Low complexity digital back propagation method using phase linear approximation for nonlinear distortion compensation for long haul transmission systems

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Abstract: We propose a novel nonlinear distortion compensation technique using linear phase transition approximation for long haul optical communication system. Self-phase modulation (SPM) for phase change, and amplified spontaneous emission (ASE) from erbium-doped fiber amplifier (EDFA) are taken into consideration in simulation. It is confirmed that phase of 28Gbaud return to zero-quadrature phase shift keying (RZ-QPSK) signals compensated with split-step Fourier methods (SSFM) change almost linearly. They can be estimated using linear approximation with only two points of any span for low computational complexity. The performance of this technique is evaluated by bit error rate (BER) with numerical analysis. The achieved reduction of computational amount is 70.0% when transmission distance is 2500 km and input power is 0 dBm at a FEC limit of BER $< 10^{-3}$.

Keywords: digital signal processing, nonlinear distortion, SPM, chromatic dispersion

Classification: Fiber-Optic Transmission for Communications

References


1 Introduction

The reach of optical communication is greatly limited due to impact of nonlinear effect such as intra-channel (SPM: Self Phase Modulation, IXPM: Intra-channel Cross-phase Modulation, XPolM: Cross Polarization Modulation) and inter-channel interference (XPM: Cross Phase Modulation, FWM: Four Wave Mixing) to increase the total transmission capacity with advanced modulation format and several kinds of multiplexing. In order to use multilevel modulation format and to transmit the signal over a long distance, it is necessary to increase the input power and improve the SNR. However, it is difficult to increase the optical power and improve SNR for high capacity communication because of nonlinear Shannon limit [1]. To mitigate the impact of nonlinear distortion, several advanced digital back propagation (DBP) technique have been mainly proposed based on split-step Fourier methods (SSFM) [2, 3] and Volterra series nonlinear equalizers (VSNE) [4, 5]. Although nonlinear compensation with DBP has been regarded as a highly iterative technique, however, complex computation performance is required in real time, and that has been one of the significant technical issues to be overcome. Therefore, development of a novel DBP algorithm that relaxes the computation complexity is an important mission.

In this paper, we propose a novel efficient technique for nonlinear distortion compensation based on SSFM calculation. We confirmed that the phase transition of symbols per span with SSFM compensation changed almost linearly under the simulation with SPM, IXPM, and ASE taken into consideration. Hence, we estimate the phase of symbols on input point from the simulation results between two points from phase information per span obtained from linear approximation. We numerically assess the performance of linear approximation, and compared against the standard SSFM for 28Gbaud single polarized RZ-QPSK signals. We also evaluate the bit error rate (BER) performance, and report the number of steps per span and target spans for effective calculation, as well as the effect of reduction of computational amount.

The structure of this paper is as follow. First, we introduce the principle of the linear approximation technique based on SSFM. Next, the numerical results of the proposed linear approximation performance evaluated by BER are presented. Finally, the conclusions of this study are stated.
2 Principle of linear approximation based on SSFM

The SSFM is based on the concept that original signals are evaluated from the received signals propagating in the opposite direction against the transmission one. This technique divides the fiber link into small steps and the linear and nonlinear distortion are taken into consideration in frequency and time domain. SSFM is useful for solving the nonlinear Schrödinger equation,

\[
\frac{\partial E}{\partial z} = -\frac{\alpha}{2} E + i\frac{\beta_2}{2} \frac{\partial^2 E}{\partial t^2} - i\gamma |E|^2 E,
\]

where \(E\) represents the electric field intensity. \(\alpha, \beta_2,\) and \(\gamma\) are the propagation loss, group velocity dispersion and nonlinear coefficient of the fiber, respectively. \(z\) and \(t\) are the variables of propagation direction and time.

Considering the attenuation term in electric field intensity of Eq. (1),

\[
A = e^{-\alpha z} E,
\]

\[
\frac{\partial A}{\partial z} = i\frac{\beta_2}{2} \frac{\partial^2 A}{\partial t^2} - i\gamma e^{-\alpha z} |A|^2 A,
\]

and consider only the phase change amount \(\varphi_L\) when transmitting the distance \(L\) is

\[
\varphi_L = \int_0^L \Delta \varphi \, dz = \int_0^L \left( \frac{\alpha^2 \beta_2}{2} - \gamma e^{-\alpha z} |A_0|^2 \right) \, dz
\]

\[
= \frac{\alpha^2 \beta_2}{2} \cdot L - \gamma \frac{1 - e^{-\alpha L}}{\alpha} |A_0|^2,
\]

and in our proposed scheme, in order to make the analysis of the phase change easier, we compensate the linear distortion in advance, then the linear distortion term from Eq. (4) can be neglected and the phase change amount between EDFAs is

\[
\varphi_{EDFA} = -\gamma \frac{1 - e^{-\alpha L_{EDFA}}}{\alpha} |A_0|^2.
\]

Hence, the phase change amount between EDFAs are same as \(\varphi_{EDFA}\) in the whole transmit of optical signals, and the phase change can be estimated with linear approximation.

Fig. 1(a) shows the simulation results of the phase transition of a symbol. Setup parameters for the simulation are as follows; propagation loss of SMF of \(\alpha = 0.16\, \text{dB/km}\), dispersion parameter of \(D = 16\, \text{ps/nm/km}\), nonlinear coefficient of \(\gamma = 1.5\, \text{W}^{-1}\text{km}^{-1}\), and core radius of 5.0 \(\mu\text{m}\). An EDFA (erbium-doped fiber amplifier) that gain of 16 dB is used for compensating fiber transmission loss. The noise figure of 3 dB is used, and it is installed every 100 km. A 28Gbaud single-polarized RZ-QPSK signal at a wavelength of 1550 nm consists of pseudo-random bit sequences with word length of \(2^{16} - 1\). The presented symbol whose phase information is assigned to \(\pi/4\) at the launched point of the RZ-QPSK signal. It is transmitted through 5000 km, then compensated with SSFM using linear approximation. The number of steps per span is 10. Here, a step is one section when fiber is separated by a small distance, a span is distance between EDFAs. We can see that the simulated points are changing almost linearly. Hence, the phase in each span of symbols with SSFM can be approximated linearly and estimate the symbol phase on input point. Hence, the symbol phase on launched point can be
estimated from the extension of the straight line. Linear approximation on Fig. 1(a) represents the linear line utilizing the results in span No. 1 and 2, or in span No. 1 and No. 10. No. 1 is defined as the first span from the received side. We can see that the point and the approximation line are almost agreed with the input point in the graph of No. 10. On the other hand, as in the graph using No. 1 and No. 2, the phase is shifted far from the exact phase at the transmission point, thus it is necessary to find the suitable conditions.

In the calculation of standard SSFM in Fig. 1(a), we set the number of steps per span: \( N_{\text{step}} = 10 \), the number of target spans: \( N_{\text{span}} = 10 \), and the number of steps (required SSFM computational amount) to be calculated is \( N_{\text{step}} \times N_{\text{span}} = 100 \) steps. From this calculation, it is found that the phase of each step changes nonlinearly as superimpose, and the SPM phase change increases at the points of launched side where the light intensity is high.

Fig. 1(b) shows the analysis flow of SSFM for linear and nonlinear distortion compensation including the linear approximation. The first three blocks are conventional SSFM processing. CDC and NLC stands for chromatic dispersion and nonlinear dispersion compensator, respectively. As mentioned above, SSFM compensates linear and nonlinear distortion per divided steps (distance \( \Delta z \)). In the middle two blocks, the obtained phases of the symbols of each span are stored in the memories in the system. This process is performed until counting the number of spans \( N' \) reaches the target span \( N_{\text{span}} \) that is the number of spans required for linear approximation. In the last block, the phases of the memorized symbols are linearly

![Phase transition of a symbol](image1)

(a) Phase transition of a symbol

![SSFM with linear approximation](image2)

(b) SSFM with linear approximation

Fig. 1. Principle of linear approximation
approximated and phase estimation is performed. Phase information for each span is stored in the previous blocks. Linear approximation can be performed using two of these points, and the phase of the symbols at the input point can be estimated. In this report, one of two points is fixed to the first span.

3 Numerical simulation results

Fig. 1(b) circuit is likely to get the great performance of compensating phase noise, however, intensity noise is concerned to limit the performance. We investigated the limit of the proposed scheme performance. Fig. 2(a) shows the performance of linear approximation evaluated by BER under the conditions of transmission length \( L = 5000 \) km, and input power \( P = 0 \) dBm. We considered the number of target spans ranging from 5 spans up to 45 spans when \( N_{\text{step}} = 2, 3, \) and 4. The horizontal dotted line is FEC limit of \( \text{BER} = 10^{-3} \). As can be seen in Fig. 2(a), there is a trade-off between the number of steps, target spans and the compensation performance.

Figs. 2(b) to (d) indicates the number of steps per span as a function of number of spans to reach \( \text{BER} = 10^{-3} \). These results are obtained from transmission distances \( L = 2500, 5000, \) and \( 10000 \) km, input power \( P = 0, 3, \) and \( 6 \) dBm, respectively. For \( N_{\text{span}} = 2, 5, 8, 10, 20, \ldots, 60, N_{\text{step}} \) satisfying \( \text{BER} = 10^{-3} \) are plotted. The circles in the graph is marked when the calculation load \( N_{\text{step}} \times N_{\text{span}} \) becomes the minimum of this technique in each transmission distance and power. For example, \( N_{\text{step}} = 5, N_{\text{span}} = 3 \) are obtained in Fig. 2(b) when \( L = 2500 \) km, and \( 0 \) dBm.
Table I shows the maximum reduction of the calculation load by the proposed method compared with standard SSFM. The minus sign indicates that computational amount increase. For this result, At \( P = 0 \) dBm, this technique has great effect of the computational amount reduction and in the case that the input power is up to around 3 dBm, this technique can be expected to reduce the computational amount of SSFM.

### Table I. The maximum reduction of the calculation load by linear approximation compared with standard SSFM

<table>
<thead>
<tr>
<th>Input power [dBm]</th>
<th>Distance [km]</th>
<th>0</th>
<th>3</th>
<th>6</th>
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<tr>
<td></td>
<td>2500</td>
<td>70.00%</td>
<td>30.00%</td>
<td>−6.70%</td>
</tr>
<tr>
<td></td>
<td>5000</td>
<td>55.00%</td>
<td>40.00%</td>
<td>4.00%</td>
</tr>
<tr>
<td></td>
<td>10000</td>
<td>36.30%</td>
<td>13.30%</td>
<td>−145.50%</td>
</tr>
</tbody>
</table>

4 Conclusion

We numerically assessed the proposed low computational complexity SSFM technique in 28Gbaud RZ-QPSK transmission system. The novel technique is performed by using linear approximation that uses two phase information of symbols in each span for estimating the phase at the transmission point, and reduce the computational amount by not performing the full span calculation with SSFM. The resultant reduction of the computational complexity of SSFM. Particularly, when input power is 0 dBm, this technique achieves reduction of the computational amount considerably. The reduction of computational amount is 70.0%, 55.0%, and 36.3% when transmission distance are 2500, 5000, and 10000 km, respectively.
The film antenna for capsular endoscope

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**Abstract:** Recently, the capsular endoscope that has extra function has been investigated. Although the power consumption tends to increase as the ability increases, it is difficult to increase the capacity of the battery due to the volumetric problem. In this study, we proposed an antenna for both wireless power transmission and image transmission to be mounted on a capsule endoscope. In this paper, we discussed the requirements desired to antennas to be mounted in capsular endoscope.

**Keywords:** capsular endoscope, wireless power transmission, FPC

**Classification:** Antennas and Propagation

**References**


1 Introduction

Recently, various studies of capsular endoscopes have been investigated [1, 2, 3, 4, 5, 6, 7, 8, 9, 10]. Capsular endoscopes are kind of medical device that have about 20 mm length and 10 mm diameter. They are used instead of ordinary endoscopes, because they are able to examine small intestine that have not able to be examined by conventional endoscopes. Capsular endoscopes can be examined while reducing burdens on patients, but they have several problems due to being battery driven. First, batteries installed in capsular endoscope has limited capacity, thus sometimes examination is over with not being enough. Second, harmful matter contained in batteries may leak and hurt to patient. These problems can be solved by wireless power transmission to the capsular endoscope. If capsular endoscope operates by wireless transmission instead of cell batteries, we do not have to worry about improving the driving time of the capsule endoscope and no disaster due to leakage of harmful substances does not occur. Furthermore, it is not necessary to mount a battery in the capsule endoscope, so that the capsule endoscope can be miniaturized.

This paper presents a film antenna made by flexible printed circuit (FPC) for capsular endoscope. By using FPC, the risk of deformation of the antenna can be reduced. Moreover, it is easy to mass-produce it. The antenna works with microwave method wireless power transmission not resonant inductive coupling method nor inductive coupling method. By using microwave method, no coil is necessary against the resonant inductive coupling method or inductive coupling method. Furthermore, the antenna can be used by not only wireless power transmission but also captured images transmission. In the other words, it is unnecessary to mount two kinds of antennas in a capsule, and only one antenna should be mounted. The operating frequency is 433.92 MHz in the ISM band.

2 System

Several antennas have already been proposed [4, 5, 6, 7] for transmission from inside the capsule, but few antennas are intended to receive power. Many of the antennas for power reception uses coils [8, 9, 10], which is not suitable from the viewpoint of miniaturization. In this research, the antenna mounted in capsular endoscope can not only receive the power but also transmit pictures taken in intestine of patient. By radiating power from antennas installed on a body of patient
and receiving by the antenna inside of capsule, capsular endoscope can take picture and transmit pictures taken in a body to outside of the body. By linking with two functions with one antenna, it is possible to be reduced space inside of capsular endoscope. Antennas installed in capsular endoscope should correspond various rotational state since capsular endoscope moves around the inside of body, in order to receive power stably. We designed an antenna with equal radiation characteristics around the axis.

2.1 Internal antenna
We analyzed the characteristic of the antenna put on a simplified human body model by using FDTD method. Power receiving antennas used in wireless power transmission currently using a coil have already been proposed and are mainstream, but antennas for transmitting images from inside of capsule to outside of body are separately required. In addition, although existing antennas for transmitting images are mainly deformed dipole antennas, in this research we adopt a loop antenna. The designed antenna is constructed by loop antenna and open stub structure. Normally, when we design loop antenna, its length should be one wave length, however it is impossible to install such a long element in a capsular endoscope, therefore by using open stub structure, we realized impedance matching. As a result, we succeeded in designing the antenna which is as small size as a thumb nail. By adopting the loop antenna as the basic structure, it becomes possible to receive regardless of the polarization plane of radio waves. In addition, since the shape is annular, radiation in the axial direction of the capsule can be suppressed compared to the dipole antennas, and it is assumed that strong radiation is obtained around the axis of the capsule. The antenna is assumed rolled in and inserted in capsular endoscope for radiating equally around axial rotation. This meets the condition that the antenna installed in capsular endoscope should adapting various rotation because capsular endoscope may be various rotational states.

The antenna model is shown in Fig. 1(a). This antenna is constructed by 0.4 mm thickness metal covered with FPC. Each dimension is shown in Fig. 1(a). The antenna is installed in a model of capsular endoscope shown in Fig. 1(b) constructed by Circuit Case (ABS resin: $\varepsilon_r = 3.0$, $\sigma = 0.0$ S/m), Capsule Case (PTFE: $\varepsilon_r = 2.0$, $\sigma = 0.0$ S/m), and Capsule Dome (acrylic resin: $\varepsilon_r = 3.5$, $\sigma = 0.0$ S/m). Properties of antenna is shown in Fig. 1(c). The portion shown in red is a loop antenna, and the portion surrounded by blue are stub structure. Turn the antenna into a tubular shape and power the point where A and A’ are connected.

2.2 External antenna
In order to convey information taken in intestine to outside of body, it is necessary to install another antenna on the body. In this study, we referred to and improved an antenna proposed by the past study [2]. The size of this antenna is adjusted for use in our research. This antenna is shown in Fig. 1(c). The antenna is constructed by 2 mm width metal on substrate ($\varepsilon_r = 2.17$, $\sigma = 0.0$ S/m). Since the antenna has very wide bandwidth, we adopted for transmission antenna in this study. In the other words, this antenna get less effect from human body than any other type of antenna because it is possible to corresponding to displacement of resonance.
frequency by dielectric constant. Furthermore, the antenna emits circular polarization wave, namely by using the antenna, it is possible to transmit power regardless of the rotational state of the capsular endoscope.

2.3 Whole model

In order to calculate the characteristics of the antenna, we used a simplified human body model instead of real human body. In this study, we used simplified two-layered model constructed by intestine and muscle.

Calculation model is shown in Fig. 2. The human body has simplified shape in order to shorten the calculation time. Calculation model is constructed by only muscle \((\varepsilon_r = 57.7, \sigma = 0.83 \text{ S/m})\) and intestine \((\varepsilon_r = 65.3, \sigma = 1.92 \text{ S/m})\). The external antenna is on gel \((\varepsilon_r = 80.0, \sigma = 0.0 \text{ S/m})\) attached to the simplified human body model.

3 Results

Calculated and measured reflection coefficient of external antenna and proposed internal antenna are shown in Fig. 3(a). The external antenna maintains wideband characteristics even in the vicinity of the human body model in this research. The calculated reflection coefficient of the antenna was \(-8.5 \text{ dB}\). On the other hand,
measured reflection coefficient was $-17.2 \text{ dB}$. Compared with the calculation results, baseline decreases depended on dielectric loss caused by FPC were observed in the actual measurement result, but the tendency was generally consistent. As a factor of errors between actual measurement and simulation, it is conceivable that errors due to electric constant of phantom, soldering and proximity of the coaxial cable and the phantom are considered, the greatest factor is the difference between the antenna model and the real antenna. In model, the film and the metal are arranged on the same plane and the same thickness due to constraints of calculation resource. However, in reality, metal thin films are arranged on the film, and these do not completely match. Since the actual antenna is strongly influenced by the dielectric loss tangent, the reflection coefficient of the measured value is deteriorated. The baseline drop in the measured value is about $-3 \text{ dB}$, and even if this is taken into consideration, the actual antenna is not inferior to the simulation.

Calculated electric field intensity distribution is shown in Fig. 3(b). By the electric field intensity distribution, we observed that the antenna radiates equally around axial rotation of capsular endoscope. It satisfying the condition desired to antennas installed in capsular endoscope.

Calculated transmission efficiency from the external antenna to the internal antenna was 0.7%. For practical use, it is required 2% of transmission efficiency. To solve this problem, we are staying addressed to improve external antenna and devise arrangement of the antenna. The external antenna used in wireless power transmission has no ground plate, thus it radiates both ways to simplified body and opposite. By attaching the ground plate to the antenna, the radiation direction of electric power is restricted. Thus leads to increasing transmission efficiency.
4 Conclusion

We proposed an antenna that can be mounted in capsular endoscope. The antenna is expected not only transmit images taken in digestive organs but also receive power emitted by external of body. We observed that matching calculated reflection coefficient with measured one. As a result of calculation by FDTD method, the antenna radiates power equally around the axis. By wireless power transmission simulation, we acquired 0.7% transmission ratio and reflection coefficient at 433.92 MHz was $-8.5$ dB. External antenna, it is necessary to properties to dissipate to one side. For practical use, it is required further improvement of the external antenna that is needed to dissipate to one side.

Acknowledgments

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A dual-band coil antenna for MHz band on-body wireless body area networks

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Abstract: A dual-band coil antenna is proposed for MHz band on-body Wireless Body Area Network (WBAN) applications by near-field coupling, with relatively low attenuation from the human body. The proposed antenna has two operating frequencies at 48.2 and 57.6 MHz with the measured −10-dB bandwidth 47.9–48.6 MHz and 57.1–58.2 MHz, respectively. Matching circuits are used to reduce the antenna size and tune the resonant frequencies. A human phantom is used for the antenna simulation and measurement. Reflection and transmission coefficients and electric and magnetic field distributions of the proposed antenna are shown, and 1-g and 10-g average Specific Absorption Rate (SAR) and the maximum safety input power are calculated to evaluate the antenna’s radiation safety.

Keywords: wireless body area networks (WBAN), coil antenna, near-field coupling, matching circuit, specific absorption rate (SAR)

Classification: Antennas and Propagation

References


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1 Introduction

In recent years, Wireless Body Area Networks (WBANs) for medical applications are improving rapidly thanks to the advance of integrated circuits and wireless communications. WBAN enables long-term dynamic health monitoring and real-time updates of patient status information [1]. Heart rate, blood pressure, oxygen saturation and other vital signs are sampled and processed by sensors and communicated with external hubs or networks for diagnosis and treatment [2].

WBAN communication links can be divided into off-body, on-body, or in-body links, and different links have different antenna design methods [3]. For on-body links, antennas are typically designed as Planar Inverted-F Antennas (PIFAs) [4], patch antennas [5], folded antennas [6], slot antennas [7], or just electrodes [8]. In order to attach to the human body, the antenna structure should be small-size and low-profile, and its communication performance should be low interference with other antennas [9].

WBAN communications typically use Medical Implant Communication Service (MICS) band, Wireless Medical Telemetry Service (WMTS) band, Industrial, Scientific and Medical (ISM) band, or ultra-wideband (UWB) [10]. However, since the human body is a lossy dielectric material, signal attenuation increases with frequency. Low attenuation can be realized by lowering the antenna resonant frequency, and communication in the MHz band is a possible option.
By using the MHz band impulse-radio (IR) communication system, WBAN communication can realize good performance in the MHz band. In [11, 12], a 10–60 MHz IR transceiver was proposed and used for Electrocardiogram (ECG) based human body and implantable medical communication. Due to the transceiver’s wide transmission band, high data rates and interference immunity can be realized.

In this paper, a dual-band coil antenna is proposed for 10–60 MHz band on-body WBAN applications. The proposed antenna communicates with other antennas via near-field coupling with operating frequencies of 48.2 and 57.6 MHz, as two signal strong frequencies of the IR transceiver. The antenna size is $60 \times 60 \times 1 \text{mm}^3$, can be integrated with the IR transceiver for high data rate communications. The matching circuit is used to miniaturize the antenna and tune the resonant frequency. The ground plane is used to reduce the interference with other devices, since WBAN communications typically use multiple devices. Reflection and transmission coefficients and electric and magnetic field distributions of the proposed antenna at 48.2 and 57.6 MHz are demonstrated, and 1-g and 10-g average Specific Absorption Rate (SAR) and the maximum safety input power are calculated to evaluate the antenna’s radiation safety.

2 Antenna structure and design

Fig. 1(a) shows the proposed antenna geometry and dimensions. The antenna consists of two coils, a ground plane and two matching circuits. Coils and matching circuits are connected through metal via holes (0.5 mm in diameter). The antenna is fabricated on a 1 mm thick FR-4 substrate ($\varepsilon_r = 4.3$; $\tan \delta = 0.035$; copper thickness: 0.017 mm). The antenna structure and values of capacitor are optimized by the frequency domain solver of Computer Simulation Technology (CST) Microwave Studio ver. 2017 SP5 [13]. Due to manufacturing errors and parasitic capacitances, the simulated and the measured capacitor values are slightly different. For the proposed antenna, $C_{S1}$ and $C_{S2}$ change from 5.6 and 6.5 pF (simulation value) to 6.8 and 5.6 pF (fabrication value).

Taking into account the human body’s electromagnetic energy absorption, a human phantom with dimensions of $160 \times 100 \times 48 \text{mm}^3$ is used for simulation and measurement, as shown in Fig. 1(b). The measured human phantom dielectric constant $\varepsilon_r$ and conductivity $\sigma$ are 75.4 and 1.46 S/m. In order to simplify the simulation, the measured electrical properties are used for the simulation model in the entire frequency range.

3 Simulation and measurement results

A prototype antenna is fabricated and photographs are shown in Fig. 1(c), with the measurement environment. Resonant frequencies of the proposed antenna are tuned to 48.2 and 57.6 MHz, as two signal strong frequencies of the IR transceiver. Simulated and measured reflection coefficients of the proposed antenna is shown in Fig. 2(a). The simulated and the measured results are basically the same, and the measured-10-dB bandwidth of two bands are 47.9–48.6 MHz and 57.1–58.2 MHz, respectively.
Fig. 1. (a) Geometry and dimensions of the proposed antenna. $l_1 = 55$, $l_2 = 45$, $l_3 = 35$, $l_4 = 13$, $l_5 = 1$, $l_6 = 1$, $l_7 = 1$, $l_8 = 1$, $l_9 = 60$, $l_{10} = 60$, $l_{11} = 5$, $l_{12} = 16$ (unit: mm); $C_{S_1} = 5.6$, $C_{S_2} = 6.5$ (simulation value), and $C_{S_1} = 6.8$, $C_{S_2} = 5.6$ (fabrication value), $C_{P_1} = 0.5$, $C_{P_2} = 0.5$ (unit: pF) (b) Simulation and measurement environment. $d_1 = 1$ (unit: mm) (c) Fabricated antenna and measurement environment.
Air gap exists between the antenna and the phantom due to the equipment packaging thickness in practical devices [7]. In the measurement, air gap width can be adjusted by covering low dielectric constant plastic film on the phantom. The resonant frequency difference of different air gap width $d_1$ is simulated, and results are shown in Fig. 2(b). It can be seen that increasing air gap width will shift the resonant frequency higher due to the decreasing effect from the human body.

The antenna communication performance is simulated and measured, and results are shown in Figs. 2(c) and (d). Although implantable antennas are not proposed in this paper, and communication performance is evaluated by two identical on-body antennas, compact and low profile implantable antennas for

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**Fig. 2.** (a) Simulated and measured reflection coefficient. (b) Simulated results with different air gap width. (c) Simulated S-parameter by two identical antennas. (d) Measured S-parameter by two identical antennas. (e) Electric field and magnetic field distributions at 48.2 and 57.6 MHz.
MHz band WBAN communications will be designed in the future. As can be seen, measured results are consistent with simulated results. However, due to the weak coupling between coaxial cables at low frequencies, measured transmission coefficients are slightly higher than the simulated results. The cable coupling can be decreased by reducing the length of coaxial cables.

The electric field and magnetic field distributions of $y$-$z$ plane at 48.2 and 57.6 MHz are shown in Fig. 2(e). It can be seen that the communication is mainly carried out by magnetic field and electric field attenuates quickly in the phantom due to the high dielectric constant characteristics of the phantom. Moreover, the radiation of magnetic field at 48.2 MHz is stronger than at 57.6 MHz, which is consistent with the transmission coefficient results in Figs. 2(c) and (d).

SAR is calculated for the evaluation of radiation safety. The IEEE C95.1-1999 standard and the IEEE C95.1–2005 standard restrict the 1-g and 10-g averaged SAR over any tissues to be less than 1.6 and 2 W/kg [14]. The simulated 1-g and 10-g averaged SAR values at 48.2 and 57.6 MHz are calculated with 1 W net input power, as shown in Table I. The maximum safety net input power can be calculated as 0.2 W.

### 4 Conclusion

In this paper, a dual-band coil antenna for MHz band on-body WBAN applications has been demonstrated. The proposed antenna has a compact size of $60 \times 60 \times 1$ mm$^3$, and has two operating frequencies at 48.2 and 57.6 MHz with the measured-10-dB bandwidth 47.9–48.6 MHz and 57.1–58.2 MHz, respectively. Reflection and transmission coefficients are simulated and measured by two identical proposed antennas with 50 mm intervals. The electric field and magnetic field distributions at 48.2 and 57.6 MHz are shown and the 1-g and 10-g averaged SAR and the maximum safety net input power are calculated.

### Acknowledgment

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**Table I.** Simulated maximum SAR (net input power = 1 W), and maximum safety net input power

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Max SAR [W/kg]</th>
<th>Max safety net input power [W]</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1-g avg.</td>
<td>10-g avg.</td>
</tr>
<tr>
<td>48.2 MHz</td>
<td>5.57</td>
<td>2.10</td>
</tr>
<tr>
<td>57.6 MHz</td>
<td>7.94</td>
<td>3.35</td>
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</table>
A low-profile antenna with multi-directional beam pattern using loop elements

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Abstract: A low-profile antenna having four directional beam patterns using loop elements has been proposed. The proposed antenna is vertically polarized and has four directional beams for sensor nodes. The thickness of the antenna is only 3.2 mm (0.11 effective wavelength) resulting in a low-profile structure. Simulated and measured average gain of 8.0 dBi and 5.6 dBi are obtained, respectively.

Keywords: low-profile antenna, vertical polarization, multi-directional beam, wireless sensor network, sensor node

Classification: Antennas and Propagation

References


1 Introduction

Wireless sensor networks have been recently used for such as industries, agriculture applications, medical and home networks. At the same time, applications of electromagnetics for sensing are getting attractive for wireless engineers and industrial society. Usually, the sensor nodes of wireless sensor networks include some integrated circuits, such as micro processor, A/D converter. Therefore, the antennas for sensor nodes require a small size and a simple structure. Several antennas have been recently proposed in [1, 2] for the above purposes for consideration of achieving simple designs and small sizes. However, we should note that most of the proposed antennas have omni-directional radiation patterns for now.

On the other hand, it is important to reduce total energy consumption for enlonging the lifetimes of network systems [3, 4, 5] because some of them uses solar panels and batteries. As a technique for solving this purpose, directive antennas have been proposed [6, 7, 8] in order to save the energy consumption for the system with the directional beam by focusing the radiated energy effectively on the required directions of sensors. However, there have been a new kind of problem in which the number of links are decreased by using directional antennas compare to omni-directional antennas. Therefore, it is important to have a number of directional beams. Moreover, the antennas of previous studies [6, 7, 8] have complex structures and large sizes.

Considering these situations, we propose a directional antenna having a simple and low-profile structure in this letter. Originally, surface wave antennas have been known to have keen beams. Based on this background, some low-profile antennas radiating bi-directional beams by surface wave have been proposed by the authors’ group [9, 10]. Low-profile structure can make antennas be inconspicuous in applications such as avoiding destroying scenery. Being developed from the previous structures, the proposed antenna has a low-profile structure and a radiation beam which is directed to four directions. This behaviors including mechanisms are discussed in the following sections.
2 Antennas structure

Fig. 1 shows the top and side views of the proposed structure consisting of two loop elements, a feeding patch, two dielectric substrates and a ground plane. In this structure, total thickness $h$ is 3.2 mm (0.11 $\lambda_g$, where $\lambda_g$ is the effective wavelength), and the radius of the substrate is 34.5 mm. In the proposed antenna, dielectric substrate is Taconic substrate that has a dielectric constant of $\varepsilon_r = 2.2$ and a loss tangent of 0.001. The 8 mm-square feeding patch is capacitively coupled with the two elements considering impedance matching. The feeding patch is installed between two dielectric substrates. A loop element has a dimension of 1 mm in width. A length of the long side $L$ is 66.79 mm (2.07 $\lambda_g$), and the length of short side $W$ is 11 mm (0.34 $\lambda_g$).

![Antenna structure](image)

(a) Top View  (b) Side View

3 Antenna performances

Fig. 2 shows the simulated and measured results of $|S_{11}|$ characteristics and radiation patterns. According to Fig. 2(a), the simulated resonant frequency is 6.28 GHz, and $-11.68$ dB of $|S_{11}|$ has been obtained. On the other hand, the measured resonant frequency is 6.44 GHz, and measured $|S_{11}|$ at the resonant frequency is $-12.46$ dB. Comparing the simulated and measured resonant frequencies, we can find a discrepancy between them. This reason is probably due to an air gap between two dielectric substrates in the fabricated antenna.

Fig. 2(b) shows simulated and measured radiation patterns at the respective resonant frequencies. As seen from this figure, four beams are directed at $\phi = 0^\circ$, $90^\circ$, $180^\circ$ and $270^\circ$ in the $xy$-plane, and the simulated and measured average gains are about 8.1 dBi, and 5.6 dBi, respectively. The gain of the measured results is lower by 2.5 dB than the simulated results. Similarly, the discrepancy between the simulated and measured resonant frequencies is found. Furthermore, the directional beams are weakly radiated at also $\phi = 45^\circ$, $135^\circ$, $225^\circ$ and $315^\circ$ in the $xy$-plane even though the radiation is attenuated in such directions according to the simulated result. The discrepancies in regard to the gain and the radiation pattern will be discussed in the next section.
4 Analysis

As seen from Fig. 3(a), the electric fields show strong amplitude along the x-axis at four locations. Now, we label the four locations as A, B, C and D, respectively as shown in the figure. Fig. 3(b) also shows the distance in wavelength between each location. This shows that the distance between A and B is 0.87 λg, A and C is 1.18 λg and A and D is 2.04 λg. Therefore, referencing the distances and considering the variation in phase with the locations including A–D in Fig. 3(c), we can understand that the electric fields at respective locations of A–D enhance each other as the radiations from the locations are almost in-phase. We can also have the same understanding along the y-axis. As a result, we can figure out that the four beams are directed in φ = 0°, 90°, 180° and 270° in the xy-plane.

On the other hand, the electric fields shows strong amplitude along the line of the φ = 45° with respect to x-axis at four locations in the Fig. 3(d). We call the four locations i, ii, iii and iv, respectively. The Fig. 3(e) shows the distance between each location of i–iv. As seen from Fig. 3(f), the electric field at respective locations of i–iv are in phase. Since the distance between i and iii is 0.40 λg and that between i and iv is 0.44 λg, we can understand that the electric fields at i–iv
cancel out each other considering that the interval between i–iv are close to \( \lambda_g/2 \). As a result, we can conclude that the directional beams are weakly radiated in \( \phi = 45^\circ, 135^\circ, 225^\circ \) and \( 315^\circ \) in the xy-plane. As discussed here, the diagonal length around the locations should be \( \lambda_g/2 \) to cancel the radiation in these angles. In the present design, we consider that the current design around the structure in Fig. 3(e) is also involved in a good impedance matching.

5 Conclusion

In this letter, we have discussed a low-profile antenna using loop elements for the purpose of reducing the energy consumption and four links for the sensor nodes. Using planar elements, the proposed antenna has a low-profile structure and multi-
directional beams by combining loop elements. The beams are directed at every 90° in $xy$-plane so that high gain can be achieved even though the antenna dimension is relatively small. However, the cross-polarization yields energy loss. Decreasing the cross polarization is a future work.
Field experiment of 400-Gbps transmission in C+L-band over dispersion-shifted fiber

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**Abstract:** Dual-carrier 400-Gbps channels in C+L-band are transmitted over dispersion shifted fiber (DSF) in field for the first time. Distributed Raman amplification not accompanied by EDFA (all-Raman amplification) is used to suppress the nonlinear distortions, which strongly occur in the zero-dispersion region in C-band. The transmitted distance is 200 km and the maximum distance is expected to be 480 km, while the Raman pump power is lower than a safety level for commercial use. The field experiment demonstrates that all-Raman amplification is feasible for effective use of existing DSF links in regional optical networks.

**Keywords:** Raman amplifier, C+L-band transmission, field trial, dispersion shifted fiber, nonlinear effect

**Classification:** Fiber-Optic Transmission for Communications

**References**


1 Introduction

To cope with rapidly increasing data traffic, technologies for beyond-100 Gbps (B100G) transmission have been intensively developed. We recently reported on successful field trials of 400-Gbps transmission [1, 2]. As well as B100G, using multiple bands (e.g. C+L-bands) has attracted much attention as a solution to increase transmission capacity [3, 4]. However, in the case of dispersion-shifted fiber (DSF), which is used in backbone networks in Japan for historical reasons, the use of wavelengths around the zero-dispersion wavelength ($\lambda_0$) is usually avoided because nonlinear effects markedly lower the achievable distance [5]. This is an issue to be addressed for effective use of optical frequency resource with existing DSF links. The distributed Raman amplifier (DRA), which NTT has been commercially deploying in its national backbone network for more than 10 years, is a promising solution because its low-noise characteristic enables low-power transmission and accordingly low nonlinear distortions [6, 7].

In this paper, we report on the transmission of dual-carrier 400-Gbps channels using C+L-bands over 200-km of aged DSF in the field [8]. To the best of our knowledge, this is the first field trial of B100G transmission using C+L-bands over DSF. The C-band channels are chosen to overlap $\lambda_0$, so as to evaluate the worst case. We used DRAs that are not accompanied by erbium-doped fiber amplifiers (EDFAs) as repeaters to minimize the nonlinear distortion.

2 Experimental setup

Fig. 1(a) shows the configuration of the experiment. We used ten 20-km DSF links between two buildings. On the transmitter side, optical signals from C-band transmitters and those from L-band transmitters were multiplexed separately by using wavelength-selective switches (WSSs), amplified separately by using EDFAs, combined by using a C-band/L-band combiner, and launched to the transmission fiber. The signals of each band were comprised of a dual-carrier, dual-polarization (DP)-16QAM 400-Gbps channel for measurement and eight background channels modulated at 200-Gbps DP-16QAM, as shown in Fig. 1(b). The C-band channel wavelengths are ranged from 1544.128 nm (ch1) to 1547.715 nm (ch10) and the
L-band channels from 1572.476 nm (ch11) to 1576.196 nm (ch20) with 50 GHz spacing in ITU-T G.694 grid. The signals were amplified with a DRA unit after every single span. Each span loss was fully compensated by the DRA. The DRA unit consisted of a signal-pump coupler, a pump combiner, pump sources for the C-band (1425 and 1450 nm), and those for the L-band (1460 and 1480 nm). On the receiver side, the optical signal was demultiplexed into the C- and L-bands, amplified by using EDFAs, demultiplexed into channels by using the WSSs, and detected at the receivers.

The span losses of the fibers ranged from 5.4 to 6.3 dB. The average and the standard deviation of $\lambda_0$ were 1546.02 nm (between ch5 and ch6) and 2.2 nm, respectively. The pump power of each DRA was 380 mW in total, which is sufficiently lower than a safety level in our safety guideline. Our safety measures and guidelines on commercial DRA systems are discussed in a previous study [9].

Fig. 1. (a) Experimental setup and (b) channel wavelength arrangement.

3 Results and discussions

In the experiment, we focused on the performances of the C-band signals because the nonlinear distortions at the zero-dispersion region were of interest. The launch power was therefore optimized for ch5 (1545.720 nm), which is the closest to $\lambda_0$. Fig. 2(a) shows the Q-margins to the forward-error correction (FEC) limit of approximately 5.5 dB and the Q penalties after 200-km transmission at different
signal powers. The Q penalty is defined as the difference in Q-factors between back-to-back and after transmission at the same optical signal-to-noise ratio (OSNR). The Q-factors at $-19$ dBm/ch and $-13$ dBm/ch were worse than those in between because of a poorer OSNR and higher nonlinearity, respectively. Though the Q-factor of $-15$ dBm/ch was slightly better than that of $-17$ dBm/ch, we concluded that $-17$ dBm/ch is optimum for pursuing lower Q penalty.

We used the same launch power for the L-band for simplicity, although the optimal power for the L-band would be higher.

We confirmed that the Q-margins were above the FEC limit for all of the channels after 200-km transmission, as shown in Fig. 2(b). Fig. 2(c) shows the optical spectrum after 200-km transmission. There was a slight tilt because we did not use gain equalizers.

![Figure 2](image_url)

**Fig. 2.** (a) Variation of Q-margin and Q penalty for different launch powers (ch5), (b) Q-margins of all the channels after 200-km transmission, and (c) Spectra of DRA-repeated signals after 200-km transmission.

We then measured the Q-factors at intermediate spans by connecting the outputs of even-numbered DRAs directly to the C/L splitter on the receiver side at building A. For comparison, we also measured the signal performances up to
eight spans under the condition that the span loss is compensated by only EDFAs. This was done by replacing the DRAs with C-band and L-band EDFA pairs. In this case, the launch power was optimized to be $-12 \text{ dBm/ch}$. Fig. 3(a) shows the Q-penalty transition along the distance. In this comparison, we measured ch5 for the C-band, which is the closer to $\lambda_0$ of the two carriers of the 400 Gbps signal, to evaluate the worst case. We also measured ch15 as an L-band counterpart. The penalty of ch5 signal repeated with the DRAs was only 0.05 dB/span, while that repeated with the EDFAs was 0.17 dB/span. The nonlinear distortion was markedly suppressed thanks to the low-power transmission with the DRAs. The penalties for ch15 signal were low for both DRA and EDFA cases.

Fig. 3(b) shows the Q-margin transition of ch5 signal along the transmission distance. The Q margin of the DRA-repeated signal after 200-km transmission was 2.3 dB, while that of the EDFA-repeated signal after 160-km transmission was only 1.6 dB. The achievable distances of the DRA and EDFA cases were extrapolated to be 480 and 260 km, respectively.

Fig. 3. (a) Comparison in the Q penalties between C- and L-bands, (b) Comparison in the Q-margin of ch5 between DRA and EDFA repeaters.

The reason for the short reach of the EDFA case is not only the high nonlinearity but also that the noise characteristic of the EDFA is not designed for such a short span length. The gap between the EDFA and the DRA cases can be
reduced if a longer span length is used. The span length of the all-Raman transmission can exceed 20 km but is capped by the allowable maximum pump power. Further consideration is required to determine the practical limit.

We also evaluated the stability of the DRA-repeated C-band signal by measuring the Q-factor variation for 15.5 hours and it was as low as 0.22 dB.

4 Conclusion

In this article, we reported on the transmission of dual-carrier 400-Gbps channels in C+L-bands over 200-km DSF in the field under the condition that the C-band wavelengths fall on the zero-dispersion region. We used all-Raman amplification to minimize the nonlinear effects. The results showed that the Q penalty at the zero-dispersion region was 68% lower than that amplified using EDFAs, and the achievable transmission distance was expected to be 480 km. All-Raman amplification is an effective and feasible solution to optimize the use of existing DSF links.

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Identification of periodic structure target using broadband polarimetry in terahertz radiation

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Abstract: This study demonstrates precision non-destructive inspection of periodic linear targets using polarimetry technique used in radars with frequency bands from P to W (0.3 to 110 GHz) and broadband terahertz pulse. The polarimetry technique is effective to identify targets in radar field, and terahertz waves are effective for non-destructive inspection due to its broadband and special penetration characteristics. In this study, we combine the both techniques to enhance the inspection accuracy. Our analytical and experimental results confirmed that the combination of the polarimetry technique and the broadband terahertz pulse wave was effective for the precision inspection of the targets.

Keywords: terahertz, broadband, radar polarimetry, non-destructive inspection

Classification: Sensing

References

1 Introduction

At present, imaging technology used mainly remote sensing and underground radar has been developed to obtain high resolution and identification of complicated shapes. In particular, polarimetry technique is effective for identification of targets in the field of remote sensing [1]. The technique is a method which transmits and receives two orthogonal polarization components, and identifies targets by the analysis of the components. In general, the technique is used to survey the material and the size of the target, and is carried out by selecting specified narrow band. However, if the analysis by polarimetry is expanded to a broadband, it permits the identifications of the material and the size of the target from the characteristics all over the wide frequency.

We focused on the uniqueness of broadband terahertz sensing, and expand the narrowband polarimetry technique to broadband. In general, terahertz waves have been using for the identification of target materials and communications which transmits enormous data in short range [2, 3, 4, 5]. For the terahertz sensing, terahertz time domain spectroscopy (THz-TDS) which acquires information in a wide band by radiating a short pulse is often used.

In this study, we introduce the radar polarimetry to the THz-TDS measurement, and verified the effectiveness of the method in the precision identification of a target shape.

2 Target identification

In radar polarimetry, the scattering matrix is used for target identification. The scattering matrix $[S]$ represents the scattering state of the target by four components, $S_{HH}$, $S_{HV}$, $S_{VH}$, and $S_{VV}$, and is obtained from the following equation

$$
\begin{bmatrix}
E'_H \\
E'_V
\end{bmatrix} = \begin{bmatrix}
S_{HH} & S_{HV} \\
S_{VH} & S_{VV}
\end{bmatrix}\begin{bmatrix}
E'_H \\
E'_V
\end{bmatrix},
$$

(1)

where $E' = [E'_H, E'_V]^T$ is the electric field vector for transmitting, $E' = [E'_H, E'_V]^T$ is the vector for reception, $H$ and $V$ denote of the horizontal and vertical polarization components, respectively, and $T$ represents the transpose. In addition, the $S_{HV}$ is the scattering coefficient received as the horizontal component when vertical polarized wave is transmitted.

The obtained scattering matrix is converted into the polarization signature to evaluate the target clearly. The polarization signature shows the received power characteristic from the target when the transmitting and the receiving electric field vectors, $E'$ and $E''$, are assumed to be the arbitrary polarizations, and it is possible to identify the rough shape of the target from the signature. When assuming $E'' = E'$, the received power $P$ in the co-polarization is given by

$$
P = |E'' \cdot [S] \cdot E'|^2 = |V|^2,
$$

(2)

where

$$
V = [\cos \epsilon, j \sin \epsilon] \begin{bmatrix}
\cos \tau & \sin \tau \\
-\sin \tau & \cos \tau
\end{bmatrix} \begin{bmatrix}
S_{HH} & S_{HV} \\
S_{VH} & S_{VV}
\end{bmatrix} \begin{bmatrix}
\cos \tau & -\sin \tau \\
\sin \tau & \cos \tau
\end{bmatrix} \begin{bmatrix}
\cos \epsilon \\
-j \sin \epsilon
\end{bmatrix},
$$

(3)

$\tau$ is the tilt angle, $\epsilon$ is the ellipticity angle in the transmitting electric field vectors $E'$. In this study, the polarization signatures calculated by Eq. (2) are used to identify the target shapes.
3 Polarimetric calibrations

On radar polarimetry, three major error factors are well known, the antenna alignment, antenna crosstalk, and channel imbalance. The inaccuracy of the antenna alignment causes the phase difference on the received wave mainly between the different orthogonal polarization components. The antenna crosstalk produces the interference to the other polarization component. So it affects \( S_{HH} \) and \( S_{VV} \). The channel imbalance causes the level difference between the orthogonal polarization components. So it mainly affects \( S_{HH} \) and \( S_{VV} \).

Calibration is performed to compensate the errors. In this study, the difference in the scattering matrix between the measured and theoretical values of the known target is used as the calibration quantities. To decide the calibration quantities, we used a paper printed metallic ink uniformly. Since the scattering characteristic of the paper printed metallic ink is almost the same as that of the metal plate, the components of the theoretical scattering matrix is assumed as \( S_{HH} = S_{VV} = 1 \), and \( S_{HV} = S_{VH} = 0 \). Furthermore, the scattering matrix measured in the printed paper is assumed as

\[
\begin{bmatrix}
S'_{HH} & S'_{HV} \\
S'_{VH} & S'_{VV}
\end{bmatrix} = \begin{bmatrix}
A'e^{j\theta_{HH}} & B'e^{j\theta_{HV}} \\
C'e^{j\theta_{VH}} & D'e^{j\theta_{VV}}
\end{bmatrix},
\]

where \( A', B', C', \) and \( D' \) are the amplitudes, and \( \theta'_{HH}, \theta'_{VH}, \theta'_{HV}, \) and \( \theta'_{VV} \) are the phases of the components in the measured scattering matrix. As \( S_{HH} \) is equal to \( S_{VH} \), and \( S_{HV} \) is equal with \( S_{VH} \) in the theoretical scattering matrix of the metallic plate, the calibration quantities are defined as

\[
\begin{align*}
\phi_{HV} &= \theta'_{HH} - \theta'_{HV}, \\
\phi_{VH} &= \theta'_{HH} - \theta'_{VH}, \\
\phi_{VV} &= \theta'_{HH} - \theta'_{VV}, \quad \text{and} \\
X &= A' - D'.
\end{align*}
\]

These quantities are decided beforehand by preliminary experiments. The scattering matrix applied the calibration quantities, Eq. (5)–(8), is obtained as

\[
\begin{bmatrix}
S_{HH} & S_{HV} \\
S_{VH} & S_{VV}
\end{bmatrix} = \begin{bmatrix}
A'e^{j\theta_{HH}} & (B - B'\sqrt{\sigma})e^{j(\theta_{HH} - \phi_{HV})} \\
(C - C'\sqrt{\sigma})e^{j(\theta_{HH} - \phi_{VV})} & (D - X\sqrt{\sigma})e^{j(\theta_{VV} - \phi_{VV})}
\end{bmatrix},
\]

where \( A, B, C, \) and \( D \) are the amplitudes, and \( \theta_{HH}, \theta_{VH}, \theta_{HV}, \) and \( \theta_{VV} \) are the phases of the components in the scattering matrix measured the target, and \( \sigma \) is the scattering cross section ratio.

4 Measurement setup and target

A landscape of the measurement equipment is shown in Fig. 1(a). The measurements are performed by using a THz-TDS. The distance from both the transmitter and receiver to the target is 150 mm. The incident angle of the wave is 30 degrees from the normal direction of the planar surface, and the receiver is located in the direction of the specular reflection. Also, a linearly polarized pulse wave tilted 45 degrees to the ground is radiated from the transmitter, and is divided into the vertical and horizontal components by a polarizer. The frequency range of pulse is...
from 0.2 to 1 THz. The radiated wave is transformed to the plane wave by a collimator. The vertical and horizontal components of the scattering wave are separately received by placing another polarizer in front of the receiver. The targets are parallel line patterns printed on glossy paper with metallic ink as shown in Fig. 1(b). The target is set in a state that the parallel lines are inclined −45 degrees from the baseline. The line widths are equal to the blank widths. Two types of the line width are tested, one is 0.3 mm and the other one is 0.2 mm.

5 Experimental results

The polarization signatures of the measured co-polarization component at 0.2, 0.4, 0.6 and 0.8 THz were shown in Fig. 2 and Fig. 3. Fig. 2 and 3 show the results when the line widths are 0.3 mm and 0.2 mm, respectively. In Fig. 2(a), the received power is maximum at the ellipticity angle of 0 degrees and the tilt angle of about −45 degrees. Therefore, we can understand that the linear lines are inclined −45 degrees. In Fig. 2(b), it can be seen that the power is maximum at the tilt angle of about −45 degrees. The power is also slightly high at 45 degrees as indicated by the yellow region. In Fig. 2(c) and (d), the power is high at the ellipticity angle of 0 degrees and the tilt angle of from −90 to 90 degrees. Therefore, the target is observed as a metallic plate at 0.6 and 0.8 THz. Even for the same target, the shape identification depends on observing frequency as described above.

In Fig. 3(a) and (b), the powers are maximum at the tilt angle of −45 degrees, so we can understand that the linear lines are inclined at −45 degrees. Furthermore, comparing Fig. 3(b) with Fig. 2(b), the power in Fig. 3(b) is lower than that in Fig. 2(b) at the tilt angle of −45 degrees and the ellipticity angle of 0 degree. The power in Fig. 3(c) is somewhat lower than that in Fig. 2(c) at tilt angle of from −20 to 20 degrees. The signature in Fig. 3(d) shows the same characteristic as that in Fig. 2(d). It is just like the characteristic of a metallic plate.

From the above results, the frequency response of the polarization signature depends on the metal line width. It is considered that the response is related to the metal line width corresponding to the wavelength of the radiated wave. Therefore,
Fig. 2. Polarization signature in target the line width of which is 0.3 mm.

Fig. 3. Polarization signature in target the line width of which is 0.2 mm.
the broadband observation of the signature supplies us new information that does not supply in the narrowband observation.

6 Conclusion

We introduced the radar polarimetry to the THz-TDS measurement, and verified the effectiveness of the method. The measurements using the periodic linear targets were carried out, and it is demonstrated that the observed polarization signature depended on the frequency. Using the frequency dependence of the signature, it becomes possible not only to roughly identify the shape of the target but also the size or the thickness.

Although this study investigated a basic target, the results show that the combination of radar polarimetry and THz-TDS is effective for the precise identification of the targets. In the future, the promotion of the combination will bring the advance of the identification in the terahertz sensing.

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Transmission performance of OFDM-based 1024-QAM in multipath fading conditions

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Abstract: Higher-order modulation is promising technique of 5th-generation mobile systems to increase data rate within a limited bandwidth. This paper presents the transmission performance of the orthogonal frequency division multiplexing (OFDM)-based 1024-QAM in multipath fading propagation conditions by link-level simulation, as well as under static propagation condition. The BER performance is investigated for two types of propagation models, i.e., “extended pedestrian A” and “extended vehicular A” models, specified by 3GPP as parameters of the coding rates. The SNR penalties are also described with respect to the case of AWGN with phase error to meet a BER of $10^{-2}$ for OFDM-based 1024-QAM. Using these results, we validate the application of OFDM-based 1024-QAM to mobile communication systems.

Keywords: mobile communications, 1024-QAM, OFDM, multipath fading, EPA/EVA models

Classification: Terrestrial Wireless Communication/Broadcasting Technologies

References


1 Introduction

In order to improve data rates within a given bandwidth, a straightforward method is the use of higher-order modulation. In LTE and LTE-Advanced systems, quadrature phase shift keying (QPSK), 16-quadrature amplitude modulation (QAM), and 64-QAM are being used for the symbol modulation of orthogonal frequency division multiplexing (OFDM). Furthermore, from Release 12 in the standards body, 3rd Generation Partnership Project (3GPP) defines to support 256-QAM in downlink, although 3GPP introduced 256-QAM as a part of small cell enhancement (SCE) feature [1, 2, 3, 4]. The maximum bandwidth utilization of 256-QAM is in principle eight times that of QPSK, although the higher-order modulation scheme is at the cost of robustness to noise and interference. However, the combination of channel coding and the higher-order modulation, i.e., modulation and coding scheme (MCS), will be more efficient.

SCE and heterogeneous network (HetNet) have been developed to potentially increase system capacity. The small cell approach is to deploy a denser infrastructure that includes support by a low-power evolved Node B (eNB). The cell radius covered by a small cell will be short; therefore, it is expected that such a small cell environment could mitigate the fading impact [1, 5].

Three-dimensional (3D) beamforming has been also considered for enhancing system performance, which can adapt the antenna beam individually for each user equipment (UE) in the elevation domain, i.e., UE-specific elevation beamforming. The 3D beamforming directs the transmitted carrier power toward the target UE, thereby promises to potentially increase the received signal-to-interference plus noise ratio (SINR) while directing less interference to adjacent cells/sectors and other UEs [6, 7]. The use of SCE, HetNet, and 3D beamforming and so on enhances the introduction of a higher-order modulation scheme, since these technologies can increase the received SINR.

Motivated by this observation, in this paper, we focus on much higher-order modulation, i.e., the use of 1024-QAM. We demonstrate the bit error rate (BER) performance of the OFDM-based 1024-QAM with turbo coding in multipath fading propagation conditions [8, 9]. The transmission model of the proposed OFDM-based 1024-QAM and the computer simulation conditions are described in Section 2. In Section 3, we present the BER performance versus received SNR in multipath fading propagation conditions, i.e., extended pedestrian A (EPA) and
extended vehicular A (EVA) models, as parameters of the coding rates. The influence of phase error on 1024-QAM is also provided. Finally, conclusions are summarized in Section 4.

2 Transmission model

Fig. 1 shows a transmission model of the OFDM-based 1024-QAM consisting of a single antenna branch. The channel between the transmitter and receiver is modeled by an additive white Gaussian noise (AWGN) and multipath fading channel. In the transmission side, turbo encoder is applied. This employs parallel concatenated convolutional encoders with a constraint length of 4 and a pseudo-random interleaver. The encoded signals are mapped to 1024-QAM by symbol modulation. OFDM modulation computes the inverse fast Fourier transform (IFFT) of the input QAM signals. Finally, cyclic-prefix insertion is used as a typical OFDM transmission. In the receiver side, a soft decision turbo decoder is used to increase reliability of the decision.

![Fig. 1. Simulation model of the OFDM-based 1024-QAM transmission.](image)

The coding rate is varied within the range from 1/3 to 1. The OFDM bandwidth is set to 20 MHz. QAM signals are mapped by the rule of Gray-code. The main simulation parameters are summarized in Table I. The multipath fading propagation conditions specify the following two profiles, EPA and EVA, whose delay profile represent low and medium delay spread, respectively. The Doppler shift is fixed at 5 Hz.

3 Simulation results

3.1 Signal constellation and OFDM spectrum

Fig. 2 shows simulation results for OFDM-based 1024-QAM transmission in multipath fading conditions. Fig. 2(a) illustrates the signal constellations of the 1024-QAM at the transmitter side, and the OFDM spectrum composed of 1200 subcarriers, in EVA fading propagation conditions.

3.2 Influence of phase error

Fig. 2(b) shows the SNR penalties with respect to the case of AWGN with phase error to meet a BER of $10^{-2}$ for 1024-QAM. It is observed that the SNR penalty increases when the phase error increases, and of course, 1024-QAM is more sensitive to the phase error compared with conventional QAM or QPSK.
As shown in Fig. 2(b), the SNR penalty of 1024-QAM is around 6 dB for the phase error of 0.05 when the coding rate of 3/4 is used. If the SNR penalty has to be less than 2 dB, the allowable phase error is 0.1 for the coding rate of 1/3 or less.

<table>
<thead>
<tr>
<th>Table I. Primary simulation parameters</th>
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<tbody>
<tr>
<td>Symbol modulation</td>
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<tr>
<td>Signal mapping</td>
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<tr>
<td>Turbo encoder</td>
</tr>
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<td></td>
</tr>
<tr>
<td>Turbo decoder</td>
</tr>
<tr>
<td>Bandwidth</td>
</tr>
<tr>
<td>FFT size</td>
</tr>
<tr>
<td>Cyclic prefix length</td>
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<tr>
<td>Number of Tx/Rx antennas</td>
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<tr>
<td>Fading type</td>
</tr>
</tbody>
</table>

| EPA | Doppler shift | 5 Hz |
| | Path delay | 0, 30, 70, 90, 110, 190, 410 ns |
| | Path gain | 0, −1.0, −2.0, −3.0, −8.0, −17.2, −20.8 dB |

| EVA | Doppler shift | 5 Hz |
| | Path delay | 0, 30, 150, 310, 370, 710, 1090, 1730, 2510 ns |
| | Path gain | 0, −1.5, −1.4, −3.6, −0.6, −9.1, −7.0, −12.0, −16.9 dB |

As shown in Fig. 2(b), the SNR penalty of 1024-QAM is around 6 dB for the phase error of 0.05 when the coding rate of 3/4 is used. If the SNR penalty has to be less than 2 dB, the allowable phase error is 0.1 for the coding rate of 1/3 or less.

(a) Signal constellation at transmitter side and OFDM spectrum at receiver side.

(b) SNR penalty versus phase error for 1024-QAM.
3.3 BER performance

Fig. 2(c) shows the BER for 1024-QAM versus the received SNR with the coding rate of 1/2. When the EPA model is used, the required SNR to meet a BER of $10^{-2}$ is approximately 31 dB. Compared with the performance in static propagation condition, the required SNR increases to 6 dB. When the EVA model is used, 1024-QAM can no longer meet a BER of $10^{-2}$, even though SNR is increased.

Fig. 2(d) shows the BER for 1024-QAM versus the received SNR under the EPA model as parameters of the coding rate. The required SNR to meet a BER of $10^{-2}$ for the coding rates of 1/3, 1/2 and 3/4 are approximately 26, 31, and 35 dB, respectively. Compared with the performance in static propagation condition, the required SNR increases around 6 dB.

4 Conclusion

This paper presented the transmission performance of OFDM-based 1024-QAM in multipath fading propagation conditions by link-level simulations. It was clarified that the SNR penalty of 1024-QAM was around 6 dB for the phase error of 0.05 when the coding rate of 3/4 was used. We also showed the BER performance as
parameters of coding rate using two types of propagation conditions, i.e., EPA and EVA. When 1024-QAM with the coding rate of $1/2$ is used in the EPA model, the required SNR to meet a BER of $10^{-2}$ increased to 6 dB compared with that in static propagation conditions.
Wireless network optimization method based on cognitive cycle using machine learning

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Abstract: The wide spread of mobile communication devices has increased the opportunities to use wireless communication technologies, irrespective of one’s geographical location. However communication quality deteriorates due to factors such as competition for scarce radio resources and interference among nearby devices. Cognitive radio technologies have been developed recently to conquer such difficulties. In this paper, we propose a wireless network optimization method using learning algorithms based on a control model known as cognitive cycle. We implement the proposed optimization method in wireless LANs and evaluate the throughput performance. The experimental results show the effectiveness of the proposed approach in a real environment.

Keywords: cognitive radio, wireless LAN, machine learning, optimization

Classification: Wireless Communication Technologies

References

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1 Introduction

Recently, wireless traffic and the number of wireless communication devices has been increasing rapidly. However, the frequency bands suitable for current technologies have already been exploited; thus, the resource is limited. Moreover, the radio environment becomes more unpredictable because of two reasons. Firstly, not only the volume but also the type of traffic is increasing, making the usage of the radio resource more complex. Secondly, distributed wireless networks such as IEEE 802.11 wireless local area networks (WLANs) are widely deployed, and the inner-system and inter-system interactions cannot be predicted easily.

Cognitive radio technologies [1, 2] have recently been developed to improve the radio resource usage of wireless networks under such situations. The basic concept of cognitive radio technology, known as cognitive cycle illustrated in Fig. 1(a), is the adaptation of the behavior of wireless systems through the recognition and learning of the radio environment. Cognitive radio systems observe and recognize the wireless network environment, make reconfiguration decisions, and apply the corresponding action to reconfigure the network. Using this approach, various types of radio parameters can be optimized through appropriate actions. Learning the relationship between the actions and performance is achieved by increasing the number of samples. It improves the precision of the decision making for the best performance.

IEEE Std. 1900.4 [3] defines the basic architecture of such a cognitive radio system. The mobile terminals select the best radio resource to optimize the efficiency of radio resource usage. In order to make an optimal decision, the necessary information can be collected to the network reconfiguration manager. Authors have analyzed the performance optimization of such wireless networks [4]. In [5], the authors introduced machine learning in mobile terminals in order to
optimize the aggregation method for IEEE 1900.4 heterogeneous wireless networks and maximize throughput. In this paper, we extend the method in [5] and apply it to distributed wireless networks. The proposed wireless network can learn the relationship between the network performance and the status of the radio environment through machine learning and can optimize the behavior of the system. Experimental results to verify the proposed method implemented on IEEE 802.11 WLAN are also shown.

2 The concept of the proposed cognitive radio system

A number of literatures indicate that machine learning can improve the performance of wireless networks [6]. In [7], it was shown that a neural network-based channel selection in IEEE 802.11 WLAN access points (APs) can enhance the throughput in an experimental testbed. This approach is very interesting though it only considers each AP independently.

Our proposed method is, in contrast, designed to optimize all mobile terminals (MTs) in the system. Fig. 1(b) shows the concept of the proposed method. Mobile terminals obtain the status and performance of the radio environment, then build the performance model using machine learning which represents the relationship between wireless parameters and the throughput. The information regarding the throughput model is sent to the cognitive controller via access points. The cognitive controller serves as the network reconfiguration manager. It decides the optimal set of wireless parameters for all MTs, and provides these to the MTs via the access points. Mobile terminals receive the wireless parameters and reconfigure their settings.

By repeating the above cycle, MTs accumulate information about the wireless network status and performance. The more training data is collected, the more accurate the throughput model built, resulting in better network performance.
3 Learning and optimization

In order for cognitive radio systems to choose the best wireless parameter values for the current wireless network status, we propose a parameter estimation method using a machine learning algorithm. As shown in Fig. 1(b), the learning algorithm builds the estimation model $y = f(x)$ from the training samples. This model estimates the throughput $y$ from the input parameters $x$. In our proposed method, the training samples are the sets of measured quality of the radio environment ($z$), the parameters of the MTs ($p$), and the measured throughput $y$. The learning algorithm estimates the relationship between $y$ and ($p$ and $z$), represented as $y = f(p, z)$.

3.1 Parameter optimization method using machine learning

We use support vector regression (SVR) as a learning algorithm, similar to the previous research in [5]. The SVR is an analog output version of support vector machines (SVMs) [8]. In SVR, the estimation function $f$ can be expressed as follows [9]

$$f(x) = \sum_{i=1}^{l} (\alpha'_i - \alpha_i)K(x, x_i) + b,$$

where $l$ is the number of training samples, $x_i$ is the input of the training samples ($p$ and $z$), $x$ is an unknown input set for the learning algorithm, and $K$ is a kernel function, respectively. $\alpha_i, \alpha'_i$ and $b$ are unknown parameters which are obtained by the optimization technique proposed in [9], using training samples $p, z$ and $y$.

3.2 Optimization of wireless parameters

In order to decide the optimal set of wireless parameters $p^*$ for the MTs, the cognitive controller solves following optimization problem

$$\arg \max_{p} \sum_{n=1}^{N} \log(1 + f(p_n, z_n)),$$

where $N$ is the number of MTs, $p_n$ is the possible parameter set for MT-$n$, $z_n$ is the current measured quality of the radio environment at MT-$n$, and $f(p_n, z_n)$ is the estimated throughput of MT-$n$ obtained using the throughput model described above. Here, we use the logarithmic utility function of throughput considering fairness among MTs. In this formulation, MTs with lower throughputs have relatively larger gains for the objective function than those with higher throughputs.

4 Experiment and results

We implemented the proposed method in IEEE 802.11 WLAN devices. Experiments with these devices are coordinated in our university laboratory working space.

4.1 Implementation model

IEEE 802.11 WLAN APs and stations (STAs) are operated in infrastructure mode in the 2.4 GHz ISM band. Laptop PCs with Ubuntu 14.04 are used as both STAs.
and APs. In each cognitive cycle, the STA observes the delay and the packet loss ratio through pinging, the received signal strength indicator (RSSI) from its connecting AP using the iwconfig command, the number of packets around the STA using tcptrace command as the link quality (z), and the throughput (y) using the TCP iPerf command. The STA sets the transmission power, channel number (from 1 to 13), and data rate at the physical layer (from 6 to 54 Mb/s) for the current wireless parameters (p).

The STA then builds the throughput model through SVR, and sends information regarding the SVR model to its connecting AP. The AP sends it to the cognitive controller. We have setup one of the APs as the cognitive controller. The cognitive controller calculates the optimal set of STA parameters p*, returns the result to the AP, then the STA gets the result from its connecting AP. In this paper we use particle swarm optimization (PSO) algorithm [10] at the cognitive controller to reduce the calculation costs for solving the optimization problem shown in Eq. (2).

4.2 Settings and results

In the experiment, three APs and nine STAs are operated in channel 1, 6, and 11 in IEEE 802.11g as shown in Fig. 2(a). The operating channel is fixed for each AP. The locations of all APs and STAs are fixed during the experiment. We use uplink TCP throughputs to evaluate the performance since in general uplink traffic makes radio resource usage more competitive in CSMA/CA (note that the number of STAs are larger than that of APs). We also add background UDP traffic of approximately 8 Mb/s on channel 11. To verify the performance of the proposed system, the uplink throughput performance is compared with other algorithms, focusing on the selection of the connecting AP at the STA as follows: (A) selection by RSSI, (B) random selection, (C) selection by radio resource utilization, and (D) to select the number of STAs as equally as possible among channels. In algorithm (A) using RSSI, the STA selects an AP with the highest RSSI. This seems to be the popular method for devices in the market. In algorithm (C), the STA selects the AP of a channel where the minimum number of packets is observed in each cycle. In each algorithm, each cycle runs for 30 s. All STAs start iPerf traffic of 2 s at the same time in every cycle. Before starting the proposed method, the STA observes the radio environment in each channel for 1 h and utilizes this as training data.

Fig. 2(b) shows the moving average throughput by time for each algorithm. The time is expressed as the number of cognitive cycle. The throughput is averaged every 10 cycles (5 min). The proposed method shows greater throughput than other algorithms, indicating that the STAs can select APs effectively.

Fig. 2(c) compares the average throughput per channel among algorithms. Compared to the basic RSSI based algorithm (A), utilization based algorithm (C) shows much higher throughput at channel 6, where it detected as most vacant channel. However, the throughputs at the other channels are much lower. This algorithm is based on the observations of wireless environment but it does not learn, nor optimize the whole system.
In contrast, the proposed method which has a function of learning and optimization shows higher throughput at channel 1 and 6, and lower throughput at channel 11 which has higher background traffic. As a whole, the proposed method can improve network performance. These results indicate that the proposed method can build the appropriate throughput model through learning, and can select the optimized wireless parameters that improves the whole network performance.

5 Conclusion

Wireless communication qualities deteriorate owing to the widespread of mobile devices and limited radio resources. Cognitive radio technologies have been developed to resolve such difficulties. In this paper, we proposed a wireless network optimization method using machine learning algorithm based on the cognitive cycle. We implemented the proposed optimization method in wireless LANs and evaluated the throughput performance. Experimental results showed the effectiveness of the proposed approach in a real environment.

Acknowledgments

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Omni-directional small planar antenna composed of folded slots spanning over both sides

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Abstract: A small planar antenna composed of folded slots is proposed for mobile communications, especially for being mounted on unmanned aerial vehicles (UAVs). Bending slits are made rotationally symmetrically on top and bottom sides of a printed circuit board. The slits are joined in pairs via open edges of the printed board to form slots spanning over both the sides. The directivity is omni-directional and the polarization is dominantly horizontal. The antenna can be controllably miniaturized by being made of a thinner substrate because the resonant frequency decreases as the substrate thickness decreases. Influence of an adjacent conductor plate is examined for the actual usages of being mounted on UAVs and mobile terminals.

Keywords: small antenna, planar antenna, omni-directional, horizontal polarization, symmetry structure, unmanned aerial vehicle (UAV), drone

Classification: Antennas and Propagation

References


1 Introduction

Various applications using unmanned aerial vehicles (UAVs) or drones have been proposed and are expected to make our life easier [1]. Since UAV uses are increasing, the frequencies in 169 MHz band were newly allocated for UAV control in Japan. For wireless remote control of a freely-movable UAV, an omni-directional antenna is required. Since the wavelength of 169 MHz is approximately 1.8 m, compactness is also required for the antenna to be mounted on UAVs [2, 3]. Thus, we proposed a small planar omni-directional antenna. The planar antenna was chosen because it can be mounted inside of or under a UAV. In order to reduce interference with the wireless systems using adjacent frequency bands, the frequency bandwidth in which $|S_{11}| < -10$ dB was set to 400 kHz (the fractional bandwidth is 0.24%) in this paper since the channel bandwidth allocated to 169 MHz band is 400 kHz.

2 Antenna structure

Fig. 1 shows the design of proposed antenna when the number $n$ of slots is four. Bending slits are made rotationally symmetrically on top and bottom sides of a printed circuit board. The slits on both sides are the same with each other, and the slits are united in pairs to form slots spanning over both the sides. Such the method of spanning slots is effective for making slots long [4]. In the previous proposal [5], the slit pairs on both sides are shortly joined via a small open edge, thus the edge of the board is mostly closed. In this proposal, to make slots longer with keeping the antenna size, the edge of the board is mostly opened and each open edge is employed as a part of the spanning slot. The spanning folded slot is indicated by an arrow in Fig. 1(a). Up to approximately one-third of the slot can be consist of the open edge. The slots are fed from the center of the antenna plate by a coaxial cable, for example.

Fig. 1(b)-1 and -2 show the top view of the proposed antenna when $n = 4$ and $n = 2$, respectively and the definition of size parameters. Fig. 1(c) shows the values of size parameters with which impedance matching is attained, where $f_{res}$ denotes the lowest resonant frequency and $\lambda_{res}$ denotes the wavelength at the frequency. A moment method software IE3D and a finite element method software Ansys HFSS were used for optimizing and calculating the antenna characteristics. Type 1 and Type 2 were designed for parametric study with the dielectric substrate thickness $h$ set to 16 mm, and Type 3 was designed for prototype in 169 MHz band with $h$ set to 1.588 mm. It is found that all the designs of $n = 4$ and 2 optimized to attain impedance matching have the common characteristic that slot parts are positioned outer of the antenna plate as the parts are nearer to the slot ends.

3 Parametric study and experimental examination

3.1 Influence of symmetrically arranged slot number $n$

It is found from the table in Fig. 1(c) that $f_{res}$ decreases from 117.6 MHz to 69 MHz when $n$ decreases from 4 to 2. The slot length is 784 mm when $n = 4$ and $\lambda_{res}$ is 2551 mm, thus the slot length is approximately 0.31 $\lambda_{res}$. The slot length is 1587 mm when $n = 2$ and $\lambda_{res}$ is 4348 mm, thus the slot length is approximately 0.36 $\lambda_{res}$.

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Therefore, the approximately half reduction of the resonant frequency should be caused by the approximately doubling of the slot length when \( n \) decreases from 4 to 2. The area, the volume and the weight can be decreased to approximately one fourth when \( n \) decreases to the half because the antenna width \( W \) is decreased to approximately the half in order not to change \( f_{\text{res}} \). Thus, it is effective for making the antenna small and light weight to select \( n \) as a small number.

Fig. 1. Antenna structure and design.
3.2 Influence of substrate thickness \( h \)

Fig. 2(a) shows the antenna characteristics of Type 1 when \( h \) is varied. Fig. 2(a)-1 shows the frequency responses of \(|S_{11}|\) and \( \eta_{rad} \). The frequency resonance is kept to exist even when \( h \) is varied whereas \( f_{res} \) varies. The frequency \( f_\eta \) also varies equally with \( f_{res} \). It is found that \( f_{res} \) becomes lower as \( h \) becomes smaller. This implies that the antenna can be designed smaller when it is made of a thinner “substrate”.

Fig. 2(a)-2 shows the dependences of \( f_{res} \), \( f_\eta \) and \( \eta_{max} \) on \( h \). The resonant frequency \( f_{res} \) is 117.6 MHz when \( h = 16 \) mm, thus \( W = 300 \) mm is 0.118 times of \( \lambda_{res} = 2551 \) mm and \( \eta_{rad} \) is −1.6 dB. The total slot length is 784 mm and is approximately 0.31 \( \lambda_{res} \). When \( h \) decreases to the half, 8 mm, \( f_{res} \) decreases to 0.82 times, 96.3 MHz, and the ratio of \( W \) to \( \lambda_{res} \) decreases to approximately 0.096 whereas \( \eta_{max} \) decreases to −3.0 dB.

The frequency bandwidth \( f_w \) in which \(|S_{11}|\) is below −10 dB is 0.42% when \( h = 16 \) mm. Thus, the target bandwidth 0.24% is covered by Type 1. Fig. 2(a)-3 shows \( h \) dependence of \( f_w \). It is found that \( f_w \) becomes narrower as \( h \) becomes “smaller”.

Fig. 2(a)-4 shows the directivity of the absolute gain \( G_a \) in the \( zx \) plane. The horizontal polarization is approximately 10 dB higher than the vertical polarization, and the gain decreases and the cross-polarization discrimination increases with decreasing \( h \).
(b)-1 Conductor plate inserted in the center between top and bottom sides of the antenna (Cu, thickness is 0.02 mm)

(b)-2 Frequency response of $|S_{11}|$ using $L_O$ as a parameter when $h = 8$ mm, $W = 300$ mm, $L_t = 170$ mm (Type 1)

(c)-1 Simulation and measurement results of frequency response of $|S_{11}|$ (Type 3)

(c)-2 Simulated horizontal pattern of absolute gain $G_a$ at 169.7 MHz (Type 3)

(c)-3 Measured horizontal pattern of relative gain $G_r$ at 170.5 MHz (Type 3)

Fig. 2. Influence of $h$, conductor plate insertion, and experimental result.
3.3 Influence of inserting a conductor plate between surfaces

We investigated the influence of inserting a ring-shaped conductor plate shown in the Fig. 2(b)-1 in the center between the top and bottom surface conductors. Fig. 2(b)-2 shows the frequency response of $|S_{11}|$ when the outer size $L_O$ is varied with the inner size $L_I$ of the ring-shaped plate fixed to 170 mm. The investigated antenna is Type 1 but $h$ is 8 mm. It is found that $f_{\text{res}}$ increases when the inserted plate becomes larger. The resonant frequency $f_{\text{res}}$ increases also when the thickness $h$, which is equal to the distance between surface conductors, increases as shown in Fig. 2(a)-1 and -2. Because electrical interaction between the slot parts on the top and bottom sides decreases in both the cases, the interaction is considered to cause the reduction of $f_{\text{res}}$.

3.4 Experimental examination

The antenna width $W$ of Type 3 is 120 mm which is 0.068 $\lambda_{\text{res}}$, thus Type 3 is electrically small compared with Type 1. Fig. 2(c)-1 shows the frequency responses of $|S_{11}|$ simulated by HFSS and measured. The simulation and measurement results agree well each other. The fractional bandwidth is 0.25% in the measurement result, thus 400 kHz can be covered by the prototype.

The simulated and measured patterns are shown in Fig. 2(c)-2 and -3, respectively. The absolute gain $G_a$ at $f_{\text{res}}$ of 169.7 MHz is shown for the simulated pattern and the relative gain $G_r$ at $f_{\text{res}}$ of 170.5 MHz is shown for the measured pattern. It is demonstrated that the omni-directional directivity is attained by the proposed antenna. The measurement result agrees with the calculation result in that the horizontal polarization is higher than the vertical polarization. Whereas the cross-polarization crimination of simulation result is 19.3 dB, that of the measurement is 8.2 dB. The low crimation of the measurement result should be caused by insufficient electromagnetic wave absorption of the anechoic chamber and electrical influence of feed cables due to the low measurement frequency.

4 Influence of adjacent conductor plate

Whereas the proposed antenna can be set inside of a UAV thanks to its planar shape, antennas are put on some objects in many cases. Therefore, the electrical influence when the antenna is located closely to a plate as shown in Fig. 1(a) was investigated. Whereas UAVs are not generally made of conductor, the case where the plane is made of cupper is investigated as the worst case.

The influence of distance $H'$ from a conductor plate was examined in the condition that the conductor plate is infinite. Fig. 3(a) shows the frequency response of $|S_{11}|$ of Type 1 when $H'$ is varied. It is found that $|S_{11}|$ at $f_{\text{res}}$ first increases, then $f_{\text{res}}$ shifts to higher when $H'$ decreases. It is found that even when $H'$ is decreased to 400 mm (approximately 0.16 $\lambda_{\text{res}}$), the impedance matching of $|S_{11}|$ being less than $-6$ dB is kept at 117.6 MHz which is $f_{\text{res}}$ of the isolated state. Fig. 3(b) shows $H'$ dependences of $f_{\text{res}}$ and $|S_{11}|$ at $f_{\text{res}}$, and Fig. 3(c) shows $H'$ dependences of $f_{\eta}$ and $\eta_{\text{max}}$. It is found that not only $|S_{11}|$ but also $\eta_{\text{max}}$ deteriorates when $H'$ decreases and $\eta_{\text{max}}$ reduces to $-4.7$ dB when $H'$ is 200 mm. When $H'$ decreases less than 200 mm, $f_{\text{res}}$ and $f_{\eta}$ steeply increases.
5 Conclusion

An omni-directional antenna was proposed to be mounted in or under UAV for remote control and its basic performance was experimentally demonstrated. The antenna is planar and small of approximately 0.1 wavelengths of the resonant frequency. The compactness is achieved by spanning folded slots over both sides of the antenna plate. Its polarization is dominantly horizontal. The resonant frequency decreases when the thickness $h$ becomes small whereas the radiation efficiency and the frequency bandwidth decrease. Thus, the proposed antenna can be designed smaller when it is made of a thinner substrate.

It was found that the impedance matching of $|S_{11}| < -6 \text{ dB}$ can be maintained even when the antenna is located closely to a conductor plate up to approximately 0.16 wavelengths of the resonant frequency. Because the dominant cause of the degradation of impedance matching is the increase of the resonant frequency, it is effective to design the resonant frequency lower to cancel the increase.

The electrical interaction between slot parts on the top and bottom sides is considered to be the cause of the resonant frequency decrease. Clarification of its mechanism is a future work. The enhancement of the impedance matching frequency bandwidth is regarded as an important subject of investigation to extend the application scope.