Heart-rate monitoring of moving persons using 79 GHz ultra-wideband radar sensor

Masahiro Shibao and Akihiro Kajiwara

Graduate School of Environmental Engineering, Kitakyushu University, Hibikino, Wakamatsu-ku, Kitakyusyu, Fukuoka 808-0135, Japan

Abstract: This paper presents a 79 GHz ultra-wideband sensor monitoring the instantaneous heart-rate (HR) of moving persons. The sensor uses three approaches for the instantaneous HR monitoring: block-based motion/movement compensation of multiple persons, multiresolution analysis and Burg method for instantaneous HR analysis. The first approach normalizes and/or compensates for motion/movement of moving persons where the consecutive received signal, called radar range-profile, is stored on a block-by-block basis and the motion/movement within each block is then normalized as stationary. The second approach uses multi-resolution analysis (MRA) in order to remove most of the breathing signal from the received vital-sign waveform (HR and breathing). And the third approach estimates the instantaneous HR using Burg method. Measurement was conducted in order to investigate the usefulness of our suggested HR monitoring system including the above three approaches. The estimated HR is also compared with an optical pulse wave sensor (finger PPG) in order to confirm the accuracy.

Keywords: millimeter-wave, radar sensor, heart-rate, health-care

Classification: Sensing

References

1 Introduction

Heart-rate signal, especially instantaneous heart-rate (HR), is one of very important vital-signs which indicates daily health state of person such as stress evaluation and falling asleep forecast [1]. The HR signal would be variable at every beat of our heart due to the influence of the autonomic nervous. For example, stress evaluation and falling asleep forecast can be estimated from the instantaneous HR. Therefore, HR monitoring systems such as wearable and wireless RF sensor have attracted considerable attention. However, the wearable sensor may require repeated placement and removal of the sensor device including the battery placement, thereby limiting the long-term use. The wireless RF sensor would be preferable, but it is challenging to monitor the instantaneous HR without body contact because the HR component included in the received signal is much weaker as compared with the breathing component [2, 3, 4, 5]. There have been so far several papers regarding the HR monitoring system with ultra-wideband (UWB) radar, but these assumes relatively ideal conditions, that is, on a single person who is motionless (seated or lying down, for example) [3, 5].

In this paper, we suggest a millimeter-wave UWB sensor which can be used to monitor the instantaneous HR of moving persons. It uses three approaches for instantaneous HR monitoring: block and storage-based motion/movement compensation of moving persons, multi-resolution analysis (MRA) and instantaneous HR analysis using Burg method. The first approach normalizes and/or compensates for motion/movement of moving persons where the consecutive received signal, called radar range-profile, is stored on a block-by-block basis and the motion/movement within each block is normalized as stationary. The second approach uses an MRA in order to remove most of the breathing signal from the received vital-sign waveform (HR and breathing). And the third approach estimates the instantaneous HR using Burg method. The use of FFT is not generally appropriate for instantaneous HR since the HR is variable. A normal, healthy heart does not tick evenly like a metronome, but instead, when looking at the milliseconds between the HR interval, there is constant variation. Therefore, the motion/movement normalized signal is separated into HR and breathing component by the MRA and the instantaneous HR is then estimated accurately using Burg method. Measurement was conducted in order to investigate the usefulness of our suggested HR monitoring system. The estimated HR is also compared with an optical pulse wave sensor (finger PPG) in order to confirm the accuracy. It has been found from the measurement results that the estimated HR is in good agreement with the finger PPG sensor (the correlation coefficient is 0.96~0.98).
2 Heart-rate monitoring sensor robust to motion/movement

Fig. 1 shows the block diagram of our suggested monitoring scheme. The sensor uses three approaches for instantaneous HR monitoring: block-based motion/movement compensation of multiple persons, multi-resolution analysis for removing the respiration signal and Burg method for instantaneous HR analysis.

The consecutive received signal is divided into blocks each of which is stored at the data size of a block (block time duration of approximately 1 second) and the motion/movement within each block of radar range-profiles (radar range-profile as a function of time) is tracked and then compensated and/or normalized as stationary as shown in Fig. 1. The compensation process is repeatedly executed until the error is less than a given value as shown in Fig. 1. And the above operation is recursively conducted from block to block. Therefore, the sensor is capable of compensating for the body motion/movement of each moving person and monitoring the HR. For multiple persons, the above operation is conducted one by one. The motion compensated signal is then applied to the following MRA. It is noted that the use of FFT is not appropriate for the HR monitoring since the HR signal is not constant. A normal, healthy heart does not tick evenly like a metronome, but instead, when looking at the milliseconds between heart beats, there is constant variation. The monitoring accuracy is largely affected by breathing signal since the HR component included in the received signal is much weaker than the breathing component. Therefore, the MRA is employed to remove the breathing component where that orthogonal wavelets are employed. The HR and breathing signals generally exist at the frequency components around 0.75 Hz to 1.6 Hz and around 0.1 Hz to 0.5 Hz respectively. Therefore, the HR signal is reconstructed using detail coefficients from level 6 to level 7 to remove the breathing signal [3]. And the HR interval is estimated from the power spectral density (PSD) calculated by Burg method to calculate the AR coefficient at the approximately 4 second data window.

The AR model is shown as

\[ y_n = - \sum_{i=1}^{M} a_i y_{n-i} + x_n \]

where, \( y_n \) is observed signal, \( a_i \) is AR coefficients, \( x_n \) is white Gaussian noise and \( M \) is order of the AR model.

The relationship between the AR coefficients \( a_k \) and the power spectral density (PSD) is shown as

\[ P(\omega) = \frac{\sigma^2}{\left|1 + \sum_{k=1}^{M} a_k e^{-j\omega k}\right|^2} \]

where \( \sigma^2 \) is the estimated variance of residuals.

It is important to determine an optimum order of the AR model. Some algorithms for determining the optimum order have been suggested such as Akaike’s Information Criterion (AIC) and Final Prediction Error (FPE), but it is not easy to determine the optimal order of AR model every time. Therefore, we select the order using the histogram generated by the optimal order of the AR model of the 1000 sets of received signal [3].
### 3 Measurement result

#### A. Measurement set-up

The measurement was conducted using a 79 GHz UWB sensor with 3 GHz bandwidth, 0 dBm transmitted power and 30° antenna beam-width (Texas Instruments IWR1443 radar module). The sensor offers high range-resolution and anti-multipath fading capability as compared with 24 GHz and 76 GHz sensor. The sensor is based on a FCM (fast-chirp modulation) architecture where it transmits a series of 1000 FCM waveforms per a frame time, while the echo signals received from some object are then mixed with the transmitted signal to produce beat signal which will give the distance and Doppler of the object. The resulting beat signal, called IF signal, is transformed into time-domain range-profile data using a FFT demodulator. A photoplethysmography (PPG) sensor for continuous monitoring beat-to-beat pulsation is attached to a finger of each subject and the measured HR is used as a reference.

#### B. Measurement result

The measurement was conducted for three scenarios as shown in Fig. 2: (a) two subjects remained seated in each chair without backrest as usual, therefore some body movements were seen, (b) subject was typing on PC keyboard involving body movement, and (c) subject was sitting a few seconds, and afterwards walking slowly toward the sensor antenna. The sensor antenna was directed towards subjects. The measurement was conducted for three graduate-student (subjects) under a protocol approved by the University of Kitakyushu and the informed consent was obtained for all subjects.

The estimated HR intervals for scenario 1 to 3 are shown in Fig. 2(a)–(c), respectively, where each PPG result is also superimposed. Fig. 2(a) presents the HR results measured simultaneously for two subjects at 1 m and 1.1 m away from the sensor respectively. It is seen that their HR intervals are in good agreement with the PPG as a reference. Fig. 2(b) shows the HR interval of subject typing on PC keyboard. The result indicates that the HR interval also falls in line with the PPG.
(a) Scenario 1 (two subjects seated)
(b) Scenario 2 (typing on PC keyboard)
(c) Scenario 3 (start walking from a sitting position)

Fig. 2. Measurement scenarios and results of HR intervals
Next, Fig. 2(c) shows the HR of a walking subject. It is seen that the movement was compensated successfully in the block by block operation and the result approximately matches the PPG. It may be interesting to show the scattered diagram of the HR interval which are estimated from our suggested sensor and PPG data. Fig. 3 shows the scattered diagram of HR intervals derived from our suggested sensor and PPG for the three scenarios. It is seen from Fig. 3 that the correlation coefficient is observed to be 0.96~0.98 and our suggested monitoring scheme is found to be effective for moving person and multiple persons.

![Scatter diagram of HR intervals](image)

**Fig. 3.** Scatter diagram of HR intervals

### 4 Conclusions

This paper has presented a 79 GHz UWB sensor monitoring the instantaneous heart-rate of moving persons. The sensor uses three processes for instantaneous HR monitoring: block-based motion/movement compensation of moving persons, multi-resolution analysis and Burg method for instantaneous HR analysis. The first process normalizes and/or compensates for some motion/movement where the consecutive radar range-profile is stored at the data size of a block and the motion/movement within each block is normalized as stationary. The second one uses a MRA in order to remove most of the breathing signal from the received vital-sign signal (HR and breathing). The third one estimates the instantaneous HR using Burg method. The measurements for three scenarios including multiple persons and moving person were conducted. As a result, it has shown that our suggested sensor can estimate the instantaneous heart-rate accurately.

We envision that the sensor can be applied to various practical situations, such as home health-care, medical care and driver monitoring system.
Adaptive bandwidth control for multi-band OFDM transmission with spectrum sharing

Yoshiki Inuzuka\textsuperscript{1a)}, Shigeru Tomisato\textsuperscript{1,2}, Kazuhiro Uehara\textsuperscript{1}, Satoru Shimizu\textsuperscript{2}, and Yoshinori Suzuki\textsuperscript{2}

\textsuperscript{1} Graduate School of Natural Science and Technology, Okayama University, 3–1–1 Tsushima-Naka, Kita-ku, Okayama-shi, Okayama 700–8530, Japan
\textsuperscript{2} Advanced Telecommunications Research Institute International, 2–2–2 Hikaridai, Keihanna Science City, Kyoto 619–0288, Japan

\textsuperscript{a)} inuzuka@s.okayama-u.ac.jp

Abstract: This paper proposes an adaptive bandwidth control method in a transmitter for multi-band OFDM transmission with spectrum sharing. The proposed method controls the bandwidth used by a multi-band system to reduce out-of-band distortion noise which interferes to other spectrum sharing systems. In addition, the method uses clipping followed by in-band and out-of-band filtering to satisfy required interference power conditions on other system bands. The evaluation results by computer simulations show that adaptive band selection in the proposed bandwidth control can reduce out-of-band distortion noise power so that the required conditions can be satisfied in other spectrum sharing systems.

Keywords: spectrum sharing, multi-band system, out-of-band noise

Classification: Wireless Communication Technologies

References


1 Introduction

Spectrum sharing is one of promising technologies to increase the capacity of whole network, and multi-band signal transmission which simultaneously uses plural frequency bands is an effective technology to realize flexible and efficient frequency band use in wireless communication systems. On the other hand, OFDM transmission which is very effective in broadband transmission in multi-path fading channels is widely used in wireless systems while it generates excessive peak power which causes in-band and out-of-band noise by non-linear distortion of a transmission power amplifier. In addition, multi-band simultaneous use causes serious distortion noise by inter-modulation distortion. Therefore, effective in-band and out-of-band noise reduction for multi-band transmission have been studied [1].

This letter proposes an adaptive bandwidth control method in a transmitter for multi-band OFDM transmission with spectrum sharing. The proposed method controls the bandwidth of frequency bands used by a multi-band system to reduce out-of-band distortion noise which interferes to other systems shared the same frequency bands. The bands applying bandwidth control are adaptively selected among all usable bands to effectively reduce interference according to the bands used by an interfered system. The method also employs clipping followed by in-band and out-of-band filtering to satisfy required interference power conditions on interfered system bands. This letter clarifies the effect of the proposed adaptive bandwidth control by computer simulations.

2 System model and bandwidth control

Fig. 1 shows the system model and the proposed bandwidth control method in this letter. Multi-band and other wireless systems with OFDM transmission shares the same frequency bands as shown in Fig. 1(a). The multi-band system simultaneously utilizes three bands of B1, B2, and B3. This multi-band simultaneous use causes serious out-of-band noise by inter-modulation distortion, and noise power on adjacent and third inter-modulation (IM3) bands becomes larger. Such out-of-band noise interferes to spectrum sharing systems. When the received power of the multi-band system is extremely larger than that of an interfered spectrum sharing system, the influence of the out-of-band noise cannot be ignored. In this letter, two cases are considered where an interfered system is assigned on the adjacent or IM3 band as shown in Fig. 1(a).

Fig. 1(b) illustrates adaptive bandwidth control of the proposed method to reduce out-of-band noise in a multi-band system which narrows used bands. The bands applying bandwidth control are adaptively selected among all usable bands to effectively reduce the noise on the interfered system bands. In addition, because the method yields unused bands in usable bands, it generates out-of-band noise reduction signals on the unused bands by iterative clipping and filtering [2, 3].

Fig. 1(c) shows an OFDM transmitter with the proposed method for a multi-band system. OFDM signals assigned to multiple frequency bands by adaptive bandwidth control are generated by IFFT processing. After the processing, clipping followed by in-band and out-of-band filtering is performed to reduce out-of-band noise. Fig. 1(d) shows the detection of peak power components exceeding the set...
clipping level in OFDM signals by clipping. These detected peak power components are transformed into in-band and out-of-band clipping noise of the frequency domain by FFT. In-band and out-of-band filtering removes the noise on used and outside bands, and only the components on unused bands remain as shown in Fig. 1(e). The clipping and filtering are iteratively performed to accurately generate out-of-band noise reduction signals. These filtered signals are added to the narrowed multi-band OFDM signals by bandwidth control, and out-of-band noise are reduced on the bands used by the interfered system.

![System model and bandwidth control](image)

### 3 Evaluation results

Computer simulations were conducted to clarify the effect of the proposed adaptive bandwidth control method in multi-band OFDM transmission. Table 1 shows simulation conditions. The modulation scheme was 64QAM. The used band numbers of multi-band and interfered systems were set to 3 and 1, respectively. When the proposed method is used in wireless LAN systems with 5 GHz band, the center frequency can be set as shown in this table. The sub-carrier number of each
band was 64, and the FFT point number was 4096. In this letter, the typical model of a transmission non-linear amplifier (NLA) was used [4], and its input back-off value was set to be 6 dB. The clipping level of clipping and filtering was set to be 3 dB, and iteration number was 5. The total power ratio of own thermal noise and interference from unknown external systems to received signals of the interfered system was set to be 15 dB.

Figs. 2(a) and (b) show SINR performance of the interfered system on the adjacent band of the case 1 as shown in Fig. 1(a). In these figures, the received power ratio of the multi-band system to the interfered system, $P_I$, is set to be 20 dB and 30 dB, respectively. They clarify the SINR of the interfered system to band utilization rates of the multi-band system according to bandwidth control band selection where the control with all three bands, two adjacent bands of B1 and B2, and one of the adjacent bands is performed. The band utilization rate of three band control can be set to be 0 to 100%, in which the use of 0% is only the interfered system without the multi-band system. The SINR without the multi-band system becomes 14.3 dB because of distortion noise by the NLA used in the interfered system. The range of the rates by two-band and one-band control is 33.3 to 100% and 66.7 to 100%, respectively.

The results show that the band use of 100% in the multi-band system significantly degrades the SINR of the interfered system with lower received power, and bandwidth control needs to satisfy required interference conditions for spectrum sharing. In bandwidth control, the control of all three bands is the most effective in interference reduction of the adjacent band. It achieves the SINR of 14 dB at the use of 50% and $P_I$ of 30 dB which is nearly equal to the SINR without the multi-band system. On the other hand, the rate of two band control decreases to less than 40% and one band control cannot achieve the SINR. This is because it can reduce distortion noise by narrowing adjacent bands and the noise caused by IM3 among three bands on the considered adjacent band.

Figs. 2(c) and (d) show SINR performance of the interfered system on the IM3 band of the case 2 as shown in Fig. 1(a). $P_I$ is set to be 20 dB and 30 dB. The control is with all three bands, two IM3 bands of B2 and B3, one of the IM3 bands.

### Table 1. Simulation conditions.

<table>
<thead>
<tr>
<th>Used Band Number (Center Frequency)</th>
<th>Multi-Band System</th>
<th>Interfered System</th>
</tr>
</thead>
<tbody>
<tr>
<td>3 (5.18 GHz, 5.22 GHz, 5.26 GHz)</td>
<td>3</td>
<td>1 (Case1: 5.2 GHz, Case2: 5.3 GHz)</td>
</tr>
<tr>
<td>Sub-Carrier Number</td>
<td>64 × 3</td>
<td>64</td>
</tr>
<tr>
<td>FFT Point Number</td>
<td>4096</td>
<td></td>
</tr>
<tr>
<td>Input Back-off of NLA</td>
<td>6 dB</td>
<td></td>
</tr>
<tr>
<td>Clipping Level</td>
<td>3 dB</td>
<td></td>
</tr>
<tr>
<td>Iteration Number</td>
<td>5</td>
<td></td>
</tr>
<tr>
<td>Total Noise Ratio by Own and External System</td>
<td>–</td>
<td>15 dB</td>
</tr>
</tbody>
</table>
In the case of IM3 band use, the one band control is the most effective, and it can achieve the SINR of 14.3 dB at 66.7%, while the rate of two band control becomes 40%. In the control of all three bands, the rate decreases to 30%. This is because distortion noise by IM3 is directly reduced by narrowing one of the bands causing IM3.

The above results confirm that the appropriate selection of bandwidth control bands according to the bands used by spectrum sharing systems can increase the effect of out-of-band noise reduction which can reduce inter-system interference in spectrum sharing.

![Fig. 2. Simulation results.](image)

4 Conclusion

This paper has proposed an adaptive bandwidth control method in a transmitter of a multi-band system with spectrum sharing to reduce inter-system interference. The evaluation results by computer simulation show that the proposed bandwidth control by adaptive band selection can reduce out-of-band distortion noise power so that the required interference power conditions can be satisfied in other spectrum sharing systems.

Acknowledgments

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Digital acoustic communication in air using parametric loudspeaker

Riku Fukuda\textsuperscript{1,a)}, Tadashi Ebihara\textsuperscript{1,2,a)}, Koichi Mizutani\textsuperscript{1,2}, and Naoto Wakatsuki\textsuperscript{1,2}

\textsuperscript{1} Graduate School of Systems and Information Engineering, University of Tsukuba, 1–1–1 Tennodai, Tsukuba 305–8573, Japan
\textsuperscript{2} Faculty of Engineering, Information and Systems, University of Tsukuba, 1–1–1 Tennodai, Tsukuba 305–8573, Japan
\textsuperscript{a)} ebihara@iit.tsukuba.ac.jp

Abstract: We propose an acoustic communication system using parametric loudspeaker that emits a communication signal into a limited area. We found that the use of minimum shift keying is suitable, since it has the potential and second harmonic signal that occur when the emit signal is distorted during propagation due to the effect of air. Experiments revealed that the proposed system emits an audible signal to a limited area, achieves a BER of $10^{-3}$ at $E_b/N_0$ of 10 dB. Thus, it outperforms benchmarks using other modulation schemes.

Keywords: parametric loudspeaker, digital acoustic communication, frequency diversity

Classification: Wireless Communication Technologies

References


1 Introduction

Digital acoustic communication in air plays an important role in establishing a wireless link between mobile devices [1, 2]. Existing acoustic communication systems tend to broadcast information using audible sounds, resulting in interference to all unintended receivers [3]. Parametric loudspeakers have the potential to address this problem, since they can create directive low-frequency sounds by exploiting nonlinear ultrasound-medium interaction effects [4, 5]. Many useful applications for such loudspeakers have been proposed, such as sound projectors for public spaces [6], high-resolution imaging [7], and sensing [8]. However, to our knowledge, no acoustic communication system that actively uses the characteristics of the nonlinear acoustic channel has been proposed.

In this paper, we investigate a digital communication scheme that is suitable for acoustic communication using a parametric loudspeaker. Specifically, we design a communication system considering a nonlinear acoustic channel, and perform experiments to evaluate the performance of the proposed system.

2 Communication system using parametric loudspeaker

Fig. 1 is a block diagram of the transmitter (Tx) and receiver (Rx) for acoustic communication using a parametric loudspeaker. The signal processing in the Tx is as follows:

1. The Tx reads a message vector of length \( N \), \( s = (s_0, s_1, \ldots, s_{N-1}) \), where \( s_n = \{0, 1\} \). Then the Tx performs pulse shaping and obtains pulse signal \( a(t) \),

\[
a(t) = \begin{cases} 
1(s_n = 0) & (nT \leq t \leq (n + 1)T), \\
-1(s_n = 1) & (nT \leq t \leq (n + 1)T), 
\end{cases}
\]

where \( T \) is the symbol time.

2. The Tx performs digital modulation using minimum shift keying (MSK)—a specific type of frequency shift keying—to obtain a modulated waveform \( b(t) \),

\[
b(t) = \cos 2\pi \left( f_c + \frac{a_n(t)}{4T} \right) t + \varphi_n, 
\]

where \( f_0, f_1, \) and \( \varphi_n \) are...
\[ f_0 = f_c + \frac{1}{4T}, \quad (3) \]
\[ f_1 = f_c - \frac{1}{4T}, \quad (4) \]
\[ \varphi_n = \frac{\pi}{2} \sum_{n=1}^{n-1} (2\delta_i - 1) \ (n > 0), \quad (5) \]

respectively, and \( f_c \) is the carrier frequency of MSK.

3. The Tx performs amplitude modulation (AM) and emits the modulated signal \([1 + mb(t)] \sin 2\pi f_2 t\) to the air using the parametric loudspeaker, where \( m \) and \( f_2 \) are the modulation index \((0 < m \leq 1)\) and carrier frequency of AM, respectively.

The emit signal is distorted during propagation due to the effect of air nonlinearity, and is received by the receiver as

\[
r(t) = \frac{P_1^2 A\beta}{16\pi^2 \rho_0 c_0^2 \alpha} \left\{ 2m^2 \frac{\partial^2}{\partial t^2} b(t) + m^2 \frac{\partial^2}{\partial t^2} b^2(t) \right\} ,
\]

\[
= \frac{P_1^2 \pi A\beta}{4\rho_0 c_0^3 \alpha} \left\{ -m \left( f_c + \frac{a_n(t)}{4T} \right)^2 \cos \left( 2\pi \left( \frac{a_n(t)}{4T} + f_c \right) + \varphi_n \right)
+ 2m \left( f_c + \frac{a_n(t)}{4T} \right) \cos \left( 4\pi \left( \frac{a_n(t)}{4T} + f_c \right) + 2\varphi_n \right) \right\} ,
\]

where \( P_1, A, \beta, \rho_0, c_0, z \) and \( \alpha \) are the pressure amplitude at source, the source radius, the nonlinearity coefficient of the medium, the density of the air, the small signal sound speed, the coordinate along the propagation direction of beams, and the dissipation factor corresponding to thermoviscous absorption, respectively [9].

From Eq. (7), we can find that the received signal \( r(t) \) contains \( b(t) \) and the second harmonic of \( b(t) \). Since the modulated signal \( b(t) \) appears at two frequency bands, the Rx can improve communication quality by merging two signals with different frequencies (frequency diversity) [10]. Note that the frequency diversity may not be utilizable simply in other digital modulation schemes, such as phase shift keying (PSK), since the phase of the second harmonic of \( b(t) \) is distorted.

The signal processing in the Rx is as follows:

1. The Rx performs non-coherent FSK demodulation on the received signal \( r(t) \) [11]. Specifically, the Rx compares two signal powers \( p_0 \) and \( p_1 \), where \( p_0 \) is the sum of the signal power at the frequencies of \( f_0 \) and \( 2f_0 \), and \( p_1 \) is the sum of the signal power at the frequencies of \( f_1 \) and \( 2f_1 \).

![Fig. 1. Block diagram of transmitter and receiver for acoustic](image-url)
2. The Rx outputs a received message vector of length $N$, $r = (r_0, r_1, \ldots, r_{N-1})$, where

$$r_n = \begin{cases} 
0 & (p_0 \geq p_1), \\
1 & (p_0 < p_1), 
\end{cases}$$

(8)

3. Experiment

We evaluated the performance of the digital acoustic communication system using the parametric speaker in experiments. Fig. 2(a) shows the experimental environment. The experiment was performed in an anechoic room. The transmitter consisted of a PC with software modulator (LabVIEW, National Instruments), a digital-to-analog converter (USB-6212, National Instruments), and a parametric speaker consisting of 49 emitters (T40-16, Nicera). The receiver consisted of a PCM recorder (CD-05, TASCAM) and a PC with a software demodulator. Table I shows the parameters used in the communication. In this experiment, we measured the relationship between the energy per bit to noise power spectral density ratio
(E_b/N_0) and the bit error rate (BER). To clarify the advantage of the proposed system (MSK with frequency diversity), communication using MSK without frequency diversity and phase shift keying (PSK) was also performed. Note that the bandwidth of PSK is smaller than that of MSK if we compare them under the same communication speed, so we employ E_b/N_0 as the normalized signal-to-noise ratio.

Fig. 2(b)–2(d) show the experimental results. Fig. 2(b) shows the sound pressure distribution of the communication signal. As shown in the figure, the emit sound was focused at high intensity into a relatively small area, which has the potential to address the problem of conventional acoustic communication. Fig. 2(c) shows the spectrum of the received signal. As shown in the figure, there exist the spectrum of b(t) and its second harmonics. Fig. 2(d) shows the communication quality obtained in the experiment. As shown in the figure, we found that the proposed MSK with frequency diversity outperforms the benchmarks. Specifically, the proposed system achieved a BER of 10^-3 at E_b/N_0 of 10.15 dB, while normal MSK and PSK achieved BER of 4.22 × 10^-3 and 3.58 × 10^-3 at the same E_b/N_0, respectively. The obtained results suggest that the combination MSK and frequency diversity is suitable for digital acoustic communication using a parametric loudspeaker.

4 Conclusion

We proposed a digital communication scheme that is suitable for acoustic communication using parametric loudspeaker. Specifically, we design a communication system considering nonlinear acoustic channel, and perform experiments to evaluate the performance of the proposed system. From experiments, it was confirmed that the emit sound from parametric loudspeaker was focused at high intensity into a relatively small area, which has the potential to address the problem of conventional acoustic communication. Furthermore, it was also found that the combination MSK and frequency diversity, which utilizes harmonic distortions, outperform normal MSK and PSK, and is suitable for digital acoustic communication using a parametric loudspeaker.

**Table I.** Parameters used in experiment

<table>
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<tr>
<th>Parameter</th>
<th>Value</th>
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<td><strong>Digital modulation</strong></td>
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<tr>
<td>MSK</td>
<td></td>
</tr>
<tr>
<td>T (ms)</td>
<td>1.25</td>
</tr>
<tr>
<td>f_0 (Hz)</td>
<td>1,800</td>
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<tr>
<td>f_1 (Hz)</td>
<td>2,200</td>
</tr>
<tr>
<td>Data rate (bps)</td>
<td>800</td>
</tr>
<tr>
<td>Signal bandwidth (Hz)</td>
<td>1,200</td>
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<tr>
<td><strong>Digital modulation</strong></td>
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<td>(PSK as reference)</td>
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<tr>
<td>T (ms)</td>
<td>1.25</td>
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<tr>
<td>f_0 (Hz)</td>
<td>2,000</td>
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<tr>
<td>Data rate (bps)</td>
<td>800</td>
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<tr>
<td>Signal bandwidth (Hz)</td>
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<td>m</td>
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<tr>
<td>f_2 (Hz)</td>
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</tbody>
</table>
PC process migration using FPGAs in ring networks

Keisuke Takano¹, Tetsuya Oda², Ryo Ozaki², Akira Uejima², and Masaki Kohata²

¹ Graduate School of Engineering, Okayama University of Science,
Ridai-cho 1–1, Kita-ku, Okayama-shi, Okayama 700–0005, Japan
² Faculty of Engineering, Okayama University of Science

Abstract: This paper describes a method of process migration of PCs by FPGA-based ring networks. Here the original system configuration is given on the premise that the proposed hybrid cluster control circuit is used, and the communication protocol used on them is determined. The experiment results show that this system has the capability to perform PC process migration with FPGAs. The primary advantage of this system is that it is suitable for low-cost operations in which the PC is shut down after the process is entrusted to the FPGAs, and the result is obtained later.

Keywords: hybrid cluster, interconnect network, dynamic partial reconfiguration

Classification: Network System

References


1 Introduction

A cluster system is regarded as a high-performance computing system, and there are many cases and various forms of cluster systems in which FPGAs are partially utilized [1, 2]. We propose a PC-FPGA hybrid cluster that includes these various forms and connects PCs (CPUs and GPUs) and FPGAs equally [3].
On the other hand, process migration [4] is one of the key technologies used to improve cluster performance and availability. Here, process migration refers to the technology that moves the software processes running on one PC to another PC without interruption of any processes on either PC. In this paper, we propose a system in which FPGAs perform process migration of PCs used in laboratories (hereinafter referred to as “this system”). These FPGAs are connected to PCs and form a ring network. In this system, when PC processes are executed for a long time, they are entrusted to FPGAs in the middle of the execution. Meanwhile, the PC itself is shut down and is restarted the next day, and the execution result is received from the FPGA.

In this paper, we report the basic configuration and communication protocol for this system, and an image processing experiment conducted using them.

2 PC-FPGA hybrid cluster

2.1 Basic configuration

Fig. 1 shows the overall system configuration and the FPGA circuit configuration. FPGA boards are installed in the PCI Express slot of each PC, and these FPGA boards constitute a ring network. Each connected PC-FPGA pair is called a node. Any user circuit (corresponding to PC application software) can be reconfigured dynamically in the user module using HWICAP. In addition, control register operation from a PC or user module and the DMA controller can be activated, and data can be transferred to any address. There are two address spaces for this purpose: one mapped to PCI Express and the other mapped to the internal bus. Table I(a) and (b) lists each address space. By specifying the internal bus address space in Table I(a), the PC can access the address space in Table I(b), i.e., the components of the FPGA.

2.2 Physical layer

The FPGA Mezzanine Card (FMC) on the FPGA is connected by a coaxial cable. The FMC is used in many FPGA boards such as Xilinx and Intel. The high-speed transceiver in the FPGA is used, and 4b5b code is used for encoding. The data transfer rate of the physical layer is set to 4 Gbps (line rate is 5 Gbps).
2.3 Data link layer
A packet consisting of a 1-word (32 bit) header and 32-word data is used as the communication unit. The header contains the sender/receiver node address, sender/receiver port, and type. A router relays packets using store-and-forward. The 4b5b code has some control signals called K characters. After assigning XON/XOFF of the flow control to these K characters, the receiver controls the flow by sending XON/XOFF to the sender according to the state of the receiving buffer.

2.4 Network layer
There are five types of packets: connection request, connection permission, DMA control, DMA data, and register control. The control information which includes these types of packets is specified in a header comprising 10 fields. The details are (22:20) Packet Type, (19:16) Source Port Address and Destination Port Address, (15) Virtual Channel, (14:12) Type, (11:8) Source Port and Destination Port, and (7:0) Source Address and Destination Address.

In Fig. 1, Source Address and Destination Address specify the source and destination node numbers, respectively. Source Port and Destination Port specify the source and destination within the node, which is either the PC or FPGA. In addition, the source and destination modules are individually specified using the Source Port Address and Destination Port Address. These header configurations enable flexible data transfer in a mixed PC and FPGA environment.

To use this system for process migration, the communication procedure is determined so that reliable communication can be achieved in all possible communication paths using the router. For example, Fig. 2 shows a procedure for data transmission from PC0 to the DRAM of FPGA1. In this example, as the number of iterations of the data packet is described in the request packet, no termination processing is required. In this system, such a communication procedure is defined between PC, DRAM, HWICAP, and user module (including the same elements of

<table>
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<th>Table I. Memory map</th>
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<tr>
<td><strong>(a) PCI Express Address</strong></td>
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<tr>
<td>Usage</td>
</tr>
<tr>
<td>DMA</td>
</tr>
<tr>
<td>System Control</td>
</tr>
<tr>
<td>Internal Bus</td>
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<tr>
<td><strong>(b) Internal Bus Address</strong></td>
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<tr>
<td>Module(Unit)</td>
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<tr>
<td>ICAP</td>
</tr>
<tr>
<td>BAR2 Register</td>
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<tr>
<td>Router.PC</td>
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<tr>
<td>Router.UserModule</td>
</tr>
<tr>
<td>UserModule(Reg)</td>
</tr>
<tr>
<td>UserModule(User)</td>
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<tr>
<td>DRAM</td>
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different nodes). As described above, HWICAP is mapped on the internal bus address space and can be accessed from PCI Express, so that (for example) dynamic partial reconfiguration of FPGA1 can be performed from PC0 by executing this communication procedure with the write address as HWICAP and write data as partial reconfiguration data.

### 3 Experiment results

#### 3.1 Experiment overview

The CPU used in each PC was an Intel Core i7-3820, and Ubuntu 18.04 was used as the OS. The Xilinx KC705 evaluation board was adopted for the FPGA. DDR3-1600 was used as the DRAM on the FPGA board. SYSTEC’s SYPCIE (runs on Gen2x4) was used for the PCI Express interface. The AXI interconnect was implemented to operate at 128 bit/160 MHz, and each FPGA module was implemented to operate at 160 MHz.

As a preliminary preparation, software to apply Laplacian filters to 1000 HD images (filter software) and circuits for the user module (filter circuit) were individually configured by using C language for the PC and Verilog for the FPGA. In the filter circuit, four FPGAs cooperatively apply Laplacian filters. Pre-filtered and post-filtered HD images on the FPGA side are placed in DRAM on FPGA0.

First, PC0 and FPGA0-3 were started, and PC1-3 were shut down. Next, after the filter software was started in PC0 and filter processing of 500 images was conducted on PC0, the execution data and the filter circuit (dynamic reconfiguration data) were transferred to the DRAM of FPGA0 and HWICAP of FPGA0-3, and after the execution of the filter circuit was instructed, PC0 was shut down. Finally, PC0 was activated and the execution results were transferred from the DRAM of FPGA0 to the memory of PC0 by the interrupt from the user module of FPGA0.

#### 3.2 Results

The experiment described in Section 3.1 showed that all images were correctly filtered. The number of used slices was 39479 when the filter circuit was transferred. The communication speed between PC and FPGA (Data size 1 GB) was 332
[MB/s] for transfer from PC0 to FPGA0 and 305 [MB/s] for transfer to FPGA1. On the dynamic partial reconfiguration of the filter circuit, the reconfiguration was completed in 31 [ms] in both FPGAs. The power consumption of PC1 + FPGA1 and FPGA1 in the experiment was 60.0 [W] and 21.0 [W], respectively.

This system was evaluated using the results described above. As for the number of slices used, approximately 22.5 [%] of the area went unused in the experiment environment. Although the user module needs to be secured as a rectangular region for dynamic partial reconstruction, the largest rectangular region was investigated, and it was found that approximately 34.6% of the region could be utilized. It was proven that power consumption could be reduced by 65.0% by turning off the PCs after the processing was requested from PC to FPGA. The speed was sufficient for the transfer of images in the process migration.

Overall, it was found that the intended behavior of process migration was realized. As demonstrated in the experiment, it is possible to shut down PCs temporarily after the process is entrusted to the FPGA, and the result is obtained later; this feature is not found in conventional systems.

4 Conclusion

This paper describes an FPGA circuit configuration and communication protocol for creating a system that can realize process migration of PCs using an FPGA circuit. Moreover, it was confirmed by the experiment that the intended operation of process migration was realized.

Currently, the network shape is limited to a ring, but in future work, we want to return to the design philosophy of a hybrid cluster so that the system can support various network forms.

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Performance comparison of antenna deployment cases and precoding schemes in an ultra-dense scenario for high SHF wide-band massive MIMO in 5G

Hiroshi Nishimotoa), Akinori Taira, and Akihiro Okazaki
Information Technology R&D Center, Mitsubishi Electric Corporation, 5–1–1 Ofuna, Kamakura, Kanagawa 247–8501, Japan
a) hh@m.ieice.org

Abstract: Multiuser multiple-input multiple-output (MU-MIMO) downlink technology over massive MIMO is a key enabler for 5G systems and beyond. However, its performance may deteriorate when users are spatially correlated. Distributed antenna deployment (DAD) and nonlinear precoding (NLP) have been studied as countermeasures for this issue. In this paper, through simulations of 28 GHz-band massive MIMO downlink transmission in an ultra-dense urban scenario, it is revealed that, rather than DAD, concentrated antenna deployment (CAD) is more robust against channel transition issues up to walking speed. It is also found that combining CAD with NLP provided the best throughput performance in the evaluated scenario.

Keywords: massive MIMO, MU-MIMO, high SHF band, ultra-dense urban scenario, antenna deployment, precoding

Classification: Wireless Communication Technologies

References


1 Introduction

In the move towards the further refinement of spatial utilization in fifth generation mobile communications (5G) and beyond, multiuser multiple-input multiple-output (MU-MIMO) downlink technology over massive MIMO has been actively studied [1]. In 5G systems and beyond, the exploitation of higher frequency bands such as high super high frequency (SHF) and extremely high frequency (EHF) is a promising approach to accommodating the forthcoming huge traffic demands. In such bands, small-cellularization is envisioned for serving high capacity. One applicable use is in a dense urban scenario where many active users are closely located, causing high inter-user interference (IUI) among users in close proximity. MU-MIMO may not work well when users are spatially correlated like this.

A distributed antenna system is one approach to mitigating the high IUI issue. Wide antenna separation can reduce channel correlation. Its effectiveness has been reported through studies targeting low SHF bands below 6 GHz [2]. On the other hand, in the high SHF and EHF bands an increase in antenna panel separation may reduce the ability to handle user mobility because the beams from the overall antenna array become excessively narrow [3]. Another approach to densely populated cases is nonlinear precoding (NLP), with which in principle we can achieve IUI-free downlink transmission by canceling the IUI observed at the user ends in advance at the base stations (BSs). It has been reported that, compared with linear precoding (LP), NLP can significantly improve the system throughput, especially in a highly-correlated situation [4]. However, in general NLP is thought to be poor at handling mobility due to the inevitable constraints on IUI pre-cancellation [5].

Based on the above background, in this paper the authors evaluate the performance of high-SHF-band massive MIMO downlink transmission in an ultra-dense urban scenario, in order to compare two cases of antenna deployment, concentrated antenna deployment (CAD) and distributed antenna deployment (DAD), and two precoding schemes, LP and NLP.

2 Evaluation scenario

Fig. 1 illustrates the scenarios evaluated in this paper. Targeting a scramble crossing with thousands of pedestrians such as Shibuya scramble crossing in Japan, in the simulation it was assumed that there were 1,600 users in total in a 70 m × 70 m evaluation area, and that 25% of the users, namely 400 users, were active. The radio frequency was set at 28 GHz (wavelength, λ = 10.7 mm), and four carrier components (CCs) were used for downlink transmission, where one CC consisted of 1,584 subcarriers at 60 kHz spacing (bandwidth of 95 MHz). The remaining simulation
parameters for modulation and demodulation were the same as those in [4]. Under the simulation conditions, the maximum sum-rate system throughput reached 32.8 Gbps.

Assuming an analog-digital hybrid configuration with multiple subarrays for the massive MIMO [4], the antenna subarray associated with a TX digital port at the BS was a 64-element planar phased array. The antenna panel was composed of a pair of ideally-isolated cross-polarized subarrays, as shown in Fig. 5 in [4]. The BS had eight antenna panels, namely 16 subarrays. For CAD, as shown in the top figure in Fig. 1(a), all the BS antenna panels were placed on the roof of a building at one corner of the scramble crossing, where the panel orientation was two vertical and four horizontal. In contrast, for DAD, as shown in the top figure in Fig. 1(b), two panels were placed on the roof of each building at the four corners of the scramble crossing, where the panel orientation at each building was both horizontal. The spacing between adjacent panels was $8\lambda$ (85.7 mm) for both cases, and the spacing between adjacent buildings was 50 m for DAD. The BS antenna height was 21 m, and the vertical downtilt was 30°. Each user had one pair of ideally-isolated cross-polarized omnidirectional antennas, namely two RX ports. Both moving vehicles and walking pedestrians were considered as users. The user antenna height was set to 1.7 m and 1 m for vehicles and pedestrians, respectively. For a fair comparison of the performance, random scheduling and a round-robin algorithm were employed. Every 20 ms, eight users out of the 400 active users were selected and transmitted to in turn to perform 16-stream MU-MIMO downlink.
transmission (two streams per user). Therefore all the active users were given fair transmission opportunities with an interval of one second. For the precoding, block diagonalization and Tomlinson-Harashima precoding were employed for the LP and NLP schemes, respectively.

In the evaluation scenario, the traffic signals for both vehicles and pedestrians were controlled just as in a real crossing as time passed (see the top figure in Fig. 2). The entire evaluation ran for 120 seconds. Pedestrians were walking across the crossing for the first 20 seconds and the last 20 seconds. In the remaining periods, excluding signal change intervals, vehicles were moving north and south, and then east and west. For the vehicles, the speed of movement was set to an average of 30 km per hour except for the periods allowed for speeding up and slowing down. For the pedestrians, the walking speed was set to an average of 3 km per hour, while in the periods when the signals were flashing and red it was about 6 km per hour and 1–2 km per hour, respectively.

Fig. 2. Sum-rate throughput versus elapsed time with traffic signal control pattern (top figure and intermediate graph) and examples of user transitions with analog beam spots (bottom figures).
3 Numerical results

Figs. 2 and 3 demonstrate the sum-rate throughput performance with elapsed time and its cumulative distributions, respectively. Here, the sum-rate throughput, which is a summation of the throughputs of eight users, is equivalent to the system throughput. Note that in Fig. 2, a simple moving average over 0.5 seconds is applied to each throughput curve for visibility, whereas in Fig. 3 the instantaneous throughputs are plotted cumulatively. We see that the throughput for CAD outperformed that for DAD when the pedestrians were walking across the scramble crossing. While DAD can mitigate IUI, it has already been reported in [3] that the ability of a more widely spaced antenna array to handle mobility tends to be worse because the resulting spot beam, while providing a high signal-to-interference-plus-noise power ratio (SINR), covers a smaller area, so a moving user rapidly moves out of the beam. Therefore CAD was more robust to channel variation up to walking speed. When the traffic signals for the pedestrians remained red, however, we had the situation that many users clustered closely together at the four corners of the scramble crossing. In that case, while the performance of CAD degraded, DAD saw a slightly improved throughput thanks to channel decorrelation due to its large antenna spacing, in addition to the benefits from reduced walking speed. It is noted that, although moving vehicles were randomly scheduled as receiving users as well as pedestrians in this period, their effect was negligible because the population of vehicles was much smaller than that of the pedestrians.

Also, when comparing the precoding schemes, we find that NLP produced higher throughput than LP, especially in high IUI conditions, typified by CAD and the period with red signals for the pedestrians. This phenomenon reflects the reports in [3, 4]. Here note that, compared with DAD, CAD tends to cause high IUI due to beam overlapping while it has better ability to handle mobility. Hence, as a result, the combination of CAD with NLP provided the best overall throughput, exceeding 20 Gbps about 95% of the time.
4 Conclusions

In this paper, the authors numerically evaluated the system throughput performance of high-SHF-band massive MIMO downlink transmission in an ultra-dense urban scenario, to compare CAD and DAD configurations and LP and NLP schemes. Through the computer simulations, it was found that, in the evaluation scenario having many walking pedestrians in close proximity, CAD provided better throughput rather than DAD. Also, it was clarified that CAD with NLP showed the best performance when walking pedestrians dominated.

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