Experimental demonstration of SPM compensation using a complex-valued neural network for 40-Gbit/s optical 16QAM signals

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Abstract: We experimentally demonstrate a novel nonlinearity-mitigation scheme based on a complex-valued neural network (CVNN) which is constructed by artificial neurons with complex-valued input and output. The in-phase (I) and quadrature (Q) components of optical signal are operated as complex values in the CVNN. A 40-Gbit/s optical 16QAM signal distorted by SPM was successfully compensated, improving error vector magnitude (EVM) by about 15%. The learning speed of the nonlinear equalizer was improved by using the CVNN, compared with conventional real-valued neural network (RVNN). Furthermore, the study show that CVNN has the potential to improve the computational complexity of RVNN.

Keywords: optical fiber communications, nonlinear distortion, SPM, digital signal processing, artificial neural network

Classification: Fiber-Optic Transmission for Communications

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1 Introduction

Waveform distortion caused by nonlinear effects such as self-phase modulation (SPM) is one of the important issues to be overcome to realize higher-performance optical fiber networks. Some methods have been proposed for compensating for nonlinear effects, including digital back propagation (DBP) and Volterra series transfer function (VSTF) [1, 2]. However, these methods require enormous amounts of calculation. Digital signal processing based on real-valued neural networks (RVNNs) is one important candidate for reducing the amount of calculation. The calculation cost of an RVNN increases in proportion to the number of neurons in each layer, whereas that of the VSTF increases exponentially as the numbers of tapped delays and the order of the Volterra series increase. In optical communication systems, the nonlinear equalizers based on RVNN have been used in frequency domain for optical orthogonal frequency division multiplexing (OFDM) systems [3, 4]. We have proposed a novel nonlinear equalization method using an RVNN to compensate optical multi-level signals distorted by the non-linearity of optical fibers [5, 6, 7]. On the other hand, complex-valued neural networks (CVNNs) are being investigated in the field of machine learning as a scheme that can accelerate the learning process compared with RVNNs [8, 9]. We proposed a novel nonlinear equalizer based on the CVNN [10]. In this paper, we experimentally demonstrated the nonlinear compensation performance of a CVNN.
In this experiment, a 16-ary quadrature amplitude modulation (16QAM) signal distorted by SPM was compensated by the CVNN. We evaluated the performance in terms of learning speed and error vector magnitude (EVM). Furthermore, we demonstrate that CVNN has the potential to improve the computational complexity of RVNN.

2 Nonlinear equalization using a CVNN

Fig. 1(a) shows the construction of an artificial neuron used in CVNNs. The inner potential of the neuron, $u$, is described as

$$u = \sum_{k=1}^{n} (x_{rk} + jx_{ik}) \times (w_{rk} + jw_{ik}) + (b_r + jb_i),$$  \hspace{1cm} (1)

where $x$ is the input signal, $w$ is the weight, and $b$ is the bias. The output of the neuron, $y$, is expressed as:

$$y = f(u),$$  \hspace{1cm} (2)

where $f$ is the complex output function of the neuron. Fig. 1(b) shows the construction of a three-layer CVNN that we used in the nonlinear compensation. The input layer of the CVNN has a feedforward tapped delay line. Input signals of in-phase (I) and quadrature (Q) components are fed into the delay line. The neurons in the input and output layers have linear output functions. Neurons in a hidden layer have the sigmoidal output function described by

$$y = \frac{1}{1 + e^{-u_r}} + j \frac{1}{1 + e^{-u_i}}.$$

(a) Complex-valued artificial neuron. (b) Three-layer CVNN. (c) System setup of 16QAM transmission.

Fig. 1. (a) Complex-valued artificial neuron. (b) Three-layer CVNN. (c) System setup of 16QAM transmission.
The values of the weight and bias are calculated by the Error Back Propagation (EBP) algorithm, and the error, which is defined as the difference between the output signals and supervised signals, is minimized. The error $e$ is expressed as

$$e = \sum_{k=1}^{n} |Y_k - T_k|^2,$$

where $Y_k$ is the output signal of the $k$-th output-layer neuron, and $T_k$ is the supervised signal. To minimize the error $e$, the weight values $w_k$ are updated by the value:

$$\Delta w_k = -\mu \frac{\partial e}{\partial \text{Re}[w_k]} - j\mu \frac{\partial e}{\partial \text{Im}[w_k]},$$

where $\mu$ is the learning rate. In our investigation, the number of input-layer neurons was set to 7, and the numbers of the hidden-layer neurons and output-layer neurons were set to 10 and 1, respectively.

### 3 System setup

Fig. 1(c) shows a 20-km 16QAM signal transmission system used in our experiment. A 10-Gsymbol/s (40-Gbit/s) 16QAM optical signal was modulated by PRBS $2^{15}-1$ data and transmitted on a 20 km length of standard single mode fiber (SSMF). The wavelength of the laser diode (LD) was 1549.7 nm. The input power to the optical fiber was as large as 12.5 dBm to induce fiber-optic nonlinearity in the SSMF. After transmission, the optical signal was attenuated using a variable optical attenuator (ATT) to observe the performance versus received optical power. The noise figure of the Er-doped fiber amplifier (EDFA) was about 5 dBm. The optical signal was received by optical homodyne detection using an optical 90°-hybrid and balanced photodetectors (BPDs). Off-line digital signal processing (DSP) was performed to execute a digital coherent algorithm including carrier phase estimation. The optical power of the local oscillator (LO) was 11.5 dBm. The distorted signal after transmission was compensated using a CVNN or an RVNN in the DSP. The compensation performance was evaluated by EVM.

### 4 Result and discussion

First, we performed a learning process to find the weights and biases of each neuron for the nonlinear compensation. Fig. 2(a) shows error $e$ versus iteration steps in the cases of RVNN and CVNN. We performed the learning process 20 times, changing the random initial values of the weight and the bias each time. It should be noted that the learning speed was dramatically improved by employing CVNN. The averaged value of error $e$ at 100 iterations was 0.133 with RVNN and 0.074 with CVNN. Fig. 2(b) shows the constellation of the received 16QAM signal in a back-to-back (BtB) configuration when the received optical power was −22 dBm. Fig. 2(c) shows the constellation after the transmission. Due to the large input power to the optical fiber, the outer symbols of the 16QAM signal were rotated in the clockwise direction by SPM. Fig. 2(d) shows the constellation after the compensation using the RVNN. Fig. 2(e) shows the constellation after the compensation using the CVNN. In both cases, the distorted symbols were success-
fully compensated. EVM was improved by about 15%. Fig. 2(f) shows EVM characteristics calculated for the cases using RVNN and CVNN. In both cases, an EVM of less than 15% was achieved when the received optical power was higher than about $-34$ dBm. From the EVM characteristics, we could not find any significant difference in the compensation capabilities between the CVNN and the RVNN. Next, we investigated the effect of the number of neurons of the RVNN and the CVNN on the compensation performance. In Fig. 3, we plotted the values of 10-times trials in each condition and drew straight lines connecting each average. Fig. 3(a) shows EVM versus the number of hidden-layer neurons. The number of neurons in input-layer was kept at 7. It should be noted that only one hidden-layer neuron can work for the nonlinear compensation to some extent, whereas RVNN requires more than one neuron. Fig. 3(b) shows EVM versus the number of input-layer neurons. The number of hidden-layer neurons was kept at 10. It should be noted that the CVNN needed only about half the number of input-layer neurons to realize the same EVM values. This efficient performance of a CVNN contributes to decrease the calculation cost of DSP for nonlinear equalizer. One complex
multiplication requires four real multiplications and two real additions. However, some complex arithmetic methods have been developed to achieve complex multiplications in DSP efficiently [11]. A CVNN has the potential to improve the computational complexity of the nonlinear equalizers.

5 Conclusion
We experimentally investigated the performance of a CVNN to compensate nonlinear waveform distortion caused by SPM. The results showed that the CVNN could efficiently compensate the distortion and could improve the calculation cost of DSP in optical communication systems.
SPM and phase-noise tolerant optical self-homodyne using a polarization-multiplexed and intensity-modulated pilot-carrier

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Abstract: We propose a novel self-homodyne scheme using a polarization-multiplexed and intensity-modulated pilot-carrier. The scheme has the capability to cancel nonlinear waveform distortion caused by self-phase modulation (SPM) in optical fibers, as well as the phase-noise of a laser diode. It has been known that a conventional constant-intensity pilot-carrier can cancel the waveform distortion caused by SPM to some extent. Theoretically, however, the proposed intensity-modulated pilot-carrier can completely cancel the effect of SPM. The performance in terms of the error vector magnitude was evaluated by numerical simulations of 50-km optical fiber transmission of 16-ary quadrature amplitude modulation (16QAM) signals.

Keywords: self-homodyne, pilot-carrier, SPM, phase-noise, optical fiber

Classification: Fiber-Optic Transmission for Communications

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1 Introduction

Waveform distortion caused by optical nonlinear effects, including self-phase modulation (SPM), is becoming a more serious problem in optical transmission systems, especially when using optical multilevel modulation schemes, such as 16-ary quadrature amplitude modulation (16QAM), because the signal power varies according to the transmitted symbols, resulting in a large peak-to-average-power ratio (PAPR). Some methods have been proposed for compensating for nonlinear effects, including digital back propagation (DBP) and the Volterra series transfer function (VSTF) [1, 2]. However, these methods based on digital signal processing (DSP) need an enormous amount of calculations, which causes serious time delay and increase in cost at the receiver. Phase conjugate twin wave (PCTW) schemes are drawing attention as a new method that has the ability to cancel nonlinearities in optical domain [3]. In this scheme, a phase-conjugate lightwave is multiplexed with the modulated signal by polarization multiplexing or time-division multiplexing (TDM) [4, 5]. On the other hand, we have proposed and investigated novel schemes that have the ability to cancel SPM and phase noise at the same time. In these schemes, a pilot carrier with constant optical power is multiplexed with a modulated signal by polarization multiplexing or TDM [6, 7, 8, 9]. In this paper, we propose a novel pilot-carrier scheme. An intensity-modulated pilot-carrier is polarization-multiplexed with a modulated optical signal. The performance, in terms of the error vector magnitude (EVM), was investigated by numerical simulations. The results clarified that the SPM cancellation capability was improved compared to that achieved with conventional constant-intensity pilot-carrier schemes.
2 Principle

Figs. 1(a) and (b) schematically show the principle of the proposed polarization-multiplexing and intensity-modulated pilot-carrier [10]. The lightwave from a laser diode (LD) is split into two polarization components using a polarization beam splitter (PBS). One polarization component is modulated as a signal. On the other hand, only the intensity of the other component is modulated so that it has the same power as the signal component, as shown in Fig. 1(a). This intensity-modulated component is used as pilot-carrier, which provides an absolute optical phase reference for self-homodyne detection [7]. Fig. 1(b) shows the principle of SPM and phase-noise cancellation. The intensity-modulated pilot-carrier has the same amplitude as the modulated signal. As a result, the optical phases of the two polarization components are identically rotated by SPM, keeping the mutual relative phase angle between them. Furthermore, this relative phase angle does not fluctuate due to the phase-noise of the LD, because the pilot-carrier has the same phase-noise as the signal component. On the receiver side, self-homodyne detection is performed using the pilot-carrier as a phase reference, where the influence of SPM and phase-noise can be cancelled. The amplitude of the receiver signal after the self-homodyne varies according to the square of modulated optical signal. However, the influence can be cancelled by digital signal processing (DSP).

In the case of the conventional constant-intensity pilot carrier, however, the intensity of the pilot carrier is not the same as the modulated signal because the

![Diagram showing the principle of polarization-multiplexed and intensity-modulated pilot-carrier.](image1)

(a) Polarization-multiplexed and intensity-modulated pilot-carrier.

![Diagram showing compensation of nonlinear distortion and phase-noise.](image2)

(b) Compensation of nonlinear distortion and phase-noise.

Fig. 1. Principle of polarization-multiplexed and intensity-modulated pilot-carrier.
intensity of the modulated signal varies symbol by symbol. Therefore, the optical phases of the two polarization components are not identically rotated by SPM. Therefore, the conventional pilot carrier scheme can not keep the mutual relative phase angle between the modulated signal and the pilot carrier. This is the reason why the SPM-compensation capability of the conventional constant-intensity pilot carrier is limited [8].

3 System setup

The performance of the proposed scheme was evaluated by numerical simulation. Fig. 2 shows a 50-km 16QAM signal transmission system using a polarization multiplexed and intensity-modulated pilot-carrier and self-homodyne detection. The lightwave from an LD is split into two polarization components using a PBS. One polarization component is modulated by 10-GSymbol/s 16QAM with PRBS $2^{11}-1$ data, whereas the other component is intensity modulated so that the two polarization components have the same optical power. Here, we assumed that the electrical skew between the data signal and the pilot carrier is zero. It should be noted that this modulation can be realized using a commercially available dual-polarization quadrature phase-shift keying (DP-QPSK) modulator. The modulated lightwave is transmitted by a standard single-mode fiber (SSMF) and a dispersion compensating fiber (DCF) having a total length of 50 km, thus canceling the total chromatic dispersion. The dispersion of the SSMF and the DCF was 16.75 ps/nm/km and $-77$ ps/nm/km, respectively. The noise figure of the Er-doped fiber amplifiers (EDFAs) was 3 dB. The transmission through the optical fibers was numerically calculated using well-known split-step Fourier method. The input power to the optical fibers was 10 dBm. The linewidth of the LD was varied between 100 kHz and 30 MHz in order to confirm the phase-noise cancellation capability. It was assumed that the phase fluctuation was taken to be a Gaussian random process [11]. On the receiver side, self-homodyne detection was performed using the pilot-carrier as a phase reference with PBS and DSP. The thermal noise of the electric devices in the transmitter and the receiver was neglected. Using the
same transmission system, we also performed conventional homodyne detection without the pilot-carrier for comparison. Here, we assumed an ideal local oscillator with an optical power of 0 dBm, which was ideally synchronized to the optical signal. We also assumed an ideal clock recovery both in the self-homodyne and the conventional homodyne receivers. The received signals were evaluated by EVM which is defined as signal fluctuation from the ideal symbol position. The definition of the EVM is shown as

$$EVM = \frac{1}{N} \sum_{i=1}^{N} \frac{|E_i - \hat{E}_i|}{|E_i|} \times 100\%,$$

where $E_i$ is the complex amplitude of the $i$-th transmitted symbol, $\hat{E}_i$ is the ideal complex amplitude of the $i$-th transmitted symbol, and $N$ is the total number of the transmitted symbols. Therefore, the value of EVM includes the phase rotation by SPM and the symbol fluctuation caused by noise.

### 4 Results and discussion

Figs. 3(a) shows the constellation of the received signal using conventional homodyne detection without a pilot-carrier. The received optical power was $-15$ dBm. Here, the received optical power means the averaged optical power including both of the modulated signal and the pilot carrier. The linewidth of the LD was 100 kHz. The 16QAM signal was seriously distorted by SPM. When we employed a conventional constant-intensity pilot-carrier, we could observe SPM cancellation capability to some extent, as shown in Fig. 3(b). We observed some waveform distortion where outer symbols were slightly rotated clockwise. However, when we employed the proposed intensity-modulated pilot-carrier, the performance of the SPM cancellation was improved, as shown in Fig. 3(c). Fig. 3(d) shows EVM characteristics versus received optical power. In the case of homodyne detection, the EVM was larger than 33% due to the waveform distortion caused by SPM. When we employed the conventional constant-intensity pilot-carrier, an EVM value of less than 9% was achieved. In the case of our proposed intensity-modulated pilot-carrier, however, the EVM performance was improved by more than about 5% in comparison with the conventional intensity-constant pilot-carrier. Furthermore, even when we used an LD with a linewidth of as large as 30 MHz, the performance was not degraded seriously due to the phase-noise cancellation capability, as shown in Fig. 3(e). We plotted the EVM performance of the 16QAM signals versus the linewidth of the LD in Fig. 3(f). The EVM degradation at a linewidth of 30 MHz was less than 1%. The EVM values in Fig. 3(d) are different from that in our previous report [10]. In the previous report, we employed a dispersion-shifted fiber (DSF) and extremely large optical input power of 17 dBm. The EVM difference was caused mainly by this points in the system setup.
A novel intensity-modulated pilot-carrier scheme was proposed. Numerical simulations clearly showed the SPM cancellation capability of the proposed scheme and improved performance over the conventional constant-intensity pilot-carrier scheme. Linewidth tolerant performance was also demonstrated in the simulation. The proposed scheme requires the construction of a special modulator; however, it can be implemented using a commercially available DP-QPSK modulator.

**Fig. 3.** Results of numerical simulations.

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**5 Conclusion**

A novel intensity-modulated pilot-carrier scheme was proposed. Numerical simulations clearly showed the SPM cancellation capability of the proposed scheme and improved performance over the conventional constant-intensity pilot-carrier scheme. Linewidth tolerant performance was also demonstrated in the simulation. The proposed scheme requires the construction of a special modulator; however, it can be implemented using a commercially available DP-QPSK modulator.
SPM and phase-noise cancellation using time-division-multiplexed and intensity-modulated pilot symbols

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Abstract: We propose a novel SPM and phase-noise cancellation scheme using pilot symbols. The pilot symbols are intensity-modulated and time-division-multiplexed with a modulated optical signal. Each pilot symbol has the same optical intensity as the multiplexed modulated symbol and therefore has the same nonlinear phase shift caused by self-phase modulation (SPM). Additionally, each pilot symbol has the same phase noise as the modulated symbol because it is generated by the same laser diode (LD). At the receiver side, self-homodyne detection is performed using the pilot symbols as absolute optical phase references. The performance of the scheme was investigated by numerical simulations of 50-km optical fiber transmission of 16-ary quadrature amplitude modulation (16QAM) signals.

Keywords: nonlinear compensation, pilot carrier, SPM, phase-noise, self-homodyne

Classification: Fiber-Optic Transmission for Communications

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1 Introduction

Multi-level modulation schemes such as 16-ary quadrature amplitude modulation (16QAM) are taking an important role in optical fiber technologies to improve the transmission speed and the spectral efficiency. However, signal degradation due to optical nonlinear effects including self-phase modulation (SPM) is becoming a more serious problem because the signal power of a high-order multi-level modulation signal varies according to the transmitted symbols, resulting in a large peak-to-average-power ratio (PAPR). One attractive method to compensate for such nonlinear effects is digital signal processing (DSP), such as digital back propagation (DBP) and the Volterra series transfer function (VSTF), which can equalize the nonlinear distortion without using additional optical components [1, 2]. However, these nonlinear equalizers using DSP need an enormous amount of calculations, which causes serious time delays and increases the power consumption of the receiver. Phase-conjugated twin wave (PCTW) schemes are drawing attention as a new scheme that can efficiently cancel nonlinear effects without requiring heavy DSP [3]. In this scheme, a phase-conjugated lightwave is multiplexed with a
modulated optical signal by polarization multiplexing or time-division multiplexing (TDM) [4, 5]. In contrast, we have proposed and investigated a polarization-multiplexed pilot-carrier scheme that simultaneously exhibits SPM-tolerant characteristics and a phase-noise cancellation capability [6, 7, 8]. We also proposed a polarization-multiplexed and intensity-modulated pilot-carrier scheme that can completely compensate for the effect of SPM [9]. However, these schemes require polarization tracking, which increases the amount of calculations in the DSP [10]. In this study, we propose a novel pilot-signal scheme, where pilot symbols are time-division-multiplexed with a modulated optical signal [11]. At the receiver side, self-homodyne detection is performed by differential detection using a 1-symbol delay line. The error vector magnitude (EVM) performance was investigated by numerical simulations, and the results confirmed the nonlinear compensation capability. Furthermore, we also demonstrated the phase-noise cancellation capability by changing the linewidth of the light source from 100 kHz to 30 MHz.

2 Principle

Figs. 1(a) and (b) show the scheme of the proposed time-division-multiplexed and intensity-modulated pilot symbols. The lightwave from a laser diode (LD) is return-to-zero (RZ) carved and split into two branches. One is modulated as an RZ multilevel signal. The other one is intensity-modulated only so that it has the same optical intensity as the RZ signal pulse. This intensity-modulated RZ pulse train is used as the pilot symbols, which provides an absolute optical phase reference for self-homodyne detection. The modulated signal pulses and the intensity-modulated pilot symbols are time-division-multiplexed as shown in Fig. 1(a). Fig. 1(b) shows the principle of SPM and phase-noise cancellation. Each intensity-modulated pilot symbol has the same amplitude as the modulated signal.
As a result, the optical phases of the two RZ pulses are identically rotated by SPM, keeping the mutual relative phase angle between them. Furthermore, this relative phase angle does not fluctuate due to the phase-noise of the LD, because the pilot symbol has the same phase-noise as the modulated signal. On the receiver side, self-homodyne is performed by differential detection using a 1-symbol delay line. The influence of SPM and phase noise is compensated for because the pilot symbol has phase noise and nonlinear phase-shift identical to those of the modulated signal. The amplitude of the received signal after the self-homodyne varies according to the square of the intensity of the RZ signals. However, the influence can be cancelled by DSP. Additionally, the differential detection generates unwanted signal pulses caused by self-homodyne with neighboring signal pulses. This influence is also eliminated by DSP.

3 System setup

The performance of the proposed method was evaluated by numerical simulation. Fig. 2 shows a 50-km 16QAM signal transmission system using the time-division-multiplexed and intensity-modulated pilot symbols and self-homodyne detection. The lightwave from an LD was RZ carved into a 10 GHz pulse train with a 40% duty cycle. The generated RZ pulses were split into two branches. One was modulated by 10-GSymbol/s 16QAM with PRBS 2^{11}-1 data, whereas the other branch was intensity-modulated so that the two branches had the same optical intensity. The modulated RZ pulses and the intensity-modulated pilot-symbols were time-division-multiplexed using a 50 ps delay line and transmitted by a standard single mode fiber (SSMF) and a dispersion compensation fiber (DCF) having a total length of 50 km, thus cancelling the total chromatic dispersion. The dispersions of the SSMF and the DCF were 16.75 ps/nm/km and −77 ps/nm/km, respectively.

The transmission through the optical fibers was numerically calculated using the well-known split-step Fourier method. The input power to the optical fiber was adjusted from 0 to 10 dBm. On the receiver side, self-homodyne detection was performed using two 50 ps delay lines for in-phase (I) and quadrature (Q) signal components. Using the same transmission system, we also performed conventional homodyne detection without the pilot symbols for comparison. Here, we assumed an ideal local oscillator with an optical power of 10 dBm, which was ideally synchronized to the optical signal. The noise figure of the Er-doped fiber amplifiers
The linewidth of the LD was varied between 100 kHz and 30 MHz in order to confirm the phase-noise cancellation capability. The signal quality was evaluated by EVM.

4 Results and discussion

Fig. 3(a) shows the constellation of the received signal using conventional homodyne detection without pilot symbols in a back-to-back (BtB) condition. Fig. 3(b) shows the constellation of the homodyne detection after the transmission with an input optical power of 10 dBm. The received optical power was −15 dBm. The linewidth of the LD was 100 kHz. The 16QAM signal was seriously distorted by SPM. However, when we employed the detection using the intensity-modulated pilot symbols, the EVM performance was dramatically improved by the SPM cancellation capability of the proposed scheme, as shown in Fig. 3(c). Figs. 3(d)
and (e) show EVM characteristics versus received optical power of the conventional homodyne scheme and the proposed self-homodyne scheme, respectively. Here, the received optical power includes those of both the modulated signal and the pilot symbols. In the case of conventional homodyne, when we increased the input power to 10 dBm, EVM was degraded to 35% or worse due to the waveform distortion caused by SPM. In the case of the proposed self-homodyne, however, the EVM penalty was only less than 3%, even when we increased the input power from 0 dBm to 10 dBm. Furthermore, even when we used an LD with a linewidth of as large as 30 MHz, the same compensation performance was observed, as shown in Fig. 3(f). We plotted the EVM performance of the 16QAM signals versus the linewidth of the LD in Fig. 3(g). However, we could not observe any significant degradation of the performance even at a linewidth of 30 MHz.

5 Conclusions

A novel intensity-modulated pilot-symbol scheme was proposed. Numerical simulations showed the SPM compensation characteristics of the proposed scheme. The results also verified the phase-noise cancellation capability of the scheme, which will be necessary to realize future ultra-multi-level QAM transmission technology [12].
Quadruple-frequency optical two-tone signal generation using a DP-QPSK modulator

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Abstract: We investigate a novel frequency-quadruple optical two-tone signal generation method using a DP-QPSK modulator, which includes four Mach-Zehnder modulators (MZMs). One MZM is used to generate a basic two-tone signal. The other MZMs are used to compensate unwanted high-order harmonics and carrier components of the generated signal. In the experiment, a two-tone signal was successfully generated, and the power levels of fourth-order harmonics and the carrier were reduced to levels 36.8 dB and 40.3 dB smaller than the required signal, respectively. We also evaluated suppression ratio of the unwanted components against fluctuations of amplitude and bias voltages for the MZMs.

Keywords: two-tone signal, Mach-Zehnder modulator, radio-over-fiber

Classification: Wireless Communication Technologies

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Radio-over-fiber (RoF) is regarded as one of the important technologies that can meet the increasing demand for higher transmission speeds on broadband wireless communication networks. One key aspect of RoF is the generation of a two-tone lightwave signal. An optical two-tone signal can be generated by using a Mach-Zehnder modulator (MZM); however, one problem with this method is that nonlinearity and an asymmetric splitting ratio in the MZM increase unwanted high-order harmonics of the generated two-tone signal, which cause waveform distortion. Some methods for suppressing unwanted high-order harmonics have been studied [1, 2, 3, 4, 5, 6, 7, 8]. One simple method employs MZMs connected in series to suppress the third-order harmonics component [1]. Some methods employ specially designed MZMs with a high extinction ratio [2, 3, 4], and some use dual-parallel MZMs (DP-MZMs) to suppress unwanted third-order harmonics components [5, 6, 7, 8]. In our previous work, we proposed a novel frequency-quadruple two-tone signal generation method using a commercially available dual-polarization quadrature phase shift keying (DP-QPSK) modulator that consists of four MZMs [9]. In this method, second-order harmonics are used to generate a quadruple-frequency optical two-tone signal. The unwanted carrier and first- and fourth-order harmonics components are suppressed using the DP-QPSK modulator at the same time. In this paper, we demonstrate the effectiveness of the proposed method by numerical simulation and an experiment. The principle is clarified using Bessel functions. Suppression performance of the unwanted carrier and harmonics components was evaluated by the experiment. Furthermore, we evaluated suppression ratio of the unwanted components against fluctuations of amplitude and bias voltages for the MZMs.

2 Principle

Fig. 1(a) schematically shows our proposed frequency-quadruple two-tone signal generation method in which a DP-QPSK modulator is used [9]. The DP-QPSK modulator includes four MZMs, namely MZM1, MZM2, MZM3, and MZM4. The radio frequency (RF) signals used to drive the four MZMs are expressed as
\[ v_1(t) = A_1 \sin(2\pi f_s t), \]
\[ v_2(t) = A_2 \sin(2\pi f_s t), \]
\[ v_3(t) = A_3 \sin(4\pi f_s t), \]
\[ v_4(t) = 0, \]

where \( A_n \) is the amplitude of the RF signal used to drive the \( n \)-th MZM. MZM3 is driven by an RF signal with a frequency \( 2f_s \), whereas MZM1 and MZM2 are driven with frequency \( f_s \). MZM4 is used to suppress the carrier, controlling only the bias point. The two-tone signal is basically generated by MZM1. An input lightwave is modulated by an RF signal \( f_s \) in MZM1, which is biased at the point of top. Here, MZM1 is driven by an RF signal with an amplitude of \( V_x \), where \( V_x \) is the half-wave voltage, to achieve maximum optical power. However, the two-tone signal generated by MZM1 includes a carrier component and high-order harmonics, which distort the generated sinusoidal waveform. Fig. 1(b) schematically shows the optical spectrum generated by MZM1. MZM2 is biased at the bottom point and is driven with frequency \( f_s \). Fig. 1(c) shows the optical spectrum of the output of MZM2. The first-order harmonics of MZM1 are canceled by the output of MZM2, while adjusting the driving amplitude of MZM2. Here, the phase of the output of MZM2 is adjusted by controlling the bias point of Phase 2 in Fig. 1(a). The condition for canceling the first-order harmonics is that the following relation is satisfied:

\[ J_1 \left( \frac{\pi A_2}{V_x} \right) = (1 - 2\gamma)J_1 \left( \frac{\pi}{2} \right), \]

where \( J_n \) and \( \gamma \) are the \( n \)-th order Bessel function of the first kind and the splitting ratio of the MZMs, respectively. MZM3 is used to suppress the fourth-order harmonics. MZM3 is biased at the top point and is driven with frequency \( 2f_s \), which is generated by a frequency doubler. The optical spectrum of the output of MZM3 is shown in Fig. 1(d). The driving amplitude and the phase of MZM3 is
adjusted so that the fourth-order harmonics of MZM1 are canceled. The condition is that the following relation is satisfied:

\[ J_2 \left( \frac{\pi A_1}{V_r} \right) = J_4 \left( \frac{\pi}{2} \right) \]  

(6)

Finally, we use MZM4 to suppress the unwanted carrier component generated by MZM1, MZM2, and MZM3. Fig. 1(e) shows the optical spectrum of the output of MZM4. The intensity and the phase are controlled by adjusting the bias voltage of MZM4 and Phase 4. The outputs of MZM1, MZM2, MZM3, and MZM4 are polarization multiplexed as x- and y-polarization components by a polarization beam combiner (PBC) after the polarization rotation section in the DP-QPSK modulator, and added together into a linearly polarized optical signal using a polarizer. A frequency-quadruple two-tone lightwave signal is generated, suppressing the first- and fourth-order harmonics and the carrier components, as shown in Fig. 1(f). Here, it should be noted that we can use a commercially available DP-QPSK modulator and a frequency doubler. Usually, the RF amplitude needed for the \( 2f_s \) signal is much smaller than \( V_r \), because the fourth-order harmonics component is about 20-dB smaller than the required two-tone signal. Therefore, an expensive modulator and frequency doubler with wideband frequency characteristics are not required in our proposed method.

3 System setup

The numerical simulation and the experiment was performed using the system setup shown in Fig. 2. In the figure, we omitted bias-controll electric lines for simplicity. A lightwave with a wavelength of 1549.5 nm (193.47 THz) was modulated by a X-cut DP-QPSK modulator. The modulator had a polarizer at the input so that only TE-polarized lightwaves were fed to the four MZMs through the power splitters in that. The lightwave from the LD was polarization-controlled (PC) to maximize the optical output power of the modulator. The lightwaves from the MZMs were added together as described above. MZM1 and MZM2 were driven by 10 GHz RF signals. MZM3 was driven by a 20 GHz RF signal generated by a frequency doubler. The phases of the RF signals for MZM2 and MZM3 were controlled using RF phase shifters (PSs). Attenuators (ATTs) were used for adjusting the amplitudes of the driving RF signals. The driving RF amplitudes for the DP-QPSK modulator and bias points were adjusted while observing the output.
optical spectrum using an optical spectrum analyzer (OSA; Anritsu, MS9710C, 0.05 nm resolution). The extinction ratio of the MZMs was assumed to be 20 dB in the simulation.

4 Results and discussion

First, we performed a numerical simulation to investigate the ideal performance of our proposed scheme. Fig. 3(a) and (b) show the results of the numerical simulation. Fig. 3(a) is the optical spectrum when only MZM1 was used. Among the many harmonics, the second-order harmonics component is needed to realize the

![Optical spectrum](image)

Fig. 3. Optical spectra of output lightwaves and suppression of unwanted high-order harmonics and carrier components.
quadruple-frequency two-tone signal. Fig. 3(b) is the optical spectrum when all the MZMs were active. In the numerical simulation, we could completely cancel the first- and fourth-order harmonics and the carrier components at the same time, as shown in Fig. 3(b). Figs. 3(c) and (d) show the results of the experiment. Fig. 3(c) is the optical spectrum when only MZM1 was used. The power level of the fourth-order harmonics was about 20.5 dB smaller than the required two-tone signal. The carrier component was larger than that in the simulation. This may have been caused by the carrier power from MZM2, MZM3, and MZM4, which were not modulated. When MZM1, MZM2, MZM3 and MZM4 were all used, the power levels of the fourth-order harmonics and carrier components were respectively suppressed to levels 36.8 dB and 40.3 dB smaller than the two-tone signal, as shown in Fig. 3(d). The power level of the first-order harmonics was also suppressed. However, we could not evaluate the suppression because of the inadequate resolution of our optical spectrum analyzer. We also evaluated the suppression ratio of the unwanted components against fluctuations of the amplitude of the RF signal $A_n$ and bias voltages $V_{\text{bias}}$ for the MZMs by numerical simulation. Fig. 3(e) shows the suppression ratio of the unwanted components against the amplitude fluctuation $\Delta A_n$ which is normalized by $V_{\pi}$. The suppression ratio was calculated by changing the fluctuation of one parameter, keeping the other amplitude and bias voltage at the optimum. Even when we adjusted all the parameters to optimum, the suppression ratio was limited by the unwanted third-order harmonics which is not suppressed in our proposed scheme. When we change the amplitude $A_3$ for MZM3, the characteristic curve became bilaterally asymmetric. This is because the $A_3/V_{\pi}$ is as small as 0.11 $V_{\pi}$. Therefore, when $A_3$ is decreased, the change of the suppression is soon saturated. To assure the suppression ratio of less than $-20$ dB, the amplitude fluctuation $\Delta A_n/V_{\pi}$ has to be controlled to less than about 0.03 $V_{\pi}$. Fig. 3(f) shows the suppression ratio against the fluctuation of bias voltages for MZMs. To assure the suppression ratio of less than $-20$ dB, the bias fluctuation $\Delta V_{\text{bias}}/V_{\pi}$ has to be also controlled to less than about 0.03 $V_{\pi}$. These adjustments of the amplitude and the bias can be controlled by using commercially available stabilizers for MZMs.

5 Conclusion

We studied a novel method for generating a two-tone lightwave signal, in which the first- and fourth-order harmonics and the carrier components are suppressed using a DP-QPSK modulator. The performance was investigated by numerical simulation and an experiment. The simulation results clearly showed that the proposed scheme works according to the theory. In the experiment, a quadruple-frequency two-tone signal was successfully generated, and the power level of the unwanted components was suppressed.
Novel optical twin-SSB detection scheme using an electric butterfly operation

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Abstract: We propose a novel detection scheme for an optical twin single-sideband (twin-SSB) signal. The detection can be performed by one optical homodyne receiver with an electric butterfly operation which demultiplexes the lower side-band (LSB) and upper side-band (USB) signals without using optical filters. The dispersion-tolerant characteristics of the scheme were investigated by numerical simulation of transmission through 50 km and 100 km standard single-mode fiber (SSMF), in comparison with dual sideband (DSB) modulation. The optical modulation of the twin-SSB signal was performed using a dual polarization quadrature phase shift keying (DP-QPSK) modulator.

Keywords: SSB, DP-QPSK modulator, dispersion, Hilbert transform, 16QAM

Classification: Fiber-Optic Transmission for Communications

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1 Introduction

Improving spectral efficiency is one of the important issues that must be solved to take full advantage of the limited wavelength resources of optical fiber transmission systems. One efficient method is to employ multi-level modulation, and spectral efficiency exceeding 10 bit/s/Hz has already been demonstrated experimentally by using this method [1, 2]. Another candidate to improve the spectral efficiency is optical single sideband (OSSB) modulation technology, which can potentially double the spectral efficiency [3, 4, 5]. The modulation occupies only one sideband, whereas conventional double sideband (DSB) modulation occupies both the lower side-band (LSB) and upper side-band (USB). Therefore, OSSB also improves the efficiency of wavelength division multiplexing (WDM) systems, where the spectra of OSSB signals are spaced wide enough apart in wavelength. Recently, a novel type of OSSB, called optical twin-SSB, has been investigated and has attracted attention [6, 7, 8]. A twin-SSB signal conveys two different messages on the LSB and USB individually. That is, two different SSB signals are arranged side-by-side.
without any spectral spacing in wavelength, improving the potential spectral efficiency of WDM systems. In the reported schemes, LSB and USB were demultiplexed by steep optical filters, and were demodulated using two individual SSB receivers. In this paper, we propose a novel twin-SSB detection scheme which can be performed using one optical homodyne receiver with an electric butterfly operation including Hilbert transforms, thus demultiplexing the LSB and USB without using optical filters. The two important features of the twin-SSB signal, i.e. the spectral efficiency and the dispersion tolerance, were investigated by numerical simulations.

2 Principle of modulation and demodulation

Figs. 1(a) and (b) schematically show the proposed twin-SSB scheme [9]. So far, three different types of twin-SSB modulation have been investigated, using a single Mach-Zehnder modulator, a quadrature phase shift keying (QPSK) modulator, and a dual polarization QPSK (DP-QPSK) modulator [10]. Here, we employed a DP-QPSK modulator to clearly demonstrate the principle of the proposed scheme. Using this method, we can clearly confirm the condition of the modulation, because the LSB and USB are generated by conventional OSSB modulation using QPSK modulators connected in parallel and Hilbert transforms, individually, as shown in Fig. 1(a) [11]. A 45-degree polarizer on the output of the DP-QPSK modulator is used to mix the two polarization components. Fig. 1(b) shows the receiver consisting of homodyne detection and an electric butterfly operation including Hilbert transforms. When the baseband signals for the LSB and USB and the carrier are expressed as

\[ r(t) = \cos(\omega_C - \omega_{SL})t + \sin(\omega_C + \omega_{SU})t. \]  \hspace{1cm} (1)

Here, we assume that the local oscillator (LO) is ideally synchronized to the received optical signal. The two outputs of the optical 90° hybrid, \( s_1(t) \) and \( s_2(t) \), are expressed as

\[ s_1(t) = \frac{1}{2} \cos(\omega_{SL}t) + \frac{1}{2} \sin(\omega_{SU}t), \] \hspace{1cm} (2)

\[ s_2(t) = \frac{1}{2} \sin(\omega_{SL}t) + \frac{1}{2} \cos(\omega_{SU}t). \] \hspace{1cm} (3)

Hilbert transforms of the signals \( s_1(t) \) and \( s_2(t) \) are expressed as

\[ s_3(t) = \frac{1}{2} \sin(\omega_{SL}t) - \frac{1}{2} \cos(\omega_{SU}t), \] \hspace{1cm} (4)

\[ s_4(t) = \frac{1}{2} \cos(\omega_{SL}t) + \frac{1}{2} \sin(\omega_{SU}t), \] \hspace{1cm} (5)

respectively. Consequently, the demultiplexed baseband signals of the LSB and USB can be obtained from the two outputs of the electric butterfly operation as

\[ s_1(t) - s_4(t) = \cos(\omega_{SL}t), \] \hspace{1cm} (6)

\[ s_2(t) - s_3(t) = \cos(\omega_{SU}t). \] \hspace{1cm} (7)

The LSB and USB are completely demultiplexed without using any optical filters. However, the proposed scheme requires Hilbert transforms, a kind of electric
filter, in the receiver. It should be noted that high-performance electric filters can be cost-effectively realized by digital signal processing (DSP) [12].

3 System setup

Fig. 2 schematically shows the system setup used in our numerical simulation for 10-GSymbol/s twin-SSB transmission over 50 km or 100 km of standard single-mode fiber (SSMF). A laser diode (LD) with a wavelength of 1550 nm was modulated by a pair of four-level baseband signals using the twin-SSB transmitter shown in Fig. 1(a). The four-level signals were composed of PRBS $2^{11}$-1 binary data. The modulated optical signal was transmitted by a 50 km or 100 km SSMF. The received optical power was adjusted using an attenuator (ATT) and was detected using the twin-SSB receiver shown in Fig. 1(b).

4 Results and discussion

First, we confirmed the spectral efficiency of the twin-SSB scheme. Fig. 3(a) shows an optical spectrum of typical four-level DSB optical signal, namely 4-ary ampli-
tude shift keying (4ASK) with a modulation speed of 10-GSymbol/s. Figs. 3(b) and (c) show the optical spectra of LSB and USB components of the twin-SSB signal, which were calculated at the output of 1st and 2nd DP-MZMs in the DP-QPSK modulator. The halved optical spectra were clearly observed. The sideband suppression ratio (SSR) was more than 30 dB. The spectra of LSB and USB were

Fig. 3. (a) Optical spectrum of 4ASK signal. (b) Output of 1st DP-MZM. (c) Output of 2nd DP-MZM. (d) Output of DP-QPSK modulator. (e) Eye-diagrams of twin-SSB signal (Received optical power: −35 dBm). Constellations of (f) twin-SSB and (g) DSB signals before and after 50-km transmission (Received optical power: −35 dBm.). (h) EVM performance of twin-SSB and DSB signals after 50-km and 100-km transmission.
added achieving twin-SSB modulation at the output of the DP-QPSK modulator as shown in Fig. 3(d). Next, we confirmed the demultiplexing performance of our proposed twin-SSB receiver in a back-to-back (BtB) configuration. Fig. 3(e) shows the eye-diagrams of the demodulated LSB and USB signals. Clear eye openings were achieved by the demodulation. We also performed the modulation and demodulation using some different modulation formats, such as binary signals and some different PRBS data for LSB and USB. However, we did not observe any significant crosstalk, because we used ideal Hilbert transforms in the simulation.

We evaluated the signal quality by drawing constellations using the received LSB and USB signals as I- and Q-components, respectively. Fig. 3(f) shows the constellations of the twin-SSB signal before and after 50-km SSMF transmission. For comparison, 16-ary quadrature amplitude modulation (16QAM) DSB signal transmission with the same bit rate was performed. Fig. 3(g) shows the constellations of the 16QAM DSB signal before and after 50-km SSMF transmission. In the case of twin-SSB, the constellation was clear even after the transmission. Fig. 3(h) shows EVM characteristics of the twin-SSB and 16QAM DSB signals after 50 km and 100 km SSMF transmission. By using the twin-SSB, the EVM performance was improved by about 9.5% and 23.9% at a received optical power of −20 dBm for 50 km and 100 km transmission, respectively. We could not observe any sensitivity degradation due to the twin-SSB modulation. The results show the chromatic-dispersion tolerant characteristics of twin-SSB are kept, even when our proposed detection scheme is employed, taking advantage of the narrower bandwidth of the LSB and USB components.

5 Conclusion

We proposed and investigated a novel twin-SSB detection scheme. The LSB and USB are demultiplexed by an electric butterfly operation without using optical filters. In the investigation, we used a digital coherent receiver. The strong chromatic-dispersion tolerant characteristics of the twin-SSB can potentially decrease the computational complexity of the equalization by DSP in the receiver. However, it should be noted that SSB signals can be received not only by the digital coherent receiver but by a direct detection (DD) using Kramers-Kronig (KK) scheme which is attracting great attention recently [13]. The twin-SSB with DD using KK method will be the next important step of the proposed detection scheme.
Study on channel prediction for automated guided vehicle using a probabilistic neural network

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Abstract: This paper describes a technique for predicting the received signal strength from a transmitter mounted on an automated guided vehicle (AGV). The predictor is part of a factory-based wireless communications anomaly detection system. We have proposed training a probabilistic neural network (PNN) to learn sliding-window sequences of the received fading signal. After training, the PNN conducts pattern matching between trained and observed sequences and returns an index to the closest matching set. Future deep-fading is predicted by adding an index offset to a fading look-up-table memory. We further propose to improve the performance by combining the predicted indices from multiple receivers. Performance results from ray-traced channels show that the mean squared error is reduced by almost two orders of magnitude by employing four receivers at 10 dB SNR.

Keywords: fading-prediction, machine-learning, probabilistic neural network, pattern-matching, ray-tracing

Classification: Wireless Communication Technologies

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1 Introduction

Factory and medical systems require highly reliable, repeatable and predictable communication links for control of time-critical operations. As wireless communications are replacing fixed-wire links in the factory, potential for signal outage and dropped packets must be considered. Improved prediction accuracy enlarges a warning-time window and permits evasive action to be taken before failure [1]. MAC-layer scheduling and physical-layer power-control can also benefit from improved prediction accuracy. Automated guided vehicles are becoming prevalent in the factory environment, and this paper focuses on prediction of their communications channel [2].

Channel state information (CSI) is usually estimated in a wireless system by comparing the received and known transmitted symbols embedded in the signal preamble field. Polynomial curve fitting can be used to extrapolate the CSI at future points in time. The capacity-loss of a polynomial estimator in a 40 Hz Doppler channel was shown to be less than 1% using a small number of past samples in [3]. A received power predictor based on filtering past CSI samples was used to predict the best transmit antenna for future transmission slots.

It is well-known that neural networks can extract signal features and predict patterns and here we apply the technology to CSI estimation. There are some
studies relating to channel prediction using neural networks in the research literature. An adaptive neural network-based channel predictor was able to reduce the mean deviation between a measured and predicted signal to within 8 dB in [4]. Improved signal field-strength prediction was demonstrated using a carefully trained neural network in [5]. We recently proposed a procedure for predicting the fading channel using a probabilistic neural network in [6]. A window of measured received signal strength indicator (RSSI) samples contains a unique pattern which is used to identify the current transmitter (Tx) position. Improved performance could be obtained by employing multiple spatially separated receivers.

This letter develops our work on fading prediction with new results under a variety of channel and PNN settings. We propose a methodology that includes discarding an estimate if a PNN returns a likely erroneous estimate in low-SNR. An ultimate aim of the work is to characterize the receiver performance in efforts to develop a reliable real-time fading predictor forming part of an anomaly detection system. Our technique can also be applied to predicting the channel on a moving head such as found in ICT equipment e.g. printer, engine compartment or wireless harness [7].

2 Probabilistic neural network (PNN)

The PNN features a feed-forward architecture which is robust to noise and outliers and is becoming increasing popular for solving classification problems with low-complexity. Mobile terminal localization prediction through learning the signature sequences of received signal strength was investigated in [8] and a PNN-based busy/idle status predictor for WLAN was described in [9]. The PNN algorithm is based on Bayes theory and applies the principle of Parzen windows to estimate a probability density function (PDF) by summation of probability contributions from multiple observations [10, 11]. The probability for each classification, \( c \) is expressed by

\[
P_c(x) = \frac{1}{\sigma_c \sqrt{2\pi}} \sum_{j=1}^{N_{c}} e^{-\frac{\|v - v_{c,j}\|^2}{2\sigma_c^2}},
\]

where \( v \) is the observed vector, \( v_{c,j} \) is the \( j \)-th training vector, \( N_{c} \) is number of training vectors for class-\( c \) and \( \sigma_c \) represents Gaussian spread. \( \|v - v_{c,j}\|^2 \) represents the squared Euclidean distance between the input and \( j \)-th training vector of class, \( c \).

3 Proposed methods

3.1 Single-receiver based prediction

The proposed technique involves a novel pre-processing of the serial fading data into a matrix form that can be suitably processed by the PNN. Let a sliding-window of samples be represented by \( s_{sw} \{v\} = s(n - N_C + 1, \ldots, n) \) where \( N \) is the total number of samples along the route and \( n \) is the first sample. Each window forms a row-\( v \) entry in an \( Nr \) (rows) \( \times \) \( N_C \) (columns) predictor matrix (P-matrix) as shown in Fig. 1(a). Each row of a corresponding target matrix (T-matrix) contains the index that points to the start address of the fading sliding-window stored in
memory. During training the PNN learns to associate each window with its start address. After training and during normal operation, the PNN returns the address index $\hat{T}$ of the closest matching row that has the minimum Euclidean distance with the current observation vector.

$$\hat{T} = \text{mode}(\mathbf{T}) \quad \forall n$$

where $\text{mode}$ is the statistical function that returns the most often occurring sample in a data set. In low-SNR, some PNN may make unreliable estimates and improved accuracy can be achieved by selecting from receivers that have sufficient SNR. This situation can be detected when the PNN outputs a constant index despite the channel changing. An erroneous estimate can also be determined by an extrapolation check using a window of previous samples or by estimating the SNR over the window period and this will be investigated in our future work. Finally, an optional offset $k$ is added to the index to estimate the fading at a future point in time.

### 3.2 Multi-receiver based prediction

Each receiver will sample a unique waveform due to its spatial position and make an independent estimate on the current window index. The estimated indices from all $N_Rx$ Rx are then collated as $\mathbf{T} = \{ \hat{T}_1 \ldots \hat{T}_{N_Rx} \}$. In the high signal to noise ratio (SNR) region, all PNN estimate the same index, i.e. $\hat{T}_1 \approx \hat{T}_r \; (r = 2 \ldots N_{RX})$. However, the indices may differ in the presence of deep fading and high noise. Fig. 1(b) shows an architecture for combining the estimated indices of all receivers. In the majority-combiner, the most likely index $\hat{T}_M$ at sample $n$ is given by

$$\hat{T}_M(n) = \text{mode}(\mathbf{T}_r(n)) \quad \forall n$$

![Fig. 1.](image-url)
4 Performance evaluation

We evaluate the performance through a Monte-Carlo software simulation using fading waveforms generated from a deterministic ray-tracing simulation.

4.1 Ray-tracing channel simulation

Channel fading waveforms are generated from a ray-tracing model at 2.4 GHz center frequency with 2 MHz signal bandwidth. The factory comprises two rooms with the inner area of dimensions 40 m by 40 m factory and outer surrounding room of width 10 m. The number of random uniformly distributed scatters in each area was set at either \( \left\{ 4, 8 \right\} \). A multi-hop scenario is created in which the signal from one inner-room scatterer is bounced by one outer-room scatterer before impinging on the static receiver as shown in Fig. 2(a).

The transmitter starts at position \((20, 0)\) m and moves along a pre-determined route at \(v = 2.5\) m/s. There is a fixed receiver in each corner of the room at positions \{(+20, −20), (+20, +20), (−20, +20), (−20, −20)\} m. Further receiver are placed at \{(±10, ±10), (±5, ±5), (±2.5, ±2.5)\} m. The total distance between transmitter, scatterer-pairs, and receiver is computed and the corresponding signal phases from the \{4, 8\} arriving paths are summed and sampled every 12.5 ms. An example of the generated fading signal and concept of sliding-window for pattern matching is shown in Fig. 2(b).

![Fig. 2.](a) Plan-view showing path of moving Tx along Route 1, four fixed Rx’s, eight scatterers and example path of ray arriving at Rx1; (b) Concept of sliding window across generated fading signal.

4.2 Training and prediction

The fading signal sampled at each receiver is pre-processed into the predictor P-matrix as it traverses the set route and the PNN is trained. The number of columns in the P-matrix was set at \(N_c = \{100, 200\}\). In performance evaluation mode, the vehicle moves along the set-route with no position error and therefore a receiver samples the same fading signal as during the training, but this time has added additive white Gaussian noise (AWGN). In the single-Rx system, each PNN computes the maximum likely window match and then predicts future fading at each Rx by reading the contents of its fading look-up table (LUT) at a required index offset. The MSE between the estimated and actual signal level is computed.
and averaged over all positions. In the multiple-Rx system, an optimized position index is selected by the majority combiner. The performance was evaluated over four different routes and over 100 simulations per route to average over the noise.

### 4.3 Performance results

The results in Fig. 3(a) show that the average MSE for a single antenna converges towards zero as the SNR increases showing that the PNN can predict the fading index accurately. The MSE achieved using four receivers was reduced by almost two orders of magnitude compared to that of the single receiver at 10 dB SNR.

![Fig. 3.](image)

(a) MSE for $N_s = \{8, 8\}, \sigma = 0.1$ (left) $N_c = 100$, (right) $N_c = 200$;
(b) CDF of No. of Rx with correct detection versus SNR for $N_s = \{8, 8\}, \sigma = 0.1$ (left) $N_c = 100$, (right) $N_c = 200$;
(c) MSE for $N_c = 100, \sigma = 0.10, (left) N_s = \{4, 4\}$ (right) $N_s = \{8, 8\}$;
(d) MSE for $N_c = 100, N_s = 8, 8, (left) \sigma = 0.1$ (right) $\sigma = 0.15$. 

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316
Comparing Fig. 3(a) (left) number of columns, $N_c = 100$ and (right) $N_c = 200$ shows that as $N_c$ increases the required SNR to achieve a given MSE also increases. There were a large number of deep fades which increases the uniqueness of each window signature and so the PNN is able to classify each window accurately with a relatively small number of samples. As the sequence length further increases it is conjectured that over-fitting occurs and performance decreases. The computational complexity increases with $N_c$ and so also selecting a lower size is beneficial.

The cumulative density function (CDF) of number of correct estimates of PNN position versus received SNR is shown in Fig. 3(b) for the above case of (left) $N_c = 100$ and (right) $N_c = 200$. In the low-SNR region few receivers report the correct index. As the SNR increases, the success rate rises until most receivers report the correct position in the high-SNR region.

Fig. 3(c) compares case of number of scatterers (left) $N_s = 4$ and (right) $N_s = 8$. In a Rayleigh fading channel, the channel becomes more deterministic and fading becomes flatter as the number of arriving rays increases. The fading becomes less deep and samples where deep fading did occur are less affected by noise which contributes to an improved matching performance.

Comparing Fig. 3(d) (left) $\sigma = 0.1$ and (right) $\sigma = 0.15$ shows that as the PNN spread $\sigma$ increases, the required SNR is reduced. As $\sigma$ increases, the PNN makes more substantial connections between neurons and the performance can improve. A balance is required as if $\sigma$ is too small then a large number of neurons are required to approximate a smoothing function and if too large more neurons are required when the channel or function is changing [12]. In future work we will investigate a large number of diverse channels to find optimized value of $\sigma$ under each condition.

5 Conclusion

We have demonstrated a novel neural network-based algorithm to classify and predict the fading profile experienced by a moving vehicle as it follows a predefined course. It was shown that the baseline performance could be improved by combining estimates from individual receivers and, by employing four receivers, the MSE was reduced by almost two orders of magnitude at 10 dB SNR. Performance results from the ray-tracing channel showed that a PNN matrix of $N_c = 100$ columns, $\sigma = 0.15$ performed best.

A study on an optimum value for $N_c$ and $\sigma$ as a function of the fading environment will be undertaken as part of our future work. In practice, sampling jitter, scatterer movement, fixed-point arithmetic and vehicle position errors will lead to prediction errors and are being investigated. The proposed technique is not restricted to predicting the fading envelope but can equally be applied to other waveforms such as the time-varying channel utilization ratio.

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Optimally designing virtualized CDN

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Abstract: Using virtualized CDN such as Amazon CloudFront in which cache servers are provided on virtual machines, content providers can flexibly select the regions of using cache servers according to the geographical demand pattern of their users, so they can expect to reduce the cost of using CDN. However, to maximize the profit of content providers, they need to carefully select the locations of cache servers which affect the quality perceived by users as well as the total cost. We propose a method of optimally selecting the geographical regions to use cache servers maximizing the profit of content providers.

Keywords: virtualized CDN, optimum design, profit

Classification: Network

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1 Introduction

Content provider (CP), e.g., YouTube, has its special requirement for the design of Content delivery network (CDN) depending on the geographical demand pattern...
against the content items. However, CDN providers do not distinguish CPs when inserting and removing content items in cache servers, i.e., applying the identical cache-insertion and cache-replacement policy to content items of all CPs, and CPs cannot limit the locations where content items are cached to reduce the cost paid to the CDN providers. Therefore, the virtualized CDN or cloud-based CDN in which cache servers are provided by VMs on public cloud have gathered wide attention and investigated [1], and commercial virtualized CDN services, e.g., Amazon CloudFront, have been already provided by major cloud providers. In the virtualized CDN, CPs directly contract with the cloud providers, and each CP can flexibly construct its own CDN by itself. In other words, each CP becomes a CDN provider serving just its content items.

Before starting to use the virtualized CDN, CPs will face the problem of optimally design its CDN, i.e., selecting the regions of using cache servers. CPs can expect to minimize the cost of using CDN by limiting the regions for using cache servers. However, the user-perceived quality, e.g., delay and smoothness in playback of video, will be degraded for users in these regions without cache servers. The number of views of users with degraded service quality will decrease, and the revenue of CPs will also decrease. Therefore, to maximize the profit of CPs, CPs should carefully select the regions to use cache servers considering both the fee paid to the cloud provider and the user-perceived quality. In this letter, we propose a method of optimally designing the virtual CDN maximizing the profit of CPs considering both the user-perceived quality and the cost of using cache servers when geographical demand distribution is given1.

## 2 Assumptions

### 2.1 Virtualized CDN service

We assume one virtualized CDN service which provides cache servers at $N$ regions, and we consider just a single CP. Let $R_n$ denote region $n$, and we define $C_n$ as the cache servers provided in $R_n$. Let $x_n$ denote a binary variable which takes unity if CP decided to use $C_n$ or takes zero if CP decided not to use $C_n$. Moreover, we define $X$ as the set of $R_n$ with $x_n = 1$. In other words, $X$ is the set of regions where CP decided to use cache servers. The origin servers are provided in just a single region $R_o$, and CP does not use $C_o$, i.e., $x_o = 0$, because content can be delivered to users of $R_o$ from origin servers. For requests from users of $R_n$, content is delivered from $C_n$ if requested content exists in $C_n$, i.e., cache hit, and content is delivered from $R_o$ if requested content does not exist in $C_n$, i.e., cache miss. When users of $R_n$ with $x_n = 0$ request content, content is delivered from the nearest cache servers of $X$ or $R_o$ giving the minimum average RTT to the requesting users.

### 2.2 User demand

We assume that each user of CP views content $W$ times on average within a month with totally satisfying the quality of video streaming service. It is anticipated that the number of views of users will decrease as the quality of video streaming is

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1A previous version of this manuscript was presented at [2]. This letter extended [2] by reflecting the cache hit ratio in the optimization problem.
degraded. Therefore, we assume that \( w(d) \), the average view count of each user when the RTT is \( d \), is given by

\[
w(d) = We^{-\alpha d},
\]

where \( \alpha \) is a sensitivity parameter of users against delay [3]. As \( \alpha \) increases, \( w(d) \) decreases more sharply with increase of \( d \).

### 2.3 Cost of delivering content from cache servers

The unit price of delivering content from cache servers to users depends on both the region of cache servers and the total amount of data transmitted from the cache servers within one month. We define \( p_n(v) \) as the unit price when delivering content from \( C_n \) to users when the total amount of data transmitted from \( C_n \) is \( v \), and the total cost of delivering content from cache servers is given by \( \sum_{n \in X} p_n(v_n)v_n \).

### 2.4 Cost of delivering content from origin servers

Many ISPs charge transit fee to CPs based on the total amount of data transmitted on transit links within one month. According to the analysis of transit fee of 20 ISPs in USA in 2004, monthly transit fee \( T(v) \) can be approximated by \( T(v) = \epsilon v^{0.75} \) where \( v \) is the monthly traffic volume transmitted on transit links [4]. Many ISPs use the 95 percent value of data transmission rate in each five-minutes bin as \( v \), and we assume that \( v \) is three times of average data transmission rate [5]. We set \( \epsilon = 0.165 \) USD which is the average transit fee for 1 Mbps in 2018.

### 3 Optimum design method of virtualized CDN

Let \( z_u \) denote the region from which content items are delivered to users of \( R_u \), and \( z_u \) is given by

\[
z_u = \arg \min_{z \in X \cup R_u} d_{z,u},
\]

where \( d_{i,j} \) is the average RTT between \( R_i \) and \( R_j \). Let \( M \) denote the number of content items, \( B \) denote the average size of content, \( q_n \) denote the number of subscribers of CP in \( R_n \), \( E_n \) denote the storage capacity of \( C_n \), and \( r_m \) denote the request ratio for content \( m \). From (1), \( v_i \), the average amount of data transmitted from \( C_i \) of \( x_i = 1 \) to users of CP within one month, is obtained by

\[
v_i = \sum_{n \in X \setminus \{u, z_u = i\}} \sum_{m=1}^{M} q_n r_m W e^{-\alpha d_{i,u}} B h_i(m),
\]

where \( h_i(m) \) is the cache hit probability of content \( m \) at \( C_i \). Using Che’s equation [6], \( h_i(m) \) is obtained by

\[
h_i(m) \cong 1 - e^{-r_m t_c}
\]

where \( t_c \) is the solution of \( t \) in equation \( \sum_{m=1}^{M} (1 - e^{-r_m t}) = E_n \). \( v_o \), the amount of data transmitted from origin servers within one month, is...
\[ v_i = \sum_{u \in X \cap \{u; z_u = 0\}} q_u r_m W e^{-\alpha d_i} B + \sum_{u \in X \cap \{u; z_u \neq 0\}} q_u r_m W e^{-\alpha d_i} B(1 - h_z(m)). \] (5)

We assume that CP obtains \( r \) fee for each view, and \( R(X) \), the revenue of CP, is given by

\[ R(X) = \sum_{u=1}^{N} q_u r_m W \sum_{m=1}^{M} \{ e^{-\alpha d_{i,j}} r_m h_z(m) + e^{-\alpha d_{i,j}} r_m (1 - h_z(m)) \}. \] (6)

\( P(X) \), the average profit of CP within one month, is obtained by

\[ P(X) = R(X) - \sum_{i \in X} p_i(v_i) v_i - T(v_o). \] (7)

We define the optimization problem maximizing \( P(X) \):

\[ \text{max} \quad P(X) \] (8)
\[ \text{s.t.} \quad x_n = \{0, 1\}, \quad 1 \leq n \leq N, \] (9)
\[ x_o = 0. \] (10)

The computational complexity to solve this optimization problem is \( O(N^2) \), and CP can obtain the optimum \( X \) in quite short time.

4 Numerical evaluation

4.1 Evaluation conditions

As the virtualized CDN service, we assume Amazon CloudFront which provides cache servers at eight regions: North America, Europe, South Africa, Japan, Australia, Singapore, India, and South America. To obtain the setting values of \( d_{i,j} \), we measured the average RTT using ping command on the virtual machines in these regions of Amazon EC2. However, we cannot setup virtual machines in Africa on Amazon EC2, we selected the seven regions excluding Africa. We assume that delay was zero when delivering servers and users existed in the same region. We set \( p_n(v) \) according to the price list of Amazon CloudFront [7]. The ratio of users in Netflix was about 0.03, so we set \( q_n \) to the population in each region multiplied by 0.03. We set \( B \) to 3 GB, \( W \) to 15, and \( r \) to 0.6 USD which was obtained by monthly flat-fee of Netflix, 10 USD, divided by \( W \). Moreover, we set the content count \( M \) to 1,000, and we set the identical size \( E \) to \( E_n \), the storage size of \( C_n \).

4.2 Monthly profit of CP

We compare \( P \), the average monthly profit of the CP, in the proposed design method as well as the two cases: No Cache and All Cache. No Cache is the case without using any cache servers, and All Cache is the case using cache servers in all the regions excluding \( R_o \). Fig. 1 plots \( P \) in each of the three methods against the cache size \( E \) for two values of \( \alpha \) when placing the origin servers in North America. The following tendencies were also observed when placing the origin servers in other areas as well. In the proposed method, as the result of optimally selecting the
locations to use cache servers, $P$ was the largest among the three methods in the entire range of $E$ and $\alpha$. For example, when $\alpha = 3$ and $E = 1,000$, the proposed method increased $P$ about 6% and about 13% compared with All Cache and No Cache, respectively.

![Graph](image)

**Fig. 1.** Average monthly profit of CP against cache size when placing origin servers in North America

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Overlaying a motion video with data information for digital signage and image-sensor-based visible light communication systems

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Abstract: A digital signage image-sensor-based visible light communication (IS-VLC) system can transmit data information by embedding it on visual information. The use of a motion video as visual information makes it difficult to completely erase the visual information in this system. The extraction of data information is an important issue because the incomplete extraction will degrade the bit error rate performance. In this letter, we expand the IS-VLC system to overlay a motion video with data information by introducing a cell-position correction method and a noise elimination method. Experimental results clarify the effectiveness of the extended system.

Keywords: digital signage, visible light communication, image sensor, motion video

Classification: Wireless Communication Technologies

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1 Introduction

Digital signage is a display that is used to display targeted information, entertainment, advertisements, and announcements. Digital signage is placed in stations, shopping malls, and other such places. The dissemination of accurate emergency information during disasters and a guide for foreign passengers [3, 4] are advanced usage of digital signage. These applications require not only a display of visual information but also transmissions of data information from the digital signage to user equipment such as smartphones.

We focus on an image-sensor-based visible light communication (IS-VLC) system [5, 6] for the transmission of data information. In this letter, the IS-VLC system employs a display of a digital signage as a transmitter and an image sensor
as a receiver. The transmitter overlays visual information with data information, and the receiver derives data information from the frames captured by the image sensor.

In the IS-VLC system, it is important to prevent quality degradation of visual information that is overlaid with data information. There are schemes to solve this issue [7, 8, 9], but these schemes have problems such as a low data rate or requirement of a special display. To resolve these problems, we proposed color differential modulation for the IS-VLC system and clarified the capability of the transmission of data information while avoiding the quality degradation of the visual information [10]. However, still images are used as visual information. When a motion video is used as visual information, it is difficult to completely erase visual information. The extraction of data information is an important issue because the incomplete extraction will degrade the bit error rate (BER) performance.

Herein, we expand the IS-VLC system to overlay a motion video with data information, which introduces a cell-position correction method and a noise elimination method. We evaluate the BER performance of the expanded system and clarify the effectiveness of these methods.

2 IS-VLC system

Fig. 1 shows the model of the expanded IS-VLC system. In this section, we explain the conventional IS-VLC system proposed in [10]. The cell-position correction method and the noise elimination method is described in Sect. 3.

At the transmitter, bit sequence of data information is mapped onto an $M \times N$ data matrix, which is differentially encoded to obtain a coded matrix. A mono-component data frame is generated from the coded matrix, where the size of the data frame is $X \times Y$ pixels. The data frame consists of $M \times N$ rectangular cells, and the size of the cell is $X/N \times Y/M$ pixels. Each cell corresponds to a bit of the coded matrix. The pixel value of the cell is 0 and $\alpha$ when the corresponding bit is 0 and 1,
respectively, where $\alpha$ indicates a signal intensity. The color-difference modulator overlays blue-color-difference (Cb) components of visual information with the data frame because the human eyes are insensitive to color-difference components, especially the Cb component. After adding a marker, which is used at the receiver, the overlaid frame is shown on a display.

At the receiver, an image sensor captures the frame on the display. Using the marker, the region of the data frame is extracted from the captured frame, and time-synchronization is achieved. Through a color filter, a Cb-component frame is derived, and the data frame is retrieved by subtracting the successive two frames. This subtraction can decode the data differentially encoded by the transmitter. The pixel values in each cell are averaged, and the data matrix is retrieved via a threshold decision. Finally, the bit sequence of the data information is derived by demapping the retrieved data matrix.

When a still image is used as visual information, the subtraction eliminates visual information from the received frame. However, when a motion video is used, visual information cannot be eliminated completely, and the residual values of visual information will degrade the BER performance.

3 Expansion for motion video

In this section, we introduce a cell-position correction method and a noise elimination method for a motion video as visual information.

3.1 Cell-position correction

When the motion video gradually changes, the movement between successive frames will be small. We introduce a cell-position correction method based on template matching, which will be able to reduce the residual values after the subtraction by adjusting the position of the successive frames.

The cell-position correction is achieved at each cell. Let $C_{m,n}(i)$ show the cell of the $m$th row and the $n$th column in the $i$th frame. First, the cell $C_{m,n}(i - 1)$ is used as a template. For the $i$th frame, this method searches and finds the position at which the similarity between the cell $C_{m,n}(i)$ and the template is the largest. In this process, only luminance (Y) components of both cells are used because Y components do not include data information. Herein, we employ a normalized cross-correlation as similarity [11]. Following the color filter for Cb components, only the duplicated regions between the successive cells $C_{m,n}(i - 1)$ and $C_{m,n}(i)$ are subtracted after adjusting their positions so that the similarity takes the largest value.

3.2 Noise elimination

The change between the successive frames of the motion video causes residual values of visual information in the subtracted frame. These residual values are treated as noise when data information is demodulated. Particularly, an edge of an object in a frame remains intense. Conversely, visual information is overlaid with data information of weak signal intensity $\alpha$ to avoid degradation of visual information, i.e., noise has large values, while data information has small values.
In the noise elimination method, the absolute value of each pixel within a cell of the subtracted frame is compared with a pre-defined threshold $\eta$. If the value is larger than the threshold $\eta$, this pixel is regarded to contain intense noise and is omitted for averaging the pixel values within the cell.

4 Experimental results

We experimentally evaluate BER performance of the expanded IS-VLC system. Table 1 lists the parameters of the components used in the experiment. We use a SHARP PN-Y475 as a display and a Raspberry Pi 3 Model B with Raspberry Pi Camera Module v2.1 as an image sensor. The experiment was carried out in a room at night, where fluorescent light was on. The distance between a display (Tx) and an image sensor (Rx) is 1.5 m. We use an 8-bit pixel value, that is, it can take from 0 to 255. A motion video named “sunflower” is used as visual information. In this video, a bee is flying over a sunflower, and the camera follows the bee. The center region of the frame abruptly changes because of the flying bee, while the other region gradually changes. The absolute category rating (ACR) [12] is used to evaluate the quality of visual information. After watching the visual information, evaluators of 13 males and 12 females scored it from 1 to 5. The quality is evaluated in terms of a mean opinion score (MOS).

Fig. 2 shows BER performance of the expanded and the conventional IS-VLC systems. The threshold $\eta$ is adjusted to reduce the BER. In Fig. 2(a), we can find that BER becomes low as the signal intensity becomes large. The expanded system can achieve better BER performance than the conventional system. In [13], a turbo code with a code rate of one-third can correct errors till the BER reaches $10^{-1}$. The performance of the extended system satisfy the conditions of both a target BER (below $10^{-1}$) and a good MOS (over 4). For $\alpha = 5$, BER of the cell-position correction method becomes worse. This method reduces the region for averaging the pixel values for the large movement of the cell position. Therefore, it is preferable that the region of averaging the pixel values is enlarged when the signal intensity is large. Fig. 2(b) shows BER of each cell for $\alpha = 2$. The cell-position correction method can reduce BER in the peripheral region of the center. Hence, this method is effective for gradual change between successive frames. The noise

| Table 1. Parameters used in the experiment. |
|---------------------------------------------|
| Frame rate of data information | 10 fps |
| Frame rate of visual information | 20 fps |
| Capture rate of an image sensor | 20 fps |
| Size of an data frame $(X, Y)$ | 1,900 pixels × 1,000 pixels |
| Cell construction $(M, N)$ | (19, 10) |
| Resolution of a display | 1,920 pixels × 1,080 pixels |
| Resolution of an image sensor | 1,920 pixels × 1,080 pixels |
| Distance between a display and an image sensor | 1.5 m |
| Signal intensity $\alpha$ | 1, 2, 3, 4, 5 |
| Visual information | Sunflower |

In the noise elimination method, the absolute value of each pixel within a cell of the subtracted frame is compared with a pre-defined threshold $\eta$. If the value is larger than the threshold $\eta$, this pixel is regarded to contain intense noise and is omitted for averaging the pixel values within the cell.
elimination method can improve BER in the whole region. This method is more robust for the movement between the successive frames.

5 Conclusions

In this letter, we expanded the IS-VLC system to support a motion video as visual information by introducing the cell-position correction and the noise elimination methods. From the experimental evaluation, we have clarified that these methods can improve BER performance for the motion video.

We evaluated the BER performance using only a motion video as visual information. However, it is not enough to evaluate the performance of the proposed system; therefore, in the future, we intend to use motion videos with various characteristics.

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Low complexity DBP with phase linear approximation for DP-16QAM coherent optical communication systems

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Abstract: We numerically assess the performance and computational complexity of split-step Fourier method based digital backpropagation (DBP) with phase linear approximation (PLA-DBP) for reduced complexity on digital implementation. From Manakov equation, we can confirm that the phase of the received signal compensated with DBP changes almost linearly between optical amplifiers, thus the input signal phase even with dual-polarization can be estimated using linear approximation using only two points. We show the performance evaluation of 28 Gbaud single carrier dual polarization-16 quadrature amplitude modulation (DP-16QAM) format, and 20.0% reduction in complexity is obtained with transmission distance of 4000 km and input power of 3 dBm.

Keywords: digital backpropagation, digital signal processing, coherent optical communication, nonlinear effect, low computational complexity

Classification: Fiber-Optic Transmission for Communications

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Introduction

The spectral efficiency is severely limited by the impact of the nonlinear distortion due to the optical Kerr effect in optical communication systems [1]. To mitigate the impact of nonlinearities, several nonlinearity compensation methods such as digital backpropagation (DBP) based on split step Fourier method (SSFM) [2] and Volterra series nonlinear equalizers (VSNE) [3] have been developed. However, due to the highly iterative algorithm, DBP have been regarded as complex for real time processing. Therefore, it is one of the significant technical issues in optical communication systems and low complexity DBP algorithms that provide sufficient performance like conventional one is required. To overcome this problem, several advanced DBP algorithms have been proposed and reduction of the computational complexity has been investigated recently [4, 5, 6]. Despite these great efforts, there would be room for further improvement in the DBP algorithm for being implemented on a digital signal processor with low power consumption, and we proposed novel efficient nonlinear compensation method for low computational complexity, phase linear approximation-DBP (PLA-DBP) [7, 8]. In the calculation for single-polarized quadrature phase shift keying (QPSK) and 16-quadrature amplitude modulation (16QAM) signals by using SSFM based DBP, almost constant signal phase change has been confirmed between the optical amplifiers placed at equal transmission distance. From this property, PLA-DBP is a promising technique to estimate the signal phase at the transmission point with linear approximation using the signal phase at two middle points, thus, the calculation process for the remained transmission link can be omitted and the computational complexity can be reduced. However, its applicability to dual-polarized signals has not been investigated so far.

1 Introduction

The spectral efficiency is severely limited by the impact of the nonlinear distortion due to the optical Kerr effect in optical communication systems [1]. To mitigate the impact of nonlinearities, several nonlinearity compensation methods such as digital backpropagation (DBP) based on split step Fourier method (SSFM) [2] and Volterra series nonlinear equalizers (VSNE) [3] have been developed. However, due to the highly iterative algorithm, DBP have been regarded as complex for real time processing. Therefore, it is one of the significant technical issues in optical communication systems and low complexity DBP algorithms that provide sufficient performance like conventional one is required. To overcome this problem, several advanced DBP algorithms have been proposed and reduction of the computational complexity has been investigated recently [4, 5, 6]. Despite these great efforts, there would be room for further improvement in the DBP algorithm for being implemented on a digital signal processor with low power consumption, and we proposed novel efficient nonlinear compensation method for low computational complexity, phase linear approximation-DBP (PLA-DBP) [7, 8]. In the calculation for single-polarized quadrature phase shift keying (QPSK) and 16-quadrature amplitude modulation (16QAM) signals by using SSFM based DBP, almost constant signal phase change has been confirmed between the optical amplifiers placed at equal transmission distance. From this property, PLA-DBP is a promising technique to estimate the signal phase at the transmission point with linear approximation using the signal phase at two middle points, thus, the calculation process for the remained transmission link can be omitted and the computational complexity can be reduced. However, its applicability to dual-polarized signals has not been investigated so far.
In this paper, we analytically assess the performance and reduction effect of computational complexity of PLA-DBP for 28 Gbaud single carrier dual polarization-16QAM (DP-16QAM) transmission system compared with conventional SSFM based DBP (C-DBP).

2 Principle of phase linear approximation (PLA) for DP-signal

C-DBP is based on the concept that original signals are evaluated from the received signals propagating in the opposite direction against the transmission one through virtual fiber. In polarization division multiplexed (PDM) optical systems, the evolution of optical field envelopes is modeled by the Manakov equation given in Eq. (1), by averaging the vector nonlinear Schrödinger equation (NLSE) over the Poincaré sphere considering that the random birefringence variation in optical fiber occurs much faster than the nonlinear interactions [9].

\[
\frac{\partial E_{x/y}}{\partial z} = -\alpha \frac{E_{x/y}}{2} - i\beta \frac{\partial^2 E_{x/y}}{\partial t^2} + i \frac{8}{9} \gamma (|E_x|^2 + |E_y|^2)E_{x/y} \tag{1}
\]

where \(E_{x/y}\) represents the optical field envelope for two orthogonal \(x\) and \(y\) polarizations, \(\alpha\), \(\beta\), and \(\gamma\) are the attenuation coefficient, group velocity dispersion and nonlinear coefficients of the fiber, respectively. \(z\) and \(t\) are the variables of propagation direction and time. SSFM is useful to solve Eq. (1). In this technique, the fiber link is divided into small steps and the linear and nonlinear distortion are taken into consideration in frequency and time domain in each step, respectively. In this letter, C-DBP analysis is calculated with SSFM.

Let us consider the attenuation term in the optical field envelope in Eq. (1), and set the relation of Eq. (2),

\[
E_{x/y} = e^{-\frac{z}{2}}E_{x/y}'.
\tag{2}
\]

Then the following equation can be obtained,

\[
\frac{\partial E_{x/y}'}{\partial z} = -i \beta \frac{\partial^2 E_{x/y}'}{\partial t^2} + i \frac{8}{9} \gamma e^{-\alpha z}(|E_x'|^2 + |E_y'|^2)E_{x/y}'. \tag{3}
\]

Additionally assuming full linear compensation the phase rotation amount \(\varphi_L\) of \(E_{x/y}'\) after distance \(L\) propagation is

\[
\varphi_L = \int_0^L \Delta \varphi \, dz = \int_0^L \frac{8}{9} \gamma e^{-\alpha z}(|E_x|^2 + |E_y|^2) \, dz
\]

\[
= \frac{8}{9} \gamma \frac{1 - e^{-\alpha L}}{\alpha} (|E_x|^2 + |E_y|^2). \tag{4}
\]

From Eq. (4), \(\varphi_L\) is a constant at a fixed \(L\). Hence, in this system, the amount of phase change per span (between optical amplifiers) is constant, and the input phase can be estimated by using linear approximation.

Next, we analytically show the signal phase transition by using C-DBP and linear approximation for 28 Gbaud DP-16QAM in Fig. 1(a). Parameters used in this simulation are as follows; loss of standard single mode fiber (SSMF) of 0.16 dB/km, dispersion parameter of 16 ps/nm/km, nonlinear coefficient of 1.3 W⁻¹km⁻¹, core radius of 5.0 µm, word length of 2¹⁹ – 1. Loss is fully compensated by EDFAs with noise figure of 4 dB installed every 100 km, trans-
mission distance of $L = 2000\text{ km}$, and launched power of $P = 0\text{ dBm}$. Linear approximation is applied with estimated phases at the first and No. $n$ span ($n$: natural number; $n \geq 2$). The phase transition as a function of the number of calculated span is shown in Fig. 1(a). Here, the phase transition of the symbol located in the middle having phase of $\pi/4$ among the 3 series consecutive ones with the smallest or largest intensity are affected with minimum or maximum intra-channel cross phase modulation (IXPM), and they are indicated as DBP-minimum or maximum IXPM, respectively. From Fig. 1(a), the phase transition is almost linear with each span regardless of the amount of nonlinear distortion, despite the fact that the change is nonlinear within each step. In addition, all three lines of No. 2, No. 10, and No. 15 are very close to the signal phase transition obtained by C-DBP, and the accuracy of linear approximation becomes better for larger $n$. On the other hand, No. 2 with a smaller span number is closer to the assigned phase of $\pi/4$ after linear approximation. Thus, it should be investigated to find conditions with fewer errors. As in the case of SP-16QAM [8], the signal intensity changes randomly, however, no large estimation error occurs from the intensity at the point of transmission, thus, the closest intensity is estimated as that at the launched point.

Fig. 1(b) shows the block diagram of PLA-DBP, and it is divided into three groups. Processing by each group are described as follows; Group 1 is a C-DBP processing based SSFM with fiber divided by small step $\Delta z$ each step. Group 2 is the process that obtains phases of signals each span and stores them in memories in the system. These process required for linear approximation is performed until the countered number of steps $N$ and spans $N'$ reaches the target steps $N_{\text{step}}$ and spans $N_{\text{span}}$, respectively. In the Group 3, linear approximation is performed to estimate the launched signal phases with two of stored phases. We assume that the $x$ and $y$ polarizations are separated and polarization mode-dispersion compensated under the optimum condition with a coherent receiver.

(a) Phase transit and linear approximation; minimum, maximum IXPM means the conditions that minimum or maximum IXPM affects the target symbol, in 16QAM.

(b) Block diagram of PLA-DBP; LC: Linear Compensation, NLC: Nonlinear Compensation, $L_{\text{res}}$: residual Linear Compensation.

Fig. 1. Principle of linear approximation.

3 Performance and complexity of PLA-DBP

In this section, we numerically assess the performance and complexity of PLA-DBP. Used parameters are fixed and the same as those used in Fig. 1(a) except propagation distance $L$ and launched power $P$. 
Fig. 2(a) indicates $Q^2$ value versus launched power of LC, C-DBP, and PLA, and the constellations after compensation are indicated in the insets. The $Q^2$ value is defined as $Q^2 = 20 \cdot \log_{10}(\sqrt{2} \cdot \text{erfcinv}(2 \cdot \text{BER}))$ where \text{erfcinv} represents the inverse complementary error function. Regarding PLA, $N_{\text{step}}$ and $N_{\text{span}}$ are varied within almost the same total computational effort ($N_{\text{step}} \times N_{\text{span}}$). Although the peak power can be increased by PLA ($P = -1 \text{ dBm}$), the performance is lower than that of the C-DBP, hence, it is necessary to study the conditions that can obtain sufficient performance. However, the power penalty of PLA-DBP compared with C-DBP is within 0.5 dB under the maximum $Q^2$ condition.

The performance of PLA-DBP is evaluated by error vector magnitude (EVM) as shown in Fig. 2(b) when $N_{\text{span}}$ is changed. Although EVM of PLA-DBP can be expected by increasing $N_{\text{step}}$ and $N_{\text{span}}$, that saturates around $N_{\text{step}} = 5$ and $N_{\text{span}} < 15$, and it approaches to performance limit of C-DBP around $N_{\text{step}} = 5$, thus, it can be confirmed that the transmission signal may be estimated from the midpoint of PLA-DBP with a small $N_{\text{span}}$ and the reduction of computational complexity could be expected, maintaining the EVM as much as C-DBP. Fig. 2(c) indicates the combination of $N_{\text{step}}$ and $N_{\text{span}}$ reaching the target FEC limit with BER of $2.7 \times 10^{-2}$ ($Q^2 = 5.7 \text{ dB}$). Here, launched power of $\geq 3 \text{ dBm}$ and transmission distance of 4000 km are selected so that BER of only linearly compensated signals are lower than the target value. $N_{\text{step}}$ and $N_{\text{span}}$ are in a trade-off relationship to achieve the target BER. Although $N_{\text{step}}$ can be reduced with increase in $N_{\text{span}}$, it can be seen that it gradually saturates. The marked symbols are the combination in which total number of DBP steps ($N_{\text{step}} \times N_{\text{span}}$) is the smallest under the same launched power.

(a) $Q^2$ value versus launched power; $L = 2000 \text{ km}$.

(b) $N_{\text{span}}$ versus EVM; $L = 2000 \text{ km}$ and $P = -2 \text{ dBm}$.

(c) $N_{\text{step}}$ versus $N_{\text{span}}$ reaching BER = $2.7 \times 10^{-2}$; $L = 4000 \text{ km}$.

Fig. 2. Performance of PLA-DBP.
In Fig. 3, difference of computational complexity between PLA-DBP and C-DBP is presented. Fig. 3(a) indicates LC reach with FEC limit of the target BER \( \frac{2.7 \times 10^{-2}}{10^{-2}} (Q^2 = 5.7 \text{ dB}) \). When the transmission distance \( L \) is 2000 km, the launched power \( P \) should be 4 dBm or less in order to satisfy the FEC limit with only LC. Similarly, \( P \leq 2 \text{ dBm} \) when \( L > 4000 \text{ km} \). In order to extend the optical power and the reach further, nonlinear compensation is required.

Fig. 3(b) shows the reduction rate of complexity by PLA-DBP compared with C-DBP reaching the target BER mentioned above, calculated by \( \text{Reduction rate of complexity} = 1 - \frac{N_{\text{step},C-DBP} \times N_{\text{span},C-DBP} \times N_{\text{span,C-DBP}}}{(N_{\text{step},C-DBP} \times N_{\text{span,C-DBP}})_{\text{full span}}} \), where \( N_{\text{step},C-DBP} \) and \( N_{\text{span,C-DBP}} = \frac{L}{\text{span length}} \) are the number of steps and spans required for C-DBP to achieve target BER. For example, when \( N_{\text{step},C-DBP} = 5 \) and \( N_{\text{span,C-DBP}} = \frac{2000}{100} = 20 \), BER reaches FEC limit, where \( L = 2000 \text{ km} \) and \( P = 5 \text{ dBm} \). Transmission distance \( L \) varies from 2000 km up to 5000 km, in steps of 1000 km, and launched power \( P \) varies from 3 dBm up to 6 dBm. The maximum reduction rate of complexity is obtained from the combination of \( N_{\text{step}} \) and \( N_{\text{span}} \) under the condition of \( Q^2 < 5.7 \text{ dB} \) with LC as indicated in Fig. 3(a). Under conditions where the nonlinearity is relatively small, PLA-DBP has a great effect, for example, PLA-DBP allows to reduce the computational complexity of 20.0\% with the same performance as DBP when \( L = 4000 \text{ km} \) and \( P = 3 \text{ dBm} \). Even when the nonlinearity is very high, the computational complexity reduction by PLA-DBP can be obtained, for example, reduction of 10.0\% when \( L = 4000 \text{ km} \) and \( P = 5 \text{ dBm} \).

### 4 Conclusion

We presented the performance and complexity of PLA-DBP in 28 Gbaud DP-16QAM in simulation. It is indicated that the phase of optical signal propagating in the fiber varies almost linearly from Eq. (1) between optical amplifiers even for dual-polarized signals. Since the computational complexity and the compensation performance are in a trade-off depending on \( N_{\text{step}} \) and \( N_{\text{span}} \), there are multiple combination of \( N_{\text{step}} \) and \( N_{\text{span}} \) reaching target BER and the optimum combination was investigated for regions of larger nonlinearity that cannot be reached with only LC. Under the condition of \( L = 4000 \text{ km} \) and \( P = 3 \text{ dBm} \), reduction of 20.0\% is obtained in the computational complexity with maintaining the performance.
Transmission power control of terrestrial pseudolite signal for global navigation satellite systems

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Abstract: In global navigation satellite systems (GNSS), we need signals from more than four satellites to find out the location at the receiver. However, the satellite visibility is occasionally blocked especially in city areas. For augmenting the availability, placing the pseudo satellite, or pseudolite, has been proposed. A pseudolite broadcasts the signal in the same format as the satellites broadcast and acts as an additional satellite to the receiver. However, the satellite signals are often blocked by a pseudolite signal at the receiver in the close proximity to the pseudolite, and the phenomenon is referred to as the near-far problem. In this letter, a pseudolite with burst pulse transmission that the power varies pulse-to-pulse is proposed to mitigate the near-far problem.

Keywords: global navigation satellite system, pseudo satellite, burst pulse, transmission power control, near-far problem

Classification: Navigation, Guidance and Control Systems

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1 Introduction

With a car navigation showing the current location, we would arrive at destination without losing our way. It finds out the location receiving the radiowaves from satellites of Global Satellite Navigation System (GNSS) such as Global Positioning System (GPS) of the United States, GLONASS of the Russian Federation, Galileo of the European Union, BeiDou of China, and Quasi Zenith Satellite System (QZSS) in Japan. Finding out the current location needs simultaneous reception of the signals from more than four GNSS satellites.

However, high-rise buildings and elevated roads especially in city areas often block the visibility of the satellite signals, and it results in failure of finding out the location of the receiver. For enabling the receiver in such areas to find out a location, placing a pseudolite (PL) that broadcasts the signal in the same format as the satellites broadcast has been proposed. The origin of PL may be in a navigation experiment using terrestrial transmitters conducted at the United States Army’s Yuma Providing Ground in 1973, the confirmation phase of satellite navigation of GPS [1].

Though various types of PL were proposed in 1980’s, setting the PL transmission power to cover the desired service area also interferes and blocks the satellite signals. The block of the satellite signals comes from the excessive power faced the receiver in the close proximity to the PL, and the phenomenon is referred to as the near-far problem. The Radio Technical Committee for Marine Service (RTCM) proposed the burst pulse transmission for PL in 1986 [2]. In the report, the duty ratio should be less than 10%, and RTCM recommended to set the duty ratio as 1/11. A receiver with the array antenna and the signal processing was also proposed to weaken unnecessary PL signals [3]. However, the shadowing facing the receiver in a terrestrial PL environment magnifies the effect of the near-far problem. Alleviating the near-far problem in PL is a subject to be solved in PL.

In this letter, a method of alleviating the near-far problem in PL operating in city areas is proposed.

2 Analysis of near-far problem deletion of duplicate words and burst transmission in pseudolite

2.1 Dynamic range of GNSS signal reception power

Because the near-far problem due to a PL is source of the excess reception power at the receiver, considering the dynamic range of the receiver is essential to know the availability. GNSS satellites are constellated in the azimuth and elevation angles at the receiver, and it results in the received power variation. The dynamic range for receiving the satellite signals can be used as the margin of the near-far problem in PL signal reception. The reception power variation due to the antenna directivity for different elevation angles reaches 8 dB [4, chap. 8]. Therefore, we assume that power difference of less than 8 dB among satellite signals and a PL signal acceptable for receiving both of the signals.
2.2 Near-far problem analysis in cellular communications

A radiowave transmitted at the antenna is weaken as an increase in the distance between the transmission and reception antennas. On the other hand, buildings, cars, and human bodies would shadow the radiowave path and the shadowing also attenuates the signal. The shadowing randomly attenuates the reception power whose statistic is the log-normal distribution \([5]\). Even in a shadowing environment, a receiver needs signal whose power is above the minimum reception power. The method of finding out the reception power under such environment is exemplified in reference \([6]\). The reception power in decibels at the distance \(d \,[\text{m}]\) can be express as

\[
P_r(d) = \frac{P_t}{C_0} - 10 \log_{10} K - 10 \gamma \log_{10} \left( \frac{d}{d_0} \right) - \Phi_{\text{db}} \,[\text{dBm}],
\]

where \(P_t \,[\text{dBm}]\) is the transmission power, \(K\) is the propagation loss coefficient, \(d_0 \,[\text{m}]\) is the reference distance. For free space propagation within the reference distance of \(d_0 = 1 \,[\text{m}]\) from the transmitter, we can use \(K = \left( \frac{\lambda}{4\pi} \right)^2 \) (\(\lambda\) is the wavelength). \(\gamma\) is the attenuation coefficient, and \(\Phi_{\text{db}}\) is a zero-mean Gaussian random variable with the variance of \(\sigma_{\text{db}}^2\) that characterize the shadowing. The outage probability is the probability the reception signal power \(P_r(d) \,[\text{dBm}]\) is below the minimum reception power of \(P_{\text{min}} \,[\text{dBm}]\),

\[
P_{\text{out}}(P_{\text{min}}, d) = \text{Prob}(P_r(d) < P_{\text{min}}) = 1 - Q\left( \frac{P_{\text{min}} - P_r(d)}{\sigma_{\text{db}}} \right),
\]

where

\[
P_r(d) = \frac{P_t}{C_0} - 10 \log_{10} K - 10 \gamma \log_{10} \left( \frac{d}{d_0} \right) \,[\text{dBm}]
\]

is the mean reception power at \(d\), and

\[
Q(z) = \int_{z}^{\infty} \frac{1}{\sqrt{2\pi}} \exp\left( - \frac{z^2}{2} \right) \, dz = \frac{1}{2} \text{erfc}\left( \frac{z}{\sqrt{2}} \right)
\]

is the \(Q\)-function representing the probability that a zero-mean Gaussian random variable with the variance of 1 exceeds \(z\) and is also expressed using the complementary error function, \(\text{erfc}(\cdot)\).

For the serving radius of \(R\) from the transmitter, the coverage probability that one can use the system within the region is derived as \([6]\)

\[
P_{\text{cov}}(P_{\text{min}}, R) = \frac{1}{\pi R^2} \int_{0}^{2\pi} \int_{0}^{R} \{ 1 - P_{\text{out}}(P_{\text{min}}, r) \} \cdot r \, dr \, d\theta
\]

\[
= \frac{2}{R^2} \int_{0}^{R} r \, Q\left( a + b \log \frac{r}{R} \right) \, dr
\]

\[
= Q(a) + \exp\left( \frac{2 - 2ab}{b^2} \right) Q\left( \frac{2 - ab}{b} \right),
\]

where

\[
a = \frac{P_{\text{min}} - P_r(R)}{\sigma_{\text{db}}}, \quad \text{and} \quad b = \frac{10 \gamma \log_{10} e}{\sigma_{\text{db}}}.
\]
2.3 Burst pulse transmission of pseudolite

A GPS L1C/A signal conveys navigation messages at a rate of 50 bit/s and the chip rate is 1.023 Mchip/s. The impact on a receiver in the close proximity to a PL would be low if the PL transmits only a small fraction of the 50-bit/s duty cycle. The PL signal blocks the satellite signals for the duration the PL transmits. If the duration of the 11 PL burst pulse is 90.91 µs out of a half duration of the navigation message bit (10 ms). Then, the duty ratio is 1/11, and each of the burst pulse has 93 chips. The all of the 1023 chips of the PL are transmitted in 10 ms. The burst pulse transmission alleviates the near-far problem but interferes the satellite signal reception. Identification, navigation message transfer, and ranging of the PL can be accomplished by the PL signal.

3 Proposed pseudolite whose transmission power varies

3.1 Coverage and outage probabilities in conventional pseudolite

For a PL, we assume $P_t$ so that $P_r(d)$ at the fringe reaches $P_{\text{min}}$. Then, $P_t$ can be calculated by $P_{\text{min}} - 10\log_{10}(K) + 10\gamma\log_{10}(R/d_0)$. A decrease in the distance between the PL and the receiver tends to interfere the reception of the satellite signals, and an increase in the distance weaken the PL signal. Therefore, the serving region of a PL is an elliptical shape as illustrated in Fig. 1(a). The region becomes complex shapes in shadowing environments. The PL signal is available at the receiver when $P_t(d)$ is between $P_{\text{min}}$ and $P_{\text{max}}$. 

![Fig. 1. Availability of pseudolite: (a) serving region and (b) the coverage and outage probabilities.](image)
The outage probability of PL can be derived by modifying Eq. (2) as

\[
P_{\text{out}}(P_{r_{\text{min}}}, P_{r_{\text{max}}}, d) = \text{Prob}(P_r(d) < P_{r_{\text{min}}} \mid P_r(d) < P_{r_{\text{max}}})
\]

\[
= 1 - Q\left(\frac{P_{r_{\text{min}}} - P_r(d)}{\sigma_{\text{dB}}}\right) + Q\left(\frac{P_{r_{\text{max}}} - P_r(d)}{\sigma_{\text{dB}}}\right).
\]

(7)

The coverage probability is also calculated using Eq. (5) as

\[
P_{\text{cov}}(P_{r_{\text{min}}}, P_{r_{\text{max}}}, R)
\]

\[
= \frac{1}{\pi R^2} \int_0^{2\pi} \int_0^R \left(1 - P_{\text{out}}(P_{r_{\text{min}}}, P_{r_{\text{max}}}, r)\right) \cdot r \, dr \, d\theta
\]

\[
= Q(a) + \exp\left(\frac{2 - 2a_{\text{min}} b}{b^2}\right)Q\left(\frac{2 - a_{\text{min}} b}{b}\right)
\]

\[
- \exp\left(\frac{2 - 2a_{\text{max}} b}{b^2}\right)Q\left(\frac{2 - a_{\text{max}} b}{b}\right),
\]

(8)

where

\[
a_{\text{min}} = \frac{P_{r_{\text{min}}} - P_r(R)}{\sigma_{\text{dB}}}, \quad a_{\text{max}} = \frac{P_{r_{\text{max}}} - P_r(R)}{\sigma_{\text{dB}}}, \quad \text{and} \quad b = \frac{10\gamma \log_{10} e}{\sigma_{\text{dB}}}.\]

\(P_{\text{out}}\) and \(P_{\text{cov}}\) are calculated using Eqs. (7) and (8) and are shown in Fig. 1(b). In the evaluation, \(R\) of 500 m, \(\gamma\) of 3.71, and \(\sigma_{\text{dB}}\) of 3.65 dB are assumed [6]. The power variation due to the antenna directivity was substituted by the dynamic range, and then \(\alpha = 8\) dB and \(P_{r_{\text{max}}} = -122\) dBm. The maximum coverage probability was about 0.4 due to the near-far problem.

### 3.2 Transmission power control of pseudolite

For further mitigating the near-far problem, a method of transmission power control is proposed as illustrated in Fig. 2(a). The duty ratio of the conventional PL is \(\tau_1\) in the figure, and the proposed PL transmits the second pulse after the 1st pulse transmission. The duty ratio is \(\tau_2\) and the power is reduced by \(\alpha\) [dB] than the 1st pulse power. The proposed PL transmits several pulses in a cycle, and the serving region illustrated in Fig. 2(b) is superposition of serving region rings corresponding
to the pulses, and it results in a wider serving region. If we set $\alpha$ to the power difference between $P_{\text{max}}$ and $P_{\text{min}}$, serving areas $A_1$, $A_2$, and $A_3$ corresponding to the pulses are continuous as illustrated in Fig. 3(a). For $R_1 = 500$ m and $\gamma = 3.71$, $R_2$, $R_3$, $R_4$ are 304.3 m, 185.2 m, and 112.7 m. Because the PL signal is available when the receiver can use one of these pulses, the coverage probability of the proposed PL is calculated by $1 - (1 - P_{\text{cov1}}) \cdot (1 - P_{\text{cov2}}) \cdot (1 - P_{\text{cov3}})$, where $P_{\text{cov1}}$, $P_{\text{cov2}}$, $P_{\text{cov3}}$ are the coverage probabilities for 1st, 2nd, and 3rd pulses. The coverage probability is shown in Fig. 3(b). The maximum coverage probability of the proposed PL was about 0.6. The ring radius of the conventional PL was from 378 m to 500 m, whereas that of the proposed PL was from 134 m to 500 m. The serving region of the proposed PL is 2.16 times wider than that of the conventional PL.

### 4 Conclusion

In this letter, a method of transmitting pseudolite signal was proposed. The outage and coverage probabilities of the conventional burst pulse pseudolite in a shadowing environment were evaluated. Then, the pseudolite whose transmission power varies pulse-to-pulse was proposed for mitigating the effect of the near-far problem. The serving region of the proposed PL is 2.16 times wider than that of the conventional pseudolite.
Estimating the cause of frame losses in 802.11 wireless networks

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Abstract: In indoor wireless environments such as warehouses and factories, since radio propagation occurs in complicated ways and there are large moving objects like Automated Guided Vehicles (AGVs), from the perspective of network management, maintaining wireless communication quality is challenging. Because the failures in wireless communications may lead to undesirable delay or even stop in the manufacturing processes, it is desirable not only to detect degradations in link quality, but also to estimate the cause of the frame losses, to recover from the link failures as quick as possible. In this paper, we propose a sensing system that allows comprehensive analysis between physical and Medium Access Control (MAC) layer information, and allows estimation of the cause of the frame losses due to signal attenuation or interference. We implemented a prototype sensor node using an FPGA board and conducted an experiment to investigate the accuracy of the estimation. From the experiment, we confirmed that the prototype sensor node can determine whether the cause is due to attenuation or interference with high accuracy.

Keywords: wireless LAN, sensing system, cross layer

Classification: Network Management/Operation

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1 Introduction

Internet of Things (IoT) devices are recently deployed rapidly to the indoor environments such as warehouses, factories, hospitals, and malls [1, 2]. By introducing the IoT devices to these environments, it is expected that the various operations such as monitoring of the environmental data and device conditions, industrial automation, and maintenance are facilitated. On the other hand, in such environments, since signal attenuation and interference occur in complex and unpredictable manner, it is difficult to ensure the network stability. From the perspective of network management in such challenging environments, it is not enough just to detect the degradations in link quality, and is desirable to estimate the cause of the link failures, to recover from the performance degradations quickly. Signal attenuation and interference are the two of the major causes of the frame losses in the indoor environments. For example, in factories, signal attenuation will occur due to various moving objects such as AGVs. Moreover, since the indoor radio propagation is complicated, interference due to the hidden terminal situation will occur easily.

The information mainly related to the physical layer (called physical layer information) and the information mainly related to the MAC layer (called MAC layer information) are the information useful for assessing wireless link quality. The typical examples of acquiring these information are spectrum analysis [3] (physical layer information) and packet capture [4] (MAC layer information). Spectrum analysis allows the users to know that how much congestion is occurring in the target frequency band, how strong the radio signals arrive, etc. Packet capture allows the users to know that which nodes are communicating, what kind of control frames are transmitted, etc. In the existing schemes or products, physical layer information and MAC layer information are often acquired and analysed individually [3, 4]. However, as given in the above examples, since each layer contains information that is difficult to be acquired from the other layer, by acquiring and analyzing both simultaneously, it seems that we can perform more detailed assessment on the wireless link quality. Specifically, in this paper, we focus on the estimation of the cause of the frame losses due to signal attenuation or interference. It is possible to detect the occurrence of the frame losses by analysing the result of packet capture easily. However, it is difficult to estimate the cause of the frame losses from only the packet capture data. By combining information both from physical and MAC layers, we propose the procedure that determines the cause of frame losses accurately.

2 System overview

We suppose that multiple IoT devices are deployed in the target environment (e.g., factory) and they are communicating via IEEE 802.11 wireless LANs. To collect physical and MAC layer information, we introduce a sensing system that consists of multiple sensor nodes and an analysis node. Fig. 1 shows the module structure of a sensor node. Each sensor node is tuned to receive the radio signals transmitted by the IoT devices, acquires In-phase and Quadrature-phase (IQ) data, which is a time series of signals to be demodulated, and generates two types of data from the IQ
data: (i) a time series of signal strengths (called *envelope data*, hereafter) and (ii) a time series of information extracted from each MAC header (called *header data*, hereafter), which is acquired by parsing the demodulated bit sequence based on the IEEE 802.11 specification. An envelope data and a header data fall into the categories of physical and MAC layer information, respectively. Both data are acquired on each sensor node simultaneously and continuously, and uploaded to the analysis node. One naive approach to acquiring both the physical and MAC layer information is that connecting both a Software Defined Radio (SDR) device and an off-the-shelf Wi-Fi interface to a sensor node, and calculating the signal strengths from the IQ data sampled by the SDR device and executing packet capture on the Wi-Fi interface. However, in this approach, it is difficult to maintain time synchronization between physical and MAC layer information. By generating the envelope data and the header data from the same IQ samples as shown in Fig. 1, we can synchronize between physical and MAC layer information easily. In this paper, we assume that the time synchronization between sensor nodes is achieved by using a time synchronization protocol such as Precision Time Protocol (PTP). By collecting sensory data from the sensor nodes, the analysis node can observe multiple data on an identical time axis and perform estimation procedure to be described in the next section.

3 Estimation of the cause of frame losses

In this section, we explain how to estimate the cause of the frame losses by analysing the envelope data and the header data acquired from multiple sensor nodes. We use an example shown in Fig. 2(a). In the figure, we consider the situation where data frames are sent from device B to the device A, and the first transmission is succeeded and the second transmission is failed due to signal attenuation (Fig. 2(b)) or interference (Fig. 2(c)). We suppose that the sensor node 1 and 2 are located near the device A and B, respectively. The procedure for estimating the cause of the frame losses (i.e., attenuation or interference) is as follows:

**Step. 1** First, we collect the data from the sensor nodes located near the sender side (sensor node 2 in the example) and near the receiver side (sensor node 1 in the example), respectively.

**Step. 2** Next, we find the header data that is contained in the set of the header data acquired at the sender side, but is not contained in the set of the header data acquired at the receiver side. For example, in Fig. 2(b) and (c), the header data that corresponds to $h_{2,1}$ is not contained in the set of the header data acquired by sensor node 1.
Step. 3  For each header data found in Step. 2, we find the enveloped data that has an intersection on the time axis from the set of the envelope data acquired at the receiver side. For example, in Fig. 2(b) and (c), $e_{1,1}$ has an intersection with $h_{2,1}$ on the time axis (i.e., $\text{time}(e_{1,1}) \cap \text{time}(h_{2,1}) \neq \emptyset$, where $\text{time}()$ represents the time range of the data).

Step. 4  Lastly, each pair of the header data $h$ found in Step. 2 and the enveloped data $e$ found in Step. 3, we calculate the absolute value of the difference between $\text{time}(e)$ and $\text{time}(h)$, and compare the value with the predefined threshold $t_{thr}$. If $|\text{time}(e) - \text{time}(h)| \leq t_{thr}$ holds, then we conclude that the cause of the frame loss is due to attenuation, otherwise, we conclude the cause is due to interference.

4 Implementation

We implemented a prototype sensor node, which consists of a RF transceiver (AD9371), an FPGA (Kintex), a SoC (Zynq), an ordinary PC, and antennas. The target frequency of the RF transceiver is 2.4 GHz band. The FPGA generates IQ data whose sampling rate is 20 MS/s. The envelope data is generated by calculating the signal strength of each IQ sample by the formula: $S = 10 \log_{10}(I^2 + Q^2)$, where $S$ is the signal strength, and $I$ and $Q$ are 16 bit integers that represent values of an I component and a Q component of an IQ sample, respectively. The envelope data is transferred directly from the FPGA to the PC via a 10 Gigabit Ethernet interface. The FPGA also performs the demodulation based on the IEEE 802.11b/g standard, and the demodulated bit sequence is transferred to the SoC. Then, SoC generates
the header data using the libpcap [5] library and transfers the header data to the PC via a Gigabit Ethernet interface. Then, the PC transfers the envelope and the header data to the analysis node.

5 Experiment

To investigate the effectiveness of the estimation method described in Section 3, we conducted an experiment that measures the accuracy of the estimation. In the experiment, we prepared the controlled environments where the frame losses occur due to signal attenuation (called environment 1) or interference (called environment 2) by connecting between sender, receiver, and sensor nodes with RF cables and attenuators. The details of the each environment is as follows:

Environment 1 One sender node, one receiver node, and two sensor nodes (each of which is located near the sender or the receiver node) are deployed. The sender node transmits an UDP broadcast frame with 1000 bytes payload at the data rate of 6 Mbps once in 10 ms. The attenuator between the sender and the receiver is adjusted so that the receiver node can not decode the frames successfully due to signal attenuation. Fig. 3(a) shows an example of envelope data acquired at the receiver side sensor node and the lost frames identified by Step. 2 of the procedure described in Section 3. From the figure, we can see that the frames are lost at the receiver node due to signal attenuation.

Environment 2 Two sender nodes (called data sender and interferer), one receiver node, and two sensor nodes (each of which is located near the data sender or the receiver node) are deployed. The Data sender transmits an UDP broadcast frame with 1000 bytes payload at the data rate of 6 Mbps once in

Fig. 3. Example of the envelope data (blue rectangles) acquired at the receiver side sensor node and the lost frames identified (green rectangles) by the Step. 2 in the procedure describe in Section 3.
10 ms. The interferer transmits an UDP broadcast frame with 1200 bytes payload at the data rate of 1 Mbps once in a random period sampled from the normal distribution with mean 15 ms and standard deviation 5 ms. The attenuators between the two senders are adjusted so that each sender do not detect ongoing transmission by the other sender. Fig. 3(b) shows an example of envelope data acquired at the receiver side sensor node and the lost frames identified. From the figure, we can see that frames are lost at the receiver node when the interfering signals are overlapped.

For each environment, the set of envelope data and the set of header data are acquired from the sensor nodes during the period in which 100 frame losses are recorded. Then, for the total of 200 instances of the frame losses, the cause of the frame loss of the each instance is determined using the procedure described in Section 3. In this experiment, we set $t_{thr}$ to 10 microseconds. From the experiment, we confirmed that the causes of the frame losses of the all 200 instances are estimated correctly.

6 Conclusion

In this paper, we showed that, by combining the information both from the physical and MAC layers, it is possible to estimate the cause of the frame losses due to attenuation or interference with high accuracy. In the future, we focus on more detailed evaluation of the proposed estimation method by deploying the sensing system in the actual environments and acquiring the data from the communications among actual IoT devices.

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Broadband design of U-shaped folded dipole antenna for WiMAX by using characteristic mode analysis

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Abstract: In this study, we utilize an effective method called Characteristic Mode Analysis (CMA) to analyze the physical mechanisms of U-shaped Folded Dipole Antenna (UFDA), which have been proposed for WiMAX device. The proposed antenna consists of the UFDA element and small Ground Plane. As a result, it is confirmed that two types of modes in the resonance of UFDA appear, the one is the mode of the Ground Plane and the other is the mode of the UFDA element. Subsequently, from these results, an easier method for broadband design of UFDA which can cover completely two frequency bands of WiMAX was investigated.

Keywords: characteristic mode analysis, U-shaped folded dipole antenna, broadband, WiMAX

Classification: Antennas and Propagation

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1 Introduction
To enhance the communication quality for wireless devices, the built-in antennas have been required to cover a wider bandwidth with a further compact design. Therefore, the design of antennas that have broadband characteristics is one of the most important concepts in antenna development for wireless devices. In the previous study, we have proposed a U-shaped folded dipole antenna (UFDA) composed of an UFDA element and a ground plane (GP) with small size which could be installed inside a wireless device for worldwide interoperability for microwave access (WiMAX). By using a suitable objective function in a particle swarm optimization (PSO), the designed UFDA covered dual frequency bands from 2.3 GHz to 2.7 GHz and from 3.4 GHz to 3.8 GHz for $S_{11} \leq -6$ dB. Furthermore, the occupancy area of the antenna was reduced to 23.1% of the previous optimized UFDA [1]. Although the final shape of the antenna could be only obtained by the PSO design, the physical mechanism of the antenna was not clear.

In this study, the characteristic mode analysis (CMA) is utilized as an effective method for the broadband design while understanding the physical mechanism of each part of the UFDA element and the GP [2, 3]. First, the characteristic modes of the UFDA on the GP are analyzed. After considering the physical mechanism of each part of the UFDA element and the GP, the antenna is designed to cover completely dual frequency bands of the WiMAX.

2 Antenna structure and CMA result
In this study, the simulator CST-MW STUDIO is used and the antenna is made of perfect electric conductor (PEC) for the analysis. Firstly, two parts of the antenna (UFDA element only and GP) are separated and analyzed independently. After determining the resonant frequencies and characteristics (broadband or narrowband) of modes of these separated parts, we combine them to return the initial model, then analyze with CMA. A mode will be defined as a resonance when its model significance (MS) value $= 1$ [4].

Fig. 1(a) shows the GP with the overall length of 75 mm $\times$ 31 mm and its MS. The modes are tracked from 1.0 to 5.5 GHz including frequency bands of WiMAX. Two resonances appear at the frequency of 1.7 GHz (Mode1) and 3.9 GHz (Mode2) and both of them have wide bandwidth defined by $MS > 0.707$ [5]. The antenna element has a folded dipole antenna bent into U-shape with detailed dimensions and its MS are shown in Fig. 1(b). As can be seen, three resonances at the frequency of 1.8 GHz (Mode3), 2.8 GHz (Mode4) and 5.1 GHz (Mode5) appear after analyzing the UFDA element only. By comparing the bandwidth of mode between GP only and UFDA element only, we know that, three resonances of UFDA element only have narrower bandwidth than the resonances of the GP only. Fig. 1(c) shows the MS of the combined model of the GP and the UFDA element without feeding. By analyzing the combined model, four resonances are defined including 1.9 GHz (Mode1), 3.2 GHz (Mode2), 2.5 GHz (Mode3) and 4.4 GHz (Mode4). With the same way to examine the characteristics of each resonance, we can realize that Model1 and Mode2 of the combined model have wide bandwidth similar to Model1 and Mode2 of the GP only shown in Fig. 1(b). Meanwhile,
Mode 3 and Mode 4 of the combined have narrower bandwidth, similar to Mode 3 and Mode 4 of the UFDA element only shown in Fig. 1(c).

3 Consideration

By observing these bandwidth similarities in the aspect of modes, we could recognize which mode is mostly influenced by which separated part of antenna. However, the mode of each part shifted to different frequency when the UFDA element and the GP are combined. Thus, the validity of them are needed to be confirmed by comparing the surface current distribution at each resonant frequency of mode between separated parts and the combined model.

Fig. 2(a) shows the surface current distribution of GP only at 1.7 GHz (Mode1) and of the combined model at 1.9 GHz (Mode1). Fig. 2(b) shows the surface current distribution of GP only at 3.9 GHz (Mode2) and of the combined at 3.2 GHz (Mode2). It can be seen that the locations of current maxima or minima are at the same position and the direction of the currents are also similar to both of GP. This verified that mode 1 and mode 2 is the mode of the GP.

Fig. 2(c) shows the surface current distribution of UFDA element only at 1.8 GHz (Mode3) and of the combined model at 2.5 GHz (Mode3). Fig. 2(d) shows the surface current distribution of UFDA only at 2.5 GHz (Mode4) and of the combined model at 4.4 GHz (Mode4). We can see that the surface current...
distributions are similar on the UFDA element and the directions are in the same direction to each other. For that reason, we could conclude that mode 3 and mode 4 are the modes of the UFDA element.

The simulated reflection coefficient of the combined model after feeding is indicated as Fig. 2(e). The UFDA element is fed by a 50 Ω source at feeding strip, and connected to the GP at the shorting strip. As can be seen, three resonances appear at 1.8 GHz, 2.6 GHz and 3.4 GHz, and they have the same resonant frequency of Mode 1, Mode 3 and Mode 2 in CMA result, respectively. Thus, the validity of CMA result was confirmed. For convenience, we call these resonances by the first, the second and the third resonance, respectively. By comparing with the result of CMA, we know that the first and third resonance are associated with mode 1 and mode 2 of GP. Meanwhile, the second one is associated with mode 3 of the UFDA element. Moreover, the gray ranges in Fig. 2(e) present two frequency bandwidths of WiMAX and they are not covered completely for $S_{11} \leq -6$ dB. Therefore, in order to cover completely these frequency bandwidths, the shape of GP should be adjusted to the first and third resonance shifting to higher frequency, while, the shape of UFDA element should be adjusted to the second resonance shifting to lower frequency.

(a) Current distributions of the GP only (1.7 GHz) and the combined model (1.9 GHz)

(b) Current distributions of the GP only (3.9 GHz) and the combined model (3.2 GHz)

(c) Current distributions of the UFDA element only (1.8 GHz) and the combined model (2.5 GHz)

(d) Current distributions of the UFDA element only (2.5 GHz) and the combined model (4.4 GHz)
4 Antenna design

Fig. 3(a) shows the design for GP only, which mostly affected mode 1 and mode 2 of CMA result. By observing the common position of maximum current distribution on the GP shown in Fig. 2(a) and Fig. 2(b), two parts on the GP are trimmed with a dimension of 10 mm × 5 mm. Therefore, resonant frequency of mode 1 and mode 2 could be shifted to higher frequency. While, for the UFDA element only, resonant frequency of mode 3 shift to lower frequency when a center part of the UFDA element is reduced from 4 mm to 2 mm, as shown in Fig. 3(b). After obtaining two separated parts with desired design, the new UFDA is combined and its CMA result is shown in Fig. 3(c). As can be seen that the resonant frequencies of mode 1 and mode 2 shift to higher frequency (2.1 GHz and 3.3 GHz, respectively), the resonant frequency of mode 3 shifts to lower frequency (2.3 GHz) than previous UFDA shown in Fig. 1(d).

The validity of the simulated results were confirmed by measured reflection coefficient of UFDA after designing, shown in Fig. 3(d). Therefore, the bandwidth with $S_{11} \leq -6$ dB is wider than previous result from 2 GHz to 2.7 GHz and from 3.3 GHz to 3.9 GHz, which could cover completely the WiMAX frequency bands. From Fig. 3(e), a good agreement is observed between the simulated and measured radiation pattern in the $yz$ plane.
5 Conclusion

In this study, by utilizing the advantageous features of CMA to analyze each part of the antenna, we gave further illustrations of the physical mechanism of UFDA. It is confirmed that there are two types of modes in resonances. The first one consists of two modes at 1.9 GHz and 3.2 GHz, affected mostly by the GP. While, the other includes two modes at 2.5 GHz and 4.4 GHz, mainly affected by the UFDA element. From these results, we investigated an easier method to help UFDA have broadband characteristic. By trimming and reducing suitably position of the maximum surface current on the GP and UFDA element, respectively, two frequency bands of WiMAX could be completely covered for $S_{11} \leq -6$ dB.

Fig. 3. Design for UFDA to cover WiMAX bandwidths
Penetration loss of outer wall materials for co-existence of indoor and outdoor-use sensors at 79 GHz

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Abstract: This paper presents penetration loss measurements of housing outer wall materials at 79 GHz to investigate the co-existence between indoor-use and outdoor-use radar sensors. The penetration loss was measured for building outer wall materials such as window and concrete wall. The penetration loss is also compared with 24 GHz-band because the band is available as indoor-use (e.g., non-contacted health-care monitoring) and outdoor-use (e.g., automotive radar) sensor systems. The loss of indoor-to-outdoor systems is experimentally investigated. As a result, the propagation path through outer wall material, at a distance of 5 m from the house, is found to be approximately 50 dB attenuated, which is 15 dB larger as compared with 24 GHz.

Keywords: co-existence, 79 GHz, indoor sensor, automotive radar

Classification: Sensing

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1 Introduction

24 GHz and 77 GHz frequency bands are currently assigned to automotive radar applications in Japan, and 79 GHz-band with up to 4 GHz bandwidth will become available soon. The 79 GHz-band automotive radar system is therefore on the market as a key technology for safety and self-driving systems because it offers all-weather capability, high range-resolution and wide-angle detection for peripheral monitoring [1, 2]. The 79 GHz-band has also attracted considerable attention as indoor-use applications such as non-contact vital (heart and respiration) monitoring because of the very high range-resolution and low privacy invasion [3, 4, 5]. To realize the co-existence with automotive applications for a same frequency band, however, the mutual interference must be understood accurately. Indoor to outdoor penetration loss is one of the most important factors of affecting the co-existence. Generally, the penetration loss at mm-wave frequencies is known to be larger relative to micro-wave frequencies, depending on the material types and incident angle [6, 7]. In this paper, penetration loss of housing outer wall materials such as window and concrete walls was measured to discuss the mutual interference at 79 GHz. Especially, outer wall windows will be crucial since the penetration loss should be lower than other wall materials. The penetration loss is also compared with 24 GHz because the 24 GHz band is available as indoor-use (e.g., security and industry robot sensor) and outdoor-use (e.g., automotive and unmanned aerial vehicle radar) sensor systems. As a result, the 79 GHz signal is found to be largely attenuated by glass window walls.

2 Measurement set-up and wall penetration loss

A. Measurement setup and penetration loss

Measurement was conducted in a frequency range of 79–80 GHz using a VNA (Agilent: E8363B) with millimeter modules in an anechoic chamber room. The measurement was also conducted in a range of 24–25 GHz for comparative study of penetration loss. Housing outer walls are generally made of concrete, PB (plasterboard), glass window walls. Especially, modern residential window wall includes double-paned glass and coated glass with a thin UV-cut seal to improve indoor comfort in summer and to prevent indoor thermal loss in winter. The former may have a frequency selective penetration loss depending on the air-space thickness, while the latter some insertion loss. Frosted window glass is also used widely which allows visual privacy, while still allowing penetration of natural light. We therefore consider (i) clear window glass of 5 mm thick, (ii) frosted window glass of 5 mm thick, (iii) UV-cut window glass of 6.6 mm thick, (iv) double-paned window glass of 13.8 mm thick, (v) PB wall of 70 mm thick, and (vi) concrete wall of 80 mm thick as illustrated in Fig. 1(a). The transmit and receive antennas with a gain
20 dBi were always boresight-aligned with the material on turn table placed in between both the antennas as shown in Fig. 1(b).

Penetration loss was calculated as the difference with the received power, $P_{FS}$, in free space with the same Tx-Rx separation distance. The penetration loss $L$ is given by,

$$ [L] = [P_{FS}] - [P_m] $$

$$ [P_{FS}] = [P_t] + [G_t] + [G_r] + 20 \log_{10} \left( \frac{c}{4\pi f d} \right) $$

where $P_t$ is the transmit power, $G_t$ and $G_r$ are the transmit and receive antenna gain respectively, $c$ is the speed of light, $f$ is the frequency, and $P_m$ is the received power for a material.

B. Measurement result
The measurement of each material was conducted as a function of incidence angle. Measurement results of a single clear and frosted glass are shown in Fig. 2(a) and (b) respectively. The penetration loss of clear glass for low incident angle is not almost seen at 24 GHz, while approximately 3 dB at 79 GHz. And increased penetration losses have been observed for larger incident angle. For the frosted glass, the effect of surface roughness on loss is not seen for the incident angle of less than 30°. This is because the surface roughness is much less than the wavelength. Measurement results for laminated and double-paned window glass are shown in Fig. 2(c) and (d) respectively. Some insertion loss of metallic seal is seen for 24 GHz, but small for 79 GHz. The penetration loss of double-paned window glass at 79 GHz is approximately 14 dB and is like to decrease with the incident angle unlike 24 GHz. As predicted, therefore, the doubled-paned glass is seen to have some frequency selective penetration loss for 79 GHz because of the resonance. Please note that the air space between window panes is approximately half
of the wavelength of 79 GHz. The penetration loss for PB wall (plaster-board plus insulating material) and concrete wall are shown in Fig. 2(e) and (f) respectively. The penetration loss of the PB wall is relatively small which is similar to the single window glass. And the loss for the concrete for 79 GHz is 80–100 dB which is double of 24 GHz. It is found from the above results that the penetration loss of 79 GHz is large relative to 24 GHz.

3 Propagation loss

Interference from an indoor-use sensor at 79 GHz to outdoor-use sensor is investigated where the measurement was conducted in a house (3LDK, 91 m²). The Tx antenna with a height of 0.6 m was placed 1.5 m away on the inside from the window glass wall and the outside received powers were recorded at a total of 196 locations for an area of 5 × 5 m² as shown in Fig. 3. The thickness of exterior concrete wall is 250 mm and the window wall is double-paned window glass with...
of 13.8 mm-thin. Fig. 3(a) and (b) show the top-viewed received power distribution through outer walls at 79 GHz and 24 GHz respectively, where the transmit power is 0 dBm and the Tx and Rx antenna height is 60 cm. The window glass leakage is approximately 20–30 dB at 79 GHz, while 40–50 dB at 24 GHz. And the sensor radio isolation at 79 GHz is seen to be high because of the high penetration loss and

Fig. 3. Outdoor leakage of indoor-use RF sensor through outer walls
diffraction loss. Another loss factor is the path attenuation of the signal as a function of two material space. It is interesting to investigate the path loss coefficient for line-of-sight area through the outer walls because it is a major component in the design of the link budget or interference effect. The coefficient for 79 GHz and 24 GHz are approximately 2.12 and 1.61 respectively. And the propagation path through outer wall material at 79 GHz is found to be approximately 50 dB attenuated at a distance of 5 m from the house, which are 15 dB larger as compared with 24 GHz. Suppose an acceptable interference level of 70 dBm, for example, the probability that the mutual interference will occur is low when the outdoor-use system is 30 m away from the house.

4 Conclusion

In this paper, we conducted the measurement of indoor-to-outdoor attenuation of 79 GHz-band sensor signal where some concrete wall and several types of window wall are considered. And the penetration loss was also compared with 24 GHz-band. The interference level of indoor-to-outdoor sensor systems at 79 GHz is shown to be significantly attenuated by outer walls and window glasses. As a result, the effect of indoor-use sensor on vehicular radar system is found to be relatively small at 79 GHz and the probability that mutual interference will occur is low.
Localized electric field generating for rats head using a patch antenna in 26.5 GHz band

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Abstract: In this paper, two types of patch antenna that can be used for animal experiments are proposed.

The proposed antenna can generate a localized electric field at the head of the rat. The operating frequency of the system is 26.5 GHz. This frequency is used the fifth generation of mobile communication (5G). The localization of the electric field is evaluated by the ratio between target area field and whole body field.

Keywords: patch antenna, reflection coefficient, SAR

Classification: Antennas and Propagation

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1 Introduction

In recent years, the fifth generation of mobile communication (5G) has been actively developed. In this communication system, relatively higher frequency ranges such as 28 GHz will be used. On the other hand, while the mobile
communication device using, a part of electromagnetic energy is absorbed by the human body including the head. The guidelines [1] are provided to protect human safety. However, guidelines for higher than 10 GHz have not been provided, and animal experiments have not been conducted to evaluate the effects of health. In order to investigate the health effect, localized electric field should be realized for rat head to simulate human mobile device use. Then, it is necessary to develop an antenna for animal experiments. In this research, we develop two types of antennas to be used for animal experiments at 26.5 GHz. These antennas are designed by numerical simulation using the FDTD method.

2 Localized field generating by the patch antenna

In this animal experiment, the electromagnetic field should be concentrated on the rat head by the antenna. The patch antenna that satisfies both conditions is used in this research [2]. For further miniaturization, a shorted patch antenna is utilized. It is expected that shorted patch antenna is less affected to the reflected waves from the rat. In the antenna design, the inner conductor radius of the coaxial cable is considered for modeling realistic model. The antenna parameters are decided based on ref. [3]. Fig. 1(a) shows patch antenna, Fig. 1(b) shows a shorted patch antenna [3, 4]. The reflection characteristics of the antennas are calculated. In this calculation, the 8 weeks old rat model is placed near the antennas as shown in Fig. 1(c). The aim of this research is to generate localized electric field at the target area. In this paper, the target area is defined as shown in Fig. 1(c). The distance between the antennas and the target area is 4 to 5 mm. The radiating element size of the shorted patch antenna is about a quarter of the size of the normal patch antenna. Fig. 1(d) and Fig. 1(e) show the calculated reflection coefficients of the normal patch antenna and the shorted patch antenna, respectively. In these calculations, the antenna is placed above the rat and the distance \( H \) is 4 and 5 mm. In these results, both antennas satisfy \(-10\) dB at 26.5 GHz at any distance. As comparing with the normal patch antenna, the shorted patch antenna has less change in resonance frequency.

Next, the localization of the electric field in the rat is evaluated. The SAR (Specific Absorption Rate) is used to evaluate the localization.

\[
SAR = \frac{\sigma E^2}{\rho}
\]

The SAR is defined by eq. (1).

\[
Localization = \frac{SAR_{TA}}{SAR_{WB}}
\]

In this paper, the localization is defined by eq. (2).

In eq. (2), \( SAR_{TA} \) is SAR average of target area, \( SAR_{WB} \) is SAR average of whole body of the rat.

Figs. 2(a), (b), (c), (d) are projection views of the SAR when the distance \( H \) is 5 mm. The target areas are indicated as a circle in these figures. The electromagnetic waves are intensively exposed to the head in both antennas. Next, the localization and average SAR in the rat model is indicated in Table I. The input power of the antennas is 1 W. Regarding the value of SAR in target area, as distance \( H \) is
decreased, the value of $SAR_{\text{average}}$ increases both antennas. The localities of both antennas are a quite high value compared with other proposed antennas [5]. In these calculations, the localization of the shorted type patch antenna is lower than the normal patch antenna, however, the localization is high enough. Therefore, both antennas can be used for our purpose.

Fig. 1. Rat model, antenna parameters and reflection coefficient
3 Conclusion

In this research, two types of patch antenna that can be used for animal experiments are developed. The SAR and the localization are calculated by the simulation. The developed antennas can be generating localized field for rat head.

Acknowledgments

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![Projection views of the maximum value of SAR](image)

**Fig. 2.** Projection views of the maximum value of SAR

**Table 1.** Evaluation of electromagnetic field exposure

|                | SAR\text{average} [W/kg] |
|----------------|--------------------------|
|                | Patch antenna            | Shorted patch antenna |
|                | 4 mm                     | 5 mm                   |
| Fat            | 0.00627                  | 0.00551                | 0.0126                  | 0.0135                  |
| Muscle         | 0.146                    | 0.159                  | 0.178                   | 0.187                   |
| Bone           | 2.54                     | 2.37                   | 1.72                    | 1.59                    |
| Organ          | 0.00699                  | 0.00734                | 0.00400                 | 0.00461                 |
| Eye            | 67.9                     | 64.9                   | 46.0                    | 47.4                    |
| Brain          | 13.5                     | 12.6                   | 8.05                    | 7.57                    |
| Skin           | 13.7                     | 13.4                   | 12.1                    | 12.0                    |
| Target area    | 776                      | 631                    | 271                     | 254                     |
| Whole body     | 1.83                     | 1.78                   | 1.59                    | 1.58                    |
| Absorption rate [-] | 0.54                | 0.53                   | 0.42                    | 0.42                    |
| Localization [-] | 425                  | 354                   | 170                     | 161                     |
A robust detection method for micro-Doppler feature of rotating blades under low SNR

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Abstract: Rotation of rotor blades is typically micro-motion and would induce micro-Doppler modulation. However, accurate detection will become extremely difficult when the echo signal is corrupted by strong noise. This paper proposes a method to realize extraction of micro-Doppler feature from the rotating rotor blade echo signals with low signal-to-noise ratio (SNR). The proposed method is characterized by applying low-rank decomposition for saliency detection of spectrogram. MATLAB simulation results verify that the proposed method is effective in saliency detection of the spectrogram from rotor blade echo signals and is robust in estimating the maximum micro-Doppler shift under low SNR.

Keywords: spectrogram, micro-Doppler, saliency detection, Hough transform

Classification: Sensing

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1 Introduction

Micro-motion could cause side-band Doppler frequency shifts around the centre of Doppler shifted carrier frequency; which is called micro-Doppler effect [1]. The frequency modulation of micro-Doppler effect contributes mono-/multi-component of sinusoidal frequency modulation or multi-frequency pattern combination.

Extraction of the micro-Doppler features from echo signals can provide a basis for detection of flying targets. The empirical mode decomposition [2], compressive sensing [3] and variational mode decomposition [4] are the typical algorithms aimed to separate micro-Doppler contributions from multicomponent signals. These methods suffer huge computation loads when estimated data are high-dimensional. With applying time-frequency analysis, micro-Doppler parameters could also be estimated by mapping the time-frequency distribution onto the parameter space by a pattern recognition tool, such as Hough transform. The existing Hough transform based methods detect sinusoids produced by micro-Doppler effect [5, 6], which are easily submerged in background noise, thus they tend to lose robustness at low SNR. Instead of sinusoid detection, an envelope detection method realizes estimation of the maximum Doppler frequency by detecting location of the steepest negative slope of the envelope of the spectrum [7]. The method still suffers poor performance under low SNR without adopting effective noise suppression.

In this paper, we propose a method that estimates the maximum Doppler frequency by detecting frequency flashes, which have the strongest spectral amplitude in the spectrogram. The proposed method is Hough transform based and featured by saliency detection of the spectrogram: global low-rank decomposition is applied to obtain saliency map firstly, then the maximum micro-Doppler shift is estimated by detecting frequency flash lines. The rest of the paper is organized as follows. The rotor signal model is given and its spectrogram is obtained, followed with description of the proposed method. Simulation results validate the proposed method finally.
2 Signal model of rotating rotor blades

For a rotor with $N$ blades, the echo continuous wave (CW) signal from rotating rotor blades with a certain reflectivity could be given by [1]:

$$s(t) = L \times \exp \left( -\frac{j4\pi R_0}{\lambda} \right) \times \sum_{n=0}^{N-1} \text{sinc} \left( \frac{2L}{\lambda} \cos \beta \cos \left( 2\pi f_r t + \phi_0 + \frac{2\pi n}{N} \right) \right)$$

where $L$ is the blade length; $R_0$ is the distance between the radar and rotation center; $\lambda$ is the wavelength of the transmitted signal; $\beta$ is the elevation angle; $f_r$ is the blade rotate frequency; $\phi_0$ is the initial phase, $\phi_0 \sim U(0, 2\pi)$.

Its maximum micro-Doppler shift for a single blade is given by:

$$f_d = \frac{4\pi L f_r}{\lambda} \cos \beta$$

The echo signal $s(t)$ could be transformed into time-frequency domain through short-time Fourier transform (STFT). Its spectrogram $TF$ is a two-dimensional map of power spectral along the time axis:

$$TF(t, f) = |STFT(t, f)|^2$$

$$= \left| \sum_{m_l=\infty}^{w-1} w(m_l) s(m_l + (t - 1)(W - O_1)) e^{-j2\pi km / W} \right|^2$$

where $w(\cdot)$ is the Hamming window function with length $W$, $m_i$ is the index of window position, $O_1$ is the overlap of two consecutive analysis windows, $k$ is the index for frequency $ko$ with $\omega_0$ representing the Fourier resolution and equals to $2\pi f_s / W$ for the sampling frequency $f_s$. Thus $TF \in \mathbb{R}^{W \times Q}$, $Q = \left\lfloor \frac{N_s - O1}{W - O1} \right\rfloor$, where $N_s$ is the total number of signals and $N_s = t_s f_s$ for observation time $t_s$.

With the sinc(·) in Eq. (1) in time domain to be turned into rectc(·) in frequency domain by executing STFT, frequency flashes with a certain width are produced in the spectrograms.

Fig. 1 demonstrates two noise-free spectrograms from the rotating rotor with two blades and a single blade, respectively. When the number of blades is even, the frequency flashes occur in a symmetric spectrum and the frequency of flashes is twice the rotation rate. When the number of blades is odd, the frequency flashes occur in an asymmetric spectrum and the frequency is the same with rotation rate.

The micro-Doppler traces of rotating scatter points on the blade tip are sinusoids in general cases. While the incident ray is perpendicular to the blade, all scatter points along the blade’s surface mainly produce specular reflections, which are presented as frequency flashes. The frequency flashes range from 0 Hz to $f_d$ and have the strongest spectral amplitude in the spectrogram. Therefore, detection of frequency flashes could potentially be an effective choice.
3 The proposed method

We propose a detection method for the maximum Doppler frequency shift in this section. Processing of the proposed method mainly consists of saliency detection by low-rank decomposition and line detection by Hough transform.

Non-saliency information causes great interference to the extraction of micro-Doppler features in spectrogram. Since a data matrix could be decomposed into a low-rank matrix and a sparse matrix, the initial spectrogram $TF$ then can be expressed as:

$$TF = TF_s + E$$

where $E$ is the gross and sparse component of the noise residual, $TF_s$ is the low-rank component of the salient regions.

The problem is described as a nuclear norm minimization approach:

$$\min_{L,M} \|TF_s\|_* + \varepsilon|E|_1$$

subject to $TF = TF_s + E$  \hspace{1cm} (5)

where $\| \cdot \|_*$ is the nuclear norm; $| \cdot |_1$ is the $l_1$-norm; $\varepsilon$ is a positive weighting parameter that guarantees the exact recovery and $\varepsilon = Q^{-1/2}$.

Following steps are carried out to extract the micro-Doppler features:

1. Calculate the global saliency map $TF_s$ by accelerated proximal gradient (APG) approach [8].
2. Calculate the optimum threshold value $T_0$ by iterative threshold method [9], and segment $TF_s$ based on the threshold value $T_0$ to obtain the binary saliency map $TF_b$.
3. Remove all connected pixels that are fewer than the empirical threshold $p_n$ from the binary saliency map $TF_b$. Through doing so noise pixels remained after saliency detection could be eliminated. The $p_n$ is set to the mean number of pixels that are contained in connected noise pixels in $TF_b$ without target signal.
4. Divide $TF_b$ along the zero frequency into the upper image $TF_u$ and the lower image $TF_d$ for uncertain symmetry of the spectrogram caused by the unknown rotor blade number.
5. Calculate the length set of the line segments of frequency flashes $L_t$ and $L_d$ in $TF_u$ and $TF_d$ respectively using Hough line detection [9].
Within the short observation interval, the rotor is assumed to rotate at a constant rate. Then the frequency flashes have the same length in $TF_i$ and $TF_d$. The estimated maximum micro-Doppler shift $\hat{f}_d$ is:

$$\hat{f}_d = \frac{\omega_0 L_i + L_d}{C^2}$$

(6)

4 Simulation results

Simulations are conducted to demonstrate performance of the proposed method using MATLAB. The echo signal is generated based on Eq. (1) by setting $\lambda$ to 0.015 m, $R_0$ to 7 km, $f_s$ to 200 KHz and $t_s$ to 0.05 s. For the rotating rotor blades, $f_r = 300$ Hz, $\beta = 0^\circ$, $L = 0.2$ m. We set the blade number to 2 for model M1, M2 and M3, and the blade number to 1 for M4, to demonstrate the influence of frequency flash density on feature extraction. We then set the blade length to 0.1 m, 0.2 m, 0.3 m and 0.2 m for M1, M2, M3 and M4 respectively to demonstrate the influence of frequency flash length to feature extraction. To realize trade-off between time resolution and frequency resolution, we set $W$ to 512 and $O_l$ to 8 in STFT, $p_n = 8$ in the step 3.

Fig. 2(a) illustrates the original spectrogram $TF$ for M2 at SNR $= -5$ dB. The micro-Doppler sinusoid produced by scatters on the blade tip are buried in strong noise and the frequency flashes are also polluted by strong noise pixels. Fig. 2(b) demonstrates the binary saliency map $TF_b$ after saliency detection. The extracted salient regions of the frequency flashes are very clear in the spectrogram.

Furthermore, Monte Carlo simulations are run 100 times under different values of SNR and the Root Mean Square Error (RMSE) for each SNR is calculated. Performance for the proposed method is compared with two conventional methods: Envelope detection [7] and Hough detection [9]. The Hough detection binarizes $TF$ by iterative threshold method then detects lines by Hough transform. Compared to the Hough detection, the proposed method further processes $TF$ by low-rank decomposition before image binaryzation.

Fig. 3(a) demonstrates that the proposed method achieves the best performance for M2 and M4 and the performance converges into 4% at SNR $= -7$ dB and $-6$ dB respectively, followed with the performance of Hough detection and which
converges at $\text{SNR} = -2 \, \text{dB}$ and $-3 \, \text{dB}$. Envelope detection does not reach its convergence performance within the investigated SNR range. In Fig. 3(b), we notice that the target RMSE reaches 4% at SNR of $-4 \, \text{dB}$ for M1, $-7 \, \text{dB}$ for M2 and $-10 \, \text{dB}$ for M3. These results suggest that longer flashes contribute to a higher estimation accuracy.

Interestingly, Hough detection achieves superior performance for M4 than for M2, which suggests that the lower frequency flash density for M4 has produced detection gain. However, the proposed method for M2 performs better than M4, which suggests that longer frequency flash length could improve saliency detection and lead to a better performance of Hough line detection.

5 Conclusion

We propose a method to extract micro-Doppler feature from the rotor blade echo signal. The proposed method is Hough transform based and is characterized by saliency detection of the spectrogram; in particular, to improve robustness at low SNR, low-rank decomposition is applied in obtaining saliency map. Simulation results have verified the proposed method’s effectiveness in detecting saliency in the spectrogram and robustness in estimating the maximum micro-Doppler shift under low SNR.

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