

Figure 1 shows the band gap region for a range of r/a (rod radius divided by pitch) value. The horizontal solid lines indicates the channels of normalized frequencies w1 and w2, while vertical solid lines indicate rod radius for the first filter r1, main waveguide channel r, and the second filter r2. Figure 1 also shows that the PhC with rod radius r has the band gap that permits the reflectance for both frequency channel and splitter. At the same time, PhCs with rod radius r1 and r2 have the band gap only for one of two frequency channels.

In this paper we investigate the demultiplexing property of wavelengths 1.55 µm and 1.31 µm in a T-junction PhC wavelength selective splitting device.

The relative frequency and the pitch value for PhC of the first filter can be obtained through equation (1)

\[ w_1 = \frac{\omega a}{2\pi c} = \frac{a}{\lambda} \]  

We have a PhC structure with the following parameters for square lattice: the background refractive index n i = 1, rod refractive index, n r = 3.4, distance between rod centers a = 0.52 µm, r1 =d1/2 =0.109a, r =0.179a and r2 =d2/2 =0.253a. The T-junction device is shown in Figure 2.

Thereupon we present a spectral analysis (the spectral dependence of the power on each output port of the device) using finite difference time domain (FDTD) numerical method.

III. NUMERICAL RESULTS

The wavelength demultiplexing is carried out by the two filter channels, left and right, for wavelength 1310 nm and 1550 nm respectively. As shown in Figure 2, the device consists of input channel produced by the W1 waveguide that is formed by removing one row of rods from the background PhC. This waveguide connects to the splitter waveguides with two output ports SO1 and SO2.

Fig. 3 shows the transmission characteristics at monitor of filter 1, filter 3 and reflectivity detected behind W1 for wavelength 1310 nm. It is clearly shown from Fig. 3a, filter 1 only allows wavelength of 1310 nm to penetrate into filter 1.
The increase in the number of pillars will give higher transmission power inside the filter 1. Fig. 3b shows transmission inside filter 2. Power from this wavelength (1310 nm) should be reflected when entering filter 2 because it can create crosstalk. Single pillar will give high crosstalk and full pillars in line defect will almost eliminate the crosstalk. Fig. 3c shows reflectivity in the device. The monitor for reflectivity is placed behind input channel, W1. The higher number of pillars will help to reduce the reflectivity.

With this design it is found that it will help to reduce reflectivity and boost up the power transmission in the device. Computed electromagnetic field distribution in the square lattice PhC demultiplexing is shown in Fig. 5 for two transmitted signals with wavelength 1310 nm (Fig. 5a) and 1550 nm (Fig. 5b).

From the investigation of the characteristics above, we propose the 1310/1550 nm multiplexing PhC using full defect pillars with diameter $d_1$ in filter 1 and 3 pillars defect with diameter $d_2$ in filter 2.

**IV. CONCLUSION**

In this work we have investigated the properties of the rectangular PC based on finite difference time domain method. Numbers of defect pillars in the defect line will affect the transmission characteristics at the monitor of filter 1, filter 2 and reflectivity. The new design of the wavelength division multiplexer based on the integrated was also proposed.

**REFERENCES**


A Study of Three-Dimensional Fractal Cavity as a Photonic Resonator

Hidetoshi Kakinishi and Yukio Iida
Dept. of Electrical and Electronic Eng., Faculty of Engineering Science, Kansai University
3-3-35, Yamate-cho, Suita-shi, 564-8680 Japan
Tel: +81-6-6368-0790, E-mail: iida@ipcku.kansai-u.ac.jp

Abstract
Confinement of EM-field in the fractal called Menger sponge will be investigated from viewpoints of application to optical functional elements. Higher Q-value is obtained with the structure with slightly corrupted fractal symmetry instead of perfect fractal structure.

Introduction
With the cubic fractal called Menger sponge, electromagnetic field can be confined with three dimensional manner [1],[2], and with a photonic crystal (PhC), electromagnetic field can be confined into a region periphery of which is enclosed by crystals of multiple period. It is interesting that in the case of Menger sponge, confinement is possible even if peripheral structure is thin. In this paper, confinement of electromagnetic field in the cubic fractal will be investigated from viewpoints of application to optical functional elements. Although computations here are made in microwave band, results obtained may be used exactly for photonic band if the cube used is made smaller.

Resonance characteristics of Menger sponge
Figure 1 shows structure of a Menger sponge. The structure is looked from -x-direction and this is a fractal referred to as stage-3 [3]. This cube is composed of dielectric bodies having relative dielectric constant of 8.8 and is placed at the center of analytic space which is the square region of 171mm. The cube is the square of 81 mm. Simulations of the electromagnetic field will be carried out by FDTD method. FDTD cell size is dx = dy = dz =1.5 mm. Excitation is applied from a plane 22.5 mm apart in z-direction from the cube. The size of the plane is same as that of a hole of the fractal in stage-1. Excitation is performed using a wave with such a shape that sinusoidal wave is cut out to have Gaussian pulse configuration, while the spectrum having its center at 3GHz is spread.

Figure 2 shows spectrum of x-component of electric field observed at center portion of the cube. It is noticed that several frequencies are in resonance and that frequency of each resonance tends to move higher side while stage number increases from 2 to 3, and at the same time Q-value (quality factor) decreases. That resonance frequency is moved higher side is attributable to that the number of holes is increased and average
dielectric constant viewed from electromagnetic field is reduced. Reduction in Q-value is noteworthy.

Fractal structure and discussion

If it is hypothesized that fractal structure is suited for confinement of electromagnetic field, it is considered natural that the greater the number of stages, the more confinement effects tend to be obtained. However, a trend contrary to this hypothesis has been exhibited. Filter characteristics of multilayer structure are studied in the literature [4] which reports that filter characteristics are improved with asymmetrical fractal instead of perfect self-similarity. Hence, the authors attempted to use asymmetrical fractal. However, our study differs from above-mentioned one in that we place importance on how strong electromagnetic fields could be confined in a certain structure, while input-output characteristics become controversial for filter case.

Figure 3(a) shows a multilayer structure composed of dielectric bodies having dielectric constant of 8.8, which maintains fractal symmetry. Figure 3(b) shows a structure in which fractal symmetry is corrupted. In Fig.3(b), dx=dy=dz=1.35 mm is used. Electromagnetic field is calculated for these two structures. Figure 4 and 5 show spectra of the electromagnetic field observed at the center of the structure shown in Fig.3(a) and (b), respectively. It is known from this that Q-value at 6.27 GHz for the structure that maintains fractal symmetry is 24 while Q-value at 6.09 GHz for the structure in which symmetry is corrupted is 65 that is 2.7 times greater.

From the results thus obtained, it has been revealed that even the case where confinement of electromagnetic field is targeted, higher Q-value is obtained with the structure with slightly corrupted fractal symmetry instead of perfect fractal structure.

References

On the strain-induced wavelength drifts of thin DWDM thin-film filters

A. K. Chu and M. C. Lee
Institute of Electro-optical Engineering and Semiconductor Technology Research and Development Center, National Sun Yat-sen University, Kaohsiung, Taiwan

Abstract: Strain-induced wavelength drifts of thin DWDM thin-film filters (TFF) were investigated. Blue shift and red shift of the design wavelengths of the TFFs were observed by applying mechanical strain and thermal stress to their substrates, respectively.

Narrow-bandpass filters are one of the key components used in wavelength division multiplexing (WDM) systems such as add/drop filters, multiplexers/demultiplexers, and dispersion compensators. In the emerging WDM networks, dynamic operations of these wavelength-selective devices are increasingly in demand recently. Many different types of wavelength-tunable devices were proposed such as stretched fiber gratings, thermo-optical waveguides, MEMS devices, and tunable thin-film filters (TFFs) [1-3]. Among these technologies, the TFFs are competitive in terms of manufacturing cost and high effectivity. They are able to achieve excellent performance on tunability, insertion losses, and spectral responses. However, due to the nature of stress-free coating characteristics, the tunability of the TFFs induced by thermo-optical effect is relatively small. In addition, these bulk optical devices are not suitable for applications of integrated optics. In fact, low-cost and small-sized integrated devices are crucial for systems of access and local area networks.

In this paper, strain-induced wavelength drift and thermo-optical effects of thin DWDM TFFs were investigated. We created a mechanical deformation of the TFF substrate by reducing the thickness of the conventional 1-nm-bandwidth TFFs. A strain caused by substrate thinning can be used to induce a shift in the design wavelengths of the TFFs. On the other hand, the thermal-optical effects of the thin TFFs were characterized by measuring the design wavelength drift at different substrate temperatures. The experimental demonstrations of the strain-induced wavelength drifts of the thin DWDM TFFs may well lead to novel applications for integrated optics.

The thin TFFs were prepared by standard substrate thinning process used in semiconductor industry. However, to avoid deformation of the TFFs during the process, molding technique was employed to package the TFF substrates prior to the thinning process. Finally, the TFFs were polished with fine alumina powders. Great care must be taken to produce TFFs with specular surfaces. Typical dimension of the filters used for measurement is 250 µm × 250 µm. Hundreds of filters can be obtained from a 0.8 cm square TFF substrate in a single process.

The transmission spectrums of the TFFs before and after the substrate thinning processes are shown in Fig. 1. We used a tunable laser with a wavelength ranged from 1480 nm to 1600 nm as the light source. In addition, two lensed fibers with a working distance of 300 µm were used to reduce optical insertion losses of the measurement. As shown in the figure, after the thinning process, the design wavelength of the TFF was observed to drift toward short-wavelength regions with the decrease of the substrate thickness. For 200-µm-thick TFF, a maximum drift of 3.8 nm from the original design wavelength was obtained. No significant insertion losses were found.

The design wavelength drifts of the thin TFFs were caused by strains due to substrate bending after the thinning process. The transmission spectrums of the TFFs before and after the substrate thinning processes are shown in Fig. 1. We used a tunable laser with a wavelength ranged from 1480 nm to 1600 nm as the light source. In addition, two lensed fibers with a working distance of 300 µm were used to reduce optical insertion losses of the measurement. As shown in the figure, after the thinning process, the design wavelength of the TFF was observed to drift toward short-wavelength regions with the decrease of the substrate thickness. For 200-µm-thick TFF, a maximum drift of 3.8 nm from the original design wavelength was obtained. No significant insertion losses were found.

The thin TFFs were prepared by standard substrate thinning process used in semiconductor industry. However, to avoid deformation of the TFFs during the process, molding technique was employed to package the TFF substrates prior to the thinning process. Finally, the TFFs were polished with fine alumina powders. Great care must be taken to produce TFFs with specular surfaces. Typical dimension of the filters used for measurement is 250 µm × 250 µm. Hundreds of filters can be obtained from a 0.8 cm square TFF substrate in a single process.

The design wavelength drifts of the thin TFFs were caused by strains due to substrate bending after the thinning process. The transmission spectrums of the TFFs before and after the substrate thinning processes are shown in Fig. 1. We used a tunable laser with a wavelength ranged from 1480 nm to 1600 nm as the light source. In addition, two lensed fibers with a working distance of 300 µm were used to reduce optical insertion losses of the measurement. As shown in the figure, after the thinning process, the design wavelength of the TFF was observed to drift toward short-wavelength regions with the decrease of the substrate thickness. For 200-µm-thick TFF, a maximum drift of 3.8 nm from the original design wavelength was obtained. No significant insertion losses were found.

The design wavelength drifts of the thin TFFs were caused by strains due to substrate bending after the thinning process. The transmission spectrums of the TFFs before and after the substrate thinning processes are shown in Fig. 1. We used a tunable laser with a wavelength ranged from 1480 nm to 1600 nm as the light source. In addition, two lensed fibers with a working distance of 300 µm were used to reduce optical insertion losses of the measurement. As shown in the figure, after the thinning process, the design wavelength of the TFF was observed to drift toward short-wavelength regions with the decrease of the substrate thickness. For 200-µm-thick TFF, a maximum drift of 3.8 nm from the original design wavelength was obtained. No significant insertion losses were found.

The thin TFFs were prepared by standard substrate thinning process used in semiconductor industry. However, to avoid deformation of the TFFs during the process, molding technique was employed to package the TFF substrates prior to the thinning process. Finally, the TFFs were polished with fine alumina powders. Great care must be taken to produce TFFs with specular surfaces. Typical dimension of the filters used for measurement is 250 µm × 250 µm. Hundreds of filters can be obtained from a 0.8 cm square TFF substrate in a single process.
measured blue shifts of the TFFs indicate that the effective cavity lengths of the devices were reduced. On the other hand, the increases of the 20-dB passband width can be resulted from increases of the refractive index contrast of the SiO$_2$/Ta$_2$O$_5$ coatings.

Fig. 2. The measured passband widths at -3 dB and -20 dB of the TFFs before and after the thinning process.

The thermal-optical effects of the thin TFFs were characterized by first mounting them onto Si optical benches. A U-groove and V-grooves were formed on the benches by standard cutting and wet chemical etching techniques. Before heating up the benches, the thin TFFs and the lensed fibers were inserted into the U-groove and the V-grooves respectively, and were fixed using high temperature epoxy. The measured transmission spectrums of the thin TFFs with different substrate thickness to the changes of the ambient temperatures are illustrated in Fig. 3. Red shifts of the passband were observed for all TFFs. This implies that the strains caused by thermal effects will tune the design wavelengths moving in a direction in contrast to that of the TFFs prepared by substrate thinning process. Maximum tunability of 4.8 pm/K was obtained by heating up the TFFs with 200 μm-thick substrate to a temperature of 250 °C.

In conclusion, the strain-induced wavelength drifts of the thin DWDM TFFs were studied. We found that mechanical strains induced by substrate thinning process will caused a blue shift of the design wavelengths. However, red shift of the TFFs was observed when thermal strains were applied to the substrates. These findings can be applied in photonic integration of various systems for local and access fiber optical networks.

Fig. 3. The wavelength drifts of the thin TFFs with substrate thickness of (a) 500 μm, (b) 400 μm, (c) 300 μm, and (d) 200 μm to the changes of the ambient temperatures.

References:
Optimum coupling efficiency condition of quadruple-series-coupled microring resonator

Tomoyuki Kato\textsuperscript{1,2}, Yasuo Kokubun\textsuperscript{1}

\textsuperscript{1}Yokohama National Univ., Graduate school of Eng., 79-5 Tokiwadai Hodogaya-ku, Yokohama 240-8501, JAPAN
\textsuperscript{2}presently with Tokyo Institute of Technology, 4259, Nagatsuta Midori-ku, Yokohama 226-8503, JAPAN
Tel: +81-45-924-5064, Fax: +81-45-924-5977, Email: kato.t.ae@m.titech.ac.jp

Abstract
Series-coupled microring resonators can synthesize the spectrum shape. We made clear that the relationship between coupling efficiencies and the spectrum shape to obtain the optimum coupling efficiency condition of quadruple-series-coupled microring resonators.

1 Introduction
Series-coupled microring resonators can realize the boxlike transmission spectrum by optimizing the coupling efficiencies between the busline and the resonator and between resonators. Higher order series-coupling realize more flat passband and shaper roll-off. We have designed and demonstrated the Butterworth filter response of the double-series-coupled microring resonators \cite{1}, \cite{2}. For further improvement of spectrum response, higher order series coupling is required \cite{3}. The quadruple-series-coupled microring resonators (QSC-MRR) illustrated in Fig. 1 is suitable for the integrated filter circuit, which consists of the cascading of microring resonator unit.

However, the synthesis of boxlike higher order series-coupled microring resonator is complicated, since there are many coupling between waveguides. We analytically made clear the dependence of spectrum shape on the coupling efficiencies.

2 Transfer function of QSC-MRR
The transfer function of QSC-MRR is approximately expressed by
\[
P_o = \frac{1}{1 + A[f + g(\omega - \omega_r) + (\omega - \omega_c)^2]}\]  \hspace{1cm} (1)
where $A$ denotes the resonant intensity, $f$ and $g$ are the polynomial of coupling efficiencies $\kappa_0, \kappa_1$, and $\kappa_2$, which are defined as shown in Fig.1.
3 Butterworth filter condition
Butterworth filter is realized when the transfer function has a single pole, i.e., $f = g = 0$.

$$P_D = \frac{1}{1 + A(\omega - \omega_0)}$$

Figure 2 shows the Butterworth QSC-MRR filter response. The coupling efficiency condition for the Butterworth filter is obtained as shown in Fig. 4. The spectrum shape gets dull by reducing the transmittance in the resonator, $a$. Although the strong $\kappa_{i1}$ is needed for the Butterworth filter in the lossy case, $\kappa_{i2}$ does not depend on the loss in the resonator.

4 Chebyshev filter condition
Chebyshev filter is realized when the polynomial of transfer function is Chebyshev polynomial.

$$P_a = \frac{1}{1 + A C_4(\omega - \omega_0)}$$

where $C_4$ is the 4th order Chebyshev polynomial. Figure 3 shows the 4th order Chebyshev QSC-MRR filter response. The coupling efficiency condition for the Chebyshev filter was obtained as shown in Fig. 5. The ripple deepens with the increase of coupling efficiency between resonators. Figure 6 shows the dependence of the ripple depth on the coupling efficiency between resonators normalized by the Butterworth condition, $\kappa_{i1b}$. Figure 7 shows the dependence of bandwidth on the coupling efficiency between resonators normalized by the Butterworth condition. From Figs. 6 and 7, if the 1dB ripple is allowed, the 2.5 times wider bandwidth than Butterworth case can be realized.

5 Conclusion
We made clear the optimum coupling efficiency condition of quadruple-series-coupled microring resonator for Butterworth and Chebyshev filters. Butterworth filter response can be realized, even if the resonators have a loss. When the ripple is allowed, boxlike spectrum shape is realized by the Chebyshev coupling efficiency condition.

References