Representation of an equivalent circuit for capacitive wireless power transfer using a distributed-constant circuit

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Abstract: Capacitive wireless power transfer is performed using large electrodes or using a medium having high relative permittivity such as water. The transfer efficiency demonstrates a pronounced change with frequency. A lumped-constant equivalent circuit with parasitic elements represents the change. However, a relationship between conditions for generating minimum/maximum values of the transfer efficiency and structural parameters of the electrodes in the coupler has not clarified. In this study, an equivalent circuit is proposed using a distributed-constant circuit. We converted the lumped-constant circuit into two open-ended directional couplers. From the result, we derived the conditions for generating the minimum and maximum values of the voltage input/output ratio. The efficiencies calculated by the distributed-constant circuit and electromagnetic simulation were in good agreement.

Keywords: wireless power transfer, capacitive wireless power transfer, directional coupler, equivalent circuit, maximum available efficiency

Classification: Energy in Electronics Communications

References


1 Introduction

Many studies on capacitive wireless power transfer (WPT) use the lumped-constant equivalent circuit [1, 2, 3, 4, 5, 6]. The transfer efficiency shows significant change with frequency in capacitive WPT via water. The change in the transfer efficiency is represented by adding parasitic elements into a conventional equivalent circuit [4, 5, 6], and is predicted by a voltage input/output ratio derived from the equivalent circuit. However, a relationship between conditions for generating minimum/maximum values of the transfer efficiency and structural parameters of the electrodes in the coupler is not clarified. In this study, an equivalent circuit is proposed via fresh water using the distributed-constant circuit. It is elucidated from our circuit that the minimum and maximum values of the efficiency can be predicted depending on wavelength of the transfer frequency. First, the frequency characteristic of the maximum available efficiency was calculated by electromagnetic (EM) simulations. The lumped-constant equivalent circuit with the parasitic elements represents the characteristic. One study has used a general directional coupler for WPT [7]. We converted the equivalent circuit into the two open-ended directional couplers (i.e., a distributed-constant circuit). Using the couplers, the conditions for generating the minimum and maximum values of the voltage input/output ratio were derived. The efficiencies calculated by the distributed-constant circuit and EM simulation were in good agreement. Finally, it was shown that the minimum and maximum values of the transfer efficiency can be predicted from the structural parameters of the coupler electrodes.

2 Representation by lumped-constant circuit

Figures 1(a) and (b) show the coupler model and the equivalent circuit of the capacitive WPT, where #1–#3 and #2–#4 show the primary and secondary sides, respectively. The capacitances that occurred in #1–#4 and #2–#3 cannot be ignored, especially when there is misalignment between the primary and secondary sides [2]. The capacitances were ignored because this study assumes that there is perfect symmetry without misalignment. The capacitance $C_s$ and loss $R_s$ occurred at that adjacent electrodes (#1–#3, #2–#4). Furthermore, $C_m$ and $R_m$ occurred at the facing
electrodes (#1–#2, #3–#4). $L_m$ and $L_s$ are parasitic inductances in each capacitance. As shown in Fig. 1(c), the coupler model comprises four metal plates and four 50-Ω reference ports imitated subminiature connectors. Here, tap water is represented using the relative permittivity $\epsilon_r = 79$ and the conductivity $\sigma = 0.011 \text{ S/m}$. We calculated the maximum available efficiency $\eta_{\text{max}}$ [3], which indicates the achievable power transfer efficiency under conjugate matching conditions. First, we calculated the four-port single-ended S-parameters using the equivalent circuit (advanced design system) and EM simulation (CST microwave studio) and converted them into two-port differential S-parameters. Next, the results were converted into the two-port impedance matrix. Finally, $\eta_{\text{max}}$ was calculated by substituting the impedance results into the following two equations:

$$\alpha = \frac{|Z_{21}|}{\sqrt{R_{11}R_{22} - R_{12}R_{21}}}.$$  \hspace{1cm} (1)

$$\eta_{\text{max}} = 1 - \frac{2}{1 + \sqrt{1 + \alpha^2}}.$$  \hspace{1cm} (2)

Figure 1(d) shows the calculated $\eta_{\text{max}}$ values. The element values are as follows: $C_s = 72 \text{ pF}$, $R_s = 960 \Omega$, $L_s = 95 \text{ nH}$, $C_m = 840 \text{ pF}$, $R_m = 74 \Omega$, and $L_m = 8.5 \text{ nH}$. The $\eta_{\text{max}}$ values are calculated from the equivalent circuit in Fig. 1(b) and EM simulation, which are in good agreement up to around 60 MHz. However, the relationship between the conditions for generating the minimum/maximum values of the transfer efficiency and the structural parameters of the electrodes in the coupler is not clarified.

3 Representation by distributed-constant circuit

3.1 Formulation of the voltage input/output ratio

Figure 2 shows the procedure of the conversion from the capacitive coupler into the directional coupler. To simplify the figures, $L_s$, $L_m$, $R_s$, and $R_m$ are not listed. Figure 2(a) is represented as the balanced circuit topology by including the power
Fig. 2. Procedure of conversion into directional coupler.
lines for the odd mode in Fig. 2(g). The capacitances of the two modes are expressed by Eqs. (3) and (4) [8]. Equations (5) and (6) show the characteristic impedances for the even and odd modes. Note that \( R_e \) and \( R_o \) show the loss generated at the inductances \( L_e \) and \( L_o \), respectively, in each mode. Furthermore, \( G_e \) and \( G_o \) show the loss generated in Eqs. (3) and (4). Based on the results of Section 2, the conditions \( R_e > 1/(\omega C_e) \) and \( R_o > 1/(\omega C_o) \) are satisfied at 3 MHz. Therefore, we apply the lossless transmission line theory for conversion into the directional coupler as follows:

\[
C_e = C_{in-GND} = C_{out-GND},
\]

\[
C_o = C_{in-GND} + 2C_{in-out} = C_{out-GND} + 2C_{in-out}.
\]

\[
Z_{0e} = \sqrt{\frac{R_e + j\omega L_e}{G_e + j\omega C_e}} \approx \sqrt{\frac{L_e}{C_e}}.
\]

\[
Z_{0o} = \sqrt{\frac{R_o + j\omega L_o}{G_o + j\omega C_o}} \approx \sqrt{\frac{L_o}{C_o}}.
\]

The input impedance for each mode is expressed by Eqs. (7) and (8). These equations can be simplified because of the open condition of \( Z_{open} \).

\[
Z_{in}^e = \frac{Z_{0e} + jZ_{open} \tan \theta}{Z_{0e} + jZ_{open} \tan \theta} \approx \frac{Z_{0e}}{j \tan \theta}.
\]

\[
Z_{in}^o = \frac{Z_{0o} + jZ_{open} \tan \theta}{Z_{0o} + jZ_{open} \tan \theta} \approx \frac{Z_{0o}}{j \tan \theta}.
\]

The voltages at #1 for each mode are listed below:

\[
V_{in}^{e'} = V_{in}' \frac{Z_{in}^e}{Z_{in}^e + Z_0}, \quad V_{in}^{o'} = V_{in}' \frac{Z_{in}^o}{Z_{in}^o + Z_0}.
\]

Here, as long as \( Z_{in} = Z_0 \) is satisfied, we have \( V_{in}' = V_{in} \) by voltage division in Fig. 2(e) because there was no voltage reflection. Moreover, by the preceding voltage definition in Fig. 2(f), \( V_{out} \) is expressed as follows:

\[
V_{out} = V_{out}^{e'} + V_{out}^{o'} = V_{in} + (-V_{in}'') = V_{in} \left[ \frac{Z_{in}^e}{Z_{in}^e + Z_0} - \frac{Z_{in}^o}{Z_{in}^o + Z_0} \right].
\]

Equation (10) reduces to Eq. (11) using Eqs. (7) and (8). The voltage ratio of \( V_{out}/V_{in} \) is given by Eq. (12).

\[
\frac{V_{out}}{V_{in}} = \frac{Z_{0o} - Z_{0e}}{Z_{0e} + Z_{0o} + j[Z_0 \tan \theta - Z_{0e} Z_{0o}/(Z_0 \tan \theta)]}.
\]
ports were terminated at \( Z_0 \) in Fig. 2(e), we would have \( \sqrt{Z_{0e}Z_{0o}} = Z_0 \) \[8\]. However, \( \arctan \pm (\lambda / 8) \) does not hold because \( Z_{\text{open}} \neq Z_0 \). Therefore, the wavelength having the maximum values are determined by the \( \sqrt{Z_{0e}Z_{0o}} / Z_0 \) ratio.

\[
\theta_{\min} = \arctan 0, \quad \arctan (\pm \infty) = \frac{(n - 1)\pi}{2} \quad (n : 1, 2, 3, \ldots) \tag{13}
\]

\[
\theta_{\max} = \arctan \left( \pm \frac{\sqrt{Z_{0e}Z_{0o}}}{Z_0} \right) \tag{14}
\]

### 3.2 Verification

Figure 3(a) shows the equivalent circuit represented by two open-ended directional couplers. For verification, we assign concrete values to \( \epsilon_r, \sigma, S, L, W, \) and \( B \) in Fig. 2(g). Here, \( L \) is the transmission length across which the phase difference occurs. The phase difference is governed by the length of the electrode, which is affected by the wavelength shorting derived from \( \epsilon_r \). Therefore, \( L \) is the diagonal length of the electrode (141 mm) in Fig. 1(c). \( W \) is 100 × 100/141 = 71 mm, which ensures that the area size is constant. Based on the information in Section 2, we specify \( \epsilon_r = 79, \sigma = 0.011 \) S/m, and the distance between the facing electrodes \( S = 10 \) mm. \( C_s \) corresponds to \( C_{\text{in-GND}} \) and \( C_{\text{out-GND}} \), as shown in Fig. 2(g). \( C_s \) is the edge-coupled capacitance, and the distance is 50 mm. However, \( C_{\text{in-GND}} \) and \( C_{\text{out-GND}} \) are the broadside-coupled capacitances and the distance is \( (B - S)/2 \). We compared the capacitances in the broadside-coupled and edge-coupled states, and specified that the broadside > edge. Therefore, \( B \) was more than 50 mm in length, and the fitting result was 360 mm.

Figure 3(b) shows the calculated \( \eta_{\text{max}} \) values; the minimum value was 59.8 MHz, and maximum values was 35.1 MHz. When we considered the effect of the wavelength shorting, we reported that 59.8 MHz was \( \lambda / 4 \), and 29.9 MHz was \( \lambda / 8 \) of \( L = 141 \) mm. The minimum values were in good agreement. The calculated results of Eq. (12) with \( Z_0 = 1, Z_{0e} = 4 \) and the various \( Z_{0o} \) values of 0.125, 0.25, and 0.5 are shown in Fig. 3(c). \( \sqrt{Z_{0e}Z_{0o}} = Z_0 \) holds at \( Z_{0o} = 0.25 \), and the maximum value occurred in \( \lambda / 8 \) (45\(^\circ\)). Moreover, we confirmed that the center around \( \lambda / 8 \) was affected by the change in the \( \sqrt{Z_{0e}Z_{0o}} / Z_0 \) ratio. From the above, we observe that the difference between 29.9 MHz and 35.1 MHz occurred because \( \sqrt{Z_{0e}Z_{0o}} > Z_0 \).
4 Conclusion

In this study, we represented the equivalent circuit of the capacitive WPT using a distributed-constant circuit. The lumped-constant equivalent circuit was converted into the two open-ended directional couplers. Consequently, we determined the condition for the generation of minimum and maximum value of the voltage ratio. The $\eta_{\text{max}}$ values were calculated using the distributed-constant equivalent circuit and EM simulation, which were in good agreement. Finally, we clarified that the minimum and maximum $\eta_{\text{max}}$ values occurred at 0 and $\lambda/4$, respectively, and centered around $\lambda/8$. The changes in $\lambda/8$ depended on the $\sqrt{Z_{0e}Z_{0o}/Z_0}$ ratio.

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PAM-4 eye-opening monitoring techniques based on Gaussian mixture model fitting

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Abstract: Four-level pulse amplitude modulation (PAM-4) data formats are adopted to achieve next-generation high-speed data transmission standards. In this letter, a novel eye-opening monitoring technique based on machine learning is proposed to evaluate the received signal quality for the adaptive coefficients setting of a transmitter feed-forward equalizer for PAM-4 signaling. The monitoring technique employs a Gaussian mixture model (GMM) to classify the received PAM-4 symbols. Simulation and measured results of the coefficient adjustment using the GMM method are presented.

Keywords: multi-valued logic, PAM-4, eye monitoring, Gaussian mixture model

Classification: Transmission Systems and Transmission Equipment for Communications

References
1 Introduction

High-speed Input/output (I/O) data rates are continually increasing to fulfill the requirement of aggregate I/O bandwidth for chip-to-chip links, backplanes, and data-center transmission. However, as the rate increases, signal distortions due to the intersymbol interference (ISI) caused by the finite bandwidth of wire channels, and reflection, significantly limit the I/O bandwidth. Therefore, advanced signaling techniques, such as the four-level pulse amplitude modulation (PAM-4) data format, are employed to improve signal bandwidth. PAM-4 signaling can transmit data at the same baud rate using a twice-slower symbol rate. Moreover, feed-forward equalizer (FFE) and pre-emphasis are used to mitigate the ISI effects at the transmitter side [1, 2, 3, 4]. Because the FFE can be implemented using simple digital circuitry, the parameters of FFE circuits can be adaptively adjusted following the transmission line characteristics. An eye-opening monitoring (EOM) technique [5, 6] is used to evaluate the quality of the received signal. However, because PAM-4 has three eyes, the EOM algorithm and circuitry for PAM-4 tend to be complicated.

To address the above problems, we propose a machine learning method for statistical symbol detection from deteriorated transmitted PAM-4 signaling to perform adaptive FFE coefficient setting. The parameter optimization problem for the unknown transmission characteristics can be considered to be an unsupervised learning problem. Thus, an EOM method is developed, using a Gaussian mixture model (GMM) to estimate the effect of waveform distortion. Each PAM-4 symbol level exhibits four Gaussian distributions at the receiver side, and the distributions are mixed together owing to ISI. GMM fitting can classify the mixed distributions into each transmitter PAM-4 symbol. Therefore, we can adjust the FFE parameters to eliminate ISI, obtaining the PAM-4 eye-opening diagram. In this letter, simulation and measured results of the coefficient adjustment using the GMM method are presented to demonstrate the feasibility of the proposed PAM-4 EOM technique.

2 Evaluation of received PAM-4 waveform distortion using Gaussian mixture model

Eye diagrams at the receiver are widely used to evaluate the signal quality of received waveforms in serial links. Figure 1(a) shows the measured examples of PAM-4 eye diagrams and the corresponding histogram of symbols distribution in a vertical region. The vertical histogram has four distinct and separate distributions. The distribution of each symbol exhibits Gaussian distribution because of ISI. As the data rate increases, the ISI also increases, which broadens the Gaussian distribution of each symbol. Subsequently, the four Gaussian distributions overlap each other,
Fig. 1. (a) Example of PAM-4 eye diagrams, and the corresponding histogram of symbols distribution in a vertical region, (b) flowchart of EM algorithm, (c) overview of the proposed GMM-based PAM-4 EOM technique

making it difficult to detect the symbol. The conventional EOM technique cannot detect the symbol from the histogram composed of superposed distributions [6].

The estimation of the received symbols passing through an unknown transmission characteristic can be considered as unsupervised learning. Therefore, by applying GMM fitting for estimating symbol distributions against an unknown transmission line, a novel PAM-4 EOM technique relying on a statistical classification is proposed.

A linear combination of multiple Gaussian distributions can be modeled as a GMM [7], which is expressed as follows:

\[ p(x) = \sum_{k=1}^{K} \pi_k N(x | \mu_k, \sigma_k), \]  

(1)

where \( N(x | \mu, \sigma) \) has a Gaussian distribution with a mean \( \mu \) and variance \( \sigma \), \( \pi_k \) is the mixing coefficient corresponding to the weight for each Gaussian distribution, which is normalized as \( \sum_{k=1}^{K} \pi_k = 1 \). A GMM is a probabilistic model that assumes that all data points are generated from a mixture of unknown Gaussian distributions. By GMM fitting from the distribution of the observed data samples, we can estimate the mean and variance of each Gaussian distribution symbol. Accordingly, the received PAM-4 symbol can be separated and estimated.

In the case of PAM-4 symbol GMM estimation, we can assume that the received symbols are a mixture of four Gaussian distributions. Typically, the expectation maximization (EM) algorithm is used for estimating the parameters, i.e., mean, variance, and weight, of each Gaussian distribution. The EM algorithm repeatedly calculates a solution that maximizes the likelihood of the observed values, and the convergence calculation is executed using the following steps: (1) Initial value setting, (2) Expec-
tation step, (3) Maximization step, (4) Repeat step (2) until convergence is achieved (Fig. 1(b)). The initial value is set by using a $k$-means clustering method and setting appropriate initial groups from the observed values. Subsequently, steps (2)-(4) are repeated until convergence is achieved.

Figure 1(c) shows an overview of the proposed GMM-based PAM-4 EOM technique. First, the histogram of the sampling values of PAM-4 signal at the receiver is developed. Subsequently, the symbol distributions are clustered using GMM fitting. The average and variance of each PAM-4 symbol can be estimated by fitting each symbol distribution with GMM from the sampling values at the receiver. By clustering the symbol distribution with GMM fitting, we can evaluate the effect of the distortion at the receiving end even when the eye is entirely closed. In GMM, the distance between symbols can be evaluated from the average values $\mu_k$. Moreover, the effect of ISI at each symbol is evaluated from the variance values $\sigma_k$.

3 Simulation and measured results for PAM-4 eye monitoring

Figure 2(a) shows the simulation results of 2 Gb/s PAM-4 transmission over a 1 m micro-strip line (MSL) and the symbol distribution histograms at the receiving end. In the simulation, the signal response and distortion were calculated using the frequency characteristics of the measured MSL’s S-parameters. As shown in the histogram of 2 Gb/s PAM-4 received signal (Fig. 2(b)), the symbol values spread owing to the ISI effect; hence, the eye height and width decrease. However, the eye is opened enough to detect each symbol level. Figure 2(b) also shows the GMM estimation results, which are fitted as four Gaussian distributions. In this simulation, we assumed that the effect of ISI in each symbol is almost similar, and the variance of each distribution is equal. Each symbol distribution of 2 Gb/s symbol is fitted using

![Fig. 2. Simulation result of PAM-4 data transmission (a) eye-diagram at 2 Gb/s, (b) the result of GMM fitting of symbol distribution histogram at 2 Gb/s, (c) eye-diagram at 4 Gb/s, (d) the result of GMM fitting of symbol distribution histogram at 4 Gb/s](image-url)
GMM, where each symbol value is estimated as average values ($\mu_0 = -0.5678$, $\mu_1 = -0.1939$, $\mu_2 = 0.1720$ and $\mu_3 = 0.5525$) and a variance value ($\sigma = 0.0024$).

Figure 2(c) shows the simulation results of 4 Gb/s PAM-4 data transmission; these data exhibit severe ISI. The resulting closed eye diagram has a histogram of overlapped Gaussian distributions (Fig. 2(d)). Even in such a harsh condition, GMM can classify each symbol distribution. Moreover, we obtained fitting results such that the average values and variance are $\mu_0 = -0.4366$, $\mu_1 = -0.1427$, $\mu_2 = 0.1686$ and $\mu_3 = 0.4485$ and $\sigma = 0.0108$, respectively, as shown in Fig. 2(d). The results can be used to adjust the FFE coefficients’ parameters.

As a verification experiment, parameters of the FFE were adjusted, provided that no information on transmission data and transmission characteristics is available. In this experiment, we assume an FFE with two taps and two adjustment parameters. The COBYLA algorithm in the python scipy library was used for parameter optimization. The objective function for the parameter adjustment sets the variance value $\sigma$, which is correlated with the effect of ISI. The variance value was calculated from the received symbols by using a GMM fitting assuming that each symbol variances are equal. The optimization algorithm attempts to search the parameters minimizing the variance.

Figure 3(a) shows the transition of the parameter adjustment results at 4 Gb/s PAM-4 data transmission. The optimization algorithm searched the set of parameters providing the smallest variance, which was calculated using 500 sampling symbols per each iteration. Consequently, it is possible to minimize the symbol variation owing to the effect of ISI at the receiver; therefore, we can open the eye. The sampling timing is randomly changed within the range of $-40\%$ to $20\%$ UI (unit interval) from the judgment timing, and the symbol values are acquired at the receiving end. The results confirm that the eye-opening ratio improves along with
minimizing the evaluation value by iterating the GMM fitting. In particular, we can observe that even if the initial eye is entirely closed, the parameter adjustment operates efficiently for improving the eye opening.

Figures 3(b) and (c) show the measurement setup and the measured PAM-4 eye-diagram (4 Gb/s) using the adjusted FFE parameters. The FFE modulated signal was calculated by simulation and was transferred to an arbitrary waveform generator (Tektronix AWG70001A; 50 GS/s) to emulate FFE operation. The PAM-4 eye-diagram obtained from waveforms passing through 1 m MSL shows that the optimized FFE mitigates the ISI; therefore, the 4-valued signal eye is open to 0.58 UI.

4 Conclusion

In this letter, we proposed a novel PAM-4 eye-monitoring technique based on machine learning using a GMM fitting. The classification technique could optimize the coefficients of the FFE for unknown transmission characteristics. Moreover, owing to the GMM fitting, the PAM-4 eye-monitoring technique made it possible to estimate ISI effects even when the eye was closed. Therefore, the proposed EOM method effectively adjusts the parameters of the FFE even when the channel conditions are not predicted.

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Optical TDM technology for N+1 redundant systems in high resolution optical earth observation satellites

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Abstract: Optical remote sensing has been greatly expanded by the construction of satellite constellations and improving the performance of the optical sensors. As the data throughput of earth observation satellites increases, photonic technology is expected to be used for the interconnections within the payload in order to reduce its overall size, mass, and power consumption, and to improve its immunity to EMI. Here, we propose an optical time division multiplexing technology for N + 1 redundancy in the payload, which is able to improve reliability by reducing the number of active components in the satellites, and present a feasibility study on its transmission efficiency.

Keywords: optical time division multiplexing, digital payload, N + 1 redundancy, Barrel shifter, remote sensing, earth observation satellite

Classification: Sensing

References


1 Introduction

Optical remote sensing from space for earth observation has been applied widely to various fields, such as land observation, weather observation, disaster prevention, etc. Furthermore, achieving higher spatial resolution thanks to the improvement of the image quality of the charge coupled device (CCD) image sensors [1] and more frequent observations using multiple satellite constellations [2] are allowing this field to expand steadily into agriculture, fisheries, and forestry.

Recently, the Complementary Metal-Oxide Semiconductor (CMOS) image sensor, which is popular for consumer Digital Cameras, has been introduced to earth observation satellites [3]. Improvement of the readout speed and reduction of the power consumption were achieved by using CMOS image sensors, which are voltage driven devices.

Digital signal processing technology using application specific integrated circuits (ASIC) and field-programmable gate arrays (FPGA) has been in development for realizing higher data throughput for the payload as improving the spatial resolution and readout speed of the image sensors. Additionally, interconnections using optical fibers have been introduced in order to further increase the data throughput, to reduce the size, mass, and power consumption, and to improve the immunity to EMI [4]. 40 Gbps (4 channels × 10 Gbps) optical modules for space conditions have been developed for the optical interconnections [5].

Here, we propose optical time division multiplexing (TDM) technology for an N + 1 redundant system in earth observation satellites. The redundant configuration, which requires twice the number of interconnections, has been applied to payloads in satellites due to the requirements for fault tolerance and higher availability. Optical TDM is a well known and mature technology used for passive optical network (PON) systems, which reduce the cost by sharing the optical fiber among multiple users in terrestrial telecommunications [6]. The proposed N + 1 redundancy system, which can be implemented with a smaller number of active components than conventional systems, will make it possible to reduce the cost, size, mass, and power consumption, and to improve the reliability of the satellites.
2 Redundancy systems in earth observation satellites

Fully redundant payload systems with cross strapping have been applied in order to maximize the availability of satellites while maintaining flexibility [7]. Figure 1(i) shows an example of a redundant configuration for an earth observation satellite with three image sensors. The parallel image data from the three image sensors are separately converted to digital form by the A/D converters (ADC), processed by the digital signal processors (DP), compressed (Comp), stored in the memories (STR), processed for transmission (TX), and then transmitted to ground via the antennas. The units from DP to antenna for each sensor are fully redundant and the connections between the primary and standby units from DP to TX are cross strapped. The standby units are only activated in the event of failure of the associated primary units in order to reduce the power consumption.

![Fig. 1. Examples of redundant system configurations for earth observation satellites.](image)

Recently, an N + 1 redundant system with the Barrel shifter configuration shown in Fig. 1(ii) has been applied in order to reduce the size and mass of the satellites [8]. There are three primary units and one set of redundant units from DP to antenna. The connections from DP to TX are in a Barrel shifter configuration. The standby units are also inactive until a failure occurs in a primary unit. Note that a 1 + 1 redundant system is the same as the cross strapped configuration shown in Fig. 1(i).
3 N+1 redundant system by using optical TDM technology

The block diagrams of a conventional Barrel shifter configuration with optical connections are shown in Figs. 2(i) and (ii). The digital signals from the serializer (Ser) are converted to optical signals by the E/O and split out by the optical coupler at the sender side (Units 1 to 4). At the receiver side (Units 5 to 8), two optical signals are separately converted to electrical signals by each O/E and selected by an electrical switch (E-SW) for Fig. 2(i). One of the two optical signals is selected by an optical switch (O-SW) and then converted to electrical signals by the O/E for Fig. 2(ii). Either the electrical or optical switches are required for the conventional Barrel shifter configuration.

![Block Diagrams](image)

**Fig. 2.** Proposed optical N + 1 redundant system.

The block diagram of the proposed Barrel shifter configuration using optical TDM technology and the timing chart of the signals at points (a) and (b) are shown in Figs. 2(iii) and (iv). The differences between this configuration and Fig. 2(ii) are the functions for forming burst packets at the serializer for TDM signals (SerT) and gating them at the deserializer for TDM signals (DesT), and the substitution of

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optical couplers (CPL) for the optical switches (O-SW) at the receiver side (Units 5 to 8). The signals are formed into burst packets by SerT, converted to optical by the E/O, and split out by the optical coupler at the sender side (Units 1 to 4). The optical bursts from the odd and even numbered units are alternated within the duration of a given signal block. At the receiver side, the two optical bursts are combined in the optical coupler, converted to electrical by the O/E, and then gated by DesT. Note that the optical burst packets become time interleaved when combined in the optical coupler, and the primary packets or the redundant packets can be selected by the DesT at the receiver side as required. The components used for generating and gating optical burst packets in terrestrial telecommunications [9] could be used to provide these functions in the proposed system.

As the proposed optical N + 1 redundant system requires a smaller number of active optical components compared to the conventional configurations shown in Figs. 2(i) and (ii), it can be expected to reduce the size, mass, and power consumption, and to improve the overall reliability.

4 Performance of proposed optical N+1 redundant system

As the proposed optical N + 1 redundant system has chains of optical bursts, each optical burst packet should have a guard time, which includes the laser on/off times, the synchronization time, and the payload data, as shown in Fig. 3(i).

We verified the transmission efficiency, an important factor for an optical TDM system, which is defined as the ratio of total payload data within the duration of a signal block. Figure 3(ii) shows the calculated transmission efficiency as a function of the duration. The guard time and the synchronization time were 128 ns and 1200 ns, respectively [6]. The calculated transmission efficiency was greater than 96% even when the duration was only 70 μs, and it was greater than 99% when the
duration was more than $300\,\mu$s.

The difference in transmission time of the fibers in the primary and redundant paths can cause a time interleaving error, but this should be small enough because even a 1 m difference in the fiber lengths causes less than a 5 ns timing error as shown in Fig. 3(iii). Although, the data rates of the E/O and O/E are required to be twice that of the original electrical signals due to the multiplexing of two signals, 40 Gbps (4 channels × 10 Gbps) optical modules suitable for space conditions have already been developed [5], in addition to which, up to 400 Gbps products are available for terrestrial telecommunications.

5 Conclusions

We have proposed an optical N + 1 redundancy system for a satellite payload using optical TDM technology that can realize a redundant configuration with a smaller number of active components than a conventional configuration. A basic configuration and an example of its operation were introduced. A study to verify the transmission efficiency demonstrated the feasibility of this system. We showed that it has the potential to reduce the cost, size, mass, and power consumption, and to improve the reliability of high resolution optical earth observation satellites.
First demonstration of spectral domain polarimetric decomposition with airborne circular synthetic aperture radar

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Abstract: In this letter, we present the first demonstration of the spectral domain polarimetric decomposition concept with an airborne circular synthetic aperture radar (CSAR) dataset. In CSAR systems, radar-carrying aircraft moves along a circular path that encloses the desired target area. The CSAR image contains the scattering characteristics of the target with respect to the azimuth angle, and we can retrieve it by transforming the spatial domain image into its spectral domain representation. Furthermore, the application of polarimetric decomposition to the spectral domain CSAR image provides additional information about the polarimetric scattering mechanisms. We used a publicly released airborne polarimetric CSAR dataset for the demonstration.

Keywords: SAR, circular SAR, polarimetric radar, polarimetric decomposition, spatial frequency spectrum

Classification: Sensing

References

1 Introduction

Circular synthetic aperture radar (CSAR) is a relatively modern microwave imaging technique that employs a circular flight path of the radar-carrying aircraft [1, 2, 3]. The advantage of CSAR is the ability of full azimuth angle measurement to produce a high-resolution radar image. The reconstructed CSAR image contains information about the variation of the scattering intensity of the observed targets with respect to the azimuth angle. We can extract this information by transforming the CSAR image into its spectral domain image via the windowed two-dimensional Fourier transform [1, 3].

In addition, several polarimetric concepts developed for the conventional synthetic aperture radar (SAR) systems that have a linear flight track (linear SAR) can be applied to the spatial or spectral domain polarimetric CSAR (PolCSAR) images to infer the scattering mechanisms of the targets. In [3], the combination of the spectral domain decomposition and the model-based polarimetric decomposition was demonstrated using a PolCSAR dataset of a turntable measurement in an anechoic chamber.

In this letter, we present the first demonstration of the spectral domain polarimetric decomposition with real airborne PolCSAR images. For this purpose, we use the publicly available PolCSAR datasets (GOTCHA) [4] released by the Air Force Research Laboratory (AFRL). The experimental result shows the validity of the spectral domain target decomposition for the real PolCSAR dataset.

2 Polarimetric circular SAR

This section describes the system model, signal model, and image reconstruction procedure for the PolCSAR images. Moreover, we summarized the spatial domain and the spectral domain polarimetric decomposition algorithms.

Figure 1 shows the PolCSAR system model. As shown in Fig. 1(a), the radar-carrying aircraft moves along a circular path of the radius $R_g$. The radar beam points toward the center of the circular path during the data acquisition. We denote radar azimuth angle as $\phi$ and the constant radar altitude as $Z_c$.

![Fig. 1. PolCSAR system model.](image)

For simplicity, we assume that all the scatterers of interest are confined in the two-dimensional $(x, y)$-plane at the altitude $z = 0$. In this situation, the radial distance between the sensor at the azimuth angle $\phi$ and the scatterer at $(x, y)$ can be
expressed as follows:

\[ r(\phi; x, y) = \sqrt{(x - R_\phi \cos \phi)^2 + (y - R_\phi \sin \phi)^2 + Z_c^2}. \]  

(1)

We denote the horizontal and vertical polarization as \( h \) and \( v \), respectively. The radar transmits \( h \)- or \( v \)-polarized wave, and then collects the scattered wave with \( h \)- and \( v \)-polarized antennas. We denote the \( qp \)-polarized spatial domain reflectivity by \( g_{qp}(x, y) \), where \( p \in \{h, v\} \) and \( q \in \{h, v\} \) represent the transmit and receive polarization channel, respectively. Using these notations, we can express the frequency domain \( qp \)-polarized received signal \( S_{qp}(\omega, \phi) \) as follows:

\[ S_{qp}(\omega, \phi) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} g_{qp}(x, y) e^{-j2k_r r(\phi; x, y)} \, dx \, dy. \]  

(2)

In the above equation, \( k_r = \omega/c \) represents the radial wavenumber, where \( c \) and \( \omega \) are the wave propagation speed and angular frequency, respectively.

PolCSAR image reconstruction is the problem to recover the spatial domain polarimetric reflectivity \( g_{qp}(x, y) \) from the corresponding received signal \( S_{qp}(\omega, \phi) \). This can be simply accomplished by

\[ g_{qp}(x, y) = \int_{0}^{2\pi} \int_{0}^{\infty} S_{qp}(\omega, \phi)e^{j2k_r r(\phi; x, y)} r(\phi; x, y) \, d\omega \, d\phi. \]  

(3)

We carried out the above reconstruction for each polarization channel. Thus, we have four images \( g_{hh}(x, y) \), \( g_{hv}(x, y) \), \( g_{vh}(x, y) \), and \( g_{vv}(x, y) \).

At this point, we can employ the several polarimetric decomposition algorithms for the spatial PolCSAR image. The direct approach of the polarimetric decomposition for the PolCSAR images is just applying it to the reconstructed spatial images as in the conventional polarimetric linear SAR image processing [2, 3]. The simplest polarimetric decomposition is the classical Pauli decomposition defined below.

\[ \alpha(x, y) = [g_{hh}(x, y) + g_{vv}(x, y)]/\sqrt{2}, \]  

(4a)

\[ \beta(x, y) = [g_{hh}(x, y) - g_{vv}(x, y)]/\sqrt{2}, \]  

(4b)

\[ \gamma(x, y) = \sqrt{2}g_{hv}(x, y), \]  

(4c)

where \( \alpha \), \( \beta \), and \( \gamma \) can be interpreted as the surface (odd-bounce), double-bounce (even-bounce), and volume (random) scattering components, respectively. However, since the PolCSAR images contain a mixture of the target signatures viewed from various radar azimuth angles, it is more useful to decompose the CSAR image into the angular components to clearly interpret the decomposition results.

We can accomplish this by transforming the CSAR images into their spatial frequency domain images using the two-dimensional Fourier transform with an appropriate spatial window [1, 3]. The spectral domain polarimetric decomposition is just the application of the conventional decomposition algorithms to the spectral domain PolCSAR images. The combination of the windowed two-dimensional Fourier transform and the polarimetric decomposition algorithms provides the information about the angular dependence and the polarimetric scattering mechanisms of the desired target in the PolCSAR images [3].
To convert the spatial PolCSAR images into the corresponding spectral domain images, we compute the following two-dimensional windowed Fourier transform:

\[
G_{qp}(k_x, k_y) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} w(x, y) g_{qp}(x, y) e^{-j k_x x} e^{-j k_y y} \, dx \, dy, \tag{5}
\]

where \(w(x, y)\) is a window function employed to extract the target response of interest from the image. The spectral domain image can be interpreted as a function of the azimuth angle; the spatial frequency domain \((k_x, k_y)\) is related to the azimuth angle \(\phi\) via

\[
k_x(\omega, \phi; x, y) = \frac{2\omega}{c} \cos \theta_l(\phi; x, y) \cos \phi_l(\phi; x, y), \tag{6a}
\]
\[
k_y(\omega, \phi; x, y) = \frac{2\omega}{c} \cos \theta_l(\phi; x, y) \sin \phi_l(\phi; x, y), \tag{6b}
\]

where \(\phi_l\) and \(\theta_l\) are the following local azimuth angle and the local grazing angle at the location \((x, y)\), respectively.

\[
\phi_l(\phi; x, y) = \tan^{-1}(R_g \sin \phi - y, R_g \cos \phi - x), \tag{7a}
\]
\[
\theta_l(\phi; x, y) = \tan^{-1} \left[ \frac{Z_c}{\sqrt{(x - R_g \cos \phi)^2 + (y - R_g \sin \phi)^2}} \right], \tag{7b}
\]

where \(\tan^{-1}(y, x)\) is the four-quadrant inverse tangent.

We can realize the spectral domain polarimetric decomposition by applying the polarimetric decomposition algorithms to the spectral domain PolCSAR images \(G_{qp}(k_x, k_y)\). As in the spatial domain case, we employ the following Pauli decomposition.

\[
A(k_x, k_y) = [G_{hh}(k_x, k_y) + G_{vv}(k_x, k_y)]/\sqrt{2}, \tag{8a}
\]
\[
B(k_x, k_y) = [G_{hh}(k_x, k_y) - G_{vv}(k_x, k_y)]/\sqrt{2}, \tag{8b}
\]
\[
\Gamma(k_x, k_y) = \sqrt{2} G_{hv}(k_x, k_y), \tag{8c}
\]

where again, \(A\), \(B\), and \(\Gamma\) can be interpreted as the surface, double-bounce, and volume scattering component, respectively. Using the relationships defined in Eqs. (6) and (7), we can estimate the scattering mechanisms as a function of the angular frequency \(\omega\) and the azimuth angle \(\phi\) (or the local azimuth angle \(\phi_l\)).

### 3 Airborne PolCSAR experiment

In this section, we present the result from real airborne PolCSAR dataset. This dataset is publicly released by AFRL and the detail of the experiment can be found in [4]. The imaged scene contains an area around a parking lot where several commercial vehicles and simple targets (i.e., corner reflectors and a top hat reflector) are placed. Table I lists the experimental parameters derived from the dataset.

Figures 2(a) and (b) show the color composite images of the spatial Pauli decomposition, where blue, red, and green represent surface (\(|\alpha(x, y)|^2\)), double-bounce (\(|\beta(x, y)|^2\)), and volume (\(|\gamma(x, y)|^2\)) scattering components, respectively. Figure 2(a) shows the area around the simple targets (i.e., corner reflectors and a large top hat.
Table I. Experimental parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
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<tbody>
<tr>
<td>Center frequency [GHz]</td>
<td>9.6 (X-band)</td>
</tr>
<tr>
<td>Bandwidth [MHz]</td>
<td>622</td>
</tr>
<tr>
<td>Azimuth angle $\phi$ [deg]</td>
<td>$[0, 360]$</td>
</tr>
<tr>
<td>Number of azimuth angle samples</td>
<td>42,561</td>
</tr>
<tr>
<td>Radar altitude (average) $Z_c$ [m]</td>
<td>7,117</td>
</tr>
<tr>
<td>Radius of trajectory (average) $R_g$ [m]</td>
<td>7,271</td>
</tr>
<tr>
<td>Depression angle (average) $\theta$ [deg]</td>
<td>44.4</td>
</tr>
<tr>
<td>Polarization</td>
<td>$hh, hv, vh, vv$</td>
</tr>
</tbody>
</table>

Fig. 2. Pauli decomposed images.

(a) Spatial image (simple targets)
(b) Spatial image (simple targets)
(c) Spectral image (triangular)
(d) Spectral image (passenger vehicle)
(e) Averaged power (triangular)
(f) Averaged power (passenger vehicle)

Fig. 2. Pauli decomposed images.

reflector), and Fig. 2(b) shows the vehicles (i.e., passenger vehicles and a forklift) in the parking lot. The interpretation of the spatial Pauli decomposed image is not straightforward as compared to the conventional linear polarimetric SAR image, because the targets in the PolCSAR image are illuminated by the sensor at the various azimuth angles; the azimuth angle at which a certain scattering component arose is unclear from the spatial Pauli image itself.

Figures 2(c) and (d) show the spectral domain Pauli decomposition for the
triangular reflector and a passenger vehicle, respectively. We enclosed the selected test targets by the dotted rectangles in the spatial image of Figs. 2(a) and (b), and these rectangles also represent the size of the rectangular windows $w(x, y)$, which are applied to generate the spectral domain images of Figs. 2(c) and (d). In Figs. 2(c) and (d), blue, red, and green represent surface ($|A(k_x, k_y)|^2$), double-bounce ($|B(k_x, k_y)|^2$), and volume ($|\Gamma(k_x, k_y)|^2$) scattering components, respectively. We evaluate the mean of the frequency components shown in Figs. 2(c) and (d) at each azimuth angle.

From Figs. 2(c) and (e), one can recognize the dominant surface (or odd-bounce) scattering around the boresight direction of the triangular reflector. In addition, the double-bounce responses appear around the sides of the boresight response. This is because the triangular behaves like a dihedral when only two facets out of the three facets can be observable from the sensor.

We focus on the case of the passenger vehicle shown in Figs. 2(d) and (f). If an incident wave arrives at the front of the car (i.e., $-90^\circ$), the strong surface scattering from the engine hood should occur. Furthermore, the double-bounce scattering between the ground and the front bumper also exists. A similar scattering process can occur for the back of the car (i.e., $90^\circ$); however, the surface scattering becomes weaker than that of the front case. Moreover, if an incident wave impinges on the side of the vehicle (i.e., $0^\circ$ and $180^\circ$), the surface and the double-bounce scattering becomes stronger than those of the front and the back direction.

The above experimental results show that, for real airborne PolCSAR image, we can clearly understand the polarimetric scattering characteristics with respect to the frequency and the azimuth angle via the spectral domain polarimetric decomposition.

4 Conclusion

In this letter, we have demonstrated the spectral domain target decomposition concept with the real airborne PolCSAR dataset. Using this concept, we can retrieve the polarimetric scattering characteristics of the target with respect to the azimuth angle.

Acknowledgments

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Load impedance perturbation formulas for class-E power amplifiers

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Abstract: This paper considers how a class-E amplifier behaves when the load deviates from its nominal impedance. The Kirchhoff’s law and Fourier series formulate the circuit into three simultaneous equations. These equations are decomposed into zeroth- and first-order terms to apply the perturbation technique. The zeroth-order solutions meet the zero-voltage and zero-current switching conditions. The first-order formulas express the DC current consumption, RF output amplitude and phase, equivalent output impedance, and periodic turn-on power dissipation in terms of the load resistance and reactance deviations.

Keywords: zero-voltage switching, zero-current switching, turn-on capacitor voltage, load deviation, equivalent output impedance, Poincaré distance

Classification: Transmission Systems and Transmission Equipment for Communications

References


1 Introduction

In radio communication, broadcasting, and wireless power transfer systems, an RF power amplifier is indispensable. Sokal presented a switch-mode but sinusoidal-...
wave output amplification scheme named class E. This scheme effectively works to nullify the periodic turn-on power loss and to achieve a high conversion efficiency [1, 2]. However, it often suffers from load-impedance deviations because the class-E operation is carried out only when it is precisely loaded with its nominal impedance. To address this problem, Suetsugu et alia performed deliberate considerations and experiments on how a circuit behaves in case the load is offset from the nominal impedance [3, 4, 5]. Because the class-E operation involves periodic discontinuity and reactive resonance, the formulation becomes quite tough, particularly in off-nominal situations. To find elegant explicit expressions on the load deviation effects, this paper applies a perturbation technique to class-E circuit equations.

2 Circuit equations

We consider the class-E power amplifier shown in Fig. 1, where DC power is input at the left-hand port, goes through a choke coil, is converted into an RF power by a transistor, is filtered by an LC resonator, and finally outputs a sinusoidal wave at the right-hand port. The transistor is assumed to function as an ideal on-off switch exclusively controlled by the base-input signal regardless of the instantaneous collector-to-emitter voltage or even its polarity. The switch turns off at \( t = 0 \), turns on at \( t = T/2 \), and repeats this sequence for time period \( T \). The series LC resonator forces the output current to assume a sinusoidal waveform.

\[
i_{RF}(t) = I_{RF} \sin(\omega t + \theta), \quad i(t) = I_{DC} - I_{RF} \sin(\omega t + \theta) \tag{1}
\]

where \( I_{DC}, I_{RF} > 0 \) and \( \omega = 2\pi/T \). Immediately after the switch turns off at \( t = 0 \), the current \( i(t) \) starts to charge the capacitor \( C \), resulting in capacitor voltage

\[
v(t) = \frac{1}{C} \int_0^t i(t) \, dt = \frac{I_{DC}}{C} t + \frac{I_{RF}}{\omega C} \{\cos(\omega t + \theta) - \cos \theta\} \tag{2}
\]

in the OFF state until the switch turns on at \( t = T/2 \). The voltage is kept at zero during the ON state because the capacitor is short-circuited by the transistor.

Because this on-off sequence is periodically repeated, the voltage waveform is a periodic function of time, which can be expanded into a Fourier series

\[
v(t) = V_{DC} + RI_{RF} \sin(\omega t + \theta) + XI_{RF} \cos(\omega t + \theta) + \cdots \tag{3}
\]
where . . . implies higher-order harmonic terms. Coefficient $X$ denotes the series LC resonator’s reactance: $X = \omega L_o - 1/\omega C_o$. According to the Fourier series theory, the DC term in (3) calculates

$$V_{DC} = \frac{1}{T} \int_0^{T/2} v(t) \, dt$$

(4)

The fundamental-harmonic coefficients in (3) calculate

$$RI_{RF} = \frac{2}{T} \int_0^{T/2} v(t) \sin(\omega t + \theta) \, dt, \quad XI_{RF} = \frac{2}{T} \int_0^{T/2} v(t) \cos(\omega t + \theta) \, dt$$

(5)

Note that the integrals are all truncated halfway because $v(t)$ vanishes right after the switch turns on at $t = T/2$. Substituting (3) into (4) and (5) yields

$$\pi^2 I_{DC} - 2(\sin \theta + \pi \cos \theta)I_{RF} = 4\pi \omega CV_{DC}$$

(6)

$$(\pi \cos \theta - 2 \sin \theta)I_{DC} - (2 \cos^2 \theta + \pi \omega C)I_{RF} = 0$$

(7)

$$2(2 \cos \theta + \pi \sin \theta)I_{DC} - (\pi + 2 \sin 2\theta - 2\pi \omega CX)I_{RF} = 0$$

(8)

These equations dominate the class-E amplifier’s behavior. They hold true even if the circuit does not satisfy the soft-switching condition. The problem is that they involve manifold trigonometry, and are thus too difficult to be solved at once for three unknowns $I_{DC}$, $I_{RF}$, and $\theta$.

3 Decomposition for perturbation

To solve the above three circuit equations in an analytical manner, we employ a perturbation technique. All the functions and variables are decomposed into their zeroth- and first-order terms as

$$v(t) = v_0(t) + v_1(t), \quad i(t) = i_0(t) + i_1(t)$$

(9)

$$I_{DC} = I_{DC0} + I_{DC1}, \quad I_{RF} = I_{RF0} + I_{RF1}, \quad \theta = \theta_0 + \theta_1$$

(10)

$$Z = Z_0 + Z_1 = R + jX = (R_0 + R_1) + j(X_0 + X_1)$$

(11)

where subscript “0” implies a nominal class-E operation, and subscript “1” indicates a possible deviation from the nominal operation. Each deviation may be upward or downward, but it is assumed to be much smaller than the original magnitude. Considering practical system applications, we should be able to keep $\omega$, $C$, and $V_{DC}$ constant.

By applying the above-mentioned decompositions to (1), (2), (6), (7), and (8), we extract their zeroth- and first-order terms as follows:

$$i_0(t) = I_{DC0} - I_{RF0} \sin(\omega t + \theta_0)$$

(12)

$$i_1(t) = I_{DC1} - I_{RF1} \sin(\omega t + \theta_0) - I_{RF0} \theta_1 \cos(\omega t + \theta_0)$$

(13)

$$v_0(t) = \frac{I_{DC0}}{C} t + \frac{I_{RF0}}{\omega C} \{\cos(\omega t + \theta_0) - \cos \theta_0\}$$

(14)

$$v_1(t) = \frac{I_{DC1}}{C} t + \frac{I_{RF1}}{\omega C} \{\cos(\omega t + \theta_0) - \cos \theta_0\} - \frac{I_{RF0}}{\omega C} \theta_1 \{\sin(\omega t + \theta_0) - \sin \theta_0\}$$

(15)
\[
\begin{align*}
\pi^2 I_{DC0} - 2(2 \sin \theta_0 + \pi \cos \theta_0)I_{RF0} &= 4\pi \omega CV_{DC} \\
\pi^2 I_{DC1} - 2(2 \sin \theta_0 + \pi \cos \theta_0)I_{RF1} &= 2\theta_1(2 \cos \theta_0 - \pi \sin \theta_0)I_{RF0} \\
(\pi \cos \theta_0 - 2 \sin \theta_0)I_{DC0} - (2 \cos^2 \theta_0 + \pi \omega C \theta_0)I_{RF0} &= 0 \\
(\pi \cos \theta_0 - 2 \sin \theta_0)I_{DC1} - (2 \cos^2 \theta_0 + \pi \omega C \theta_0)I_{RF1} &= \theta_1(\pi \sin \theta_0 + 2 \cos \theta_0)I_{DC0} - (2\theta_1 \sin 2\theta_0 - \pi \omega C \theta_1)I_{RF0} \\
2(2 \cos \theta_0 + \pi \sin \theta_0)I_{DC0} - (\pi + 2 \sin 2\theta_0 - 2\pi \omega C \theta_0)I_{RF0} &= 0 \\
2(2 \cos \theta_0 + \pi \sin \theta_0)I_{DC1} - (\pi + 2 \sin 2\theta_0 - 2\pi \omega C \theta_0)I_{RF1} &= 2\theta_1(2 \sin \theta_0 - \pi \cos \theta_0)I_{DC0} + 2(\theta_1 \cos 2\theta_0 - \pi \omega C \theta_1)I_{RF0}
\end{align*}
\]

4 Zeroth-order solutions

Imposing the zero-current switching (ZCS) and zero-voltage switching (ZVS) conditions upon (12) and (14) at \( t = T/2 \), we obtain

\[
I_{DC0} + I_{RF0} \sin \theta_0 = 0, \quad \pi I_{DC0} - 2I_{RF0} \cos \theta_0 = 0
\]

Feeding these relationships back into (18) and (20) and then eliminating \( \theta \) from the resulting equations yields

\[
(\omega CR_0)^2 + \left( \omega CX_0 + \frac{2}{\pi^2} - \frac{1}{2} \right)^2 = \left( \frac{2}{\pi^2} \right)^2
\]

\[
(\omega CR_0)^2 + (\omega CX_0 - 1)^2 = \frac{4}{\pi^2} + \frac{1}{4}
\]

Impedance \( Z_0 = R_0 + jX_0 \) that satisfies these equations plots circular arcs marked “ZCS” and “ZVS” as shown in Fig. 2. These arcs are called geodesic lines in hyperbolic geometry. They intersect with each other at

\[
\omega CR_0 = \frac{1}{\pi} \cdot \frac{8}{\pi^2 + 4} = 0.184, \quad \omega CX_0 = \frac{1}{2} \cdot \frac{\pi^2 - 4}{\pi^2 + 4} = 0.212
\]

This point specifies the nominal load for class-E operation. Although \( R_0 \) and \( X_0 \) are normalized with \( \omega C \), their ratio becomes a dimension-free constant

Fig. 2. Load impedance loci for class-E operation: \( 1/\omega C = 50 \Omega \).
This is called the series-resonator loaded $Q$. The constant-$Q$ contour is plotted and shown as a broken arc in Fig. 2, along which the ZVS-ZCS intersection moves according to $\omega C$. From (12), (14), (16), and (22), the zeroth-order variables and waveforms are found as

\[
I_{DC0} = \pi \omega CV_{DC}, \quad I_{RF0} = \frac{1}{2} \sqrt{\pi^2 + 4 I_{DC0}}, \quad \tan \theta_0 = -\frac{2}{\pi}
\]

\[
i_0(t) = \left(1 + \cos \omega t - \frac{\pi}{2} \sin \omega t\right) I_{DC0}
\]

\[
v_0(t) = \pi \left(\omega t + \sin \omega t + \frac{\pi}{2} \cos \omega t - \frac{\pi}{2}\right) V_{DC}
\]

### 5 First-order solutions

Once the above zeroth-order solutions are known, we can regard (17), (19), and (21) as equations only for first-order unknowns $I_{DC1}$, $I_{RF1}$, and $\theta_1$. Because the equations are all linear with respect to the three unknowns, we analytically solve them as

\[
\frac{I_{DC1}}{I_{DC0}} = -\frac{\pi^2 - 4}{\pi^2 + 4} \left(\frac{R_1}{R_0} + \frac{\pi^2 X_1}{4 X_0}\right) = -0.423 \frac{R_1}{R_0} - 1.04 \frac{X_1}{X_0}
\]

\[
\frac{I_{RF1}}{I_{RF0}} = -\frac{\pi^2}{\pi^2 + 4} \left(\frac{R_1}{R_0} + \frac{\pi^2 - 4 X_1}{8 X_0}\right) = -0.712 \frac{R_1}{R_0} - 0.522 \frac{X_1}{X_0}
\]

\[
\theta_1 = -\frac{\pi^2 - 4 X_1}{32} \frac{X_0}{R_0} = -0.576 \frac{X_1}{X_0} [\text{rad}] = -33.0 \frac{X_1}{X_0} [\text{deg}]
\]

These results appear simple, but are so informative that we can predict the amplifier’s basic behavior by focusing on its DC input and RF output currents with respect to the load impedance deviation. The above three formulas tell us some significant points.

The DC current consumption (30) increases if the load resistance or reactance decreases. Note that the load-reactance deviation is identical to the resonator-reactance deviation because they are connected in series as shown in Fig. 1. This information is useful for consideration of the DC power supply as well as the heat generation due to the transistor’s on-state resistance. The RF output current magnitude (31) behaves in a manner similar to that in (30). In contrast, the RF output phase (32) is not affected by the resistance deviation, but is merely delayed by the reactance increment. In other words, the amplifier’s equivalent output impedance $Z_{out}$ must be real. Therefore, we can simply formulate $Z_{out}$ as follows.

The Ohm’s law on the load enables us to decompose the RF output voltage $V_{RF}$ into

\[V_{RF0} = R_0 I_{RF0}, \quad V_{RF1} = I_{RF1} R_0 + I_{RF0} R_1\]

Applying (31) to this decomposition, the output impedance is expressed as

\[
Z_{out} = \frac{V_{RF1}}{-I_{RF1}} = -R_0 - \frac{I_{RF0}}{I_{RF1}} R_1 = \frac{4}{\pi^2} R_0 = 0.405 R_0
\]

where the first $I_{RF1}$ has a negative sign because the output impedance should be observed backward from the load.
Above $Z_{out}$ is particularly crucial when we consider the load deviation as a standing-wave problem. Imagine, for any reason, part of the transmitted power is coming back as a wave from the load to the amplifier. This returning wave, in turn, detects above $Z_{out}$ at the amplifier’s output port as shown in Fig. 3. Because $Z_{out}$ differs from $R_0$, the wave reflects back to the load again.

By considering this phenomenon from a wave-engineering viewpoint, we deduce that amplifier output reflectance $\Gamma$ (or $S_{22}$), standing-wave ratio $\rho$, and Poincaré distance $D$ between $Z_{out}$ and $R_0$ as

$$\Gamma = \frac{Z_{out} - R_0}{Z_{out} + R_0} = \frac{4 - \pi^2}{4 + \pi^2}, \quad \rho = \frac{1 + |\Gamma|}{1 - |\Gamma|} = \frac{\pi^2}{4} = 2.47$$

$$D = \ln \rho = 2(\ln \pi - \ln 2) = 0.903$$

This result is apparently different from the typical behavior of linear RF circuits, from which we expect that $Z_{out}$ matches the nominal load $R_0$, and hence $D$ becomes zero. This is exactly a special feature of the class-E operation.

One final note we can find out from the aforementioned first-order waveform is the shunt capacitor’s residual voltage at the moment of turn-on. Applying (30) and (31) to (15), we reach

$$v_1(T/2) = \frac{\pi}{2} \left( \pi R_1 - \frac{\pi^2 - 4}{4} X_1 \right) I_{DC0} = (4.93 R_1 - 2.30 X_1) I_{DC0}$$

This undesired voltage remains due to the combination of $R_1$ and $X_1$, namely the complex load deviation, which would soft-land onto zero volt if the amplifier were loaded with its nominal impedance given by the zeroth-order solution in (25) or at the intersection shown in Fig. 2. Given $I_{DC0}$ in (27), it is so simple for (36) to estimate the turn-on excess power dissipation in the transistor using the well-known formula $P_{sw} = \omega C v_1^2(T/2)/4\pi$.

6 Conclusion

A perturbation approach has successfully derived easy-to-use formulas for class-E power amplifiers. The zeroth-order solution locates the nominal load impedance at the intersection of the ZVS and ZCS geodesic arcs projected onto a Smith chart. The amplifier’s equivalent output impedance is found out to be 40% of the nominal load resistance, which is clearly distinct from that of linear RF circuits used with a matched load. The shunt capacitor’s residual turn-on voltage formula enables us to quickly predict the transistor’s excess heat generation from both resistance and reactance deviations. The formulas presented in this paper provide an intuitive and
lucid view on the class-E operation, which will offer a constructive bridge between power-electronics engineers and radio-wave engineers, through which we can gather momentum toward future works on high-power high-frequency transmission system development.

Acknowledgments

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Effect of early stopping on error performance of iterative MIMO equalization: An experience in reality

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Abstract: Terminal-collaborated multiple-input multiple-output (MIMO) reception with adaptive terminal selection schemes has been studied experimentally. In the experimental system, frequency-domain iterative MIMO equalization was employed, and its performance was evaluated by making use of recorded signal waveforms in a measurement campaign. In iterative signal processing, it is known that early stopping can control unnecessary iterations. In this letter, it is revealed that early stopping can not only reduce computational complexity, but also improve the bit error ratio performance of iterative processing. This is because early stopping prevents the iteration process from causing errors.

Keywords: collaborative reception, distributed MIMO, terminal selection, early stopping, iterative equalization, measurement campaign

Classification: Wireless Communication Technologies

References

1 Introduction

Terminal-collaborated multiple-input multiple-output (MIMO) reception has been studied [1, 2]. This is a form of distributed MIMO, in which a virtual terminal with a large number of reception antennas receives multiple data streams from a base station (BS). This virtual terminal consists of multiple mobile terminals (MSs) that are close to each other in order to share their received signals from the BS with other MSs. Terminal-collaborated MIMO reception does not require precoding, thereby is expected to achieve better performance in mobile environments than that of multi-user MIMO transmission with precoding.

The performance of this system can be improved with the increase of the number of collaborated MSs. However, the amount of power consumption and traffic overhead for inter-terminal collaboration will also be increased. In order to reduce these overhead, adaptive MS selection schemes that select an appropriate set of collaborated MSs have been proposed [3, 4, 5].

In this letter, early stopping (ES) [6] is applied to frequency-domain iterative MIMO equalization for this system. The only expected benefit of ES is reduction of computational complexity. However, it is revealed by a measurement campaign that ES can offer better bit error ratio (BER) performance.

2 System description

A BS transmits $M$ spatially-multiplexed independent signal streams to $N$ MSs on the same carrier frequency at the same time. On the receiver side, each MS equipped with a single antenna shares the received signals with other MSs. $L$ MSs out of $N$ candidate MSs are selected for terminal-collaborated MIMO reception. Let $\mathcal{L} \subseteq \mathcal{N}$ denote the set of selected MSs where $\mathcal{N}$ is the set of the candidate MSs.

2.1 Frequency-domain iterative equalization

Frequency-domain (FD) iterative equalization is applied to received signals [7]. This scheme combines three processes: i) MMSE frequency-domain equalization, ii) soft decoding of low-density parity-check code (LDPC) by belief propagation (BP), and iii) soft cancellation.

Let $\mathbf{y}_L(k) = [y_1(k), y_2(k), \ldots, y_L(k)]^T \in \mathbb{C}^{L \times 1}$ where $y_l(k)$ is the received signal at the $k$th symbol of $l$th MS in $\mathcal{L}$. The frequency-domain received signals...
\( Y_L(f) \in \mathbb{C}^{L \times 1} \) are equalized by a minimum mean square error (MMSE) filter. The equalized signals \( \hat{X}(f) \in \mathbb{C}^{M \times 1} \) are converted to the signals \( \hat{x}(k) \in \mathbb{C}^{M \times 1} \) in TD. Then, BP decoding calculates log likelihood ratios (LLRs) \( L(c_{m,k,i}) \) where \( c_{m,k,i} \) is the \( i \)th bit of the \( k \)th symbol of the \( m \)th transmitted stream. In an FD soft replica generator, soft-decision symbols \( \hat{x}(k) = [\hat{x}_1(k), \hat{x}_2(k), \ldots, \hat{x}_M(k)]^T \in \mathbb{C}^{M \times 1} \) are generated as follows in the case of quadrature phase-shift keying (QPSK) modulation:

\[
\hat{x}_m(k) = \frac{1}{\sqrt{2}} \left( \tanh \left( L(c_{m,k,1})/2 \right) + \tanh \left( L(c_{m,k,2})/2 \right) \right). \tag{1}
\]

The symbols \( \hat{x}(k) \) are converted to signals \( \hat{X}(f) = [\hat{X}_1(f), \hat{X}_2(f), \ldots, \hat{X}_M(f)]^T \in \mathbb{C}^{M \times 1} \) in FD. Next, soft-decision replicas \( \hat{Y}_{L,m}(f) \in \mathbb{C}^{L \times 1} \) are generated as follows:

\[
\hat{Y}_{L,m}(f) = g_{L,m}(f) \hat{x}_m(f), \tag{2}
\]

where \( g_{L,m}(f) \in \mathbb{C}^{L \times 1} \) is a channel transfer function.

Equalized signals by an MMSE filter with soft-cancellation can be expressed as

\[
\hat{X}_m(f) = w_{L,m}^H(f) \left( Y_L(f) - \sum_{i \neq m} \hat{Y}_{L,i}(f) \right), \tag{3}
\]

where \( w_{L,m}(f) \in \mathbb{C}^{L \times 1} \) is an MMSE filter with a priori information. This filter utilizes residual error coefficients \( \beta_m (0 \leq \beta_m \leq 1) \) shown below [7]

\[
\beta_m = \begin{cases} 
0, & \text{all the parity-check equations are satisfied} \\
1 - \frac{1}{K} \sum_k |\hat{x}_m(k)|^2, & \text{otherwise,}
\end{cases} \tag{4}
\]

where \( K \) is the number of data symbols. As shown in this equation, if all the parity-check equations of the \( m \)th stream are satisfied for hard decisions formed on the a posteriori LLRs, let \( \beta_m \) be 0. These three processes are repeated up to \( Q \) times as an outer loop.

### 2.2 MS selection

Three MS selection schemes are considered. In a maximum product of singular values (MPoSV) MS selection scheme [3], \( L \) MSs are selected frame by frame based on the singular values of the estimated channel matrix. The following two schemes assume BER information available at the receiver. Therefore, their performance is studied for comparison purpose. In a fixed MS selection scheme (denoted as Fixed), \( L \) MSs are selected and remain unchanged. We focus on the best selection pattern \( L_q^* \) in terms of average BER, which can be given by

\[
L_q^* = \arg \min_{L \subseteq N} \sum_j p(j, L, q), \tag{5}
\]

where \( p(j, L, q) \) is the BER averaged over all streams at the \( q \)th outer iteration in the \( j \)th frame. In a perfect MS selection scheme (denoted as Perfect), the best \( L \) MSs are selected frame by frame among \( \binom{N}{L} \) selection patterns in terms of BER. The selected MSs can be given by

\[
L_{j,q}^* = \arg \min_{L \subseteq N} p(j, L, q). \tag{6}
\]
2.3 Stopping criterion

The residual error coefficients shown in Eq. (4) are employed as a stopping criterion of ES. The outer loop is repeated until the following inequality holds or until the maximum number of outer iterations is reached.

\[ \sum_{m=1}^{M} \beta_m \leq \varepsilon \] (7)

By adjusting the iteration control threshold \( \varepsilon \), the number of iterations \( q \) can be reduced. The BER performance with ES is denoted as \( p_{ES}(j, L, Q) \), where \( Q \) is the maximum number of iterations.

In Fixed with ES, the BER \( p_{ES}(j, L, Q) \) is used instead of \( p(j, L, q) \) in Eq. (5). In Perfect with ES, we jointly optimize \( L \) and \( q \) as follows:

\[ L^q_{j, Q} = \arg \min_{L \leq N, 1 \leq q \leq Q} p(j, L, q). \] (8)

3 Experimental setup

Four transmit antennas \( (M = 4) \) were arranged in a 3.8 m \( \times \) 2.5 m square-shape as shown in Fig. 1 (a). Each transmit antenna was a horizontal-plane omnidirectional vertical antenna (5.8 dBi) and mounted on the roof of a building at 25.5 m above the ground. The BS transmitted spatially multiplexed packets by QPSK modulation every 50 ms frame. The transmit power was 1 W per antenna, the carrier frequency was 427.2 MHz, and the symbol rate was 312.5 kilo symbols per second.

Six receive antennas \( (|N| = N = 6) \) were arranged in a uniform circular array as shown in Fig. 1 (b). Each receive antenna was a 2.15 dBi quarter-wave and horizontal-plane omnidirectional monopole antenna and mounted on the roof of a vehicle (2.1 m height). A subset \( L \) (\( |L| = L = 4 \)) was selected from six received signals and used for equalization/demodulation. In order to examine BERs of all
possible signal combinations, received signal waveforms from the BS were recorded at each MS and used for offline processing. Therefore, there was no inter-MS communication for collaboration in this letter.

The packet included a synchronization word of 15 symbols, a training sequence of 39 symbols, a control word of 15 symbols, a cyclic prefix of four symbols, and a data sequence of 192 symbols. Timings and frequencies of the entire system were based on 1-pulse-per-second signals and 10 MHz signals of global positioning system receivers.

We drove the vehicle twice along Shirakawa-dori Street, Sakyo-ku, Kyoto, heading north as shown in Fig. 1 (c). In this driving course, the average received power was greater than −80 dBm. The received signals while the vehicle stopped at traffic lights were not used in offline processing.

The noise variance for the MMSE filter and the LPDC decoder was optimized according to the average BER without ES. The number of iterations of the inner loop (BP) and the outer loop were eight and \( Q = 3 \), respectively.

### 4 Experimental results

The BER performance of iterative MIMO equalization with and without ES was evaluated by making use of the recorded signals. The iteration control threshold \( \epsilon \) for ES was determined based on the average BER over the entire course.

Empirical cumulative distribution functions (CDF) of BER averaged over four streams in a frame are shown in Fig. 2. Note that absolute values of empirical CDFs are different between two trials due to different traffic conditions (e.g. vehicle speed, lane position, other vehicles). In Fixed, the best performance among those of 15 selection patterns is shown.

The BER performance of all three MS selection schemes was improved by ES as shown in Fig. 2. These improvements were confirmed in both of the two trials. Table I shows the average number of iterations and the average BER performance. As shown in this table, the BER performance without ES was improved by increasing \( q \). Interestingly, ES can offer better average BERs with much smaller average \( q \). This is because ES can control unnecessary iterations leading to catastrophic error propagation.

![Fig. 2. Empirical CDF of BER. (a) First trial. (b) Second trial.](image-url)
Table I. Comparisons of average number of iterations and BER performance.

(a) First trial.

<table>
<thead>
<tr>
<th>Scheme</th>
<th>$q = 1$</th>
<th>$q = 2$</th>
<th>$q = 3$</th>
<th>ES</th>
</tr>
</thead>
<tbody>
<tr>
<td>Perfect</td>
<td>Avg. $q$</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td></td>
<td>Avg. BER</td>
<td>$1.4 \times 10^{-4}$</td>
<td>$6.1 \times 10^{-5}$</td>
<td>$4.4 \times 10^{-5}$</td>
</tr>
<tr>
<td>MPoSV</td>
<td>Avg. $q$</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td></td>
<td>Avg. BER</td>
<td>$8.5 \times 10^{-4}$</td>
<td>$3.2 \times 10^{-4}$</td>
<td>$2.5 \times 10^{-4}$</td>
</tr>
<tr>
<td>Fixed</td>
<td>Avg. $q$</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>(Best)</td>
<td>Avg. BER</td>
<td>$3.4 \times 10^{-3}$</td>
<td>$3.4 \times 10^{-4}$</td>
<td>$2.7 \times 10^{-4}$</td>
</tr>
</tbody>
</table>

(b) Second trial.

<table>
<thead>
<tr>
<th>Scheme</th>
<th>$q = 1$</th>
<th>$q = 2$</th>
<th>$q = 3$</th>
<th>ES</th>
</tr>
</thead>
<tbody>
<tr>
<td>Perfect</td>
<td>Avg. $q$</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td></td>
<td>Avg. BER</td>
<td>$3.1 \times 10^{-4}$</td>
<td>$1.2 \times 10^{-4}$</td>
<td>$9.4 \times 10^{-5}$</td>
</tr>
<tr>
<td>MPoSV</td>
<td>Avg. $q$</td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td></td>
<td>Avg. BER</td>
<td>$1.3 \times 10^{-3}$</td>
<td>$6.3 \times 10^{-4}$</td>
<td>$5.1 \times 10^{-4}$</td>
</tr>
<tr>
<td>Fixed</td>
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<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>(Best)</td>
<td>Avg. BER</td>
<td>$5.5 \times 10^{-3}$</td>
<td>$7.5 \times 10^{-4}$</td>
<td>$6.2 \times 10^{-4}$</td>
</tr>
</tbody>
</table>

ES by parity-check equations (not $\beta_m$) can offer almost the same but slightly degraded BER performance and increased average $q$ in this measurement campaign. Note that ES can reduce the errors that cannot be avoided even by Perfect.

5 Conclusion

This letter has presented an effect of ES on the error performance of iterative MIMO equalization. By using ES, the average number of iterations was reduced significantly. Moreover, it is shown that the BER performance was also improved. This is beyond our expectations, and currently under investigation. The results thus far suggested that a possible cause was phase rotation in a packet.

Acknowledgments

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Proposal for optical TDM system for digital payload in next generation high throughput satellites

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Abstract: The digital payload for next generation high throughput communication satellites is expected to reduce the operators’ CAPEX and OPEX. Photonic technologies, which have a great affinity with digital signals, have been studied in order to increase the data throughput and reduce the mass of the satellites. Here, we propose an optical time division multiplexing system for routing the signals in the digital channelizer and have conducted a feasibility study on the transmission efficiency. This system has the potential to reduce the cost, mass, and power consumption, and to improve reliability by cleaning up the switches used for routing the signals within the satellites.

Keywords: optical time division multiplexing, digital payload, digital channelizer, matrix switch, high throughput satellite

Classification: Satellite Communications

References


1 Introduction

Next generation high throughput satellites (HTS) are expected to reduce the cost per bit for the operators and to improve the flexibility in order to make it possible to support changes in existing services, and to provide new services on a timely basis. Digital signal processing in the payload using application specific integrated circuits (ASIC) and field-programmable gate arrays (FPGA) is a candidate for realizing both high throughput and flexibility, and much work on this has been reported [1, 2, 3]. However, the power consumption and heat generation of the digital signal processors, especially those used for the signal routing function in the channelizer, have become a critical problem as the number of antennas (links) and the data throughput of the payload increase [4].

Photonic technologies, which have a great affinity with digital signals, have the potential to increase the data throughput, to reduce the size, mass, and power consumption, and to improve the immunity to EMI of the satellites. Therefore, many studies have been reported in recent years; the feasibility of a photonic payload was introduced in [5], 25 Gbps (4 channels x 6.25 Gbps) optical modules were developed using high speed ADCs and DACs [6], and an optical switch with 24 x 24 ports was developed and demonstrated for signal routing in the channelizer [7]. Here, we propose an optical time division multiplexing (TDM) system for routing the signals in the digital channelizer. Optical TDM is a well known and mature technology used for passive optical network (PON) systems, which reduce the cost by sharing the optical fiber among multiple users in terrestrial telecommunications [8]. Streamlining the routing in the digital channelizer by adopting the proposed system will make it possible to reduce the cost, size, mass, and power consumption, and to improve the reliability of the satellites.

2 Digital payload in communication satellites

A digital payload configuration based on a multi-beam antenna (MBA) system as introduced in [1] is shown in Fig. 1(i). The uplink signals from the MBA are amplified by the low noise amplifiers (LNA), downconverted (D/C) to a suitable intermediate frequency (IF), converted to digital form by the A/D converters (ADC), demultiplexed, and then routed to any desired downlink direction by the routing switch. The downlink sub-channels as rearranged in the routing switch are remultiplexed,
converted to analog form by the D/A converters (DAC), upconverted (U/C) to radio frequency (RF), amplified by the power amplifiers (PA), and transmitted to ground via the MBA.

When photonic technology is applied to this process, electrical to optical converters (E/O) follow the DMUXs, optical to electrical converters (O/E) are inserted before the MUXs, and an optical matrix switch is substituted for the electrical routing switch, as shown in Fig. 1(ii) [5].

![Block diagram of MBA system in [1]](image1)

![Block diagram of optical digital channelizer in [5]](image2)

**Fig. 1.** Digital payloads in an MBA system.

### 3 Proposed optical TDM system for digital payload

The block diagram of the proposed optical TDM system is shown in Fig. 2(i). A digital channelizer with 4 input and 4 output ports between the ADCs and DACs is illustrated for simplicity. The digital uplink signals are formed into burst packets by the DMUX, converted to staggered optical bursts by the E/Os, combined in the first optical coupler, split out again by the second optical coupler, converted back to electrical by the O/E s, and then gated by the MUX to reroute them ready for downlinking.

Figure 2(ii) shows the timing of the signals at points (a) to (g). Each of uplink signals #1 to #4 is synchronized, formed into burst packets, converted to optical, and then allocated to each defined time slot, noting that the data rate of the optical burst packets is 4 times that of the original uplink signals due to the multiplexing of 4 uplink signals within the same duration. The optical burst packets become time interleaved when combined in the optical coupler, and the time interleaved optical burst packets are split across the 4 channels via the second optical coupler. The time
Fig. 2. Proposed optical TDM system for digital channelizer.
interleaved optical burst packets are converted back to electrical and signals #1 to #4 are selected as required. Any of signals #1 to #4 can be selected for downlinking by gating the time slot to be recovered. Note that the sub-channels of the burst packets at points (b) and (f) are not illustrated, in other words, in Fig. 2(ii) the uplink signals are not shown demultiplexed at the DEMUX, and the downlink signals are not shown multiplexed at the MUX, just for simplicity.

As the proposed optical TDM system does not require an optical switch for routing the signals, it can be expected to be smaller and lighter, and to achieve lower power consumption and higher reliability than the previously proposed optical digital channelizers.

4 Performance of proposed optical TDM system

As the proposed optical TDM system has chains of optical bursts, each optical burst packet should have a guard time, which includes the laser on/off times, the synchronization time, and the payload data, as shown in Fig. 3(i).

We verified the transmission efficiency, an important factor for an optical TDM system, which is defined as the ratio of total payload data within the duration of a signal block. The data rate, the duration of the optical burst packet, the guard time, and the synchronization time were 25 Gbps, 500 µs, 128 ns, and 1200 ns, respectively [8].

Figure 3(ii) shows the calculated transmission efficiency and the data rate of the

![Fig. 3. Model of optical burst packet and calculated transmission efficiency.](image)
uplink and downlink signals as a function of the multiplexing factor (N), which is the same as the number of antennas (links). For 64-way multiplexing, the calculated transmission efficiency and the data rate were 83% and 334 Mbps, respectively. The transmission efficiency will be improved by using a longer duration. The data rate can be increased by using higher speed E/Os and O/Es: currently 400 Gbps products are available, and ones for 800 Gbps or more will become available in future. In addition, both parameters can also be improved by adding in other multiplexing technologies such as wavelength and space multiplexing (using multiple fibers).

5 Conclusions

We have proposed an optical TDM system that can take the place of the routing switches in a digital channelizer. The basic configuration and an example of its operation were introduced. A study to verify the transmission efficiency and the data rate of the uplink and downlink signals demonstrated the feasibility of this system. It has the potential to reduce the cost, size, mass, and power consumption, and to improve the reliability of the next generation HTS by substituting the proposed system for conventional routing switches in the digital channelizer.
Extended-loss-budget pluggable transceiver for 10G/1G compatible PON with N:1 redundant OLT protection

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Abstract: We develop the first OLT-pluggable extended-loss-budget transceiver necessary for 10G/1G compatible PON with N:1 redundant OLT protection. This paper newly define the requirement of the OLT transceiver supporting protection by assuming an additional loss of redundant function. The average output power of 6.95/8.35 dBm and minimum receiver sensitivity of −32.02/−34.53 dBm for 10G/1G are attained, respectively. Experiments also confirm 40-km-SMF transmission with no dispersion penalties. It achieves the 36.02/35.53-dB loss budget as 10G/1G OLT transceiver, where ONU transceiver’s specification is based on 10GBASE-PR40/1000BASE-PX40, successfully meeting required loss budget for N:1 redundant OLT protection.

Keywords: optical communication, PON, OLT transceiver, redundant OLT protection

Classification: Transmission Systems and Transmission Equipment for Communications

References

[8] Physical layer specifications and management parameters for extended Ethernet passive optical networks, IEEE Std. 802.3bk, 2013. DOI: 10.1109/IEEESTD.
1 Introduction

The explosive growth in internet data traffic has been driving the demand for high capacity and cost-effective optical access networks, resulting in the commercial development of the 10-Gbps-class passive optical access network (PON), that is, the 10G-EPON system standardized in IEEE802.3av [1]. It is essential to consider cost-effectiveness and compatibility with legacy GE-PON systems [2] because the optical line terminal (OLT) must accommodate both 10G optical network units (ONUs) and 1G ONUs on the same PON branch at the initial stage of adoption. In addition, recently emerging PON applications such as mobile backhaul and business subscribers have raised the importance of optical access network reliability [3], and protection schemes have been discussed and standardized [4, 5]. Redundant OLT protection is attractive because it can also reduce total system operating expense (OPEX) owing to its automatic and fast switching to the redundant OLT transceiver.

In terms of realizing its redundant OLT architecture, it is necessary to define the specification of OLT transceiver for protection by assuming an additional loss imposed by protection components, which must meet its requirement. However, the loss-budget requirement of redundant-protection-based PON architecture has not been adequately examined, and no argument have addressed this problem up to this point.

In this paper, we develop an extended-loss-budget optical transceiver pluggable to current OLTs for a 10G/1G compatible PON with redundant OLT protection; its wavelength-division multiplexing (WDM) filters are thinned sufficiently to minimize its insertion loss and thus maximize output power and improve receiver sensitivity. An internal booster semiconductor optical amplifier (SOA) is set at only the 10G transmitter to reduce overall volume and permit the realization of a truly compact pluggable package.

2 Required loss budget for PON with redundant OLT protection

This chapter describes the loss budget needed to realize the 10G/1G compatible PON with redundant OLT protection by summarizing the loss of each component such as optical fiber line, optical coupler and optical switch (OSW).

Four types of PON protection topologies are standardized in Ref. [4], whereas Ref. [5] develops type B and N:1 protection schemes using N:1 OSW. “Type B with N:1” in Ref. [5] is an example of PON with redundant OLT protection, which we think is one of the best approaches because of its low cost and OPEX. Its architecture is schematically illustrated in Fig. 1, where optical subscriber unit (OSU) for protection is linked to working OSU via OSW and optical coupler. The protection system, composed of optical coupler and OSW, causes additional loss...
between OLT and ONUs, which must be considered when estimating required loss budgets. IEEE standards allow 29/26 dB channel insertion loss between OLT and ONUs for an optical link based on 10G/GE-PON respectively [1, 2], so a 10G/1G compatible PON system must work with optical link losses of at least 29 dB. The maximum insertion loss of N:1 OSW and that of 2 × 2 optical coupler are specified as 1.5 dB in [6] and 3.6 dB in [7], respectively, so the additional insertion loss of the protection system is assumed here to be 5.1 dB. Accordingly, channel optical link loss is increased from 29 dB to 34.1 dB. The extended EPON, which was standardized in 2013, provides 10GBASE-PR40 and 1000BASE-PX40 [8], but neither has a budget sufficient for the PON with redundant OLT protection described above. In Ref. [9], of 10.3-Gbps PON with 37.6-dB loss budget is developed based on SOA-embedded OLT, but none of the components were assembled in any package. Furthermore, an optical band-pass filter (OBPF) with pass bandwidth of 4.2 nm is inserted after the preamplifier SOA, making 1G signal reception impossible because GE-PON system allows upstream lasers to operate in the 100 nm wavelength range [2].

To meet the loss-budget requirement described above, we introduce here the OLT-pluggable extended-budget optical transceiver. An inset located in the right side of Fig. 1 illustrates its functional configuration. 10G transmitter (10G-Tx) and 1G transmitter (1G-Tx) are multiplexed by WDM filter#1, and WDM filter#2 is inserted to de-multiplex the received upstream signal. Each WDM filter is thinned sufficiently to minimize their insertion losses, which yields high output optical power and improved minimum receiver sensitivity. 1G-Tx is composed of laser diode (LD) and external modulator (EM), and booster SOA is set only in the 10G-Tx for the purpose of reducing the entire volume so that all components can be assembled in the compact package pluggable to the current OLTs. Received signal is input into an avalanche photodiode (APD), and its output is amplified by a trans-impedance amplifier (TIA).

3 Experimental evaluations

Experimental procedure for transmitter evaluation is as follows. Pulse pattern generator (PPG) output bipolar electrical signals with electrical power of 500 mVpp. 10.3125-Gbps signal of $2^{31}$ pseudorandom binary sequence (PRBS31) was used for 10G tests while 1.25-Gbps signal of PRBS7 was used for 1G tests. Modulated
signal propagated through standard single-mode fiber (SMF) and input into ONU receiver after passing an optical variable attenuator (ATT). SMF lengths were 0, 20, 40 and 80 km, and received optical power at ONU is changed by tuning ATT level. Received optical signal is converted into an electrical signal and input to error detector (EDT) for bit-error ratio (BER) measurement. Eye diagrams captured at each bitrate are shown in Figs. 2 (a) and (b). They confirm the extinction ratios of 7.82/10.87 dB, and average output power of 6.95/8.35 dBm for 10G/1G transmitter, respectively. Figure 2 (c) shows measured BER performance of the transmitter. Even after 40-km-SMF transmission, no dispersion penalties are observed at the point of $BER = 10^{-3}$ for 10G signal and at $BER = 10^{-12}$ for 1G signal.

![Fig. 2. Transmitter evaluation. (a) 10G Eye diagram. (b) 1G Eye diagram. (c) Transmitter BER performance](image)

Similarly, receiver test is as follows. The each output of 10G and 1G ONUs are coupled after each ATT and input to the transceiver, so the EDT could measure BER. The bitrates, extinction ratios, wavelengths of ONU transmitters of 10G/1G were 10.312 Gbps of PRBS31/1.25 Gbps of PRBS7, 6.5/9.0 dB and 1270/1310 nm. The detailed frame configurations of this burst-mode test are depicted in Figs. 3 (a) and (b), where the length of preamble data, dummy data and guard time are described. Figure 3 (c) shows the results of the receiver sensitivity evaluation. At the point of $BER = 10^{-3}$ received optical power of the 10G signal was $-32.02$ dBm, and at the point of $BER = 10^{-12}$ that of the 1G signal was $-34.53$ dBm, resulting in the minimum receiver sensitivity of $-32.02/-34.53$ dBm for 10G/1G, respectively.

10GBASE-PR40 and 1000BASE-PX40, which we assume as specifications of ONU transceiver, define ONU minimum-launch-powers of $+6/+2$ dBm and maximum receiver sensitivities of $-29.5/-30$ dBm. These means 36.45/38.35-dB available power budgets for 10G/1G downlink and 38.02/36.53 dB for 10G/1G uplink, respectively. They also define maximum transmitter and dispersion penalties as 2/1 dB for 10G/1G uplink; our evaluations confirm the transceiver has zero dispersion penalty at the transmitter. Consequently, these results validate the 36.02/35.53-dB
loss budgets for 10G/1G optical links, successfully meeting loss-budget requirements of 34.1 dB we summarized for the redundant OLT protection PON in the previous chapter.

4 Conclusion

This paper defined the loss budget needed to realize the 10G/1G compatible PON with redundant OLT protection by summarizing the loss of each component, and we developed the extended-loss-budget pluggable optical transceiver to meet its loss-budget requirement. All components such as EM, SOA and APD are assembled in a compact package pluggable to existing commercial OLTS, and internal WDM filters are thinned enough to minimize their insertion loss and thus maximize output power and improve receiver sensitivity. As for 10G performance, our transceiver attained average output power of 6.95 dBm and extinction ratio of 7.82 dB; its receiver sensitivity at BER = 10^{-3} is −32.02 dBm. Similarly, 1G average output power is 8.35 dBm, and receiver sensitivity is −34.53 dBm at BER = 10^{-12}. The loss budget of 36.02/35.53 dB are attained for 10G/1G optical link, respectively, where ONU specifications are assumed as 10GBASE-PR40 and 1000BASE-PX40. These values achieve the OLT transceiver requirements we defined for 10G/1G compatible PON with N:1 redundant OLT protection.
Radio propagation prediction using deep neural network and building occupancy estimation

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Abstract: In this paper, we propose a radio propagation prediction method using machine learning and building occupancy estimation. There have been learning-based researches using aerial photographs and building occupancy images as spatial information. However, the availability of building occupancy images are often limited to urban areas. We assume the situation that only the aerial photographs are available, and aim to improve radio propagation prediction accuracy by estimating building occupancy images from the given aerial photographs.

Keywords: radio propagation prediction, machine learning, neural network

Classification: Antennas and Propagation

References

1 Introduction

Radio propagation prediction plays an important role in high-speed wireless communication. There are two representative but different methods in radio propagation prediction; experimental methods and theoretical methods. Recently, the prediction methods using machine learning have been developed [1, 2, 3]. The prediction model in [1] evaluates simulation data obtained from ray-tracing method, while the model in [2, 3] evaluates the data acquired in urban areas. The method in [2] employs the input data of the spatial data like map images, and the system parameters like base station specifications and Tx-Rx (transmitter-receiver) distance. It has been reported that using building occupancy images as spatial information can provide higher accuracy than using aerial photographs [2]. However, the building occupancy images are often provided only in urban areas.

In this paper, we propose a novel method to improve the accuracy when using only aerial photographs. We generate building occupancy images from aerial photographs using the Convolutional Neural Network (CNN) model called U-Net [4]. In addition, we also study the effect of using the midpoint image between Tx and Rx as additional spatial data.

2 Radio propagation environment prediction by neural network

Two of the authors have already developed a receiver power prediction method based on CNN and fully connected neural network (FNN) [2]. The spatial data and system parameters were used as the input data in [2], where aerial photographs and building occupancy images are used as spatial data. The distances between Tx and Rx, base station specifications and some other data are used as system parameters.

They employed the CNN based on AlexNet [5] whose model has four convolutional and two pooling layers. The FNN consists of input, output and six hidden layers, where the number of the weight parameter $W_{\ell}$ in the $\ell$-th layer is defined by

$$W_{\ell} = 4096 \times \left( \frac{1}{4} \right)^{\ell-1}.$$  \hspace{1cm} (1)

The final output of the prediction model is given as the received power in dBm, and is evaluated by the Root Mean Squared Error (RMSE) between the measured and predicted values.

3 Proposed method

We propose a radio propagation prediction method basically based on [2] but to improve prediction accuracy when using only aerial photographs, i.e., the building occupancy images are not available. We try to estimate building occupancy images from aerial photographs and use them in radio propagation prediction.

3.1 Occupancy image generation

It is indicated in [2] that the prediction accuracy becomes higher when using the building occupancy images as a spatial data, than using only aerial photographs. The Building occupancy images can make clear boundaries of buildings and roads. However, the areas where the building occupancy images are provided is limited
only to urban areas and it is often difficult to get the latest image data. Therefore, we try to estimate building occupancy images from aerial photographs using the CNN and use them in radio propagation prediction. We adopt U-Net [4] for building occupancy image generation.

3.2 Radio propagation prediction

We performed radio propagation prediction using spatial data and system parameters as input data. As the spatial data, we use the rectangular images at three points, where the center of the images are given by the Tx point, the Rx point and the midpoint of Tx and Rx. The midpoint image compensates the spatial data between Tx and Rx. There are three types of images: aerial photographs, building occupancy images, and estimated building occupancy images generated by U-Net. As system parameters, we use 9 kinds, 12 parameters such as Tx antenna height, Tx antenna direction and directivity, Tx antenna gain, transmission power and Tx-Rx distance.

The prediction model consists of CNN and FNN. The CNN model has the convolution and pooling layers and they extract feature values. Its structure is same as that of [1]. In the FNN, the feature values of images and system parameters are inputted, and the final output is the predicted value of the received power in dBm. In network learning, MSE (Mean Squared Error) was used for the loss function and Adam was used for the optimization algorithm.

4 Simulation

We evaluated the proposed method using the measured LTE signals in 2.1GHz band obtained in Tokyo metropolitan area. The height of the transmission points are within 13 to 115m high, installed on the top of buildings. We obtained the measurement data by a receiver antenna installed on a roof of a running vehicle.

4.1 Dataset

After the data is measured by the above method, the median value within 5 x 5 m² area is extracted to remove instantaneous value fluctuation. When the received data contain the signals from multiple Tx points, the signal with the highest receiving power is taken as the data at that point. As a result, the total number of measurement data was 41,650 points. The aerial photographs and building occupancy images are created based on the location of the measured data. The image size is 64 x 64 pixels and covers an area of 256m². We use 4,000 points data for building occupancy image generation, and the remaining 37,650 points for radio propagation prediction. For both the building occupancy image generation and the radio propagation prediction, we use 90% of the total data as learning data, and the rest for test.

We prepared two types of data sets: Dataset 1 and Dataset 2. The test data in Dataset 1 is randomly extracted from all the measurement data, means that both the training and test data covers the whole area. Dataset 1 is constructed in the same way as in [2], but the generalization performance of the predictive model cannot be properly evaluated because of the mixture of training and testing data in the measurement area. Therefore, Dataset 2 is prepared in addition. Dataset 2 separates
4.2 Occupancy image generation

We used the U-Net model described in Section 3.1 to train the building occupancy image generation model. The input image size is $64 \times 64 \times 3$ and the output image size is $64 \times 64 \times 1$. The network structure is similar to that of [4], with the upsampling and downsampling four times each. In addition, the number of channels in the input/output section is changed so that the input is an RGB image and the output is a grayscale image. Also the padding is done so that the image sizes of the input/output become same. The test data (400) is 10% of the total 4,000 points while the rest is used as the training data (32,400). In training the model, the mini-batch size is 32, the maximum number of epochs is 150, and the early stopping is applied when no improvement is seen in 50 epochs. After learning the model, the building occupancy images were generated from the aerial photographs using for radio propagation prediction.

Figure 1 shows the examples of (a) aerial photograph, (b) building occupancy image, and (c) estimated building occupancy image generated from the aerial photograph by U-Net. From Fig. 1, we see that the estimated image in Fig. 1(c) has a similar feature with the real building occupancy image in Fig. 1(b). Note that the building occupancy image in Fig. 1(b) is originally a binary image in which the place where the building is located is white while the other areas are black. The estimated image in Fig. 1(c) is a grayscale image that represents the probability of the building existence in each pixel. The colored aerial photograph in Fig. 1(a) is converted to grayscale when it is used for building occupancy image generation and radio propagation prediction.

4.3 Radio propagation prediction

We evaluated the proposed method by performing radio propagation prediction using Dataset 1 and Dataset 2 shown in subsection 4.1. The evaluation index is the RMSE between the measured and the predicted receiver powers, and is expressed calculated by

\[
RMSE = \sqrt{\frac{1}{N} \sum_{i=1}^{N} (p_i - \tilde{p}_i)^2}
\]  

(2)
where \( N \) is the number of data, \( p_i \) is the measured receiver power value, and \( \tilde{p}_i \) is the predicted receiver power value. In training the model, the mini-batch size is 648, the maximum number of epochs is 1000, and the early stopping is applied when no improvement is seen in 50 epochs. We also compared the RMSE values of the input image with and without the midpoint image to evaluate the effect of adding the midpoint image. The results for each type of input data is shown in Table I, where the methods 1, 2 and the proposed method used aerial photographs, true building occupancy images, and the estimated building occupancy images, respectively. We confirmed from Table I that the RMSE value became smaller when the spatial information (images) was added to the system parameters than when only system parameters were used for prediction. Besides in Dataset 2, the RMSE value was improved by using the estimated building occupancy image and adding the midpoint image.

<table>
<thead>
<tr>
<th>Dataset</th>
<th>Input type</th>
<th>Method 1</th>
<th>Method 2</th>
<th>Proposed Method</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Rx, Tx images + System Parameter</td>
<td>4.10</td>
<td>3.93</td>
<td>4.01</td>
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<td>4.06</td>
<td>3.93</td>
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<td></td>
<td>System Parameter Only</td>
<td>4.49</td>
<td></td>
<td></td>
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<td></td>
<td>Rx, Tx images + System Parameter</td>
<td>8.99</td>
<td>8.19</td>
<td>8.36</td>
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<tr>
<td>Dataset 2</td>
<td>Rx, Tx, Mid images + System Parameter</td>
<td>7.81</td>
<td>7.66</td>
<td>7.54</td>
</tr>
<tr>
<td></td>
<td>System Parameter Only</td>
<td>9.06</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Table I.** Comparison of RMSE (in dB)

![Example of received power prediction result](image-url)

**Fig. 2.** Example of received power prediction result
image, compared to the case where the original aerial photographs were used as spatial information. This is because we could capture the positions of buildings from the estimated building occupancy images.

As an example of the radio propagation prediction results, Figs. 2(b), 2(c), and 2(d) show the difference between the estimated and measured received powers on a map of Dataset 2, while Fig. 2(a) shows the measurement data as a reference. Here, the error is defined as the difference of the measured value from the predicted value. We see from Figs. 2(b) and 2(c) that there are many red and blue points (with large estimation error) but many white points (with small estimation error) in Fig. 2(d). That means, the use of the midpoint image is effective in the received power prediction.

5 Concluding remarks

We proposed a method of estimating building occupancy images from aerial photographs using U-Net for accurate radio propagation prediction. The RMSE of the received power estimation became better when using the images generated by U-Net than the case of aerial photographs only.
Interoperator channel management for dynamic spectrum allocation between different radio systems

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Abstract: For the efficient use of frequency resources, dynamic spectrum allocation (DSA) between different radio systems is promising. In DSA, our approach is to allow different mobile network operators to use the same shared channel without the inter base station (BS) interference. However, this approach is computationally very expensive. Therefore, we formulate this reuse channel allocation as a 0-1 mixed-integer linear programming problem and propose an algorithm to improve the satisfaction of each BS and the time and frequency continuities, which can be used to solve the problem in practical time. Our evaluation demonstrates that our proposed algorithm can improve the satisfaction with the time and frequency continuities simultaneously.

Keywords: dynamic spectrum allocation, spectrum management system, channel allocation algorithm, 5G

Classification: Wireless Communication Technologies

References

1 Introduction

For the efficient use of limited frequency resources, dynamic spectrum allocation (DSA) between mobile network systems and different existing radio systems is promising. In DSA, mobile networks are allocated and utilize unused sharable channels (USCs) that are detected based on a time, area, and frequency that are not in use by existing radio systems. Concrete spectrum management systems such as the citizens broad radio service (CBRS) [1] in the USA and licensed shared access (LSA) [2] in Europe for the social implementation of DSA among different radio systems have been examined. In Japan, the government is also considering the possibility of DSA in the 2.3 GHz, 26 GHz and 38 GHz bands [3].

Recently, to improve the spectrum efficiency, DSA methods designed according to the mobile network operator (MNO) demand have been proposed [4, 5]. However, these existing methods allocate USCs on an MNO-by-MNO basis; thus, even if an MNO is not using their allocation channel in some area, other MNOs cannot use the channel allocated to that MNO, resulting in a spatial waste of spectrum resources. In addition, to reduce the number of channel switch operations and improve the utility of shared channel, the demand-based DSA requires time and frequency continuities of channel mapping in the frequency and time directions for the same BS. The existing methods [4, 5] do not address these channel continuities.

This paper’s approach is to improve the spectrum efficiency with these channel continuities simultaneously by allowing different MNOs to use the same shared channel when the inter base station (BS) interference is below a noise level. However, this channel reuse between BSs with these channel continuities has not been implemented due to the difficulty in calculating the allocation in practical time to meet the BS demand from the myriad combinations of mutual interference between BSs and channel mapping. In this paper, we formulate this reuse channel allocation as a 0-1 mixed-integer programming (MIP) and propose an algorithm to improve the satisfaction of each BS and the time and frequency continuities simultaneously, which can be used to solve the problem in practical time.

2 System model

We show the system model of DSA between different radio systems in Fig. 1(a). This model consists of three systems: the incumbents, the mobile networks and the spectrum management system. Incumbents are nonmobile radio systems, such as satellite systems and fixed wireless access, have been assigned a licensed band and have priority to use the spectrum. The spectrum management system detects the incumbent’s position and the USC by the usage registration information from the incumbents or the sensing data of an incumbent and dynamically allocates the USC to BSs of MNOs at a predetermined timeslot according to their demand. Our proposed channel allocation algorithm is one of the functions of the spectrum management system.

system. Here, the transmission power of the BS is assumed to be pre-calculated as a value that does not interfere with the incumbents in reference to the CBRS system [1].

Next, we show the proposed allocation procedure in Fig. 1(b). In Fig. 1(b), USC B consisting of multiple allocation units b and the demand spectrum bandwidth \( f_{d,n,t} \) of BS index \( n \in N \) at timeslot \( t \in T \) are input into the proposed allocation algorithm, which then outputs the allocation channel \( x_{n,f,t} \) of BS \( n \) at timeslot \( t \) in channel index \( f \in F \). Here, \( x_{n,f,t} \) is a binary variable defined as

\[
x_{n,f,t} = \begin{cases} 
1 & \text{channel } f \text{ is allocated to BS } n \text{ at } t \\
0 & \text{otherwise}
\end{cases}.
\]  

(a) System model \hspace{1cm} (b) Proposed allocation procedure

Fig. 1. System model and proposed allocation procedure.

3 Design optimization problem

To judge whether the same channel can be used at the same time between BSs of different MNOs, we determine whether interference occurs depending on the operational status \( BS_{n,p} \) of each BS, which is defined by:

\[
BS_{n,p} = \begin{cases} 
1 & \text{BS } n \text{ is up and running} \\
0 & \text{BS } n \text{ is down}
\end{cases}.
\]  

Here, \( p \in P \) is an index indicating the combination set \( BSset_p \) of the operational status \( BS_{n,p} \), and \( BSset_p \) is as follows:

\[
BSset_p = \{BS_{1,p}, BS_{2,p}, \ldots, BS_{N,p}\}.
\]  

Let the interference between two BSs from BS \( a \in N \) to BS \( b \in N \) be \( I_{a,b} (a \neq b) \). Then, the amount of interference \( IF_n \) to BS \( n \) in the combination set \( BSset_p \) can be expressed as the following as an aggregation of the interference from the set \( n' \in N' \) of other BSs:

\[
IF_n = \sum_{n' \in N'} I_{n',n} BS_{n',p}.
\]
Let $P_{th}$ be the allowable interference power of BS; then, $\text{BSset}_p$, including a BS with $IF_n > P_{th}$, is determined to be the combination set $\text{BSset}_q$ with interference influence.

Next, we express the problem formulation to determine the allocation channel $x_{n,f,t}$. In this paper, our objective is to maximize three indices simultaneously: the satisfaction of each BS and the channel frequency continuity (FC) and channel time continuity (TC) of the same BS. Satisfaction at $t$ is expressed as a ratio of the allocation spectrum bandwidth to the demand spectrum bandwidth $f_{d_{n,t}}$ using the channel allocation $x_{n,f,t}$ as follows:

$$\text{Satisfaction} = \min \left( 1, \frac{\sum_{f \in F} x_{n,f,t}}{f_{d_{n,t}}} \right).$$  \hspace{1cm} (5)

FC along the frequency axis and TC along the time axis are binary variables defined as

$$\text{FC} = x_{n,f,t} x_{n,f+1,t} = \begin{cases} 1 & \text{frequency continuity in the same BS} \\ 0 & \text{otherwise} \end{cases},$$  \hspace{1cm} (6)

$$\text{TC} = x_{n,f,t} x_{n,f,t+1} = \begin{cases} 1 & \text{time continuity in the same BS} \\ 0 & \text{otherwise} \end{cases}. $$ \hspace{1cm} (7)

These formulas are 1 only when the same BS $n$ is assigned to adjacent frequency or time channels. To improve these three indices, we express the problem of determining the channel allocation $x_{n,f,t}$ of the USC to the BSs as the following optimization problem:

$$\max_{x_{n,f,t}} \sum_{n \in N} \sum_{t \in \tau} \left( \frac{\sum_{f \in F} x_{n,f,t}}{f_{d_{n,t}}} \right) + \alpha_2 \sum_{n \in N} \sum_{t \in \tau} \sum_{f \in F} (x_{n,f,t} x_{n,f+1,t}) + \alpha_3 \sum_{n \in N} \sum_{f \in F} \sum_{t \in \tau} (x_{n,f,t} x_{n,f,t+1})$$ \hspace{1cm} (8a)

s.t.  

$$f_{d_{n,t}} \geq \sum_{f \in F} x_{n,f,t}, \quad (\forall n,t)$$ \hspace{1cm} (8b)

$$\sum_{n \in N} B_{s_{n,q}} x_{n,f,t} \leq \sum_{n \in N} B_{s_{n,q'}} - 1, \quad (\forall q', f, t)$$ \hspace{1cm} (8c)

$$x_{n,f,t} \in \{0, 1\} \quad (\forall n, f, t)$$ \hspace{1cm} (8d)

Here, $\tau$ is the time slot window for solving the optimization problem. In objective function (8a), the first term counts the satisfaction, the second term counts the FC, and the third term counts the TC, where $\alpha_1$, $\alpha_2$ and $\alpha_3$ are the weights of each index. Constraint (8b) prohibits allocation in excess of demand to prevent overallocation to a particular BS. Constraint (8c) expresses that one or more BSs must be down in the interference combination set $\text{BSset}_{q'}$ to prevent interference between BSs. The optimization problem (8), which is a second-order mixed-integer nonlinear programming, can be transformed into the following mixed-integer linear programming (MILP) by using the auxiliary variables $x_{n,f,t} x_{n,f+1,t} = y_{n,f,t}$ and $x_{n,f,t} x_{n,f,t+1} = z_{n,f,t}$.

$$\max_{x_{n,f,t}} \sum_{n \in N} \sum_{t \in \tau} \left( \frac{\sum_{f \in F} x_{n,f,t}}{f_{d_{n,t}}} \right) + \alpha_2 \sum_{n \in N} \sum_{t \in \tau} \sum_{f \in F} (y_{n,f,t}) + \alpha_3 \sum_{n \in N} \sum_{f \in F} \sum_{t \in \tau} (z_{n,f,t})$$ \hspace{1cm} (9a)

s.t.  

$$f_{d_{n,t}} \geq \sum_{f \in F} y_{n,f,t}, \quad (\forall n,t)$$ \hspace{1cm} (9b)

$$\sum_{n \in N} B_{s_{n,q}} x_{n,f,t} \leq \sum_{n \in N} B_{s_{n,q'}} - 1, \quad (\forall q', f, t)$$ \hspace{1cm} (9c)
4 Proposed algorithm

Although optimization problem (9) is NP-hard, a general-purpose solver [8] that provides effective approximate solutions to practical scale problems has been developed. However, as the numbers of BSs and USCs and the size of the decision variable \( x_{n,f,t} \) increase, an increasingly larger number of computing resources are required in (9). Therefore, we propose the following Algorithm 1 to solve optimization problem (9) in practical time by reducing the size of the decision variables \( x_{n,f,t} \) and the number of constraints (9c) of the optimization problem.

Steps (1)–(5) show the extraction of the interfering combination set \( BSset_q \) from the operating status combination set \( BSset_p \). In steps (6)–(9), to reduce the number of \( BSset_q \) values entered into constraint (9c), we determine the inclusion relation by calculating the inner product between \( BSset_q \) and output the combination set \( BSset_q' \) that has no inclusion relation. Here, \( q \) denotes all combination sets except \( q \). For example, if \( BSset_q^1 = \{0, 0, 1, 1\} \) and \( BSset_q^2 = \{0, 1, 1, 1\} \), then \( BSset_q^1 \) is included in \( BSset_q^2 \), and therefore, only \( BSset_q^1 \) is entered into constraint (9c), while \( BSset_q^2 \) is removed.

In steps (10)–(18), to reduce the size of decision variable \( x_{n,f,t} \), we divide the
optimization time window into small portions and solve the optimization problem while repeatedly shifting the time window. The allocation channel $x_{n,f,0}$ at $t = 0$ is determined by solving problem (9) in steps (10)–(12). In steps (13)–(18), we repeatedly solve optimization problem (9) with the most recent channel allocation results added as a constraint until the range mapping of $t \in T$ is completed. Here, $\tau$

<table>
<thead>
<tr>
<th>Parameters and Assumptions</th>
<th>Value</th>
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<tr>
<td>Center frequency (f_c) (\text{GHz})</td>
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<tr>
<td>ULC (\text{GHz})</td>
<td>150</td>
</tr>
<tr>
<td>Allocation unit (\text{GHz})</td>
<td>10</td>
</tr>
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<td>Time interval (\tau)</td>
<td>5 minutes</td>
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<tr>
<td>BS location</td>
<td>Random (10 km x 10 km two-dimensional space and 100 pattern BS location)</td>
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<td>Transmit power of BS (\text{dBm})</td>
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<td>Antenna pattern</td>
<td>Omni</td>
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<td>Pathloss (d_B)</td>
<td>UMa model (Loss of site ratio was set to 50%)</td>
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<td>Allowable interference (P_{\text{acc}}) (\text{dBm})</td>
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<td>Urban dimension</td>
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<td>Weight of $x_{n,f,0}$ and $x_f$</td>
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<td>Optimisation time window (\tau)</td>
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<td>Standard deviation caused by peak traffic</td>
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Fig. 2. Comparison of the average satisfaction, time continuity, frequency continuity and computational time between the proposed method and existing methods.
is the optimization range of timeslot $t$, and $\delta$ is the time window size used in solving problem (9).

5 Evaluation

Common parameters and assumptions used for the system-level evaluation are given in Fig. 2(a). To compare with our algorithm, we used the existing channel allocation method presented in [4], which allocates the USCs for the average demand by the MNO, not by the BS, starting with the least amount of allocation in the past. In the existing method, the channel mapping along the frequency and time axes is allocated in ascending order of the BS index $n$ or determined by classic random mapping.

Figures 2(b), 2(c) and 2(d) show a comparison of the satisfaction, TC and FC for the number of BSs. Here, the TC and FC are the total values for all timeslots in equations (6) and (7) divided by the total number of allocated channels. The proposed algorithm can improve the satisfaction by 31%, TC by 101% and FC by 8%. Figure 2(e) shows a comparison of the time required to solve problem (9). The proposed algorithm reduces the computational time compared to that of the case without steps (6)–(9) and steps (13)–(17) of proposed Algorithm 1. Validating three weights $\alpha_1$, $\alpha_2$ and $\alpha_3$ in the objective function can offer insights and improvements, but we leave that for our future work.

6 Conclusions

In this paper, we have presented an optimization problem for channel allocation that improves the spectrum efficiency by allowing different MNOs to use the same shared spectrum channel and then proposed an algorithm to solve the problem in practical time. Our numerical evaluation demonstrates that our proposed algorithm can improve the satisfaction, TC and FC simultaneously and reduce the computational time required to solve the problem.

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