Multi-stage adaptive equalization for all-optical-aggregated 16QAM signal

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Abstract: We propose a multi-stage adaptive equalizer for the demodulation of an optical aggregated 16QAM. A successful demodulation of the 100 Gbps class 16QAM signal was demonstrated without an OSNR penalty when the input PAM4 signals to the optical aggregation were not symbol-synchronized.

Keywords: all-optical signal processing, modulation format conversion, adaptive equalizer, data center networks

Classification: Fiber-Optic Transmission for Communications

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1 Introduction

The transmission capacity in metro data center (DC) interconnects has been rapidly increasing to meet the demand of the traffic between distributed DCs [1]. To accommodate such a high-capacity transmission demand, the quadrature amplitude modulation (QAM) format for aggregating the traffic of the intra-DC network is expected to serve as the main modulation format in the metro DC interconnects. In contrast, pulse amplitude modulation over four levels (PAM4) has been employed in the intra-DC network, which requires simple and low-cost transceivers [2]. Therefore, a modulation format conversion from PAM4 to QAM is necessary to connect DCs, as shown in Fig. 1.

Optical modulation format conversion is one of the more promising technologies for realizing low-latency and efficient connection of networks with different modulation formats [3]. For inter-DC networks, optical format conversions from PAM4 to QPSK and 16QAM using highly nonlinear fibers (HNLFs) have been proposed [4, 5]. However, the PAM4 to 16QAM conversion, which includes an aggregation of two PAM4 signals, requires an adjustment of the input timing between two PAM4 signals. This decreases the flexibility of the optical signal processing.

In this study, we propose to apply a multi-stage adaptive equalizer to a demodulation for the 16QAM signal aggregated from PAM4 signals without a timing adjustment. We demonstrate by numerical simulations that a 100 Gbps class aggregated 16QAM signal can be demodulated regardless the timing shift of PAM4 signals.

Fig. 1. Metro inter-DC network and format conversion

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2 Optical aggregation from 2×PAM4 to 16QAM

Figure 2 (a) shows the overview of the optical aggregation from two PAM4 signals to a 16QAM signal using a nonlinear optical loop mirror (NOLM) configured with HNLF [5]. We use the effect of cross-phase modulation (XPM) in the HNLF for the aggregation. The continuous wave (CW) probe light is coupled with the PAM4 signal and propagates in the HNLF, rotating the phase of the probe light by XPM. At the output of the HNLF, extracting only the probe light component using an optical band-pass filter (OBPF) with rectangular-shape passband obtains the 16QAM signal.
converted QPSK signal [4]. Applying this conversion to the NOLM configuration allows us to convert the two PAM4 signals to clockwise and counter-clockwise QPSK signals. The two QPSK signals are coupled with a power ratio of 1 : 4 since the probe light passes twice through the 1 : 2 optical coupler (OC) at the input and output of the NOLM. Eventually, the 16QAM signal is generated at the output of the OBPF [5]. However, in this aggregation it is difficult to synchronize the timing of the two PAM4 signals. Therefore, we propose a method of separating the signals through disaggregation without any timing adjustment in the optical aggregation.

3 Multi-stage adaptive equalizer for aggregated 16QAM signal

In the field of optical communications, multi-stage adaptive equalizers have been proposed for the removal of inter-carrier interference components and signal disaggregation in non-orthogonal multiple access (NOMA) systems [6, 7]. In this study, we employ a multi-stage adaptive equalizer to demodulate the optical aggregated 16QAM signal. Figures 2 (b), (c) illustrates the overall and detailed block diagrams of digital signal processing (DSP) for the disaggregation from 16QAM to 2×data. The received signal is input into the first equalizer trained by a known training sequence of the QPSK signal 1. The first equalizer then treats the components of the QPSK signal 2 as noise and outputs the QPSK signal 1. The obtained QPSK signal 1 is decoded and remodulated for generating a replica of the QPSK signal 1. The replicated QPSK signal 1 and the received 16QAM signal are input to a 2×1 multiple-input and single-output (MISO) adaptive equalizer (second stage). The second equalizer then subtracts the components of the QPSK signal 1 from the received 16QAM signal and outputs the QPSK signal 2. By repeating these processes of equalizing and replicating each QPSK signal, the original QPSK (PAM4) signals before the optical aggregation are obtained. This method equalizes the QPSK signal by canceling another QPSK signal as an interference component, thus enabling the demodulation of the two QPSK signals without the need for symbol synchronization between them. Therefore, the 16QAM signal can be demodulated regardless of the time difference between the two PAM4 signals input to the converter.

For the QPSK signal replication, we generate the QPSK signal by emulating the phase rotation of the probe light by XPM during the all-optical PAM4/16QAM conversion. The equalized signal is decoded by forward error correction (FEC) to remove the bit errors. Then, we generate the PAM4 signal by encoding, modulating, upsampling, and filtering using a Bessel filter. The PAM4 signal is converted to the QPSK signal through power-to-phase conversion in the digital domain.

4 Numerical simulations

We performed a numerical simulation to investigate the feasibility of the proposed multi-stage equalizer. Figure 3 (a) depicts the simulation model. The PAM4 signals of 26.6 GBAud were generated as input signals using random bit sequences. Assuming the asynchronous situation, one of the two PAM4 signals was shifted by $\Delta T$ using an optical delay line, where $\Delta T$ is the amount of time shift normalized by the symbol length. $\Delta T = 0$ represents symbol synchronization and $\Delta T = 0.5$ represents a 0.5 symbol shift between the two PAM4 signals. We set the wavelength...
of the probe light as $\lambda_0$ to 1550 nm and that of the PAM4 signal $\lambda_1$ to 1540 nm. We coupled the PAM4 signals and CW probe light and launched them into the PAM4 to 16QAM converter [5]. In the optical domain, we calculated the electric field of the optical signals and the probe light in HNLF consisting of the format converter using the split-step Fourier method [8]. The parameters of the HNLF were as follows: length of 2.5 km, dispersion parameter of 0.6 ps/nm/km, dispersion slope of 0.027 ps/nm$^2$/km, nonlinear coefficient of 13 W$^{-1}$/km, and fiber loss of 0.54 dB/km. At the receiver, the signal received by the coherent receiver with a sampling rate of
4 samples/symbol was input to the multi-stage adaptive equalizer shown in Fig. 2 (c). Each equalizer was trained using a decision-directed least mean squares (DD-LMS) algorithm. We set the tap size of the equalizer to 61, and employed the optimized step size that achieved the smallest error vector magnitude (EVM) for the training. We prepared the symbol sequences of a random sequence of length $2^{18}$ and pseudo random bit sequence (PRBS) with a length of $2^{17} - 1$ for the training and test, respectively.

Figure 3 (b) shows the EVM of the disaggregated QPSK signals under a varying number of equalizer stages. A three-stage equalizer including two stages for the QPSK signal 1 and one stage for the QPSK signal 2 was sufficient to disaggregate the 16QAM signal even when the QPSK (PAM4) signals were not symbol-synchronized. Figure 3 (c) shows the constellation diagrams of the disaggregated QPSK signals with the three-stage equalizer. The interference components of another QPSK signal were removed from the received signal, which was demodulated as a QPSK signal. Figure 3 (d) shows the bit error rate (BER) of the 16QAM signal (average BER of two disaggregated QPSK signals) for $\Delta T = 0, 0.25, \text{ and } 0.5$. As a comparison, the BER of a 16QAM signal with single-stage equalization is also shown. We assumed the use of a hard decision FEC with a 7% overhead, which has a BER $= 3.8 \times 10^{-3}$ limit for the output BER of $1.0 \times 10^{-13}$ [9]. The proposed multi-stage equalizer could improve the BER of the aggregated 16QAM signal compared to the single-stage equalizer. This is because the error bits of the equalized signal were eliminated by the FEC and then the replica was generated and used as input to the equalizer. The single-stage equalizer was unable to demodulate the received signal for $\Delta T = 0.5$ because it did not include a disaggregation process. In contrast, the proposed multi-stage equalizer could disaggregate the received signal into QPSK signals and demodulate them without an OSNR penalty at the FEC limit, even when the input PAM4 signals were not symbol-synchronized.

5 Conclusions

This study proposed a multi-stage equalizer for demodulating the optical aggregated 16QAM signal. We demonstrated a successful demodulation of the 100 Gbps class 16QAM signal without an OSNR penalty when the input PAM4 signals to the optical aggregation were not symbol-synchronized.

Acknowledgments

This work was partially supported by the Japan Society for the Promotion of Science KAKENHI (JP20K14733, JP20H04178), the Telecommunications Advancement Foundation, and the Fujikura Foundation.
Accurate beam tracking scheme for V2N transmission using vehicle logs

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Abstract: In 5G vehicular communications, beamforming technologies for vehicles, which compensate for large propagation losses at high carrier frequencies, are being considered. However, because the deflected beamforming loss degrades the communication quality, an accurate beam tracking system that follows the movement of the vehicle is required. In conventional schemes using a periodic codebook, the main lobe often deviates from the direction of the base station owing to fast movement. Therefore, in this study, a scheme for adaptively generating beam candidates using beam selection logs and the steering angle of the vehicle is proposed. Numerical results show that the scheme has high accuracy.

Keywords: 5G, beamforming, millimeter wave, autonomous driving

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1 Introduction

In recent years, there has been much research on vehicle-to-everything communications. The 28-GHz millimeter-wave (mmWave) band, which makes large-capacity communications possible, is supposed to be used for vehicular communications. However, the propagation loss of such a high-frequency band is high, resulting in a narrow coverage per base station (BS). Therefore, beamforming technology has been studied in BSs and vehicles experimentally [1, 2] to improve performance. However, in beamforming technologies, the signal strength decreases when the communication target is away from the main-lobe direction because of the fast mobility, and a beam tracking (BT) system that switches the beam with the movement of the vehicle is necessary. Conventional schemes using codebook-based beam selection have low accuracy [2]. Therefore, a vehicle-side high-accuracy beamforming scheme is proposed using beam selection logs and the steering angle of the vehicle. The proposed scheme predicts changes in the BS direction and adaptively generates beam measurement (BM) candidates. To the best of the authors’ knowledge, there has been no proposal for vehicle-side adaptive BT in mmWave communications. Numerical results show that the proposed scheme can achieve high accuracy compared with conventional schemes.

2 Proposed beam tracking scheme

2.1 System model

It is assumed that a vehicle and BS communicate, and the beam selection at the BS is ideal. The vehicle is assumed to be equipped with a square $8 \times 8$ uniform planar array (UPA) of antennas using half-wavelength spacing microstrip antennas (MSAs) as elements. First, $N$ candidate beam angles $b = \{b_0, b_1, \cdots, b_{N-1}\}$, $b_n \in \mathbb{R}$ are set at the vehicle side. Assuming the general frame format, as shown in Fig. 1 and described in detail in Section 3, one candidate beam is formed in the uplink-BM (ULBM) slots starting from $b_0$, and a reference signal (RS) is transmitted. At the BS, the average RS received power (RSRP) is measured, the beam index $r$ of $b_r$ that maximizes the average RSRP is fed back to the vehicle, and the vehicle uses $b_r$. The above operation is repeated at the ULBM slots to perform BT.

2.2 Proposed beam tracking scheme using vehicle logs

In a UPA, the beam steering vector $\psi(\theta, \phi)$ can be represented by the ground-based elevation $\theta \in [0^\circ, 90^\circ]$ and the azimuth $\phi \in [-180^\circ, 180^\circ]$, where $\phi$ is positive clockwise and negative counterclockwise at $0^\circ$ in the vehicular front direction. When performing BM, one of $\theta$ and $\phi$ is updated, and the other is fixed. Because the beam angle between the BS and vehicle varies more with $\phi$ than with $\theta$, 15 measurements of $\phi$ and one measurement of $\theta$ are considered as one search set.

The configurations described in the following section are empirically determined using various simulation courses to have versatility. The proposed scheme sets $b_n$ adaptively based on the past two $\psi(\theta, \phi)$ values and the current steering angle of the vehicle in the following three steps.

**Step 1:** The angle spacing parameter is set as $\omega (> 0^\circ)$ between adjacent $b_n$. Here, $\theta$ and $\phi$ are determined using the same algorithm. The stored angle measurement
logs of the second-to-last \( \psi_a \in \mathbb{R} \, [\text{deg}] \) and the last \( \psi_b \in \mathbb{R} \, [\text{deg}] \) (in the initial state, \( \psi_a = \psi_b = 0 \)) are used. When \( \psi_i \in \mathbb{R} \, [\text{deg}] \) is the exploration result for the \( i \) (\( i \in \mathbb{N} \))-th time, \( \psi_a \) and \( \psi_b \) are updated as follows:

\[
\begin{cases}
\psi_a = \psi_a, \psi_b = \psi_b & (\psi_{i-1} = \psi_b) \\
\psi_a = \psi_b, \psi_b = \psi_{i-1} & (\psi_{i-1} \neq \psi_b)
\end{cases}
\]  

(1)

Subsequently, the adaptive angle spacing parameter \( \Delta \psi \) is set by

\[
\Delta \psi = |\psi_a - \psi_b|.
\]

(2)

Given that the vehicle tends to change direction continuously, \( \omega \) is set adaptively using \( \Delta \psi \). However, if \( \psi_a \) and \( \psi_b \) are not updated continuously, there may be a slight fixed-angle error and insufficient resolution. The insufficient resolution is caused when \( \Delta \psi \) is large and the case of \( \psi_{i-1} = \psi_b \) lasts, even if more resolution is needed, e.g., the vehicle found the BS direction with a large \( \Delta \psi \) and was getting away from the BS. To address this, the count \( c_\delta \) in which \( \psi_a \) and \( \psi_b \) are fixed is introduced as a forgetting factor for \( \omega \), as shown in Equation (3). Furthermore, when \( \omega \) is extremely large or small, a quantization error may occur, even if a correct beam is selected. Therefore, \( \omega \) is set in a limited range.

\[
\omega = \begin{cases}
\Delta \psi(0.8^\omega) & (\text{for } \theta) \\
\Delta \psi B(0.96^\omega) & (\text{for } \phi)
\end{cases}
\]

\[
0.01 \leq \omega \leq 0.5 & (\text{for } \theta), \\
0.2 \leq \omega & (\text{for } \phi),
\]

(3)

where \( B \in \mathbb{R} \) is the azimuth bias obtained from the steering angle \( \delta_i \in \mathbb{R} \, [\text{deg}] \) in the \( i \)-th measurement, which is given by

\[
B = \begin{cases}
10.0 & (\delta_{i-1} = 0, \delta_i \neq 0) \\
1.15 & (\delta_{i-1} \neq 0, \delta_i \neq 0) \\
0.1 & (\delta_{i-1} \neq 0, \delta_i = 0) \\
1.0 & (\delta_{i-1} = 0, \delta_i = 0)
\end{cases}
\]

(4)

where \( \delta_i \) is positive on the right and negative on the left. When \( \delta_{i-1} = 0 \) and \( \delta_i \neq 0 \), the steering wheel starts to turn, and the change in the BS direction can be expected to increase rapidly. When \( \delta_{i-1} \neq 0 \) and \( \delta_i \neq 0 \), the steering wheel is still turning, and the change in the BS direction is large. Therefore, the measurement range is expanded by setting \( B \) to be larger. However, when \( \delta_{i-1} \neq 0 \) and \( \delta_i = 0 \), the steering wheel returns to neutral, and the amount of change in the BS direction is expected to decrease rapidly. When \( \delta_{i-1} = 0 \) and \( \delta_i = 0 \), no adjustment by \( B \) is made because the steering wheel is kept in neutral.

**Step 2:** Here, \( b_n \) is determined at equal angular spacing \( \omega \). Because one of the \( b_n \),\( s \) should be \( \psi_b \), the number of remaining beam candidates is \( N - 1 \); if \( N \) is an even number, the beam candidates cannot be allocated equally to the right and left of \( \psi_b \). Therefore, \( b_n \) for \( \theta \) and \( \phi \) is configured as

\[
b_n = \begin{cases}
\psi_b + \left( n - \frac{N}{2} \right) \omega & (\text{Case } 1, 3, 6, 7, 9) \\
\psi_b + \left( n - \left( \frac{N}{2} - 1 \right) \right) \omega & (\text{Case } 2, 4, 5, 8, 10)
\end{cases}
\]  

(5)
Here, $\psi_{b\phi}$ is the BM memory in the $\phi$ direction, and Cases 1 to 10 are as follows.

For configuration of elevation $\theta$:

Case 1: $i = 0$, $|\psi_{b\phi}| > 90$
Case 2: $i = 0$, $|\psi_{b\phi}| \leq 90$
Case 3: $i > 0$, $\psi_{b\phi} \leq 0$
Case 4: $i > 0$, $\psi_{b\phi} > 0$
Case 5: $i < 0$, $\psi_{b\phi} \leq 0$
Case 6: $i < 0$, $\psi_{b\phi} > 0$

For configuration of azimuth $\phi$:

Case 7: $i = 0$, $\psi_{b\phi} \leq 0$
Case 8: $i = 0$, $\psi_{b\phi} > 0$
Case 9: $i > 0$
Case 10: $i < 0$

In Cases 1, 3, and 6, the vehicle is expected to move away from the BS, whereas, in Cases 2, 4, and 5, it is expected to approach it. Therefore, additional $b_n$ are set to the expected direction — a decremental and an incremental $\theta$, as shown in Equation (5).

In Cases 7 and 9, the BS direction changes counterclockwise, whereas, in Cases 8 and 10, it changes clockwise. Hence, additional $b_n$ are set to decremental and incremental $\phi$.

**Step 3:** The beams are swept from $b_0$ to $b_{N-1}$ by transmitting RS at the ULBM slots from the vehicle. The BS measures the average RSRP, selects the largest RSRP beam $b_r$, and feeds back index $r$ to the vehicle. Subsequently, the uplink data are transmitted using $b_r$. This adaptive beam search process is periodically iterated according to the frame shown in Fig. 1, and the BT is achieved.

### 3 Numerical results

The frame structure used, as shown in Fig. 1, is based on the literature [1]. One frame consists of five subframes of 1 ms, and each subframe consists of 40 slots of 25 $\mu$s. The slot types defined are UL/DL BM slots, UL random access channel slots, UL/DL data slots, and the guard time, which is the preparation time for switching between UL and DL. One ULBM slot is inserted into subframes #1 to #4. In the proposed scheme, Steps 1–3 are performed in one frame to reduce the BM overhead. Therefore, $N = 4$ was set.

![Fig. 1. Frame structure in vehicular communications.](image-url)
The performance of the proposed scheme was evaluated using the driving course shown in Fig. 2(a). This is the central block of the nine blocks of the urban Manhattan grid model used in the third-generation partnership project (3GPP). The vehicle speed was set to a maximum of 60 km/h and a minimum of 15 km/h when going straight and at a constant speed of 15 km/h when turning. An environment without buildings was assumed. Figure 2(b) shows the vehicle antenna layout with four UPAs [3]. The antenna of \{0\} was placed at a tilt angle of 30° to the horizontal plane, and those of \{1\}–\{3\} at a tilt angle of 90°. The antennas were switched when the sweep range of the antennas shown in the figure was exceeded. The antenna switching interval was assumed to be 2 s. For the MSA design, the frequency was set to 28.0 GHz, the element length and width were set to 2.77 mm, and the substrate thickness and relative permittivity were set to 0.5 mm and 2.6, respectively, according to the literature [4]. The transmitting power was 23 dBm, feed loss was 5.0 dB, and transmitter antenna height was 1.6 m [5].

Figure 3 shows the cumulative distribution function (CDF) of the received power at BS. The proposed system using vehicle logs has a lower probability of being received at low power than the conventional codebook-based system [2], and the proposed scheme improves the received power by approximately 4 dB at a 50% CDF value. This is because the proposed scheme makes a highly accurate BT and low switching delay possible by generating BM candidates according to the change in the target direction. The figure also shows the case in which the MSA single element is mounted on the rooftop of a vehicle for reference. This result indicates that the application of beamforming technology is essential for mmWave communications.

4 Conclusions

A highly accurate BT system was proposed that exploits vehicle logs. In conventional systems, BT is performed according to a codebook, which sometimes results in a decrease in the received power owing to the deviation between the main lobe and the
target direction. In contrast, with the proposed system, the change in beam direction was predicted based on vehicle logs and adaptively generated fitting BM candidates. Through numerical simulations, it was shown that the proposed method is suitable for mmWave BT.
Multi-channel FM transmission of vibrotactile signals on 2-D communication textile

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Abstract: This paper proposes a frequency-division multiplexed vibrotactile signal transfer system on conductive textiles for implementing wearable tactile suits. For multiplexing, many carrier frequencies are in one-to-one correspondence with many receivers. Each carrier signal is frequency-modulated with analog vibrotactile waveforms, and each receiver demodulates it and drives its built-in actuator. The analog waveform transmission achieves multi-channel real-time vibrotactile actuation, which is suitable for applications such as virtual reality (VR) games. A commercially available frequency modulation (FM) radio receiver chip achieves high sensitivity and superior channel selectivity. Carrier frequencies can be located in a 32 MHz bandwidth, from 76 MHz to 108 MHz, every 200 kHz, and 160 channels of vibrotactile waveforms can be transmitted simultaneously.

Keywords: two-dimensional communication, wearable system, haptics, virtual reality

Classification: Wireless Communication Technologies

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1 Introduction

With the widespread virtual reality technologies, tactile sensing/actuating has attracted much interest. We can find some research works investigating wearable tactile suits in the literature [1, 2]. Many vibrotactile actuators distributed over the whole body require many cables for power supply and control signal transfer.

Conductive-textile-based power and data transfer methods have been proposed to eliminate one-to-one wirings [3, 4, 5, 6]. Those research works include synchronous serial data transfer [6, 7] and asynchronous transfer [5]. They are all digital communication schemes, and no analog communication on conductive textile has been reported to the best of the author’s knowledge.

The main contribution of this paper is to demonstrate the feasibility of simultaneous transmission of many vibrotactile waveforms to antenna-less receivers over conductive textiles. The previous works developed the analog front-end specialized for over-textile communication, while the digital interface was integrated into a commercially off-the-shelf (COTS) microcontroller chip. The high-frequency analog circuit was composed of some discrete and integrated components. Implementing a sophisticated signal processing circuit that achieves high spectral efficiency under the restriction of circuit footprint was impractical. Contrarily, in the system presented in this paper, a COTS single-chip FM receiver completes the high-frequency signal processing, thus enabling spectrally efficient multiplexing with a small circuit.

The proposed method can achieve real-time vibrotactile waveform transmission without digital data buffering or signal compression/expansion. Receivers implemented with a COTS one-chip FM receiver IC achieve a small footprint, high sensitivity, and high reliability. The superior channel selectivity of the receiver chip enables dense carrier location with a 200 kHz interval. Using a receiver IC compatible with worldwide broadcast FM radio frequencies from 76 MHz to 108 MHz, 160 channels of independent vibrotactile signals can be transferred simultaneously.

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2 Vibrotactile signal transfer over textile transmission line

VR applications require higher tactile sensation fidelity than cell phone vibration alerts. An eccentric rotating mass (ERM) vibration motor, commonly used in mobile devices, vibrates at a fixed frequency when a constant voltage is applied. It is cost-effective for notifying phone calls, but other actuators such as voice coil actuators (VCAs) are more suitable for applications requiring higher fidelity of tactile sensation.

In a previously reported inter-IC for wearables (I²We)-based wearable tactile suit [8], the VCA-loaded receivers play back vibrotactile waveforms preset in the flash memory. The controller transmits only short digital frames specifying the waveform to be played back, instead of the analog waveforms. This approach reduces the data transferred over the textile, but the tactile representation is limited to only the combinations of the preset waveforms.

The analog signal transfer system presented in this paper, shown in Fig. 1, supports transferring arbitrary waveforms to drive such actuators in real-time. The system enables real-time generation and display of arbitrary vibrotactile stimuli without being limited to preset waveforms. The frequency range of vibrotactile signals is up to about 1 kHz [9]. The audio signal bandwidth of FM receiver IC, 15 kHz [10], is sufficient for the vibrotactile signal transfer.

3 Multi-carrier FM signal transfer

In the proposed FM transmission system, a one-chip FM radio receiver IC completes the entire signal processing from the radio frequency (RF) input to the vibrotactile waveform output. The special-purpose IC optimized for FM receiver yields the advantages of high sensitivity, high immunity to fading, and high channel selectivity with a small footprint. The selectivity between two adjacent channels with a 200 kHz interval is 50 dB [10]. The remarkably high selectivity enables many signal

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**Fig. 1.** Experiment system of the proposed FM vibrotactile signal transfer and prototyped RX module closeup. RX module is 20 mm in diameter. The FM radio receiver IC completes the signal processing from RF to vibrotactile waveform. The RX module is composed of the FM RX IC, microcontroller (MCU) to configure the IC, and a class-D amplifier to drive the VCA.
multiplexing with high carrier density.

In the previously reported I²We systems, two carriers were modulated by two pulse sequences of I²C data and clock. Simple resonant bandpass filters were composed of discrete chip inductors and capacitors to distinguish the two carriers with a small footprint. Due to the broad passband characteristics of the filters with relatively low-quality-factor components, two carriers must be chosen to have a significant interval [6]. Therefore, adding another I²C bus on the same textile transmission line by frequency-division multiplexing (FDM), i.e., using additional carriers, requires occupying an order of magnitude broader frequency range. Thus, increasing the communication channel capacity by FDM was impractical.

Using the FM carrier frequencies as many as the RX modules, the system can uniquely assign one carrier frequency to one RX module. Based on this scheme, each RX can be simply tuned to a fixed carrier frequency, and no dynamic frequency exchange among many RXs is required. When the transmitter sends the modulated signal with one of the carrier frequencies, the corresponding RX immediately receives the signal, without any overhead process such as listen-before-talk or addressing. This is one of unique features of the proposed scheme compared with the serial-communication-based I²We.

4 Prototype evaluation

Two-channel vibrotactile signal transfer using two carriers with the 200 kHz interval was evaluated in the system shown in Fig. 1. The schematic diagram of the system is shown in Fig. 2(a). The configuration of the FM receiver ICs including tuning frequency and output volume are preliminarily programmed into each on-board microcontroller (MCU) of the RX modules. The waveforms of the two baseband vibrotactile signals and two demodulated currents flowing through the RX VCAs are shown in the Fig. 2(b). The waveforms were obtained with the FM carrier frequencies \( f_1 = 80.0 \text{ MHz} \) and \( f_2 = 80.2 \text{ MHz} \), and carrier magnitudes \( V_1 = V_2 = -40 \text{ dBV} \), where the magnitudes are represented in open circuit voltage. The spectrogram of the FM signals actually generated under the configuration is shown in Fig. 2(c).

The result shows that the two RXs almost accurately demodulate the original vibrotactile waveforms. By looking closer, there exists crosstalk between the two current waveforms. When RX1 current reaches the peak, RX2 current fluctuates, and vice versa. This crosstalk is due to the DC power supply voltage fluctuation at each RX. All RXs share the same power supply voltage applied to the textile transmission line; therefore, when one of the RXs instantly draws a significant current for actuator operation, the DC voltage on the textile drops, and all the RXs experience the supply voltage drop. To clarify this effect, when RX1 is removed from the system and the two carrier strengths are kept equal, i.e., \( V_1 = V_2 = -40 \text{ dBV} \), no crosstalk appears on the RX2 current waveform as shown in Fig. 3. Therefore, the crosstalk seen in Fig. 2(b) is independent of the interfering adjacent channel carrier at the same signal level. This is a problem with the DC voltage regulating, but not with the mechanism of FM transmission. It can be resolved by implementing an on-board voltage regulator on each RX module.

The RX2 current waveforms with increased interfering adjacent channel strength
Fig. 2. Experiment system (a) and measured waveforms (b). Baseband signals for FM modulator and demodulated current flowing through VCAs in RX1 and RX2 are shown in (b). Two actuators are independently controlled. With these baseband signals, FM-modulated RF signals are generated as shown in a spectrogram (c). The spectrogram was obtained by connecting a spectrum analyzer at point A shown in (a).
Fig. 3. Crosstalk between RX1 and RX2 caused by the power supply voltage fluctuation due to RX1/RX2 load currents disappears when RX1 is removed from the textile transmission line. Interference of the adjacent 80.0 MHz channel to the desired 80.2 MHz channel, observed on the RX2 load current waveform, is almost negligible for the interference wave of up to about 50 dB greater than the desired wave.

are also shown in Fig. 3. The result shows that the interference to RX2 load current waveform is almost rejected for the interference wave up to about 50 dB greater than the desired wave.

Thus, interference-free demodulation under a wide range of interfering carrier strength has been confirmed. Therefore, the proposed multi-channel FM transmission system enables transferring a large number of vibrotactile waveforms simultaneously using densely located carriers. In practical textile two-dimensional communication systems, standing waves are generated in the textile because the edges are not terminated with a matched load. Therefore, the carrier strength varies depending on the receiver position. Even with such a spatial fading, the wide-range interfering carrier rejection capability will provide robust and stable demodulation.
performance.

5 Conclusion and prospects
We have presented a multi-channel vibrotactile actuator control system using frequency modulation for real-time analog waveform transfer instead of digital signal transfer. Using an FM radio receiver IC, up to 160-channel vibrotactile waveforms can be transferred simultaneously. In the experiment, the signal transmitter was built with benchtop signal generators. Future work will be to integrate the multi-channel FM transmitter into a small, battery-driven circuit module.

Another critical issue of battery-driven wireless wearable system is the battery life. The measured quiescent power consumption of the RX module was approximately 150 mW, which is slightly higher than the 110 mW power consumption of the I2We RX. Additionally, stronger and more power-consuming actuators will be suitable in applications requiring higher fidelity and a wider dynamic range of tactile sensation. Reducing power consumption for longer battery life will be one of the critical issues in future work.

Acknowledgments
This work was supported in part by JSPS KAKENHI Grant Number JP20H04182, MIC/SCOPE #JP205006003, and JST PRESTO Grant Number JPMJPR18J7. The jackets made of double-sided conductive textiles were provided by Teijin Limited, Tokyo, Japan.
Rolling-shutter-sensor-based visible light communication for multi-user long-range communication and positioning

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Abstract: We propose a rolling-shutter-sensor-based visible light communication (RS-VLC) system using a cross-screen filter suitable for intelligent transport system (ITS) scenarios. The proposed system uses a cross-screen filter to diffuse the received light in the scanning direction of the rolling-shutter sensor, thus improving the intensity of the received signal and increasing the communication range. The experimental results show that the proposed system can improve the communication range. We also show that the proposed system is capable of multi-user communication and positioning.

Keywords: visible light communication, ITS, rolling-shutter sensor, cross-screen filter, phase-shift keying

Classification: Wireless Communication Technologies

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1 Introduction

Visible light communication (VLC), which utilizes an LED as a transmitter (Tx) and an image sensor as a receiver (Rx), has been attracting attention as a new communication technology in intelligent transport system (ITS) scenarios (e.g., vehicle-to-vehicle communication), since it offers the advantages of multi-user communication and positioning by image information [1, 2]. A VLC system that achieves a high data rate using the rolling-shutter effect of a commercial camera (rolling-shutter-sensor-based VLC; RS-VLC) can not only enable these functions but also use a smartphone camera as a VLC Rx [3]. However, it is challenging to achieve both long-distance and high-speed communication in existing RS-VLC systems, and the current communication range is limited to the order of several meters [4]. Therefore, to utilize an RS-VLC system for ITS, it is necessary to achieve long-distance and high-speed communication while enabling multi-user communication and positioning.

To address this issue, we proposed a single-user visible light communication system using a cross-screen filter [5]. In [5], it was found that this filter can improve the communication range by diffusing the received light in the scanning direction of the rolling shutter sensor and improving the received signal power. In this paper, we study the feasibility of multi-user communication and positioning using the proposed system, which was not clarified in [5]. In the following section, we design the proposed system, perform experiments with it, and evaluate its performance.

2 System overview

Figure 1 shows a block diagram of the proposed RS-VLC system using a cross-screen filter. Below we explain signal processing in the Tx and Rx (the detailed configuration of the transmitter and receiver is described in [6]).

![Fig. 1. Block diagram of RS-VLC system using cross-screen filter and vehicle-to-vehicle communication in ITS scenario.](image-url)
2.1 Transmitter
Let us define a binary data block \( b_l = (b_{l,0}, b_{l,1}, \ldots, b_{l,N-1}) \), where \( b_{l,n} = \{0, 1\} \), \( l = 0, 1, \ldots, L - 1 \), \( n = 0, 1, \ldots, N - 1 \), \( l \) is the block number, \( L \) is the total number of data blocks, and \( N \) is the block length. The Tx converts the binary data \( b_{l,n} \) to the symbol data \( d_{l,n} \), as

\[
d_{l,n} = \begin{cases} 
1 & (b_{l,n} = 1) \\
-1 & (b_{l,n} = 0)
\end{cases}.
\]  

(1)

Next, the Tx performs phase-shift keying (PSK) modulation by multiplying the symbol data \( d_{l,n} \) and the carrier signal by the frequency of \( f_c \). The PSK modulated payload signal \( p_l(t') (0 \leq t' \leq T_d, T_d: \text{length of payload signal in time}) \) can be expressed as

\[
p_l(t') = a \Re \left[ d_{l,n} e^{j 2\pi f_c t'} \right] + a_{\text{off}} \left( \frac{n}{N} T_d \leq t' < \frac{n + 1}{N} T_d \right),
\]

where \( a \) and \( a_{\text{off}} \) are the amplitude and DC offset of the signal, respectively. Then, the Tx calculates a transmitted signal block \( s_l(t') (0 \leq t' \leq T_b, T_b: \text{length of signal block in time}) \) by adding a header (a known sequence of data used to identify the start of a data block) on \( p_l(t') \) and transmits a signal as a fast-blinking light from an LED.

2.2 Receiver
The Rx diffuses the incoming light by a cross-screen filter, captures the incoming light using a rolling-shutter sensor as images, and converts the images to a signal. Then, the Rx obtains a received payload signal that contains information on the transmitted signal block \( s_l(t') \) by detecting a header, performs PSK demodulation, and obtains binary data block \( b_l \).

Figure 2(a) shows the relationship between the transmitted and received signals in the previous RS-VLC system (without a cross-screen filter). As the figure shows,
in the previous system the incoming light is captured directly by the rolling-shutter sensor. In a rolling-shutter sensor, only the area of the light source captured on the sensor has signal power strong enough to be suitable for communication. Therefore, the previous system is used mainly for short-range communication, where the light source exists over the entire sensor area (the longer the communication range, the smaller the area with large signal power, resulting in poor communication quality).

Figure 2(b) shows the relationship between the transmitted and received signals in the proposed RS-VLC system (with a cross-screen filter). In the proposed system, the incoming light passes through the cross-screen filter before it is captured by the rolling-shutter sensor. This filter is an optical lens with a slit in one direction, and it diffuses the incoming light in the scanning direction of the sensor. Therefore, it is possible to have a large signal power area on the sensor even if the communication range becomes long, resulting in the improvement of the bit error rate (BER). Moreover, since the proposed system achieves communication while focusing on the transmitted light sources and suppressing their interference, it would be possible to realize multi-user communication and positioning simultaneously.

3 Experiment

3.1 System setup

We evaluated the performance of the proposed system in an indoor experiment. Figures 3(a) and 3(b) show the experimental environment and parameters, respectively. As the figure shows, both the light source of Tx#1 and that of Tx#2 were placed horizontally because we assume that the transmitter is located on a specific plane considering the vehicle-to-vehicle communication scenario. In this experiment, both Tx#1 and Tx#2 convert random binary data of 10,000 bits, perform PSK modulation, and emit a signal from each LED. The Rx captures images by the sensor with a frame rate of 30 fps and detects the light sources of Tx#1 and Tx#2 by image processing. The Rx then reads the pixel values of the pixel rows above the detected light source and converts them into signals from Tx#1 and Tx#2, performs PSK demodulation, and calculates the BER for each communication range $d$ (Tx#1-Rx distance). Note that the Tx#2-Rx distance is fixed and that the experimental parameters (e.g., transmitted signal power and angle of view) of the previous system and the proposed system are equalized.

3.2 Experimental results and discussion

Figures 3(c)–3(e) show the experiment results. Figures 3(c-1) and 3(c-2) show an example of a received image (when $d = 1,000$) using the proposed system and using the previous system, respectively. The figure also indicates the center coordinates of the transmitted light source detected by image processing [7]. As the figures show, both the proposed and previous systems successfully detect the coordinates of the transmitted light source from the images.

Figures 3(d-1) and 3(d-2) show the relationships between a pixel and the pixel values of Tx#1 and Tx#2, respectively (when $d = 1,000$). As the figures show, the pixel value of the proposed system (blue line) is larger than that of the previous system (red line). This is because the cross-screen filter can diffuse the transmitted
light in the scanning direction of the rolling-shutter sensor, as described in Section 2.

Figure 3(e-1) shows the relationship between the communication range $d$ and the BER of Tx#1. In the previous system, the BER of Tx#1 (red line) is less than $10^{-3}$ when $d$ is less than 700 mm. In the proposed system, on the other hand, the BER of Tx#1 (blue line) is less than $10^{-3}$ when $d$ is less than 1,400 mm. Therefore, we found that the proposed system can improve the communication range compared to the previous system.
Figure 3(e-2) shows the relationship between the communication range $d$ and the BER of Tx#2. As the figure shows, the BER of Tx#2 decreases when $d$ increases. This is because when $d$ is large, the interference of the Tx#1 signal with Tx#2 is small, resulting in an increase in the signal-to-noise ratio of Tx#2. In the proposed system, the BER of Tx#2 (blue line) is less than $10^{-3}$ at all distances except 500 mm, which is lower than that of the previous system (red line). This is because the cross-screen filter improves the power of the received signal. Therefore, we confirmed that the proposed system is tolerant against interference from other light sources and is capable of multi-user communication.

4 Conclusion

We proposed an RS-VLC system using a cross-screen filter suitable for ITS applications. The previous systems can achieve only short-range and single-user communication, and long-range multi-user communication and positioning are challenging. To address these issues, we proposed a VLC Rx with a cross-screen filter that diffruses the received light in a specific direction. We evaluated the performance of the proposed system in experiments. The experimental results suggested that the proposed system outperforms the previous system in terms of the received signal power and BER while improving communication range compared to the previous system. We also showed that the coordinates of the transmitted light source can be calculated accurately by processing the images captured by the proposed system and that multi-user communication is possible. Consequently, we conclude that the proposed system can become a viable means of establishing wireless communication links and is suitable for ITS scenarios including vehicle-to-vehicle communication. On the other hand, this research is still an initial study on the RS-VLC system suitable for ITS scenarios. Therefore, it is necessary to conduct a more detailed study through experiments in real environments in the future.

Acknowledgments

This work was supported by JSPS KAKENHI Grant Number JP18K19774.
Improved DFT-based channel estimation for spatial modulated orthogonal frequency division multiplexing systems

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Abstract: In current cellular networks, Spatial Modulation with Orthogonal Frequency Division Multiplexing (SM OFDM) is a newly designed transmission method that substitutes Multiple Input Multiple Output (MIMO) OFDM transmission. In practical frameworks, estimation of the channel is vital for recovering the transmitted data. In a conventional DFT-based channel estimation algorithm, impact of noise is reduced in the time domain. It retains the significant values with some sort of threshold value equal to the Cyclic Prefix (CP) length that differentiates the signal values and noise terms of Channel Impulse Response (CIR), thereby improving the performance in OFDM systems. Here we require information on CIR length in advance. But this results in performance degradation if the path delays are less for a channel. Hence this work proposes improved DFT-based channel estimation where the threshold value is set equal to the estimated CIR length using RMS delay spread approximation in SM OFDM systems. Symbol Error Rate (SER) and Mean Square Error (MSE) are the performance metrics considered in Rayleigh fading channel employing ITU- Vehicular A channel model for 16 QAM SM OFDM systems. Simulation output shows performance improvement of SER and MSE by the implementation of improved DFT-based channel estimation. At an SNR of 40 dB, the proposed method can enhance SER performance by about 2.8 dB and MSE performance about 4dB compared with conventional DFT estimation.

Keywords: MIMO, SM OFDM, DFT, channel model, RMS delay spread

Classification: Wireless Communication Technologies

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1 Introduction

The upcoming cellular networks requirements include high throughput, less complexity, and flexibility. Combining OFDM with MIMO is favorable for future generation wireless systems that improve spectral efficiency and link reliability [1]. MIMO transmission capacity gain relies on transmitting and receiving antenna separation, synchronizing transmit antennas, and algorithms to minimize Inter Channel Interference (ICI) at the receiver input [1]. The SM OFDM technique is an alternative approach to alleviate the MIMO-OFDM problem [2]. It completely avoids ICI at the input of the receiver. It employs the transmit antenna index to convey information, the spatial dimension, and the two-dimensional signal constellation. It saves implementation cost by minimizing the count of radio frequency chains [2]. Channel estimation is vital for the systematic implementation of a coherent receiver since perfect Channel State Information (CSI) is always impractical. Non-selectivity assumption over frequency selective fading channels is not suitable for SM OFDM systems. Generally, pilot symbol-aided channel estimation techniques where pilot symbols are inserted at regular intervals along with data symbols are commonly
used to get CSI at the receiver side for multipath fading channels [3]. Least Square (LS) estimation is done to obtain the estimates at pilot locations. Authors Acar Yusuf et.al discusses pilot symbol aided estimation of the channel for a SM OFDM system using LS estimate along with different types of interpolations [4]. To enhance the accuracy of channel estimates obtained through LS estimates, DFT-based channel estimation is used [5]. The conventional rectangular window DFT spreads energy to its neighboring lobes and increases leakage due to discontinuities at the edges of window [6]. Authors A.M Soman et.al discussed an improved DFT-based channel estimation technique by using frequency domain data windowing and time domain weighting function concept in minimizing leakage and channel estimation error and improving estimation performance [7]. This work proposes an improved channel estimation using DFT with CIR length estimation using RMS delay spread approximation in SM OFDM systems

2 System model

Figure 1 illustrates system model of SM OFDM. Let $V(k)$ be a $b \times N$ data matrix to be transmitted. Here $b$ denotes the total number of bits/symbols that belongs to a subcarrier out of $N$ total subcarriers. This $V(k)$ data matrix then gets mapped to a $N_T \times N$ matrix $Z(k)$ using an SM mapper [4]. Here $N_T$ represents the total number of antennas at transmitter side. The entries of $Z(k)$ matrix remain zero excluding the mapped location indices of antenna used for transmission. So, the total transmitted bits on each sub channel are given as [4].

$$p = \log_2(N_T) + \log_2 Q$$  (1)

where $Q$ is the modulation degree.

Fig. 1. System model

Each row vector from $Z(k)$ matrix is then modulated using an OFDM modulator. It includes pilot insertion which supports estimation of channel at the receiver side,
Inverse Discrete Fourier Transform (IDFT) operation and CP insertion to avoid distortion produced by Inter Symbol Interference (ISI) in the channel. The resultant OFDM symbol is then converted to serial data and gets transmitted through MIMO channel. The time-domain channel output response between \( j^{th} \) transmit antenna and \( r^{th} \) receive antenna in a multipath fading environment is written as

\[
h_{jr}(n) = \sum_{l=0}^{L-1} h_{jr,l}(n - \tau_{jr,l})
\]  

(2)

where \( L \) represents the number of channel paths, \( h_{jr,l} \) represents time-varying channel coefficients and \( \tau_{jr,l} \) represents the delay of the \( l^{th} \) path. At receiver, demodulation takes place. This includes CP removal, pilot extraction, channel estimation, equalization and detection so that the data is obtained as transmitted. The received signal matrix is written as

\[
Y(k) = H(k)x_j(k) + W(k)k = 1,2,\ldots N
\]  

(3)

Where \( x_j(k) = \begin{bmatrix} 0 & \ldots & x_q(k), & \ldots & 0 \end{bmatrix}^T \) \( x_q(k) \) is the \( q^{th} \) active antenna symbol stream and \( W(k) \) is Additive White Gaussian Noise (AWGN) having mean zero and variance \( \sigma_w^2 \). The Maximum Likelihood (ML) detection process involves finding the transmit antenna index and the transmitted symbol sends on it. The ML detection rule is

\[
\hat{j}_{ML}, \hat{q}_{ML} = \arg \max_{j,q} \| h_{jr}(k)x_q(k) \|_F^2 - 2\Re \{ y^*(k)h_{jr}(k)x_q(k) \}
\]  

(4)

for \( 1 \leq j \leq N_t, 1 \leq q \leq M \). Here \( h_{jr}(k) \) denotes the \( j^{th} \) column of \( H(k) \) and \( x_q(k) \) denotes \( q^{th} \) active antenna symbols from constellation diagram.

3 Channel estimation for SM OFDM

Channel Estimation is essential for the demodulation process. If the CSI is not available, using LS Estimator, the frequency response of known pilot position is obtained and using interpolation methods the channel frequency responses at data positions are obtained. The pilot-based frame format of SM OFDM is such that the pilot symbols are periodically placed over subcarriers. As a result, each transmit antenna transmits pilot symbols for each OFDM symbol number. Therefore, for each OFDM symbol, the received signals at pilot subcarrier \( (k_p) \) is written as

\[
Y(k_p) = H(k_p) \rho + W(k_p)
\]  

(5)

where \( \rho \) represents the pilot symbol. From Eq. (5) initial channel estimate is derived for pilot subcarriers using LS estimate as

\[
\hat{H}(k_p)_{LS} = Y(k_p)/\rho
\]  

(6)

Generally, DFT based Channel Estimation is used to increase LS estimation accuracy. Applying orthogonal transformation to LS estimates, the transform domain has \( L \)
number of most significant values. Retaining the most significant values and treating non-significant values as zeros the noise effect outside the maximum CIR length $L_{\text{max}}$ will be eliminated and hence accuracy is improved [5].

Here we assume that $L$ is known in advance. Hence some kind of threshold is required to differentiate between signal values and noise terms. Generally, this threshold value is set to be equal to the CP length that differentiates noise components and most significant values of signal in CP length of CIR and improves the performance. But this results in performance degradation if the path delays are less for a channel. The block diagram representation of DFT estimator is illustrated in Fig. 2. Mathematically, if $\hat{H}(k)$ denote channel estimate obtained at $k^{th}$ subcarrier by LS estimation, then IDFT of the channel estimate is written as [5]

$$\text{IDFT}\{\hat{H}(k)\} = h(n) + \bar{w}(n)\Delta h(n)n = 0, 1, 2 \ldots N - 1$$  \hspace{1cm} (7)

Here $\bar{w}(n)$ denote noise term in time domain. Neglecting coefficients of $h(n)$ that contain noise term by depicting the coefficients for maximum CIR length $L_{\text{max}}$, $\hat{h}_{\text{DFT}}(n)$ is written as [5]

$$\hat{h}_{\text{DFT}}(n) = \begin{cases} h(n) + \bar{w}(n)n = 0, 1, 2 \ldots \ldots L_{\text{max}} - 1 \\ 0, \text{otherwise} \end{cases}$$  \hspace{1cm} (8)

Lastly, DFT based estimation is written as

$$\hat{H}_{\text{DFT}}[k] = \text{DFT}\{\hat{h}_{\text{DFT}}(n)\}$$  \hspace{1cm} (9)

\begin{figure}[h]
    \centering
    \includegraphics[width=\textwidth]{channel_estimation_diagram.png}
    \caption{Channel estimator based on DFT}
\end{figure}

4 Proposed channel estimation

This letter proposes an improved DFT-based channel estimation method for SM OFDM systems which estimates CIR length $L$ using RMS delay spread estimation ($\sigma_T$). This estimated CIR length is set as the threshold in DFT based estimation. The result is compared with that of conventional DFT method. The improved DFT-based channel estimation comprise of the following steps.

Step 1: Initial channel estimate is derived for pilot subcarriers using LS estimate based on Eq. (6). This information on channels frequency selectivity can be used to estimate $\sigma_T$. 

Step 2: Carry out IDFT to convert estimated frequency response coefficients to time domain impulse response. The time domain channel taps obtained are assigned into a sub region. Outside the sub region, the terms corresponding to noise are made zeros and only most significant taps are retained.

Step 3: In the proposed work, the length of significant tap is calculated to be equal to the estimated CIR length using RMS delay spread estimation. From [8] RMS Delay spread is

$$\sigma_\tau = \left( \frac{\sum_{l=0}^{L-1} P_l (\tau_l - \overline{\tau})^2}{\sum_{l=0}^{L-1} P_l} \right)^{1/2}$$

(10)

where $P_l$, $\tau_l$ and $\overline{\tau}$ represents the power of $l^{th}$ path, arrival time of $l^{th}$ path and average delay of multipath. Here the frequency selective channel is expected to be uniform over an OFDM symbol and fluctuates between OFDM symbols in order to estimate the RMS delay spread. Then, the instantaneous channel frequency correlation values is determined from the LS channel estimates as

$$\hat{R}_H(l) = E_{k_p} \{ \hat{H}_{LS}(k_p) \hat{H}_{LS}(k_p + l) \}$$

(11)

where $E_{k_p}$ is the mean with respect to $k_p$, $l$ is the subcarrier separation. As these estimates are noisy it needs to be averaged over several OFDM symbols. After mathematical analysis, the averaged correlation estimate is derived as

$$\tilde{R}_H(l) = \begin{cases} R_H(l) & if \ l \neq 0 \\ R_H(0) + \sigma_w^2 i f l = 0 \end{cases}$$

(12)

The Power Delay Profile (PDP) estimate is then derived by taking IDFT of averaged correlation estimate.

$$P_l = IDFT \left\{ \tilde{R}_H(l) \right\}$$

(13)

RMS delay spread is then estimated using Eq. (10). The CIR length is estimated based on RMS delay spread of channel.

$$\text{Length of CIR} = \beta \times \sigma_\tau$$

(14)

where $\beta$ is constant in the range between 2 to 4 as a rule of thumb [9, 10].

Step 4: Set estimated CIR length as the threshold in DFT channel estimation and compare with that of conventional method.

Step 5: The noise reduced signal is converted back to Frequency domain by DFT.

5 Performance analysis

This section quantifies the SER and MSE performance of SM OFDM systems based on the proposed channel estimation algorithm under Rayleigh channel, employing ITU Vehicular A Channel model [11] compared to conventional DFT method. The simulation parameters are set for a System bandwidth 10MHz, carrier frequency 2.4GHz, Number of subcarriers as 256, Pilot ratio 1/8, Actual channel length (L) as 6 and Number of transmitter and receiver antennas as 4. Figure 3 shows the
simulation results where Fig. 3a) shows the initial estimates of channel done by LS estimation of pilot signals, Fig. 3b) shows PDP of ITU Vehicular A channel model and estimated PDP through frequency correlation of LS estimated channel and Fig. 3c) and Fig. 3d) shows SER and MSE performance. It is observed that the estimated channel consists of channel coefficients and noise coefficient where both are statistically independent. Also the SER performance of proposed method and conventional DFT method is increased gradually for an SNR range from 0 to 40dB. SER performance for perfect CSI is also shown for reference. For perfect CSI, channel estimation does not require pilot symbols. But comparing the proposed DFT based estimation method from perfect CSI case strongly depends on pilot symbols. Increasing pilot symbols in subcarrier increases the SER performance to perfect CSI case but with less data rate. Initially, for low SNR values, conventional DFT performs better, but as SNR is increased, the SER performance of the proposed method is enhanced remarkably compared to conventional method. For 16 QAM, the SER performance of the proposed method is improved by about 2.8 dB for SNR value 40 dB. From the MSE graph, the proposed DFT method performance is more acceptable than the conventional DFT method. When SNR is low, both shows similar values of MSE. When SNR is higher than 10dB, there is reduction in values of MSE for the proposed method and it is improved up to 4dB at 40 dB SNR compared to the conventional method.

![Fig. 3. Simulation results](image-url)
6 Conclusions

This work addressed improved channel estimation using DFT for SM OFDM communication systems. We have focused on estimating the Channel Impulse Response (CIR) length based on RMS delay spread approximation. In practice, the length of CIR is much smaller and is proportionate to the number of delay taps in PDP of the channel. To obtain RMS delay spread, initially, PDP is estimated from the channel frequency correlation function of the initial channel estimate. Then the estimated CIR length is calculated. It is observed that SER performance and MSE performance of the proposed estimation method is more acceptable and can produce justifiable results than the conventional estimation method and hence shows great possibilities to satisfy the requirements of new wireless communication systems.
Capacity enhancement on optical camera communication with clock divider-mounted multi-cameras

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Abstract: This paper proposes a technique of sampling-rate retrieval using multi-cameras with a lower frame rate than the symbol rate on an optical camera communication (OCC) with a clock divider. This proposed technique applies the characteristics of the integral sampling method that occurs when an image sensor-based camera receives the optical signal. The technique realizes the demodulation of the optical signal having a larger bit rate than the frame rate of the camera. Employing the proposed technique can achieve the transmission capacity of the number of cameras times frame rate per camera.

Keywords: visible light communication (VLC), optical camera communication (OCC), light fidelity (LiFi), image sensor, integral sampling, wavelength division multiplexing (WDM)

Classification: Wireless Communication Technologies

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1 Introduction

Many types of visible light communication (VLC) systems have been proposed [1, 2]. In particular, a CMOS image sensor-based VLC system has been getting attention [3]. The system is termed as optical camera communication (OCC) systems [4]. The OCC consists of an LED light as a transmitter and a commercially supported image sensor-based camera as the photodetector of a receiver. The detection of the optical signal is not complicated because the camera can detect multiple luminous bodies simultaneously and separate the image processing. However, a frame rate of the camera limits the symbol rate of the VLC link, which is 60 frame-per-second (fps) as an example.

The techniques to increase the bit rate is that the wavelength division multiplexing (WDM) scheme using an RGB color LED, the transmitter consisting of multiple luminous bodies, and using the high-speed camera of more than 1,000 fps [5]. Moreover, a rolling shutter method using the characteristics of the CMOS image sensor is an interesting solution [3].

In addition to these techniques, this paper proposes the demodulation and decoding method by using synchronous multiple image sensors. The multiple sensors are synchronized and adjusted the timing of the signal sampling. Each sensor has its shooting timing shifted symbol-by-symbol. The optical signal can be demodulated even when the frame rate per image sensor is lower than the symbol rate of the transmitter. We indicate the concept of this method for the first time and reveal the fundamental characteristics of bit error rate (BER) versus signal-to-noise power ratio (SNR).

2 Modulation and demodulation on OCC

This section introduces an integral sampling scheme to demodulate the OCC signal. In general, the camera continues to concentrate lights during the exposure time. That is, when we use the camera with the frame rate of 60 fps and set the exposure time to 1/60 sec, all concentrated lights for each interval are overwritten and displayed on 1 sampled data. The formula of the integral sampling is given by,

\[ r_v[n] = \int_{nT + t_0}^{nT} r(t)dt, \]  \hspace{1cm} (1)

where \( n \in \mathbb{N} \) is the frame number. \( T \in \mathbb{R} \) is the start time of recording for each
frame. \( t_e \in \mathbb{R} \) is the exposure time. Note that this formula can be expressed only linear region. Even if the exposure time is extended, the amount of light that can be displayed is limited. Therefore, there is a saturation region. In the saturation region, we cannot demodulate the signal with no error.

If an increase in the bit rate is required, we can employ a pulse amplitude modulation (PAM) scheme on the OCC. This is a general approach and the optical fiber transmission system has adopted the PAM scheme to the short reach link [6]. However, when PAM is employed on the OCC to increase the bit rate, the intensity of the light changes depending on the amplitude of the signal. It affects the human’s eye.

In the OCC, by using the characteristics of the integral sampling a pulse width modulation (PWM) has been reported [7]. The PWM signal through the OCC link is converted into the PAM signal at the CMOS image sensor. In the OCC link, the intensity of the signal does not change. Therefore, the effects on the human’s eye can be eliminated. To more remove the flicker, we should use the high-speed data rate. The high-speed camera or the rolling shutter method is needed; however, the proposed method can detect the high data rate signal with the low frame rate camera.

3 Proposed method

3.1 Concept

This section describes the construction of the proposed scheme and the overview of the operation. The proposed scheme is installed in the receiver of the camera. Figure 1 shows the configuration when employing the proposed scheme. The receiver consists of an external trigger, delay lines to give clock offsets, and two cameras or more. The motivation of the configuration is to detect the signal of the higher symbol rate using the camera with the lower frame rate. Thus, we use multiple cameras with a lower frame rate than the symbol rate from the transmitter. The number of the cameras is determined by \( R_s T_c \) where \( R_s \) is the symbol rate [bps] of the transmitter and \( T_c \) is the shutter speed per a camera [s]. This paper assumes the shutter speed is same as the exposure time. A clock divider consisting of the external trigger and the delay lines controls the start time of the recording. The clock signal from the external trigger is delayed by \( T \) for each camera. For example, the second camera is given a delay of \( T \) and the third is done that of \( 2T \). The variable “\( l \)” in Fig. 1 is the number of cameras. The variable “\( T \)” is the time of the clock offset [s] and it is related to the symbol rate \( R_s \).

Figure 2(a) shows the operation example of the proposed scheme. In the example
of Fig. 2(a), the symbol rate of the transmitted signal and the exposure time of each camera were set to $R_s = 100$ bps and $T_c = 20$ ms, respectively. Thus, we need two cameras ($R_sT_c = 2$). Each camera adjusts the shutter timing by the clock signal from the clock divider and conducts the integral sampling. The sampling time of the second camera is delayed by the clock offset $T$. One frame includes two symbols. The integral sampled signals combined and we obtain the received signal $r_v[n]$.

The actual desired bit sequence cannot be obtained from the received signal $r_v[n]$. The receiver performs the following process: First, the received signal $r_v[n]$ is quantized to $q[n]$. Second, the symbols included in one interval of the frame rate are selected. In this example, 2 symbols are extracted. Using the extracted symbols, the bit decision is performed by,

$$b[n] = I(q[n])I(q[n + 1]),$$

where $I(x)$ is given by,

$$I(x) = \begin{cases} 1, & x \neq 0 \\ 0, & x = 0. \end{cases}$$

The output bit $b[n]$ is subtracted from $q[n + 1]$ as follows,

$$q[n + 1] = q[n + 1] - b[n]$$

This differential scheme is repeated every symbol. Consequently, we can obtain the desired bit sequence.

### 3.2 Formulation

This section formulates the proposed scheme. Figure 2(b) shows the functional block of the proposed demodulation and decoding scheme. The received signal $r_v[n]$ can be obtained by combining the integral sampled signals from the multiple cameras.
Equation (1) is the case of the sampling with one camera. When we use the multiple cameras, then the equation is expressed as,
\[ r_v[n + i] = \int_{nT}^{nT + t_v} r_j(t) dt, \]  
(5)
where \( j = (i \mod l), j \in \{0, 1, \ldots, l - 1\} \) is the camera number. \( r_j(t) \) is the signal received by the camera \( j \). Second, the received signal is quantized and the signal \( q[n] \) is given by,
\[ q[n] = r_v[n] + \Delta, \]  
(6)
where \( \Delta \) is the quantization noise because the error occurs to perform the thresholding. The bit decision is performed by using \( q[n] \). The bit \( b[n] \) is expressed as,
\[ b[n] = \prod_{k=n}^{n+l-1} I(q[k]), \]  
(7)
where \( l \) is the number of the cameras. After outputting the bit \( b[n] \), the quantized signal \( q[m] \), \( m = \{\forall m \in \mathbb{N} \mid n < m < n + l\} \) is updated as follows,
\[ q[m] = q[m] - b[n]. \]  
(8)
Finally, we obtain the bit sequence from these formulations. The proposed scheme does not require a massive number of pixels on an image, such as the rolling shutter scheme. That is, the proposed method and spatial multiplexing schemes can also be used in combination.

4 Numerical simulation

The numerical simulation revealed the characteristics of the proposed scheme as shown in the previous section. We evaluated the bit error rate changing the signal-to-noise power ratio (SNR). First, the 1-sample/symbol intensity-modulated signal was transmitted to the camera receiver via the additive white Gaussian noise (AWGN) channel. The camera receiver conducted direct detection and integral sampling. That is when the number of the cameras is \( N \), the frame rate of the camera is \( R/N \) [fps] with the transmitted data rate of \( R \) [bps]. Figure 3(a) shows the simulation result. The number of the cameras \( N \) changed 2 to 10. When \( N = 1 \), the transmitted data
rate is equivalent to the frame rate of the camera, for the proposed scheme was not employed. We confirmed that all results have the BER curves without the error floor. When the number of cameras increased, the SNR penalty also increased but a rate of increase was not linear. When $\text{BER} = 10^{-3}$, 10 dB of the SNR penalty between $N = 1$ and $N = 10$ occurred. Figure 3(b) shows the desired SNR changing the number of cameras with $\text{BER} = 10^{-3}$ and $10^{-2}$. The proposed scheme requires a sufficiently high SNR to increase the throughput, however the required SNR is saturated rather than increasing linearly. That is, when there is the environment where we can obtain the sufficient SNR, an increase in the number of cameras is not difficult. Generally, OCC system has the larger SNR environment. This is because the LED gives both functions of the lighting to illuminate and the transmitter. In other words, the OCC is performed in darkness if the lighting is required. As other issues, we should note that it is expected that the format conversion from PWM to PAM affects the characteristics.

5 Conclusion

This paper proposed the technique of the capacity enhancement using multiple image sensor-based cameras. While the technique to improve the sampling rate using multiple down-sampler is popular in the field of communication theory, OCC was not able to adopt such a technique because the integral sampler is employed. The clock-synchronized multiple cameras were able to sample the light signal with a lower rate than the bit rate of a transmitter. We described the technique of the proposed demodulation and decoding in detail. The numerical simulation revealed the SNR characteristics.

Acknowledgments

This work was supported by JSPS KAKENHI Grant Numbers JP20H04178 and GMO Foundation, Japan.
Performance and design of uplink multiuser massive MIMO with low-resolution ADCs

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Abstract: This letter investigates the performance and the design of fully-digitized, uplink multiuser massive multiple-input multiple-output (MIMO) systems with low-resolution analog-to-digital converters (ADCs). We analyze the effects of quantization errors when low-resolution ADCs are applied to uplink multiuser massive MIMO systems for reducing power consumption. We derive appropriate decoding parameters to estimate channel bit log-likelihood ratios (LLRs) by taking the power of quantization errors and channel estimation errors into account. The simulation results reveal that lower bit error rates (BERs) can be achieved by enhancing the number of receiving antennas and reducing the number of quantization bits for given spatial correlation matrices.

Keywords: massive MIMO, low-resolution ADC, spatial correlation, channel estimation, bit LLR

Classification: Wireless Communication Technologies

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1 Introduction

Fully-digitized multiple-input multiple-output (MIMO) systems using a massive number of antennas at base stations (BSs) are being researched and developed to establish higher-rate and higher-reliability mobile communications services in mobile-cellular networks. Low-resolution analog-to-digital converters (ADCs) are useful in suppressing hardware cost and power consumption for fully-digitized massive MIMO systems. However, they reduce the fidelity of spatially multiplexed received signals transmitted from dozens of user terminals (UTs).

Dong et al. analyzed the achievable rate of multiuser massive MIMO systems with low-resolution ADCs under spatially correlated channels due to limited physical space at a BS, taking imperfect channel state information (CSI) into account [1]. Ding and Lian revealed the effectiveness of a mixed-ADC architecture to enhance the uplink achievable rate and energy efficiency in massive MIMO systems [2]. Gao and Sanada showed an appropriate quantization range for low-resolution ADCs to enhance the achievable rate in massive MIMO systems [3]. Azizzadeh et al. analyzed the bit error rate (BER) performance of the single carrier modulation for massive uplink MIMO systems using low-resolution ADCs [4].

In this letter, we apply a modulation and coding scheme (MCS) employed in fifth-generation (5G) wireless communications to uplink multiuser massive MIMO systems with low-resolution ADCs and analyze the performance for spatially correlated channels from the viewpoint of BER including the effects of imperfect CSI. We derive appropriate parameters to estimate channel bit log-likelihood ratios (LLRs) for decoding of low-density parity-check (LDPC) codes by taking the power of quantization errors and channel estimation errors into account. The simulation results reveal that higher data reliability can be achieved by enhancing the spatial resolution, i.e., the number of receiving (RX) antennas at a BS and reducing the magnitude resolution, i.e., the number of quantization bits.

2 System model

2.1 Uplink multiuser massive MIMO

We assume uplink multiuser massive MIMO communications with \( K \) single-antenna UTs and \( M \) RX antennas at a BS where \( M \gg K \). Orthogonal frequency division multiple access (OFDMA) and LDPC codes are employed for uplink communications and all \( K \) UTs transmit data frames to BS by the same resource blocks. Open-loop power control is supported for UTs such that the signal power received from a UT is the same at an antenna element of the BS. To simplify the formulation, let us assume that the total RX signal power at an antenna element is normalized by
1 so that the signal power received from a UT is $1/K$. The power delay profile at RX antennas is defined as $R_n$ at delay $n$ for $n \in [0, \tau]$ and $\sum_{n=0}^{\tau} R_n = 1$. The maximum delay $\tau$ is assumed to be smaller than the cyclic prefix (CP) length $N_{CP}$. The channel impulse response (CIR) matrix at delay $n$ is defined as $M \times K$ matrix $f_n$ where the $(m, k)$-entry is CIR between $m$-th RX antenna and $k$-th UT $[f_n]_{m,k} = f_n^{(m,k)}$. We employ $M \times M$ exponentially RX-correlation matrix $r$ with $[r]_{i,j} = 99.7\%$ for a spatial correlation ratio $0 \leq \eta < 1$ to generate a spatially correlation matrix

$$f_n = \sqrt{R_n} r^{1/2} h,$$

where $r^{1/2}$ is the square root of $r$ and $h$ is an $M \times K$ circularly symmetric Gaussian distributed matrix [5]. The CIR and channel frequency response (CFR) are assumed to be invariant within a resource block.

The transmit (TX) signal at UT $k$ and time $n$ is expressed as $x_n^{(k)}$ with $E[x_n^{(k)}(x_n^{(k)})^*] = 1/K$ where $E[X]$ is the mean of a random variable $X$ and $a^*$ is the conjugate of a complex number $a$. The RX signal $y_n^{(m)}$ at antenna $m$ and time $n$ with thermal noise $z_n^{(m)}$ is expressed as

$$y_n^{(m)} = \sum_{k=1}^{K} \sum_{i=0}^{\tau} f_i^{(m,k)} x_{n-i}^{(k)} + z_n^{(m)},$$

with $E[z_n^{(m)}(z_n^{(m)})^*] = 2\sigma_z^2$. The per-frame RX signal power without noise at antenna $m$ is expressed as

$$p_t^{(m)} = \frac{1}{K} \sum_{k=1}^{K} \sum_{i=0}^{\tau} |f_i^{(m,k)}|^2,$$

with $E[p_t^{(m)}] = 1$. The RX power $p_t^{(m)} + 2\sigma_z^2$ is measured through a power detector equipped in each receiver of the BS [3].

### 2.2 Low-resolution ADC

The RX signal at an antenna can be modeled to be complex Gaussian distributed because it is a composite signal transmitted from multiple UTs and passed through multiple wireless paths and multiple subcarriers. The quantization range at antenna $m$ for in-phase and quadrature components is from $-3\sqrt{(p_t^{(m)} + 2\sigma_z^2)/2}$ to $3\sqrt{(p_t^{(m)} + 2\sigma_z^2)/2}$ because $99.7\%$ of the Gaussian-distributed data is within 3 standard deviations of the mean. Through uniform-quantized ADC with quantization bit depth $b_q$, the RX signal is re-expressed as

$$y_n^{(m)} = \sum_{k=1}^{K} \sum_{i=0}^{\tau} f_i^{(m,k)} x_{n-i}^{(k)} + q_n^{(m)} + q_n^{(m)},$$

using a quantization error $q_n^{(m)}$. The power $E[q_n^{(m)}(q_n^{(m)})^*] = 2(\sigma_q^{(m)})^2$ is approximately expressed as

$$2(\sigma_q^{(m)})^2 \approx \frac{3}{2b_q} (p_t^{(m)} + 2\sigma_z^2).$$

The TX symbol at UT $k$ and subcarrier $\ell$ is defined as $X_{\ell}^{(k)}$ and the RX symbol at antenna $m$ and subcarrier $\ell$ is defined as $Y_{\ell}^{(m)}$. After the processing of CP removal
and discrete Fourier transform (DFT) with $N_{\text{DFT}}$ points at BS, the RX time-domain signal $y_n^{(m)}$ is converted to the RX frequency-domain symbol

$$Y_{\ell}^{(m)} = \sum_{k=1}^{K} F_{\ell}^{(m,k)} X_{\ell}^{(k)} + Z_{\ell}^{(m)} + Q_{\ell}^{(m)},$$

(6)

where CFR $F_{\ell}^{(m,k)}$ is the Fourier transform of CIR $f_n^{(m,k)}$. $Z_{\ell}^{(m)}$ and $Q_{\ell}^{(m)}$ are the thermal noise and the quantization error components at antenna $m$ and subcarrier $\ell$, respectively, and can be modeled as complex Gaussian distribution due to the DFT process. In the matrix representation, Eq. (6) is expressed as

$$Y_{\ell} = F_{\ell} X_{\ell} + Z_{\ell} + Q_{\ell}$$

(7)

at subcarrier $\ell$. In practice, CFR $F$ needs to be estimated at the receiver for retrieving the transmitted data symbol vector [6].

Channel estimation errors due to low-resolution ADC can be modeled as additive complex Gaussian noise as well as quantization errors. The estimated CFR is expressed as $\hat{F}_{\ell} = F_{\ell} + \Delta F_{\ell}$ with a channel estimation error matrix $\Delta F_{\ell}$. The frequency-domain power $E[\Delta F_{\ell}^{(m,k)}(\Delta F_{\ell}^{(m,k)})^H] = (N_{\text{SC}}/N_{\text{SYM}}) \cdot 2(\sigma_{\ell})^2$ is approximately derived as

$$2(\sigma_{\ell}^2)^2 \approx \Lambda \left( 2\sigma_{\ell}^2 + \frac{3}{2b_d}(p_{\ell}^{(m)} + 2\sigma_{\ell}^2) \right),$$

(8)

where the non-negative real number $\Lambda$ is a coefficient that represents the infidelity for the channel estimation, $N_{\text{SC}}$ is the number of data subcarriers, and $N_{\text{SYM}} = N_{\text{DFT}} + N_{\text{CP}}$ is the number of samples in an OFDM symbol. The channel estimation infidelity $\Lambda$ is roughly inversely proportional to the number of training OFDM symbols used for the channel estimation. The first term in the parenthesis in Eq. (8) is due to thermal noise and the second term is due to quantization errors.

The zero-forcing (ZF) weight matrix is applied to the received symbol vector as

$$\hat{X}_{\ell} = \hat{W}_{\ell}^T Y_{\ell} - \hat{W}_{\ell}^T \Delta F_{\ell} X_{\ell} + \hat{W}_{\ell}^T Z_{\ell} + \hat{W}_{\ell}^T Q_{\ell},$$

(9)

and it is given by the pseudo-inverse matrix of the estimated channel matrix as $\hat{W}_{\ell} = \hat{F}_{\ell}^+ = (\hat{F}_{\ell}^H \hat{F}_{\ell})^{-1} \hat{F}_{\ell}^H$, where $A^T$ and $A^H$ respectively stand for the transpose and the conjugate transpose of a matrix $A$, and $A^{-1}$ stands for the inverse of a regular matrix $A$.

By normalizing the power of TX symbol $X_{\ell}^{(k)}$ for UT $k$ and subcarrier $\ell$, i.e., $E[X_{\ell}^{(k)}(X_{\ell}^{(k)})^H] = 1$, the power of thermal noises in Eq. (9) for UT $k$ and subcarrier $\ell$ is expressed as

$$P_{\ell}^{(k)} = E \left[ \left| \left( \hat{W}_{\ell}^{(\cdot,k)} \right)^T Z_{\ell} \right|^2 \right] = \frac{K \left( \hat{W}_{\ell}^{(\cdot,k)} \right)^H \hat{W}_{\ell}^{(\cdot,k)}}{R_c \gamma}$$

(10)

with signal-to-noise ratio (SNR) $\gamma$ and code rate $R_c$, where $A^{(\cdot,k)}$ is the $k$-th column vector of a matrix $A$. The power of quantization errors in Eq. (9) for UT $k$ and
subcarrier \( \ell \) is expressed as
\[
P_{Q,\ell}^{(k)} = \frac{3K}{2^{2b_q}} \sum_{m=1}^{M} \left( \frac{N_{SC}}{N_{SYM}} \cdot p_t^{(m)} + \frac{1}{R_c \gamma} \right) |\hat{W}_{(m,k)}^{(\ell)}|^2. \tag{11}
\]
The power of channel estimation errors in Eq. (9) for UT \( k \) and subcarrier \( \ell \) is expressed as
\[
P_{E,\ell}^{(k)} = \lambda K \sum_{m=1}^{M} \left( \frac{1}{R_c \gamma} + \frac{3}{2^{2b_q}} \left( \frac{N_{SC}}{N_{SYM}} \cdot p_t^{(m)} + \frac{1}{R_c \gamma} \right) \right) |\hat{W}_{(m,k)}^{(\ell)}|^2. \tag{12}
\]
The powers \( P_{Z,\ell}^{(k)}, P_{Q,\ell}^{(k)}, \) and \( P_{E,\ell}^{(k)} \) can be computed from the power detector outputs \( \{p_t^{(m)} + 2\sigma_z^2 | m \in [1,M]\} \) and the ZF weight matrix \( \hat{W}_{\ell} \).

The LDPC decoder needs the input of channel bit LLRs obtained from estimated subcarrier SNRs, i.e., estimated subcarrier noise powers normalized by subcarrier signal powers. For high-resolution ADCs, the quantization error components in the estimated noise power are negligibly small because the quantization error power \( P_{Q,\ell}^{(k)} \) and the quantization error component in the channel estimation error power \( P_{E,\ell}^{(k)} \) asymptotically approaches to zero as the quantization bit depth \( b_q \) approaches to infinity. Meanwhile, the quantization error components have a significant impact on reliable LDPC decoding for low-resolution ADCs. The LDPC decoding using only the thermal noise components is denoted by DEC-ZE and that using the thermal noise power, the quantization error power, and the channel estimation error power is denoted by DEC-ZQE.

### 3 Simulation results
Monte-Carlo simulations are conducted to evaluate the performance of uplink multi-user massive MIMO systems with low-resolution ADCs by randomly generating TX data bits and fading channel coefficients. A tapped delay line (TDL)-A model in [7] is employed as the power delay profile and an MCS in [8] is employed as shown in Table I.

Figure 1 illustrates the BER performances of DEC-ZE and DEC-ZQE with the number of RX antennas \( M = 128 \) at BS, the quantization bit depth \( b_q = 1 \), channel estimation infidelity \( \lambda = 0, 0.25, \) and \( 0.5 \), and spatial correlation ratio \( \eta = 0.3 \) and \( 0.7 \) for given SNR \( \gamma \). The DEC-ZQE achieves a slightly better performance than

| Table I. Simulation specification |
|-------------------|------------------|
| Parameter          | Scheme/Value     |
| Number of UTs, \( K \) | 16               |
| RX power delay profile, \( R_n \) | TDL-A            |
| Modulation         | OFDM             |
| Subcarrier modulation | 16-QAM          |
| Number of DFT points, \( N_{DFT} \) | 2048            |
| Number of data subcarriers, \( N_{SC} \) | 1200            |
| Number of CP points, \( N_{CP} \) | 144              |
| Error-correcting code | LDPC           |
| Code rate, \( R_c \) | 1/2              |
DEC-ZE in a low-SNR regime whereas the performance of DEC-ZE destructively deteriorates in a high-SNR regime. This is because the estimated noise power is underestimated to calculate accurate channel bit LLRs by not taking quantization errors into account. Thus, the estimation of the quantization error power at BS has a significant impact on its decoded data reliability.

Figure 2 illustrates the BER performance of DEC-ZQE with the total number of quantization bits $M b_q = 128$, channel estimation infidelity $\lambda = 0$ and 0.25, and spatial correlation ratio $\eta = 0.3$ and 0.7 for given SNR $\gamma$. The larger the number of RX antennas is, the higher the spatial resolution is, whereas the larger the quantization bit depth is, the higher the magnitude resolution is. In addition, the increase in the number of RX antennas leads to a larger diversity in RX signals. The simulation results show that higher spatial resolution results in higher RX-diversity and coding gains, even with employing one-bit ADCs in an uplink multiuser massive MIMO system for the given channel estimation infidelities and spatial correlation ratios. Note that the degree of BER degradation increases as the channel estimation infidelity increases for lower-resolution ADCs.
4 Conclusion

In this letter, we have investigated the impacts of quantization errors on the performance and the design of uplink multiuser massive MIMO systems with low-resolution ADCs. The channel bit LLRs for LDPC decoding should be computed by taking the power of quantization errors into account. The simulation results show that higher decoded data reliability can be achieved by enhancing the spatial resolution and reducing the magnitude resolution under the constraint of the same total number of quantization bits. In future work, we will attempt to reduce the power of quantization errors using Lloyd-Max quantization [9].
Stabilization of phantom fabrication by degassing using ultrasonic vibrations

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Abstract: Bioequivalent phantoms are used to evaluate the effects of the human body in electromagnetic waves propagation for small wireless communication devices such as smart phones and wearable devices used in close proximity to the human body. The conventional phantom fabrication process requires fabrication skills and experience to avoid air bubbles in the phantom that affect the electrical characteristics. This letter proposes a new phantom fabrication method using ultrasonic vibrations for degassing to stabilize the electrical constants of the fabricated phantom and demonstrates its effectiveness by experiments.

Keywords: phantom, complex permittivity, ultrasonic vibration

Classification: Electromagnetic Compatibility (EMC)

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1 Introduction

As wearable devices have been becoming more popular common and the size of information and communication devices will be reduced further in future, there is a growing need to evaluate the effects of human body in electromagnetic wave propagation when these devices are used in close proximity to the human body. It is difficult to evaluate the effects of the human body using an actual human body for moral reasons. Therefore, phantoms have been developed to simulate the electrical characteristics (relative permittivity and conductivity) of biological tissues [1].

The phantom fabrication method is similar to a cooking recipe [2], and the conventional approach requires fabrication skill and experience to achieve the target electrical constants. One of the major problems in the conventional phantom fabrication process is air bubbles mixed in the phantom, which affect the electric constants and the quality of the fabricated phantom heavily depends on the skills and experiences of the manufacturer.

This letter proposes a new phantom fabrication method that uses ultrasonic vibrations for degassing to stabilize the electrical constants of the fabricated phantom and demonstrates its effectiveness through experiments.

2 Phantom fabrication using ultrasonic vibration

Figure 1 shows a photo of phantom manufacturing using ultrasonic vibration for degassing [3]. An ultrasonic cleaner (MCD-2) manufactured by AS ONE Corporation was used for generating ultrasonic vibration. The maximum frequency of the ultrasonic vibration is 40 kHz, and the maximum operating time is 15 minutes. Due to the specifications of the ultrasonic cleaner, the experiments have been conducted using ultrasonic vibration frequency of 40 kHz and operating time of 15 minutes.

Two types of phantoms, one with ultrasonic vibration (w/ USV: ultrasonic vibra-
Table I. Composition of muscle-equivalent solid phantoms [4].

| Sample                       | Amount [g] | Component Percentage [%] |
|------------------------------|------------|--------------------------|
| Deionized water              | 256.8      | 85.60                    |
| Agar                         | 7.95       | 2.65                     |
| NaCl                         | 2.97       | 0.99                     |
| Sodium dehydroacetate        | 0.15       | 0.05                     |
| TX-151                       | 6.42       | 2.14                     |
| Polyethylene powder          | 25.68      | 8.56                     |

Due to the ultrasonic cleaner’s tank size limitation (tank size: $150 \times 140 \times 100 \text{ mm}^3$), two cylindrical muscle-equivalent solid phantoms of 300 g were created in this experiment; the composition of the muscle-equivalent solid phantoms created in the experiment is shown in Table I [4].

It is noted that the electrical properties of the phantom are sensitive to its size. Hence, when preparing the phantom, it is necessary to consider not only the composition ratio but also the size of the phantom [5]. However, we did not prepare the same size ($15.8 \times 15.8 \times 15.8 \text{ cm}^3$) as the literature in this paper because it did not fit into the tank of a commonly available ultrasonic cleaner. As a result, we must keep in mind that the phantom may contain factors that do not correspond to the electrical properties of biological tissues. The main goal of this paper, however, is to reduce the dispersion of the electrical constants by removing the bubbles introduced during phantom fabrication. The effect of phantom size and the number of bubbles mixed in on the electrical constants will be studied further in the future.

3 Measurement of electrical constants and verification of the effectiveness of ultrasonic vibration

Ten phantoms with/without ultrasonic vibration have been fabricated to confirm the effectiveness of the proposed method. The relative permittivity ($\varepsilon_r$) and conductivity ($\sigma$ [S/m]) of the fabricated muscle equivalent solid phantoms were measured using the network analyzer (Keysight Technology E5071C) and the dielectric probe kit (Keysight Technology N1501A). The coaxial probe method is used for the measurement, and frequency characteristics of relative permittivity and conductivity of the phantoms are measured over the frequency from 300–3000 MHz, which is UHF band widely used for IoT application. Since the electrical constants vary fabrication by fabrication, the electrical constants of the five fabricated phantoms were measured. Additionally, cross sections of fabricated phantoms were observed using a microscope and the number of visible air bubbles per unit area were counted.

Figure 2(a) shows the frequency characteristics of averaged relative permittivity and conductivity of the fabricated phantoms. According to the IFAC [6], the electrical constants of muscle tissue are also shown for reference. The maximum and minimum values of relative permittivity and conductivity according to frequency are also plotted to assess the stability of phantom fabrication. According to the results, the average difference between the maximum and minimum values of relative permittivity and conductivity are...
Fig. 2. (a) Relative permittivity and conductivity of the phantom manufactured with/without ultrasonic vibration according to frequency. (b) The number of air bubbles per square meter. (c) A cross-sectional photograph of the fabricated phantoms.

Relative permittivity is 12.2, and that of electrical conductivity averaged is 0.35 S/m when ultrasonic vibration is not used. However, average of the difference between the maximum and minimum values of relative permittivity is improved from 12.2 to 4.0, and that of electrical conductivity is also improved from 0.35 S/m to 0.19 S/m. These findings indicate that ultrasonic vibration for the proposed phantom fabrication method is effective in stabilizing the fabricated phantom’s electric constants.

The number of air bubbles per unit area with and without ultrasonic vibration is shown in Fig. 2(b). The average number of air bubbles per unit area of the phantom is reduced from 0.08 to 0.02 by using ultrasonic vibration, as shown here. Figure 2(c) presents photo of a cross-sectional image of the fabricated phantoms. It is apparent
that the number of air bubbles is reduced by using ultrasonic vibration. These results reveal that the number of air bubbles in the cross section of the phantom is reduced, thus the electric constants of the fabricated phantoms are stabilized by using ultrasonic vibration.

These results also imply that the electrical constants of the phantom increases since air bubbles inside the phantom is removed by using ultrasonic vibration.

4 Conclusions

This paper proposes a new phantom fabrication method that uses ultrasonic vibrations for degassing to stabilize the fabricated phantom’s electrical constants. The results of the experiments show that ultrasonic vibration can remove air bubbles within the phantom and stabilize the electrical constants of the fabricated phantoms. In future work, we intend to quantify the effects of phantom size and the number of air bubbles on the electrical constants.

Acknowledgments

This study was supported by the Nanzan University Pache Research Subsidy I-A-2 for the 2021 and 2022 academic year.
Reducing fronthaul traffic with direct constellation mapping method

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Abstract: Centralized RAN, which connects a centralized base band unit (BBU) to multiple remote radio units (RRUs), is one of the promising architectures because it enables the BBU to coordinate with distributed RRUs by using multiple-input multiple-output (MIMO) processing without control signaling among RRUs. However, fronthaul traffic between the BBU and RRUs increases and fronthaul link capacity is to be under strain due to increasing the number of RRUs in a same fronthaul network segment. In this paper, we propose a direct constellation mapping method to reduce the fronthaul traffic. And it is also shown that our proposed method can drastically reduce the required fronthaul link capacity with slight downlink throughput degradation by Monte-Carlo computer simulation.

Keywords: C-RAN, distributed MIMO, fronthaul

Classification: Wireless Communication Technologies

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1 Introduction

Several kinds of radio access network (RAN) architectures are being discussed for 5G (5th Generation) and beyond 5G systems. Centralized RAN (C-RAN), which connects a centralized base band unit (BBU) to multiple remote radio units (RRUs) via fronthaul using passive optical network (PON), is one of the promising architectures because it enables the BBU to coordinate with distributed multiple RRUs by using multiple-input multiple-output (MIMO) processing without control signaling among RRUs [1]. This architecture can coordinate inter-RRUs interference and achieve high link capacity uniformly over the service area in a cellular system.

The O-RAN alliance specifies the fronthaul interface and function split between the entities called distributed unit (DU) and radio unit (RU) [2]. In the C-RAN, the DU and the RU are implemented in the BBU and the RRU respectively. Furthermore, the DU supports the functions of radio link control (RLC), medium access control (MAC), and the high part of physical layer (High-PHY), and the RU supports the function of the Low-PHY and radio frequency (RF) respectively [2, 3]. In the downlink, the DU transmits the data sequence after resource element (RE) mapping in an IQ sampling sequence of orthogonal frequency division multiplexing (OFDM) signal in the frequency domain. The data sequence is quantized and adopted forward error correction (FEC) for guaranteeing fronthaul transmission quality, and then symbol modulation for optical transmission in a fronthaul. The RU receives the decoded data sequence from fronthaul and performs inverse fast Fourier transform (IFFT) to generate an OFDM symbol in the time domain for radio transmission. This function split between DU and RU can reduce fronthaul traffic compared with conventional common public radio interface (CPRI) [4] in which OFDM symbol in the time domain is transmitted in a fronthaul. However, as the number of RRUs which connect to a common fronthaul increase, fronthaul traffic increases and fronthaul link capacity is to be under strain.

In this paper, we propose a direct constellation mapping method to reduce the fronthaul traffic, in which IQ sampling sequence of OFDM signal in the frequency domain is directly mapping to the symbol of optical transmission without quantization and FEC in a fronthaul. We evaluate the required fronthaul link capacity and downlink throughput achievable with direct constellation mapping by Monte-Carlo computer simulation. In addition, we evaluate the user equipment (UE) ratio which can adopt this method within less than 5% degradation of downlink throughput in a cellular system.

2 Direct constellation mapping method

Figure 1 shows the downlink transmitter structure of our proposed direct constellation mapping method in C-RAN architecture. In the BBU, according to specification in 3GPP [3], the bit sequence after RLC and MAC processing is encoded using
low-density parity-check (LDPC) code and scrambled for whitening the interference from other cells, and then modulated to quadrature phase shift keying (QPSK) or quadrature amplitude modulation (QAM) symbols. The modulated symbol sequence undergoes layer mapping and precoding for MIMO processing. In this paper, MIMO with $M$ transmit antennas of the RRU and $N$ receive antennas of the UE is considered.

The data sequence after precoding is mapped to each RE and then the downlink frequency domain OFDM signal $D^{¹}_m k$ is obtained, where $m = 0 \sim M - 1$ is the transmit antenna index and $k = 0 \sim K - 1$ is the RE index respectively.

In the fronthaul, $D^{¹}_m k$ is inputted sequentially, digital signal for optical transmission $F^{¹}t$ is obtained by applying the normalizing and direct mapping on $D^{¹}_m k$ as

$$F^{¹}t = \frac{1}{A} D^{¹}[[t/K], t \bmod K],$$  \hspace{1cm} (1)$$

where $A$ is a constant for normalizing the IQ samples and $t$ is time index. Finally, electrical-to-optical converter (E/O) converts $F(t)$ into optical signal and transmit via an optical fiber. Consequently, the fronthaul traffic can be reduced since the redundancy is reduced by not adopting quantization and FEC in the fronthaul.

At the receiver of fronthaul, received signal through the optical fiber undergoes equalization, and then soft-decision symbol sequence $D^{¹}_m(k)$ after equalization for $m$th transmit antenna can be expressed as

$$D^{¹}_m(k) = E(mK + k) + n(k),$$  \hspace{1cm} (2)$$

where $n(t)$ is a residual interference and noise component. The optical channel can be approximated as an additive white Gaussian noise (AWGN) channel by using channel equalization [5], then we assume the $n(t)$ is an AWGN component. In the RRU which has the $m$th transmit antenna, IFFT is applied to $D^{¹}_m(k)$ into the time domain signal block. Finally, the last few samples of time-domain signal block are copied as a cyclic prefix (CP) and inserted into the guard interval, and then the OFDM symbol is transmitted from the $m$th transmit antenna after up-conversion into RF. Although the transmitted OFDM signal includes the residual noise of fronthaul transmission, the UE which receives the OFDM symbol can correct some errors caused by both the fronthaul channel and the dynamic wireless channel using LDPC decoding.
3 Monte-Carlo computer simulation

Firstly, we evaluate the required fronthaul link capacity and downlink throughput performance of the direct constellation mapping. We consider the downlink OFDM transmission between the RRU and the UE. MIMO transmission with single stream using $M = 2$ transmit antennas of the RRU and $N = 2$ UE receive antennas (i.e. transmit and receive diversity transmission) is assumed. $K = 3276$ REs with 30 kHz subcarrier spacing in 100 MHz system bandwidth are used in an OFDM symbol. Link adaptation is adapted ideally, and thus it selects the modulation and coding scheme (MCS) which can achieve the highest throughput of the OFDM transmission in response to the instantaneous wireless channel condition. In the MCS table, QPSK, 16 QAM, 64 QAM and 256 QAM can be selected and the coding rate $R = 0.117 \div 0.926$ of LDPC coding can be used as specified in [3]. The $M \times N$ dynamic wireless channel model in 3GPP [6] is considered, and assuming that the UE can estimate the wireless channel ideally. In the fronthaul, average signal to interference plus noise ratio (SINR) of fronthaul transmission (i.e. $E[|F(t)|^2]/E[|n(t)|^2]$, where $E[\cdot]$ represents the ensemble average operation) is determined statically when the optical fiber has been installed. In this paper, it is set to 30 dB, in which error-free transmission over a 10 km fiber can be achieved using 256 QAM with FEC of 14% Reed-Solomon code, while bit error rate (BER) $\approx 10^{-3}$ error occurs without FEC [7]. Figure 2 shows the required fronthaul link capacity and the downlink throughput performance. From Fig. 2 (a), it can be seen that the direct constellation mapping method can reduce about 56% of required fronthaul link capacity per RRU compared with conventional method using quantization and FEC in a fronthaul. And Fig. 2 (b) plots the downlink throughput as a function of received SINR at the UE on wireless channel. The direct constellation mapping method can achieve almost the same downlink throughput as conventional method in the low region of SINR, while degrades the downlink throughput in the high region of SINR. This degradation occurs because the residual noise of fronthaul transmission

![Fig. 2. Required fronthaul link capacity and downlink throughput performance.](image-url)
becomes dominant compared to noise in the signal received from a wireless channel in the high region of SINR. When SINR at the UE on wireless channel is less than 15 dB, the degradation of downlink throughput can be suppressed less than 5%.

And then, we evaluate the UE ratio which can adopt our proposed method in a cellular system, when the allowable degradation of downlink throughput is less than 5%. A simple hexagonal cellular model with 100 m cell radius of 19 cells is assumed. RRUs are located in center of each cell and UEs are randomly located in each cell. Path loss with probability model is considered between each RRU and each UE [6]. Figure 3 shows the distribution of received SINR at the UEs in the center cell. From Fig. 3 (a), it can be seen that received SINR at the UE near the RRU is quite high, and it becomes low with distance from the RRU. And Fig. 3 (b) plots the cumulative distribution function (CDF) of received SINR at the UE. When using all REs of each RRU, the ratio of UEs which received SINR is less than 15 dB is more than 99.9% both without coordination among RRUs and with coordination among 6 adjacent RRUs. Even when the usage of REs is 50% in each RRU, the ratio of UEs which received SINR is less than 15 dB are 99.9% and 96.3% without coordination among RRUs and with coordination among 6 adjacent RRUs respectively.

Fig. 3. Distribution of average received SINR at the UEs.

4 Conclusion

In this paper, we propose a direct constellation mapping method to reduce the fronthaul traffic. The direct constellation mapping method can reduce the physically required fronthaul link capacity drastically. More than 96.3% (99.9%) UEs in a cellular system can adopt this method within less than 5% degradation of downlink throughput when 50% (100%) RE usage with coordination among 6 adjacent RRUs.

Acknowledgments

A part of this work was conducted under “R&D for further advancement of the 5th generation mobile communication system” (JPJ000254) commissioned by the Ministry of Internal Affairs and Communications in Japan.
Transmission datarate adaptation using redundant check information for IEEE 802.11ax wireless LAN

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Abstract: This letter proposes a practical transmission datarate adaptation (TDA) scheme using Q-learning applicable to IEEE 802.11ax wireless local area networks. In the proposed scheme, each basic service set (BSS) selects an appropriate transmission datarate according to the buffer statuses of adjacent BSSs which are periodically collected and the transmission results of DATA frames in the BSS. Then, the BSS conducts underlay transmissions based on the framework of spatial reuse defined in IEEE 802.11ax. The performance of the proposed scheme is evaluated through system-level computer simulation assuming downlink full-buffer traffic. It is confirmed that the proposed scheme can achieve higher average area throughput than conventional TDA schemes such as Robust Rate Adaptation Algorithm (RRAA) and adaptive modulation and coding (AMC) based on the received power.

Keywords: wireless LAN, IEEE 802.11ax, Q-learning, transmission datarate adaptation, underlay transmission

Classification: Wireless Communication Technologies

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1 Introduction

IEEE 802.11 wireless local area networks (WLANs) [1] have been widely and densely deployed, and their demand is still growing. The current WLAN generally employs a distributed and autonomous channel access mechanism based on carrier sense multiple access with collision avoidance (CSMA/CA). Therefore, increase of traffic demand in the WLAN brings severe contention among multiple basic service sets (BSSs) sharing the same radio channel. It causes frequent collision and resultant failure of frame transmission.

Since IEEE 802.11 WLANs support multiple transmission datarates, each transmitter needs to adjust its transmission datarate to a suitable one. When the signal-to-noise ratio (SNR) of a transmitted frame is insufficient, its transmitter should lower the transmission datarate. On the other hand, when transmission failure is caused by collision, the transmission datarate should be raised so as to shorten the length of the transmitted frame and to reduce the probability of collision consequently.

Therefore, in general, the transmission datarate is adjusted according to one or more metrics calculated from the results of frame transmission [2]. For example, Automatic Rate Fallback (ARF) [3] raises/decreases the transmission datarate if the number of successful/failed frame transmissions reaches a pre-determined threshold. Robust Rate Adaptation Algorithm (RRAA) [4] raises/decreases the transmission datarate if a frame error rate (FER) becomes larger/less than a given FER threshold.
However, IEEE 802.11 WLANs have no way to know directly whether a failure of frame transmission is caused by collision or by other reasons such as insufficient SNR because the transmitter recognizes its transmission failure by occurring a timeout of ACK frame reception. Several studies were recently conducted to estimate the factor of transmission failure using one or more frame sniffers to utilize the estimated cause to adjust the transmission datarate [5, 6].

Furthermore, a transmission datarate adaptation (TDA) scheme was studied to select an appropriate transmission datarate using Q-learning with the aid of side information called “redundant check information” about the frame transmission in adjacent basic service sets (BSSs) [7]. In this scheme, each BSS collects, as the redundant check information, the information whether or not adjacent BSSs will transmit their frames in near future, and then learns and selects the best action (i.e., selects the best transmission datarate or defers its frame transmission). If the best transmission datarate is selected, each BSS makes underlay transmissions against its adjacent BSSs, and thus this scheme can improve the throughput.

However, the performance evaluation conducted in [7] assumes slotted channel access. On the other hand, IEEE 802.11 WLAN employs random backoff based on CSMA/CA [1], and thus it is difficult to precisely know when each node will transmit its frame. Hence, this letter proposes a practical scheme to apply the concept of TDA in [7] to IEEE 802.11ax [8] which defines a mechanism of spatial reuse for underlay transmissions. The proposed scheme adjusts the transmission datarate of each BSS based on Q-learning at an adaptation interval. The buffer status of each BSS is collected as the redundant check information, and it is obtained using buffer status report (BSR) defined in IEEE 802.11ax.

The remainder of this paper is as follows. Section 2 introduces the concept of the TDA presented in [7] and our proposed TDA scheme. Section 3 explains the configuration of the system-level computer simulation and its results. Finally, conclusion is given in Sect. 4.

2 TDA using redundant check information

Figure 1(a) shows the concept of the TDA presented in [7] applicable to slotted channel access. We focus on BSS 0 as a target BSS of TDA, and other BSSs are adjacent BSSs. BSS 0 collects the information whether adjacent BSSs will transmit their frames or not in the next slot. As shown in Fig. 1(a), this information is encoded as “state.” BSS 0 selects the transmission datarate with the maximum Q-value on the state, or selects a transmission datarate randomly with a certain probability. (Hereafter, this probability is called as “random selection probability.”) BSS 0 makes its DATA frame transmission at the selected transmission datarate, and then updates the Q-value according to the result of frame transmission. This process is conducted slot-by-slot (in other words, frame-by-frame). This scheme can improve throughput comparing with the conventional slotted ALOHA because it can select an appropriate transmission datarate even when collision is expected.

Figure 1(b) shows the concept of our proposed TDA scheme. In our proposed scheme, the transmission datarate is adjusted at an interval \( T_c \). Since BSS 0 cannot know when adjacent BSSs will transmit their frames exactly due to the random
backoff, it collects the buffer statuses of adjacent BSSs, and selects the transmission datarate used in the next adaptation period. Here, an access point (AP) in each BSS can obtain the buffer statuses of its associating stations (STAs) by BSR. Each node checks by using BSS Color [8] defined in IEEE 802.11ax whether the received frame comes from the BSS that is expected to transmit frame(s) in the current adaptation period. (The state of such BSS is denoted by “1” in Fig. 1(b).) If the node detects a frame from such BSS, the CCA level at the node is raised so that the frame from the BSS is not detected using the framework of spatial reuse, which is also defined in IEEE 802.11ax, in order to enable underlay transmissions.

The Q-value is updated by the transmission results of DATA frames in a logging duration ($T_{\text{log}}$ from the beginning of the adaptation period) frame-by-frame as follows.

$$ Q(s_t, a_t) \leftarrow Q(s_t, a_t) + \alpha \left\{ r_{t+1} - Q(s_t, a_t) \right\} $$  \tag{1}$$

where $\alpha$ is the learning rate, $s_t$ is the state of buffers in adjacent BSSs, and $a_t$ is the selected action of the target BSS (i.e., BSS 0 in Fig. 1(b)). The reward $r_{t+1}$ of each DATA frame is calculated by

$$ r_{t+1} = \begin{cases} (\delta_{X,1} W_{\text{suc}} - \delta_{X,0} W_{\text{fail}}) RD & \text{if } R \geq 0 \\ r_{\text{notx}} & \text{if } R = -1 \end{cases} $$  \tag{2}$$

where $\delta_{a,b}$ denotes Kronecker delta, $X$ is the transmission result (“1” means SUC-
CESS, and “0” means FAILURE) of the DATA frame, \( D \) [kbyte] is the payload length of the DATA frame, \( W_{\text{suc}} \) and \( W_{\text{fail}} \) are the weights of reward for successful and failed frame transmissions, respectively. \( R \) is the used transmission datarate [Mb/s]. Here, \( R = -1 \) denotes that frame transmission is pended in the corresponding adaptation period, and the reward is set to \( r_{\text{notx}} \) in this case.

3 Computer simulation

The performance of the proposed TDA scheme is evaluated through system-level computer simulation based on IEEE 802.11ax WLAN. We compare three schemes: our proposed scheme, RRAA, and adaptive modulation and coding (AMC) based on the received power.

Table I shows the simulation parameters. The area size is assumed to be 80 m \( \times \) 80 m, and it is segmented into 4 \( \times \) 4 (thus, the segment size is 20 m \( \times \) 20 m). Each BSS is located in different segment. Propagation model is based on that of the Residential Scenario (SS1) defined by IEEE 802.11 TGax [9]. In our proposed scheme, each BSS collects the buffer statuses of the BSSs whose frame is received with the received power equal to or greater than \(-88\) dBm which is same as the frame detection limit in this simulation. The random selection probability of action \( P_{\text{rand}} \) at the \( t \)th adaptation period is set to \( P_{\text{rand}} = 1/(1 + t/N) \), and \( N \) is set to 133, which gives \( P_{\text{rand}} = 0.1 \) at 2 minutes. The parameters for the proposed scheme used in this performance evaluation are adopted because they gave high performance in the preliminary simulation. (It should be noted that \( t \) should be reset to 0 if each BSS detects the change of environment.)

In RRAA, the transmission datarate is adjusted every \( T_c \) if the number of transmitted DATA frames after the previous datarate adjustment is equal to or greater than 50. If FER is greater than 0.1, the modulation and coding scheme (MCS) index is decremented by one, and it is incremented by one if the FER is less than 0.05. Otherwise, the current MCS index is retained. In AMC, each node selects the maximum MCS index at which the received power satisfies the minimum input

| Table I. Simulation parameters |
|--------------------------------|
| Evaluation duration          | 20 minutes |
| Number of BSSs               | 4, 16      |
| Number of STAs per BSS       | 1          |
| Supported MCS index          | 0–9 for IEEE 802.11ax [8] |
| Transmission power           | 20 dBm     |
| Signal bandwidth             | 20 MHz     |
| Noise level                  | \(-94\) dBm (including 7 dB noise figure) |
| Propagation model            | IEEE 802.11 TGax Residential scenario [9] |
| Frequency channel            | Ch 1 in the 2.4 GHz band |
| Retransmission limit         | 7 times    |
| RTS/CTS exchange             | Not in use |
| Traffic                      | Downlink full-buffer traffic |
| Payload length               | 3000 bytes |
| Interval of datarate adaptation (\( T_c \)) | 100 ms |
| Logging duration (\( T_{\log} \)) | 80 ms |
| Learning rate (\( \alpha \))  | 0.1        |
| Weights of reward (\( W_{\text{suc}}, W_{\text{fail}} \)) | (1, 0.1) |
| Reward for pending transmission (\( r_{\text{notx}} \)) | \(-2\) |
Figures 2(a) and 2(b) show the average area throughput and the frame delivery rate of AMC, RRAA and the proposed scheme, respectively. In addition, the number of DATA frames transmitted at each MCS index when the number of BSSs is 16 is shown in Fig. 2(c). The proposed scheme achieves the area throughput 1.3 and 2.1 times (1.4 and 4.0 times) as high as that of AMC and RRAA when the number of BSSs is 4 (16), respectively. AMC can achieve higher throughput than RRAA even though its frame delivery rate is worse than RRAA because AMC uses much higher MCS index (in other words, higher transmission datarate) than RRAA. On the other hand, the proposed scheme achieves a frame delivery rate almost as high as RRAA, and it tends to use higher MCS index whereas RRAA tends to use lower MCS index. It means the proposed scheme can select an appropriate MCS index according to the buffer statuses of adjacent BSSs.

4 Conclusion

This letter proposed a practical TDA scheme using Q-learning applicable to IEEE 802.11ax WLAN. In the proposed scheme, each BSS selects an appropriate transmission datarate according to the buffer statuses of adjacent BSSs which are periodically collected and the transmission results of DATA frames in the BSS. Then, the BSS conducts underlay transmissions based on the framework of spatial reuse defined in IEEE 802.11ax. The performance of the proposed scheme was compared with the conventional TDA schemes: RRAA and AMC based on the received power through system-level computer simulation based on IEEE 802.11ax WLAN assuming downlink full-buffer traffic. It was confirmed that the proposed scheme achieves the area throughput 1.3 times and 2.1 times (1.4 and 4.0 times) as high as that of AMC and RRAA when the number of BSSs sharing the same frequency channel is 4 (16), respectively, by selecting an appropriate MCS index according to the buffer statuses of adjacent BSSs.

Acknowledgments

This research and development work was supported by the MIC/SCOPE #JP196000002.
Closed-form steady-state bias error of tracking filter using coordinated turn model for constant velocity target

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Abstract: This letter presents the derivation for the steady-state bias errors of a tracking filter using a coordinated turn (CT) model for tracking a target moving at a constant velocity as a tracking performance index of maneuvering target. The analysis of the tracking filter using the CT model indicates bias errors due to differences between the model-assumed motion (constant angular velocity) and target motion (constant velocity). We derive this bias error in the closed-form as expressed by the filter design parameters. Then, numerical simulations show that the derived bias errors are correct and can be efficiently used for filter design.

Keywords: tracking filter, coordinated turn model, tracking performance index

Classification: Navigation, Guidance and Control Systems

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1 Introduction

Sensing systems in autonomous vehicles, robots, and surveillance applications often track moving targets as an essential function. Tracking filters are commonly used for such purposes, and numerous tracking methods have been proposed, such as the Kalman and particle filter-based techniques [1, 2]. To use such tracking filters, it is essential to determine the dynamic models of the targets assumed in the filtering process. The most commonly used dynamic models are the constant velocity (CV) and constant acceleration (CA) models [3]. Another well-known dynamic model is the coordinated turn (CT) model [3, 4], which assumes that the target moves with a constant angular rate. Although the tracking filter using the CT model (i.e., CT tracking filter) assumes that the angular rate is known, estimation methods have been proposed for the angular rate to improve the practicality of the approach [5, 6]. In many tracking systems developed to track maneuvering vehicles, the CT model has been used along with the CV model to track both the uniform and turning motions of maneuvering targets, and this model fusion approach is commonly known as the interacting multiple model [7, 8].

To design the tracking filters, it is important to evaluate the relationship between the filter parameters and steady-state tracking errors [3, 9]. For the CV and CA models, the closed-form performance indices in the steady-state have been derived and efficiently used. For example, the steady-state bias errors in CV tracking filters when tracking CA targets owing to differences between the model-assumed and actual motions are important design features [9, 10]. However, such instances have not been sufficiently investigated for the CT tracking filter. Although many researchers have investigated the performances of CT tracking filters, they mainly rely on simulations [4, 5, 6], such that theoretical analyses of the CT filter tracking errors have not been sufficiently performed. For example, it is empirically known that the CT tracking filter outputs steady-state bias errors when tracking CV targets; these bias errors have been confirmed in simulations [5] and are known to occur because of differences between the model-assumed and target motions, similar to the bias errors of the CV tracking filter for CA targets. However, such bias errors have not been theoretically analyzed. If the steady-state bias errors of CT filters for targets exhibiting CV motions are theoretically expressed using the filter parameters, then these errors may be considered as one of the performance indices of CT filters that express the tracking performances of maneuvering targets having both CT and...
CV motions.

In this work, we derive the steady-state bias error of the CT tracking filter for a CV target in the closed form. Then, numerical simulations clarify that the derived closed-form bias errors of the CT filter for CV targets can quantitatively predict the errors for targets with CV motions in the filter design process.

2 Problem definition

This study analyzes the steady-state errors of the CT tracking filter by considering two-dimensional tracking in the $x$-$y$ plane.

The tracking process invokes iterative prediction and filtering processes to estimate the state of the target. The state of the target includes its position and velocity in the $x$-$y$ plane. The prediction process is conducted according to the CT dynamic model that the target turns with a constant angular velocity during sampling interval, which is expressed as follows [3]:

$$x_{p,k} = x_{s,k-1} + \frac{\sin \omega T}{\omega} \hat{x}_{s,k-1} - \frac{1 - \cos \omega T}{\omega} \hat{y}_{s,k-1}, \quad (1)$$

$$\hat{x}_{p,k} = \cos \omega T \hat{x}_{s,k-1} - \sin \omega T \hat{y}_{s,k-1}, \quad (2)$$

$$y_{p,k} = 1 - \cos \omega T \hat{x}_{s,k-1} + \sin \omega T \hat{y}_{s,k-1}, \quad (3)$$

$$\hat{y}_{p,k} = \sin \omega T \hat{x}_{s,k-1} + \cos \omega T \hat{y}_{s,k-1}, \quad (4)$$

where $(x_{p,k}, y_{p,k})$ is the predicted target position at time $kT$, $T$ is the sampling interval, $(x_{s,k}, y_{s,k})$ is the filtered target position, $(\hat{x}_{p,k}, \hat{y}_{p,k})$ is the predicted velocity, $(\hat{x}_{s,k}, \hat{y}_{s,k})$ is the filtered velocity, and $\omega$ is a constant parameter corresponding to the angular velocity assumed in the CT dynamic model. Note that the process noise was not considered in the dynamic model because the steady-state was assumed in our analysis. The filtering process is defined as [3]

$$x_{s,k} = x_{p,k} + \alpha_x (x_{o,k} - x_{p,k}) + \alpha_y (y_{o,k} - y_{p,k}), \quad (5)$$

$$\hat{x}_{s,k} = \hat{x}_{p,k} + \frac{\beta_x}{T} (x_{o,k} - x_{p,k}) + \frac{\beta_y}{T} (y_{o,k} - y_{p,k}), \quad (6)$$

$$y_{s,k} = y_{p,k} + \gamma_x (x_{o,k} - x_{p,k}) + \gamma_y (y_{o,k} - y_{p,k}), \quad (7)$$

$$\hat{y}_{s,k} = \hat{y}_{p,k} + \frac{\delta_x}{T} (x_{o,k} - x_{p,k}) + \frac{\delta_y}{T} (y_{o,k} - y_{p,k}), \quad (8)$$

where $(x_{o,k}, y_{o,k})$ is the measured position, and $\alpha_x, \alpha_y, \beta_x, \beta_y, \gamma_x, \gamma_y, \delta_x$, and $\delta_y$ are dimensionless filter gains. In the steady state, these filter gains are converged to fixed values. Thus, these filter gains are assumed to be constant in our analyses. The purpose of this study is to derive the bias errors of $x_{p,k}$ and $y_{p,k}$ for a target moving a constant velocity in the steady state.

3 Derivation of steady-state bias errors for CV target motion

First, the transfer functions of the CT tracking filter are derived. By simplifying the $z$-transforms of Eqs. (1)–(8), the transfer functions of the CT tracking filter, i.e., relationships between the predicted positions $(X_p(z), Y_p(z))$ and measured positions $(X_o(z), Y_o(z))$ in $z$-domain are derived as follows:

$$(X_p(z), Y_p(z)) = \frac{A(z)}{H(z)} X_o(z) + \frac{B(z)}{H(z)} Y_o(z), \quad \frac{C(z)}{H(z)} X_o(z) + \frac{D(z)}{H(z)} Y_o(z), \quad (9)$$
where

\[ A(z) = \omega z(z - 1)(\beta_x z + \beta_x \gamma_y - \beta_y \gamma_x + \alpha_x \delta_y - \alpha_y \delta_x - \beta_x)T \sin \omega T \\
+ \left((\omega^2 z^3 + \alpha_x \gamma_y - \alpha_y \gamma_x - \alpha_x)T^2 \right)
+ (\omega(z + 1)(\delta_x z + \delta_x \gamma_y - \delta_y \gamma_x - \delta_x \alpha_x \beta_y + \alpha_y \beta_x))T \\
+ 2(\beta_y \delta_x - \beta_x \delta_y) \cos \omega T + \omega^2 z^2(z + 1)(\alpha_x z + \alpha_x \gamma_y - \alpha_y \gamma_x - \alpha_x)T^2
\]

\[ B(z) = \omega(z - 1)T(\beta_y z(z - 1) \sin \omega T - z(2\alpha_y \omega T - \delta_y(z + 1)) \cos \omega T + \alpha_y \omega(z^2 + 1)T - \delta_y z(z + 1)) \\
C(z) = \omega(z - 1)T(\delta_x \sin(\omega T)z^2 - \beta_x \cos(\omega T)z + T \gamma_z \omega z^2 + \beta_x z^2 + \beta_x \sin(\omega T)z \\
- \delta_x \sin(\omega T)z + \beta_x \cos(\omega T)z - 2T \gamma_x \omega \cos(\omega T)z - \beta_x \cos(\omega T)z \\
+ T \gamma_x \omega \sin(\omega T)^2 + T \gamma_x \omega \cos(\omega T)^2) \\
D(z) = -T \omega(z - 1)(T \alpha_y \omega(z^2 + 1) - \cos(\omega T)z(-\delta_y z + 2T \alpha_y \omega - \delta_y) + \delta_y z(z + 1) \\
+ \beta_y \sin(T \omega)(z - 1)(\delta_x z^2 - \delta_y z) + \cos(T \omega)(-\beta_x z^2 - 2T \gamma_x \omega z - \beta_x z) \\
+ T \gamma_x \omega z^2 + \beta_x z^2 + \sin(T \omega)^2(\beta_x z + T \gamma_x \omega z) \\
+ \sin(T \omega)(\delta_x z^2 - \delta_y z) + \cos(T \omega)(-\beta_x z^2 - 2T \gamma_y \omega z - \beta_y z) \\
+ T \gamma_y \omega z^2 + \beta_y z^2 + \sin(T \omega)^2(\beta_y z + T \gamma_y \omega z) \]

\[ H(z) = \omega(z - 1)(\beta_x z + \beta_x \gamma_y - \beta_y \gamma_x + \alpha_x \delta_y - \alpha_y \delta_x - \beta_x)T \sin \omega T \\
+ \left((\omega^2 T^2 z^2 + \gamma_y z + \alpha_x z - 2z + \alpha_x \gamma_y - \gamma_y - \alpha_y \gamma_x - \alpha_x + 1) \right)
+ \omega T(z + 1)(\delta_x z - \beta_y z - \delta_x \gamma_y - \delta_x \alpha_x \beta_y + \beta_y + \alpha_y \beta_x) \\
+ 2z^2(\beta_y \delta_x - \beta_x \delta_y) \cos \omega T + \omega^2 T^2(z^2 + 1)(\delta_x z - \beta_y z + \delta_x \gamma_y - \delta_y \gamma_x \\
- \delta_x - \alpha_x \beta_y + \beta_y + \alpha_y \beta_x) + 2z^2(\beta_x \delta_y - \beta_y \delta_x). \]

Then, the closed-form bias errors for the CV target in the steady state are presented herein. The true target position is assumed as

\[ (x_{i,k}, y_{t,k}) = (v_x kT, v_y kT). \] (10)

where \( v_x \) and \( v_y \) are the constant velocities of the target. The bias errors for the CT filter tracking are then defined as

\[ (e_{x,k}, e_{y,k}) = (x_{i,k} - x_{p,k}, y_{t,k} - y_{p,k}). \] (11)

Then, the steady-state bias errors are:

\[ (e_{x,\text{fin}}, e_{y,\text{fin}}) = (\lim_{k \to \infty} e_{x,k}, \lim_{k \to \infty} e_{y,k}). \] (12)

Because the derivations of \( e_{x,k} \) and \( e_{y,k} \) are identical, only the derivation of \( e_{x,\text{fin}} \) is presented below. Because we consider the bias errors and not the measurement errors, \( x_{i,k} = x_{o,k} \) is assumed. Thus,

\[ (X_i(z), Y_i(z)) = (X_o(z), Y_o(z)) = \left( \frac{z}{(z - 1)^2}, \frac{z}{(z - 1)^2} \right) \]

Using the final value theorem, \( e_{x,\text{fin}} \) is calculated as
\[ e_{x,\text{fin}} = \lim_{\varepsilon \to 1} \left( z - 1 \right) \cdot E_x(z) = \lim_{\varepsilon \to 1} \frac{z - 1}{\varepsilon} \cdot (X(z) - X_p(z)). \]  

By substituting Eqs. (9) and (13) into (14), we obtain \( e_{x,\text{fin}} \) as follows:

\[ e_{x,\text{fin}} = \frac{(\delta_y v_y + \beta_y v_x)\omega T^2 - (\alpha_y v_y - \gamma y v_x)\omega^2 T^3}{(\alpha x y - \alpha y x)\omega^2 T^2 + (\delta_y y - \delta x y + \alpha x \beta y - \alpha y \beta x)\omega T + \beta y \delta y - \beta y \delta x} \]  

Using the same process as that for \( e_{x,\text{fin}} \), we obtain \( e_{y,\text{fin}} \) as follows:

\[ e_{y,\text{fin}} = \frac{(\alpha x v_y - \gamma y v_x)\omega^2 T^3 - (\delta_y v_y + \beta_y v_x)\omega T}{(\alpha x y - \alpha y x)\omega^2 T^2 + (\delta_y y - \delta x y + \alpha x \beta y - \alpha y \beta x)\omega T + \beta y \delta y - \beta y \delta x} \]  

Here, Eqs. (15) and (16) are the main results of this letter and are expressed in their closed forms using the parameters of the motion model of the CT tracking filter \((T, \omega)\) and the filter gains.

### 4 Numerical simulation

We present the numerical simulations to validate the derivation results. The target motion is set as \((x_{t,k}, y_{t,k}) = (kT, kT)\) [m] for \(kT \leq 4\) s \((k = 1, 2, \cdots)\), \((400T + \cos(-\omega_1(k-400)T + \pi)+1, 400T + \cos(-\omega_1(k-400)T + \pi)+1)\) [m] for \(4 < kT \leq 8\) s, and \((400T + \cos(-400\omega_1T + \pi)+1 - 0.2(k-800)T, 400T + \cos(-400\omega_1T + \pi)-0.3(k-800)T)\) [m] for \(8 < kT \leq 12\) s, where \(\omega_1\) is the angular velocity of the target and \(\omega_1=\pi/3\) rad/s. Figure 1 shows the true orbit of the target. The target first moves with constant velocity, then turns with constant angular velocity, and finally resumes CV motion.

We investigated and compared the three simulation cases summarized in Table I. We empirically set low- and high-gain settings of Table I to confirm the performance difference between these settings (the high-gain setting is relatively suitable for high maneuver target tracking). For all cases, initial values of the predicted target position and velocity were simply set to zero because we assumed the tracking was started at \(t=0\) without a priori information of the target and this study focused on the steady-state (not the transient state). The prediction error of each \(k\) was then defined as

\[ e_k = \sqrt{(x_{t,k} - x_{p,k})^2 + (y_{t,k} - y_{p,k})^2}. \]

Figure 2 shows the simulation results where steady-state bias errors are confirmed in the tracking of CV motions for all cases. The values of the steady-state bias errors for \(kT \leq 4\) s in Cases A, B, and C are \(1.49 \times 10^{-3}, 1.49 \times 10^{-5},\) and \(2.97 \times 10^{-3}\) m. The values of the steady-state bias errors for \(kT > 8\) s in Cases A, B, and C are \(3.71 \times 10^{-4}, 3.73 \times 10^{-6},\) and \(7.42 \times 10^{-4}\) m. These values are matched to the calculated results of \(e_{x,\text{fin}}^2 + e_{y,\text{fin}}^2\) using Eqs. (15) and (16). The steady-state errors for the CV target of Case A were 100 times larger than those of Case B; this shows the effects of the sampling interval, which is in agreement with Eqs. (15) and (16). The steady-state errors in Case C are larger than those of Case A, and these show the differences depending on the filter gains. Thus, the simulation results indicate the validity of the derived closed-form bias errors.

For all the simulation cases, although relatively small errors were achieved when the targets turned with constant angular velocities, the errors did not converge and did
Fig. 1. True motion of the target in the simulation: (a) orbit in $x$-$y$ plane; (b) time variations of $x$ and $y$.

Table I. Simulation conditions.

| Simulation case | $T$ [s] | Gain setting |
|-----------------|---------|--------------|
| Case A          | 0.01    | high-gain    |
|                 |         | $(\alpha_s=0.6, \beta_s=0.5, \beta_\theta=0.5)$ |
| Case B          | 0.001   | high-gain (same as Case A) |
| Case C          | 0.01    | low-gain     |
|                 |         | $(\alpha_s=0.3, \beta_s=0.25, \beta_\theta=0.25)$ |

Fig. 2. Simulation results.

not have large differences when the targets moved with constant velocity, especially for target velocities that were small for $kT > 8$. In our scenarios, due to the non-linearity of the process model, the achieved errors for the CT target were not small compared with the tracking of the CV target. These results show the cases where the CT tracking filter is not sufficiently effective even when the target includes turning motions and can achieve relatively small errors for CV targets.

5 Conclusion

In this work, we analyzed the steady-state bias errors of a CT tracking filter for tracking a CV target. The closed-form steady-state bias errors were derived as Eqs. (15) and (16) and expressed using filter gains, sampling interval $T$, and the CT model parameter $\omega_s$, which comprise the design parameters of the CT tracking filter. The numerical simulations show the validity of the derived steady-state bias errors for the designed filter for the maneuvering target tracking system.