Residual dispersion equalization using correlation detection in Nyquist OTDM scheme

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Abstract: We propose a wavelength-dispersion equalizing scheme that counteracts the signal degradation caused by residual dispersion in Nyquist optical time-division multiplexing. To counteract the signal degradation, this scheme uses optical correlation receivers and a distorted reference signal. At a Q factor of 6.4 dB, the permissible values of the residual dispersion increased from 14.5 ps/nm to 99.5 ps/nm at a baud rate of 160 Gbd.

Keywords: OTDM, residual dispersion, correlation detection

Classification: Fiber-Optic Transmission for Communications

References


1 Introduction

A Nyquist optical time division multiplexing (OTDM) scheme can realize low inter-symbol interference and high spectral efficiency with a high baud rate [1]. To achieve a high optical signal-to-noise ratio (OSNR) tolerance with high spectral efficiency, we previously proposed a Nyquist OTDM scheme based on correlation detection [2, 3, 4, 5]. However, the signal spread caused by residual dispersion increases with increasing baud rate [6]. The signal spread causes inter-symbol interference that degrades the signal quality. Therefore, a precise dispersion compensation is needed to suppress the signal degradation.

This paper proposes and investigates a residual-dispersion equalization scheme that detects distorted optical signals with residual dispersion using a distorted reference signal and optical correlation receivers [7]. This scheme reduces the required accuracy of the dispersion compensation. We then investigate the permissible value of the residual dispersion by numerical simulation.

2 Principle of residual dispersion equalization

Fig. 1 shows a Nyquist OTDM scheme using correlation detection. A tributary signal is generated by an optical impulse train source and an optical modulator. The spectrum of the modulated signal is limited by an optical root-raised cosine (RRC) filter with a 3-dB bandwidth of $1/T$, where $T$ is the time slot of the multiplexed signal. The tributary signal is multiplexed by an OTDM multiplexer [5]. The chromatic dispersion, which distorts the multiplexed signal in the single mode fiber (SMF), is compensated by a dispersion compensation fiber (DCF). The mismatch between the SMF and DCF causes a residual dispersion that distorts the received signal. The spectrum of the received signal is given by

$$S_{RX}(\omega) = S_{OTDM}(\omega) \exp \left\{ - \frac{j \pi c R_d}{\omega_0^2} (\omega - \omega_0)^2 \right\}, \quad (1)$$

where $S_{OTDM}$, $c$, $\omega_0$ and $R_d$ are the multiplexed signal before transmission, velocity of light, carrier frequency, and residual dispersion, respectively [6]. The lowercase $s$ and uppercase $S$ denote a signal in the time and frequency domains, respectively. To simplify the discussion, we ignore the loss and Kerr effects in the optical fibers. An optical correlation receiver, which comprises optical 90° hybrid and two balanced photo receivers (BPRs) with integrators, de-multiplexes and detects the real and imaginary parts of the received signal. The detected signal is given by

$$s_{OUT}(t) = C_{BPR} \int_{t-\frac{MT}{2}}^{t+\frac{MT}{2}} s_{RX}(\tau) s_{REF}(\tau) d\tau, \quad (2)$$
where $s^*_\text{REF}$ denotes the complex conjugate of the reference signal. When the multiplicity is sufficiently high and the influence of adjacent signals can be ignored, Eq. (2) can be approximated by an infinite time integral using Parseval’s theorem:

$$s_{\text{OUT}} \approx C_{\text{BPR}} \int_{-\infty}^{\infty} s_{\text{RX}}(\tau)s^*_{\text{REF}}(\tau)d\tau$$

$$\approx C_{\text{BPR}} \int_{-\infty}^{\infty} S_{\text{RX}}(\omega)S^*_{\text{REF}}(\omega)d\omega,$$

where the constant $C_{\text{BPR}}$ includes the photo-current sensitivity of the photo receivers and the gain of the trans-impedance amplifiers. Previously, a conventional reference signal was generated by an impulse train source and transmitted by the same optical RRC filter. The residual dispersion induces inter-symbol interference and signal amplitude reduction.

To suppress the degradation caused by residual dispersion, we propose an equalization scheme that counteracts the residual dispersion using the distorted reference signal with the same residual dispersion. The spectrum of the distorted reference signal is given by

$$S_{\text{REF-D}}(\omega) = S_{\text{REF-RRC}}(\omega) \exp\left\{-\frac{j\pi c R_d}{\omega_0^2} (\omega - \omega_0)^2\right\},$$

where $S_{\text{REF-RRC}}$ is the conventional reference signal using the optical RRC filter. The distorted reference signal can be generated by an optical short-pulse source and an optical pulse-shaping filter [8]. From Eqs. (1), (3) and (4), the detected signal with the distorted reference signal is expressed as:

Fig. 1. Nyquist OTDM scheme based on correlation detection.
\( s_{\text{OUT}} \approx C_{\text{BPR}} \int_{-\infty}^{\infty} S_{\text{RX}}(\omega) S_{\text{REF-D}}^*(\omega) d\omega, \]

\[
\approx \int_{-\infty}^{\infty} S_{\text{OTDM}}(\omega) \exp \left\{ -\frac{j \pi c R_d}{\alpha_0} (\omega - \omega_0)^2 \right\} \times \exp \left\{ j \pi c R_d \frac{\omega^2}{\alpha_0} \right\} d\omega.
\]

The distorted reference signal can counteract the residual dispersion effect. In practice, the temporal integration limits are finite [5], and the waveforms of the received and reference signals are broadened by residual dispersion. The finite integral time and signal broadening might degrade the signal quality.

### 3 Relationship between residual dispersion and Q factor

We investigated the permissible range of the residual dispersion under the influence of finite integral time and temporal broadening of the signals. The signal generation was modulated by binary phase-shift keying, and the pattern length of the pseudo-random bit sequence was \(2^9 - 1\). The signal wavelength was 1550 nm. The roll-off factor, multiplicity and baud rate were 0.1, 16 and 16 \(\times\) 10 Gbd, respectively. The OSNR of the received signal was 30 dB. The noises of the BPDs and the integrators were ignored.

Fig. 2 shows the relationship between the residual dispersion and the quality factor (Q factor) using the conventional reference signal \(S_{\text{REF-RRC}}\) and the proposed reference signal \(S_{\text{REF-D}}\). To investigate only the effect of the residual dispersion, the fiber loss and Kerr effects in the transmission line were ignored. In this case, the Q factor depended on the absolute value of the residual dispersion. The Q factor of the proposed reference signal was reduced by the limited integral time \(MT = 100 \text{ ps}\) and the temporal expansion of the signals by residual dispersion. In an ideally matched filter with zero residual dispersion, the theoretical upper limit of the Q factor is 18.92 dB. In the proposed scheme, the permissible values of the residual dispersion at a 6.4-dB Q factor increased from 14.5 ps/nm to 99.5 ps/nm. The Q factor of 6.4 dB was the forward error correction (FEC) limit of the triple-concatenated FEC in soft decision decoding [9].

Fig. 3 plots the relationship between the residual dispersion and the Q factor of signals distorted by Kerr effects in the transmission fiber. The nonlinear parameter \(\gamma\), fiber loss \(\alpha\), length \(L\), and dispersion coefficient \(D\) were 2.0 W\(^{-1}\) km\(^{-1}\), 0.2 dB/km, 100 km, and 17 ps/km-nm, respectively. In both the conventional and proposed schemes, the Q factor degraded with increasing average launch power \(P_{\text{in}}\). This degradation was caused by Kerr effects, which distort the signals in the absence of residual dispersion. At a Q factor of 6.4 dB, the permissible values of the residual dispersion within the \(P_{\text{in}}\) range 0–13 dBm increased from 14.2–14.5 ps/nm in the conventional method to 98.4–99.5 ps/nm in the proposed method. The improvement was that expected after removing the residual dispersion, leaving only the Kerr effect.
4 Conclusion

We proposed a scheme that equalizes the residual dispersion in OTDM, and thereby suppresses signal degradation. The Q factor of the detected signal was improved by distorting the reference signal with the same residual dispersion. We also investigated the effect of finite integration time and temporal broadening of the signals due to residual dispersion. At a Q factor of 6.4 dB, the proposed scheme increased the permissible value of the residual dispersion from 14.5 ps/nm to 99.5 ps/nm at the baud rates of 16 x 10 GBd. The same improvement was observed with and without signal distortion caused by Kerr effects.

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Optical-fiber-connected passive primary surveillance radar for aeronautical surveillance

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Abstract: A new radar system using a radio over fiber (RoF) is proposed. The proposed system is optical-fiber-connected passive primary surveillance radar (OFC-PPSR), which is based on a passive bistatic radar approach and uses RoF technology. A separate receiver unit uses the waves scattered from aircraft and the radar reference data transmitted by the RoF. The reference data include the radio frequency signals of the transmitter unit and the processing data of the controller unit, such as radar rotation angle. We first present the principles of OFC-PPSR and the experimental system, which was deployed at the Sendai airport in Japan. Moreover, we present some preliminary experimental results obtained with the proposed system. The proposed system is capable of detecting moving aircraft, as demonstrated by a comparison of the experimental results with real surveillance data.

Keywords: primary surveillance radar, multistatic primary surveillance radar, radio over fiber, aeronautical surveillance, air traffic management

Classification: Sensing

References

1 Introduction

Airport surveillance radar typically include both a primary surveillance radar (PSR) [1] and a secondary surveillance radar (SSR) [2]. Since the SSR uses the reply signals from an aircraft and provides the aircraft’s position, identify and altitude, it has become the main surveillance system in air traffic management. On the other hand, PSRs play an important role as backup and in improving the security of operations, because it uses the waves scattered by aircraft and is a type of independent noncooperative surveillance [3]. However, the update and detection rates of PSRs are lower than those of SSR technologies. Therefore, PSR application technologies are required to improve operational security.

Recently, multistatic primary surveillance radar (MSPSR) [4] has been expected to be used as a conventional PSR alternative. One interesting property is the selection of some signal sources, e.g., present radar signals, digital terrestrial television broadcasts, mobile communication (e.g. 3G and LTE), global navigation satellite system, and so on [5]. Our final goal is to develop a combined surveillance system using several signals. As one core technology of MSPSR, we consider passive radar using the PSR signal. The purpose of this system is to expand the present PSR coverage and to contribute to the spectral efficiency. The present PSR coverage is about 60 NM, and the required detection rate is about 70%. Since PSR depends on the waves scattered from obstacles, undetected area exists, e.g., aircraft at low altitude and the shadowed areas behind mountains or buildings. To overcome this problem, we propose optical-fiber-connected passive PSR (OFC-PPSR) to expand the present PSR coverage. One of the strengths of this proposal lies in its use of radio over fiber (RoF), which enables radio frequency (RF) signals to be transmitted to a separate receiver over a long distance by an optical fiber. Consequently, OFC-PPSR is capable of operating in the same manner as conventional PSR. In addition, because a receiver unit is connected to a transmitter unit by RoF, the receiver always computes the target position, even if the incident waves cannot be detected. The signal-to-noise ratio (S/N) would also be better than that using incident waves propagating in the atmosphere. Moreover, as one application, the existing infrastructure can be shared by collaborating with other surveillance systems such as multilateration (MLAT).

In this letter, the system concept and its operating principle are described first. Then, a prototype system deployed at Sendai airport is introduced. Finally, we show the experimental results. It is shown that the proposed system can detect aircraft through comparison with real surveillance data.

2 Optical-fiber-connected passive primary surveillance radar

In general, a radar has a transmitter unit combined with a receiver unit [1]. As the receiver unit always receives the information of the transmitted signals (transmitted
timing, antenna rotation angle, etc.), estimation of the target position is relatively easy. However, in a passive radar system, the separate receiver unit does not have this information. Hence, it requires some signal processing [6] and a mechanism to estimate the transmitted waves. In order to overcome these problems and to simplify the system, we propose a new radar concept based on passive bistatic radar [7].

The proposed system employs RoF technology, which enables the transmission of RF signals over long distances in comparison with a coaxial cable. An OFC-PPSR receiver unit is connected to the transmitter at a radar site by an optical fiber. Thus, OFC-PPSR can stably use the original RF signals as a reference. Hence, the receiver unit can be located far from the transmitter unit, even if the directed waves do not arrive at the receiver side. Consequently, similar to the present radar, the receiver unit can easily estimate aircraft positions by using the transmitted timing, radar rotation angles, trigger and others. Given that the proposed system is capable of using scattered waves that do not return to the radar site, it is expected to be capable of expanding the coverage area of the current PSR. Moreover, owing to RoF, the S/N is expected to be improved in comparison with the use of incident waves propagating in the atmosphere. It should be noted that sharing the same infrastructure with other surveillance systems such as MLAT is an advantage. Therefore, OFC-PPSR is expected to be used as a PSR distributed surveillance system, in correspondence with the common use of SSR applications.

Fig. 1 shows the system conceptual diagram of the proposed system, and an ellipsoidal curve illustrating the principles of passive bistatic radar (PBR). An RoF transmitter unit is located at the radar site, and it collects the RF signals transmitted by a rotating antenna and some information in the controller unit, and they are provided to the separate receiver unit by RoF. On the other hand, the receiver unit consists of a receiving antenna for the scattered waves, a preamplifier, a down-converter, a signal processing unit, and an RoF receiver unit.

The estimation procedure is summarized as follows:

- Adjust the signal delays corresponding to the optical fiber length from the transmitter unit to the receiver unit
- Measure the RF signals (1. scattered waves from aircraft, 2. RF signals transmitted by RoF) and collect radar information (1. timing when PSR faces north, 2. RF transmitted timing)
- Analyze the bistatic ranging from the time difference of arrival between the radar transmitted timing and the waves scattered from aircraft

In the above procedure, the signal delay \( \tau \) is given by

\[
\tau = \frac{L_1 + L_2 - L_0}{c} \quad [s]
\]

where \( L_1 + L_2 \) is the total distance from the source to the receiver via obstacles, \( L_0 \) is the direct distance from the source to the receiver, and \( c \) is the velocity of light. These relations are shown in Fig. 1b. In the proposed system, \( L_0 \) is modified by the optical fiber length and the source and receiver positions. \( \tau \) is computed by the receiver unit, as mentioned above. However, since \( L_1 \) and \( L_2 \) are unknown parameters, one of them is required to obtain a solution. \( L_1 \) is computed by...
\[ L_1 = \frac{\Delta(\Delta + 2L_0)}{2L_0(1 - \cos \theta + \Delta/L_0)} \] (2)

where \( \Delta \) is defined by \( \Delta = c\tau \), and the angle between the directions of the radar and the target is given by \( \theta \). The result computed for \( L_1 \) (or \( L_2 \)) is an ellipsoidal curve.

3 System deployment and experimental results

A prototype system has been developed and deployed at Sendai airport in Japan, and preliminary experiments were performed to confirm the basic operation of the OFC-PPSR system. Fig. 2 shows the experimental environment and setup. In this experiment, a training radar was used; it is located at the southern part of the airport where the RoF transmitter unit is also located. A general PSR produces an asymmetric beam shape known as a fan beam, and the vertical plane is the cosecant-squared elevation pattern. The RoF transmitter unit is connected to the RoF receiver unit located at the western part of the airport by an optical fiber.

The PSR emits a short pulse of 1 \( \mu \)s and a long pulse of 80 \( \mu \)s. Long pulses are modulated by FM chirp. Since the long pulse is emitted after transmitting the short pulse, the coverage area of the short pulse is limited by the time interval between the short and long pulses. The frequency is assigned to the S band. The distance from the radar site to the OFC-PPSR receiver is \( \sim 1800 \) m. The signal processing
unit separately analyzes both pulses, and a standard horn antenna is selected as the receiver.

Fig. 3 shows the experimental results. This is illustrated by overlapping five scans (= 20 s) of data, and the figure is converted from the estimated bistatic ranging to the PPI scope. In this figure, there are large echoes on the left side. We confirmed that the echoes are located at Mt. Zao on the map. On the right side of the airport, waves reflected from a moving object are also observed. In order to check whether the moving object is an aircraft, we compare the experimental results with the real surveillance data obtained by automatic dependent surveillance-broadcast (ADS-B). The right side of Fig. 3 shows a magnification and comparison. The ADS-B tracks are indicated by red circles. The positions between the echoes from the moving object and ADS-B are slightly different. However, as ADS-B has some errors and the results obtained by the proposed system disregard

Fig. 2. Experimental environment.

Fig. 3. Experimental results in comparison with ADS-B tracks.
the height, the echo of the moving aircraft and the ADS-B tracks are almost identical. Therefore, we conclude that these echoes represent a moving aircraft.

4 Conclusion

In this letter, we proposed OFC-PPSR as a new radar system by using RoF. One of the characteristics is the use of original RF signals at a separate receiver unit. This results in the same operation as conventional PSR, even if the receiving antenna is in the non-line-of-sight of the transmitter. Moreover, S/N of reference signal with RoF is higher and stable than that with monitoring antenna to receive the signal from PSR. Experiments were performed at Sendai airport in Japan. It was shown that the proposed system detects the echoes from a mountain and moving object. Comparing the experimental results with ADS-B, it was demonstrated that the moving object was an aircraft.

The proposed system is expected to expand the coverage area of present aeronautical surveillance systems. However, all signal processing in the proposed system was disregarded. Our future work will consider some radar signal processing techniques such as moving target indication (MTI) and constant false-alarm rate (CFAR) in order to suppress unnecessary signals from fixed structures. It will be our future work.

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High-resolution geomagnetic observation system using HTS-SQUID

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Abstract: Our research group reported successful observation of “co-
faulting” Earth’s magnetic field changes because of piezomagnetic effects
caused by earthquake tremors during the 2008 Iwate-Miyagi Nairiku earth-
quake of M7.2 using a geomagnetic observation system with flux-gate
magnetometers. This is an important finding: electromagnetic fields prop-
gate from a source to an observation site at the light speed in the crustal
materials. Further earthquake detection efforts can lead us to a new system
for super-early warning of earthquake detection with the geomagnetic signal.
However, the observed result with the earthquake was suggested that the
geomagnetic field change accompanying fault movement, whose sources are
the piezomagnetic effects, is very small and short term. Therefore, to develop
an extremely important high-resolution magnetometer system, we first con-
ducted long-term precise geomagnetic observations using a high-temperature
superconductor based superconducting quantum-interference device (HTS-
SQUID) magnetometer system. The HTS-SQUID magnetometer system had
never been used for high-resolution geomagnetic observations outdoors.
Since March 2012, we have observed geomagnetic field using the HTS-
SQUID magnetometer at Iwaki observation site (IWK) in Fukushima, Japan.
Comparison between the introduced HTS-SQUID magnetometer and con-
vventional flux-gate clarified that the HTS-SQUID magnetometer in our
system has higher resolution of magnetic field observation.
Keywords: HTS-SQUID, magnetometer, geomagnetic observation, high-resolution

Classification: Sensing

References


1 Introduction
1.1 Background

For decades, researchers have studied the seismomagnetic effects [1, 2, 3]. Their reports describe that the surrounding magnetic field changed before and after earthquake occurrence.

From our continuous observations, our research group also reported a successful result which is “co-faulting” Earth’s magnetic field variation due to piezomagnetic effects caused by earthquake rupturing (i.e., earthquake-piezomagnetic effects) in 2008 Iwate-Miyagi Nairiku earthquake of M7.2 [4].

However, the magnetic field changes are very small variations of 300 pT [4]. Our successful result is observed by geomagnetic observation system with a flux-gate magnetometer (10 Hz sampling rate) and a synchronized accelerometer.

The magnetic field propagates from the sources to the observation point at the light speed. Therefore, our result suggested the possibility of earthquake detection from magnetic signal. If we can detect an earthquake from magnetic field, we are
able to early warn the occurrence. Further efforts could lead us to a new system for super-early warning of earthquake detection with the magnetic signal [4].

1.2 Magnetometer for geomagnetic observation
In general geomagnetic observation, it might be said that the sampling frequency of 1 Hz is sufficient for observation, and that it is not important to observe with higher accuracy.

However, our successful observation of the Iwate-Miyagi Nairiku earthquake, the magnetic field variations had continued to change only several seconds from the earthquake occurrence time. Therefore, our system (10 Hz sampling rate) recorded magnetic field variations of only several tens of data points.

Additionally, because the highest magnetic field resolution of a flux-gate magnetometer is greater than 10 pT, such a conventional magnetometer is not suitable to observe a small signal by the earthquake-piezomagnetic (EQ-piezomagnetic) effects.

Incidentally, a LTS-SQUID magnetometer using liquid helium is considered to be more accurate as an element of sensor. Nevertheless, it is not promising for continuous long-term geomagnetic observation because it requires liquid helium. That is, the long-term geomagnetic observation demands a magnetometer which not only is easily operated in the outdoor field but also has very high resolution.

2 Geomagnetic observation
2.1 Observation system using HTS-SQUID magnetometer
Our research group has developed a new geomagnetic observation system with low running cost and higher resolution: we introduce high-temperature-superconductor based superconducting-quantum-interference-device (HTS-SQUID) as a magnetometer for a long-term geomagnetic observations. The HTS-SQUID magnetometer has much lower running costs than LTS-SQUID because it uses liquid nitrogen to maintain a superconducting state. This is the most important point. It is the first trial in the world that we had adopted the HTS-SQUID magnetometer for a geomagnetic observation system.

The sampling-frequency of our magnetometer system is 50 Hz (0.02 s) which are higher sampling frequency than our conventional observation system using a flux-gate. Our system observed the orthogonal three-vectors of geomagnetic field vibration \( (H_x, H_y, H_z) \). The clock of this system is synchronized with a GPS signal. These observed data are uploaded to the web server through the mobile network.

2.2 Observation at IWK site
Since March 2012, we have observed three geomagnetic field components using a HTS-SQUID magnetometer at Iwaki observation point in Fukushima, Japan. Here, Iwaki was assumed as an area affected by aftershocks of The 2011 off the Pacific coast of Tohoku Earthquake. Figure 1(a) shows the location of IWK site. The map (b) is a zoom up view of (a). Here, we use Generic Mapping Tool (GMT) for making the map of this figure [5, 6, 7, 8].
Figure 1 also depicts Kakioka Geomagnetic Observatory (KAK), operated by the Japan Meteorological Agency. The distance separating IWK and KAK is about 100 km. Geomagnetic observation results obtained at KAK are used as reference data.

As an example of the result observed by the system at IWK, Fig. 1(c) shows the geomagnetic signal on March 17th 2015. The geomagnetic signal at KAK [9] is also depicted in this figure. Here, the sampling frequency of KAK data was 10 Hz.

The daily K-index on March 17, 2015 was 5, where the K-index quantifies disturbances in the horizontal component of earth’s magnetic field with an integer [9].

From this figure, we can confirm that geomagnetic signals using the HTS-SQUID magnetometer at IWK are very valid by a comparison of waveforms obtained at the two observation points (IWK and KAK). It is reasonable that two waveforms are similar, because the geomagnetic longitude of two observation points is almost same.

3 Results

3.1 Geomagnetic field observed by HTS-SQUID magnetometer

We show the three orthogonal components of the geomagnetic field observed using the HTS-SQUID magnetometer in Fig. 2, where the DC offset of the data is adjusted for easy visibility. The signals are observed on October 28th 2015, whose daily k-index is the smallest 0 over this year. Hence, geomagnetic variation itself of this day was very small.
For subsequent evaluation, we use observation data of 18:00–19:00 (UTC) corresponding to the middle of night of Japan (i.e., relatively much lower artificial noise period) in Fig. 2(b). For comparison to a conventional magnetometer, the flux-gate observation result at IWK is also portrayed in this figure. The sampling frequency of the setup flux-gate magnetometer is 10 Hz.

![Graph showing observation result at IWK site]

**Fig. 2.** Observation result at IWK site

### 3.2 Evaluation

To evaluation the magnetic field resolution of the system, the mean absolute deviation (MAD) is used as an evaluation value as

\[
MAD = \frac{1}{n\Delta t} \sum_{i=0}^{n} |H_i - \bar{H}_i| \Delta t
\]

where \(\bar{H}_i\) represents the mean of \(H_i\).

Figure 3 presents results of calculating the MAD for each of the three-components of the two magnetometers (HTS-SQUID and Flux-Gate) using the data observed in 18:00–19:00UT on October 28, 2015. As shown in Fig. 3(a1)–(a6), we calculated the MAD every second because the observation target signal by the EQ-piezomagnetic effect changes during a few or several seconds. Additionally, the average of MAD per hour is displayed in Fig. 3(b).

As shown in Fig. 3, the deviation of signals observed using the HTS-SQUID magnetometer is smaller than that of the flux-gate magnetometer. Results show that the HTS-SQUID magnetometer in our system has higher resolution of magnetic field observation than the usual flux-gate magnetometer. The HTS-SQUID magnetometer in our system provides geomagnetic field observations with a high accuracy and a high sampling rate.
4 Conclusion

For this study, we develop a geomagnetic field observation system with high accuracy and high sampling rate. In this first trial, the HTS-SQUID magnetometer was used for a continuous long-term geomagnetic observation.

Comparison of the proposed magnetometer and a conventional flux-gate clarified that the HTS-SQUID magnetometer we used has higher resolution of magnetic field observations. The deviation of the HTS-SQUID magnetometer is about 3 pT to 6 pT. We can say that this value is extremely small rather than conventional magnetometer system.

Future task is to accumulate the geomagnetic field signals accompanying the EQ-piezomagnetic effect using the HTS-SQUID magnetometer and to reveal that effect.

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Implementation and user testing of personal authentication having shoulder surfing resistance with mouse operations

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Abstract: Typically, typing character strings on a keyboard is used for personal authentication for PC login and unlocking. Although some graphical and biometric-based methods have been developed, most of them have weak authentication strength, weak shoulder surfing resistance, or other drawbacks. In this paper, we propose a personal authentication method that employs mouse operations in which the mouse itself does not need to be moved. Thus, the user can hide the mouse during authentication, so the method has shoulder surfing resistance and can be used in public places. We performed user testing to validate the proposed method.

Keywords: personal authentication, mouse operation, shoulder surfing resistance

Classification: Multimedia Systems for Communications

References

1 Introduction

Currently, many people use character-based passwords entered by keyboard for personal authentication in public places, such as a classroom of office to unlock a PC or sign into a service on the user’s own PC or a public PC. This method in these situations risks leakage of passwords by shoulder surfing.

SECUREMATRIX [1] provides a graphical password as positions on three or four matrices. The memory burden is less than it is for a character string. An attacker, however, can identify the password by shoulder surfing attack by watching the keyboard and the monitor.

In the gaze-following method [2], a user performs authentication by following a moving icon on the monitor using their eyes. This method has shoulder surfing resistance, but the user must use a camera.

In the method presented in [3], the user registers extracted features of the user’s mouse movements on a desk. The user can use this method in many public places since most PCs have a mouse. However, the attacker can imitate mouse movements by shoulder surfing.

The Secret Tap with Double Shift (STDS) method [4] requires the user to register icons as a password on the screen of a smartphone. The user authenticates the icons by selecting the icons. This method has shoulder surfing resistance, but the genuine icons can be revealed by recording attack twice.

As described above, no easy-to-use method has shoulder surfing resistance for a PC in public places. In this paper, we propose a method that uses mouse clicks and mouse wheel rotations. This method can improve shoulder surfing resistance because the user does not need visual feedback and can operate the mouse under a desk.

2 Mouse authentication method

2.1 Overview

Here we define the personal authentication method. The input interface is a common mouse with right and left click, upward and downward wheel rotation, and wheel click. The output interface is an $N \times N$ matrix.

Use case: A user unlocks a PC or signs into a service on a PC in a public place while hiding the mouse such as under the desk. For example, a user signs into an e-mail service on their own laptop in a café.
**Strength:** When the user registers $m$ positions (including the registration order) on an $N \times N$ matrix, the probability that an attacker succeeds in a random attack on this method (accidental authentication probability) is $1/N^{2m}$. We are aiming for an accidental authentication probability of less than $1/10,000$.

**Registration phase:** Registration is performed as follows.
1. The screen displays an $N \times N$ matrix with an initial position randomly selected.
2. The user registers a position on the matrix using mouse operations. The current position is marked by a red circle, and the positions moves left by left click, right by right click, up by upward wheel rotation, and down by downward rotation. The user registers the current position with a wheel click.
3. The user registers $m$ positions by repeating (1) and (2). The registered order is part of the authentication information.

**Authentication phase:** Authentication is performed as follows.
1. The screen displays an $N \times N$ matrix with a randomly selected initial position.
2. The user specifies the first registered position with mouse operations. The mouse operations and its effects are the same as (2) in the registration phase. The user hides the mouse such as under the desk, and only the initial position is displayed, not the current positions, to strengthen shoulder surfing resistance.
3. The user specifies $m$ registered positions in their registered order using the same mouse operations used during registration.

**Benefits:** The mouse operations in this method are intuitive. Some people cannot operate a keyboard intuitively while hiding it. In addition, the proposed method is a challenge and response authentication system, in which the screen position and position order are hidden. Therefore, users can use this method in public places safely.

### 2.2 Implementation

We implemented the proposed method using a $5 \times 5$ matrix ($N = 5$), which struck a balance between the usability and authentication strength. A position count of $m \geq 3$ meets $1/5^{2m} \leq 1/10,000$, so we decided to make the “password” three or more positions long.

Fig. 1 shows an animation of the proposed method. It is natural to fix the initial position on the matrix, but the shoulder surfing resistance is weak against hearing the mouse sounds sometimes. Therefore, although it seems that the usability is lower than when the initial position is fixed, we decided that the initial position is determined randomly. The registered position is revealed when an initial position is near it in the authentication phase, so the user must register a position after moving at least three times.

We developed two variations in which the user (1) selects positions directly by combinations of mouse clicks and mouse movements, or (2) selects by combinations of colors and numbers on the matrix. A user can select a variation according to their taste.
3 Evaluation

We conducted two experiments on the usability and shoulder surfing resistance of the implemented method. The subjects used a desktop PC and a common mouse with mouse clicks and wheel rotations (Logicool wireless mouse M186). This mouse emits the sound of clicks and rotations clearly. In the user testing and shoulder surfing experiment reported here, we did not use the two variations, but we plan to conduct experiments of these in the future.

3.1 Usability test

We conducted the usability test as follows:

1. We explained to the subject how to use the method.
2. The subject completed a tutorial to become familiar with the mouse operations.
3. The subject registered three positions.
4. The subject performed three authentications successfully.
5. The subject answered a questionnaire on comprehension, ease of use, ease of familiarization, safety for shoulder surfing, and user needs. These items were rated from 1 to 5 (very bad to very good).

Fig. 2 shows the result of the usability tests. The subjects were 20 Kanagawa Institute of Technology students. The average time to complete the third successful authentication was 15.5 s [see Fig. 2(a)], and the average authentication success rate was 63.1%. It took time for users to get used to this authentication method, and the success rate was not enough due to the invisibility of the current positions. Fig. 2(b) shows the result of the responses to the usability testing questionnaire. All five averages are over 3, so usability of this method is confirmed.

3.2 Shoulder surfing resistance test

In the second experiment, shoulder surfers were positioned 1 m behind the user so that the surfer could see the monitor and hear the mouse sounds. We eliminated ambient noise so that the surfer could hear the mouse sound clearly. The procedure of the shoulder surfing resistance experiment was as follows:
1. We formed teams consisting of five or six people.
2. We choose a user from each team, and the user registered three positions in the method. Shoulder surfers were the other people in each team, and not allowed to see the monitor during the registration.
3. The user performed ten authentication successes while hiding the mouse under the desk. Shoulder surfers observed the authentication and tried to detect the registered positions; they were allowed to take notes.
4. We repeated steps 2–4 until everyone on the team was a user once.

The subjects were 16 Kanagawa Institute of Technology students. We performed this experiment using one team with six people and two teams with five people each. The result shows that the detection rates for one, two and three positions were 22.8%, 8.6%, and 1.4%, respectively. Only once were all three of a user’s registered positions detected (by one person). For the other 15 subjects, only one or two registered positions were detected. Under real conditions in public places, ambient sound would interfere with the shoulder surfing, so the results indicate that this method has shoulder surfing resistance.

3.3 Comparison with other methods
We compared the proposed method with related methods [1, 2, 4], as shown in Table I. The proposed method has fewer password combinations than SECUREMATRIX [1] and the gaze-following method [2], but it has shoulder surfing resistance.
surfing resistance and needs no special device. The STDS method on a smartphone [4] has shoulder surfing resistance and does not require special devices, but the proposed method has more password combinations than does STDS.

4 Conclusion

We proposed a personal authentication method with shoulder surfing resistance using mouse operations. In the proposed method, a user can hide the mouse and thereby achieve shoulder surfing resistance. We implemented the proposed method and evaluated it for usability and shoulder surfing resistance. The results show that the proposed method has good usability, shoulder surfing resistance. However, the sounds made by the mouse during operation is a possible source of password detection. A measure to protect against this will be developed in the future. Moreover, although we assumed that a user always hides a mouse such as under a desk in authentication phase, we will develop the method that have shoulder surfing resistance without hiding a mouse.

Acknowledgments

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Experimental path loss model for liver implanted wireless communication channel at ultra-wideband range

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Abstract: With the potential use of wireless implanted devices for transplanted liver monitoring applications, it is important to investigate the in-body propagation channel at liver area and to demonstrate the feasibility of the applications. Experimental measurements were performed using our developed multilayer human equivalent phantoms. Subsequently, path loss data were recorded and analyzed for various distances between the implanted and on-body antennas at lower range of ultra-wideband to assess the in-body propagation channel. Finally, our preliminary results indicate a possibility for liver implanted wireless communications using UWB technology in the example case scenario.

Keywords: ultra-wideband (UWB), liver implanted wireless communication, path loss model, transplanted liver monitoring

Classification: Antennas and Propagation

References


1 Introduction

Recently, wireless implanted device is of an interest in the area of medicine for healthcare monitoring and collecting biological parameters. One of the promising wireless implanted applications is physiological monitoring of transplanted organs. Nowadays, organ transplantation has been continuously performed during medical treatments [1]. Particularly, liver transplantation has become the primary clinical treatment for end stage liver disorders [2]. However, the technical failure rate is reported to be highest during the first two-week period after surgery [3]. Conventional techniques such as periodic blood testing and liver tissue biopsy do not offer real-time and constant monitoring after organ transplantation and are often too slow to respond to the potential loss of transplanted organ. Being one of the most common transplanted solid organs with only a limited number of organ donors, this strongly emphasizes the need to minimize liver transplantation failure [1].

![Diagram of wireless implanted monitoring system](image_url)

Fig. 1. An overview of the wireless implanted monitoring system.

Fig. 1 demonstrates an example scenario for the wireless implanted monitoring system. Wireless implanted device positioned on the liver surface monitors and transmits body parameters such as blood flow and oxygen saturation which are predictors of transplanted liver function [3] to on-body node which acts as a
relay for wireless transmissions to personal devices or hospital servers. This application can provide a reliable method of instantaneous and continuous monitoring to report the organ status of a patient to any healthcare staffs. This early detection should enable prompt medical operations before serious damage occurs to the transplanted liver [1, 3]. Moreover, together with future biodegradable materials, an implanted device will not need to be removed from a patient, thus, patient risks will be reduced. Eventually, this will potentially improve the success rate of liver transplantation.

On the other hand, it is important to select the appropriate frequency band of wireless communications between the devices that is suitable to the specific applications in consideration. Ultra-wideband (UWB) technology is a favorable option for wireless implanted applications [4]. Due to its high frequency range, the physical size of implantable antennas can be significantly reduced. Moreover, since it has a simplicity on the transmitter side, this will lead to the miniaturization of implanted devices. Its low power consumption will extend the implanted device longevity. However, at UWB range, the signals encounter severe loss inside the human body leading to the difficulty of system implementation. Therefore, it is important to evaluate the characteristics of propagation channel at UWB range to achieve reliable in-body wireless communication links. There have been various attempts to study UWB propagation channel at various locations inside human body e.g. chest [4], abdomen [5, 6], and brain [7]. Nevertheless, to the authors’ knowledge, UWB channel characteristics at liver location for liver implanted wireless monitoring system has not been reported in any open literature. Hence, to this aim, we performed experimental measurements using our developed simplified multilayer human equivalent phantoms to obtain attenuation data and proposed path loss model which would provide initial understandings of the characteristics of liver implanted wireless communications using UWB channel [8].

2 UWB antenna and measurement setup

The antennas used as the on- and in-body antennas in our measurements were Skycross UWB antennas (SMT-3TO10M-A) manufactured by Skycross Inc. as shown in Fig. 2(a). Experimental measurements were conducted using multilayer phantoms as illustrated in Figs. 2(b) and 2(c). The measurement setup consisted of the two Skycross UWB antennas (SMT-3TO10M-A) as shown in Fig. 2(a), an Agilent N5230C PNA-L vector network analyzer (VNA), two coaxial cables connecting each antenna to port 1 and port 2 of the VNA, and the human equivalent semi-solid UWB phantoms of fat, muscle, and liver tissues. The whole size of phantoms was approximately 200 mm × 130 mm × 120 mm. The cables’ frequency response was subtracted from the channel measurements by the method of VNA calibration. A broadband measurement of phantom dielectric properties was conducted to confirm their values compared to values reported in the work of Gabriel et al. [9].

$S_{21}$ data within the frequency range of 3–5 GHz for various distances between the antennas in the measurements were recorded five times and averaged. Our measurements were set up as followed. One antenna located on the liver surface
under the muscle layer (referred to as in-body antenna) was at a fixed position while the other antenna (referred to as on-body antenna) was placed directly on the fat layer and moved from the beginning point (0 mm) in steps of 20 mm in each measurement up to the ending point (100 mm). Both antennas were placed in a face-to-face orientation as presented in Fig. 2(b).

3 Results

Firstly, to confirm the performance of the antennas, we recorded the $S_{11}$ result of the on-body antenna and the $S_{22}$ result of the in-body antenna within the 3–5 GHz range as illustrated in Figs. 3(a) and 3(b) respectively. We can see that within the considered frequency range, the on-body antenna shows good behavior. On the other hand, since this UWB antenna was originally designed for free space utilization, the performance of the in-body antenna is not good but is acceptable for our research purpose here. Subsequently, with the in-body antenna at the fixed position, we moved the on-body antenna from the starting point (0 mm) to the ending point (100 mm) and obtained experimental results of $S_{21}$ parameter for various separation distances between the in- and on-body antennas. Then, to evaluate the signal attenuations inside the human body, we later processed all recorded $S_{21}$ data from the measurements to obtain the path loss (PL) data of each link as $PL = \text{mean}(|S_{21}|)$ in decibels. Consequently, we obtained the path loss model of the in-body channel as a function of the propagation distance by applying linear fitting to the path loss data at all points. The log-distance path loss model was applied as the following expression:
where \( d \) is the separation distance in millimeters between the in- and on-body antennas and \( d_0 \) is the reference depth of implantation which is 40 mm. \( \text{PL}_0 \) is the path loss at the reference location, \( d_0 = 40 \) mm. The exponent, \( n \), which is known as path loss exponent, can be used to evaluate how fast the signal power attenuates with the distance. Fig. 3(c) shows the average path loss data collected from the measurements as a function of the logarithmic antenna separation distances and the linear fitting curve of the measured data. In our case, \( \text{PL}_0 \) was 47.95 dB. The path loss exponent which is the slope of the obtained fitting curve was 6.25.

### 4 Conclusion

The propagation characteristics of in-body channels using UWB technology for liver implanted wireless communications have been investigated for the first time.
by mean of experimental measurements using simplified multilayer human equivalent phantoms. This study provides important preliminary insights for liver implanted wireless monitoring system. The in-body path loss data for various propagation distance values between the implanted and on-body antennas within the 3–5 GHz band obtained using measurement approaches were presented. It was found that path loss data were between 45 dB and 80 dB within 40 mm and 100 mm propagation range approximately. Our preliminary studies in this paper show that the realization of liver implanted wireless monitoring system using UWB channel in the example scenario case is possible. Moreover, our results reveal path loss data for liver-skin surface communication channel which is a necessary parameter for the design and evaluation of liver implanted wireless communication links. The information here can be used as guidelines for future studies of various in vivo medical applications such as wireless implanted monitoring of organs and implantable drug delivery at precise targeted locations and would also encourage other innovative applications utilizing UWB technology.

For our future work, due to the necessity to accurately characterize the in-body wireless communication channel at liver location and to further confirm the feasibility of liver implanted wireless monitoring system in the example scenario case, we will continue our further investigations using antennas designed for the implantation at liver location. In addition, we will conduct simulations using various digital human models as well as experimental measurements using realistic human phantoms including liver and other internal organs around the liver area to obtain insights of antenna performances and propagation channels for liver implanted wireless communications under more realistic environments.

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Multi-slot and multi-user joint detection scheme for CRDSA in satellite network

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Abstract: A multi-slot and multi-user detection scheme is presented for the contention resolution diversity slotted ALOHA (CRDSA) in satellite networks. The scheme can effectively resolve the deadlock loop of the collided packets of J users in J slots, and can be applied to DVB-RCS2 directly with a little change on receiver. The simulation results show the throughput of the CRDSA with the proposed scheme can improve greatly when the PLR is at 0.001, and has less latency compared with the conventional CRDSA.

Keywords: CRDSA, satellite networks, dead loop, collided packets

Classification: Satellite Communications

References


1 Introduction

Random access technique has attracted much more attentions in recent years, which can provide a large access capacity for the huge number of the machine to machine (M2M) communication terminals with low latencies [1]. Among the available RA
techniques, CRDSA [2] significantly enhances the throughput and the Packet Loss Ratio (PLR) with the iterative interference cancellation at the demodulator, and it has been employed in the DVB-RCS2 standards. However, in the CRDSA, the collisions resolution with the interference cancelling (IC) depends on the clean packets in the frame, and it performs well only for the long frame with hundreds of slots. The so-called “deadlock loop” increases as the channel load increases or the number of slots in the frame decreases, which greatly degrades the performance of the CRDSA. CRDSA++ and IRS and other variants [3, 4] are proposed to mitigate the deadlock loop and further improve the throughput with the cost of more replicas or additional signalling overhead. However, the deadlock loops still exist, and the power consumption of the user terminal is also highly increasing.

By analysing the remained collided packet signals after the IC processing in the CRDSA at the access point, we model the collided packets of J users in the J slots, one of the deadlock loops, as a new multiple inputs and multiple outputs (MIMO) signal model, and propose a multi-slot and multi-user joint detection (MMJD) scheme to recover the packets of J users. It can improve the throughput of the CRDSA without any modification and increasing the power consumption at the user terminal, and is independence on the clean packets, and is also suitable for the short frame with several tens of slots, which can reduce the latency of the user data compared with the conventional CRDSA.

2 Proposed method

2.1 Signal model

A typical satellite random access system with the CRDSA is considered, and there are M terminals try to access the common channel in the frame with N_{slot} slots. By taking advantage of the broadcasting signal in the satellite network, the user terminal employs the time advance (TA) and frequency offset pre-compensation to synchronize in the uplink. Suppose the transmitted complex data of the packet of the u-th user terminal in the slot is $s^{(u)} = \text{diag}(s^{(u)}(1), s^{(u)}(2), \ldots, s^{(u)}(L))$, where L is the number of the symbols. The u-th user terminal randomly selects the slots to send the packets according the CRDSA protocol. If there are packets of K users which are sent on the j-th slot, the discrete complex received signals can be described as

$$y_j = \sum_{u=1}^{K} h_j^{(u)} s^{(u)} + w_j$$  \hspace{1cm} (1)

where $y_j = [y_j(1), y_j(2), \ldots, y_j(L)]^T$, $h_j^{(u)} = [h_j^{(u)}(1), h_j^{(u)}(2), \ldots, h_j^{(u)}(L)]^T$ is the complex channel vector between the u-th user terminal and the access point during the j slot, $w_j$ and is the zero mean and covariance matrix with $E[w_j w_j^H] = \sigma_w^2 I_L$, and $I_L$ is the identical matrix with L order.

2.2 Multi-slot and multi-user joint detection scheme

At the access point in the satellite network, the received packets signal of the frame is firstly processed as that in the CRDSA, which searches the clean packets, and recovers the corrupted packets with the interference cancellation (IC) processing. After the IC processing, there are still remained many corrupted packets on the slots when the channel load is high. This is because the collided packets of the different
users are in the formation of the deadlock loops. By analysing packets of the deadlock loops, we find that the probability of the packets of J users colliding on the J slots is high, where J is less than the slot number of the frame. To resolve the loop and recover the packets of J users, we firstly model the received signals of the J slots, and then present the data recovery method.

In the deadlock loop of the packets of J users colliding on J slots, the indexes of the slots and users are written as \( j_1, j_2, \ldots, j_J \) and \( u_1, u_2, \ldots, u_J \) respectively. By Eq. (1), the received signals of J slots can be given as

\[
\begin{bmatrix}
  y_{j_1} \\
  \vdots \\
  y_{j_J}
\end{bmatrix}
= 
\begin{bmatrix}
  h_{j_1}^{(u_1)}, \ldots, h_{j_1}^{(u_J)} \\
  \vdots \\
  h_{j_J}^{(u_1)}, \ldots, h_{j_J}^{(u_J)}
\end{bmatrix}
\begin{bmatrix}
  s^{(u_1)} \\
  \vdots \\
  s^{(u_J)}
\end{bmatrix}
+ 
\begin{bmatrix}
  w_{j_1} \\
  \vdots \\
  w_{j_J}
\end{bmatrix}
\]

(2)

where the \( \{h_{j_1}^{(u_1)}, h_{j_1}^{(u_2)}, \ldots, h_{j_J}^{(u_1)}\}^T \) in the channel matrix is the channel vector between the \( u_i \)th user terminal and the access point on the different slots, and the elements are zero except the channel elements on the selected slots that the \( u_i \)th user terminal transmitted the packets. It is also noted that the nonzero channel elements in the channel matrix can be estimated by the preambles of the packets, and the channel matrix is a sparse matrix. For the received vector in the J slots, the \( k_{th} \) received symbol vector is

\[
Y_{j,k} = [y_{j_1}(k), \ldots, y_{j_J}(k), \ldots, y_{j_J}(k)]^T = H_{j,k}S_{j,k} + W_{j,k}
\]

(3)

From the Eq. (3), it is seen that the received signal model during the \( k_{th} \) symbol period is a MIMO signal model. The model can be dubbed the multi-slot and multi-user joint MIMO signal model. Therefore, many MIMO detection algorithms can be used to recover the data of the J symbols of the J users.

To recover the packets of J users effectively, we employ a soft MIMO detection algorithm to give the soft information of the coded bits in the symbols, and the soft decoders are employed to decode the bits of the J users. The multi-slot and multi-user joint detection (MMJD) scheme is shown in Fig. 1. The received signals of the J slots are input to the MMJD detector. If the complex symbol in Eq. (3) contains Q coded bits, which can be described as \( c_k^{(u_i)} = [c_{k,1}^{(u_i)}, \ldots, c_{k,q}^{(u_i)}] \), the log-likelihood ratio (LLR) in Fig. 1 can be represented as

\[
LLR(c_{k,q}) = \ln \left( \frac{p(c_{k,q} = 1|Y_{j,k})}{p(c_{k,q} = 0|Y_{j,k})} \right)
\]

(4)

By using Eq. (4), the MMJD detector output the soft information of the coded bits of the J packets, which are input to the corresponding de-interleavers and soft-
decoders. The soft-decoders output the bits of the $J$ users. The transmitted data of the $J$ users can be recovered from the collided packet signals in the $J$ slots.

### 2.3 Algorithm for CRDSA with MMJD

The MMJD scheme can effectively resolve the collided packets of $J$ users on the $J$ slots, one of the deadlock loops in the CRDSA. Its algorithm to process the received signals is described as follow.

#### Algorithm for CRDSA with MMJD:

**Initialization:**
Set the number of the clean packets, the number of loop, the initial number of users in the loop and the number of all the slots in the frame, as $N_{\text{clean}} = 0$, $N_{\text{loop}} = 0$, $J = 2$, and $N_{\text{slot}}$ respectively. Initialize the clean packet index set $P_{\text{index}}$ and the loop index set $P_{\text{loop}}$ to the empty sets.

**START:**
1) Scan and search the clean packets in all the slots in the frame, and set $N_{\text{clean}} = \text{the number of clean packets}$, and $P_{\text{index}} = \{\text{indexes of clean packets}\}$.
2) If $N_{\text{clean}} \neq 0$, get the indexes of the clean packets from the set $P_{\text{index}}$, demodulate the clean packets; else go to 4).
3) Using the demodulated data of the clean packets, reconstruct the signals of their replicas, and cancel them from the collided packets signals, go to 1).
4) Scan and search the loop of the packets of the $J$ users in the $J$ slots in the remained signals, set $N_{\text{loop}} = \text{the number of loops}$, and $P_{\text{loop}} = \{\text{indexes of the loops}\}$.
5) If $N_{\text{loop}} \neq 0$, get the indexes of the loops from the set $P_{\text{loop}}$, use the MMJD scheme to demodulate the packets of $J$ users.
6) Increase $J$, if $J < N_{\text{slot}} + 1$, go to 4), else, go to 7).
7) End.

### 3 Simulation results

In the simulation, the SC-FDMA packet with preamble and pilots in [5] and the 3GPP turbo code are employed at the user terminal; the channel model is ITU-R M.1225, and the perfect power control is assumed. The frame size of 100 slots is used. The traffic of user terminal is modeled as Poisson traffic sources and transmits fixed copies of packets (2 or 3) at random selected slots during a frame according to the CRDSA. An open loop transmission scheme is assumed, that is, there is no re-transmission or congestion control in simulation.
Fig. 2 shows the packet error rate (PER) curves of two detection methods. It can be found that the PLR of the MMJD is better than that of the detection of the clean packets because the user packet and its replicas are all used in the MMJD, and only one packet of the user is employed in the detection of the clean packets.

![Fig. 2. Packet error rate of two detection methods](image)

Fig. 3(a) shows the throughput of the CRDSA with and without MMJD versus channel load. It is seen that the throughput peak of CRDSA-2 with MMJD is higher 6% than the CRDSA-2, and the CRDSA-3 with MMJD is higher 23.2% than the CRDSA-3.

Fig. 3(b) shows PLR of the CRDSA with and without MMJD versus channel load. It is seen that the PLR of the CRDSA with MMJD is lower than that of the CRDSA. At the \( PLR = 10^{-3} \), the channel loads of the CRDSA-2 without and with MMJD are 0.055 and 0.35 respectively, and those of the CRDSA-3 without and with MMJD are 0.55 and 0.8 respectively. So, the MMJD scheme can improve the channel load 6.4 and 1.5 times for the CRDSA-2 and CRDSA-3 respectively.

![Fig. 3. (a) Throughput of the CRDSA with and without MMJD. (b) PLR of the CRDSA with and without MMJD](image)

### 3.1 Conclusion

By using proposed multi-slot and multi-user detection algorithm, some of the deadlock loops in the collided packets can be solved effectively. The proposed scheme can be applied to DVB-RCS2 directly with a little change on receiver,
improve the throughput and reduce the latency of the return link by using short frame size effectively.

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Wideband shielding effectiveness of laminated sheet using copper and magnetic materials

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Abstract: This paper presents conditions of a laminated sheet composed of copper and magnetic sheets for obtaining wideband shielding effectiveness below 10 MHz. The following two conditions are clarified: (1) the optimum thickness ratio of the copper sheet to the laminated sheet is 0.5–0.7, (2) the loss factor (\(\tan \delta\)) of the relative permeability of the magnetic sheet is above 1 for increasing reflection losses of the laminated sheet. These conditions are validated by measuring the shielding effectiveness of the laminated sheet. The shielding effectiveness of the laminated sheet is higher than that of the copper sheet in the frequency from 0.1 MHz to 10 MHz.

Keywords: radiation noise, shielding sheet, lamination

Classification: Electromagnetic Compatibility (EMC)

References


1 Introduction

Clean power generation is demanded owing to increasing energy consumption and greenhouse gas emission [1]. Power electronics equipment such as electric vehicles and photovoltaic power systems are expected as the technologies to meet the demand. Recently, switching frequencies of power semiconductors in the equipment have been increasing for the higher power conversion efficiency. Accordingly, the switching devices radiate high frequency and wideband noise whose levels are so high [2, 3]. The semiconductors such as insulated gate bipolar transistor (IGBT) and metal-oxide-semiconductor field-effect transistor (MOSFET) operate in low frequency (LF: 30–300 kHz) band, and the noise frequencies are up to high frequency (HF: 3–30 MHz) band. Currently, electromagnetic compatibility standards for radiation noises do not cover below 30 MHz, but suppression of the radiation noises is needed because of the above reason.

Metallic materials are mostly used for suppressing the radiation noises from LF and HF bands. Although the materials should be thin for workability and functionality, the materials are required to be thick for shielding magnetic field effectively in the wideband frequency [4]. The authors focus on laminated sheets to solve the above problem. Laminated sheets which are composed of copper and magnetic materials are expected to obtain higher shielding effectiveness than that of the metallic materials [5, 6]. However, optimum conditions of laminated sheets for obtaining wideband shielding effectiveness from LF to HF bands were not reported.

This paper presents optimum conditions of a laminated sheet for obtaining wideband shielding effectiveness below 10 MHz and discusses the thickness ratio of the copper sheet to the laminated sheet and the relative permeability of the magnetic sheet. Since material thicknesses are quite thin against the wavelength in the above target frequency, attenuation losses of the laminated sheet are negligibly small. Therefore, the discussion is focused on reflection losses of the laminated sheet.

2 Shielding effectiveness analysis

2.1 Evaluation method

Fig. 1 shows (a) a magnetic shield measurement system called Kansai Electronic Industry Development Center (KEC) method [7, 8] and (b) its electromagnetic
simulation model. As shown in Fig. 1(a), this system is composed of two shielded loop antennas embedded in two metal cavities. A part of the antenna is arranged in the center of metal cavity, and a shielding sheet is set in between the cavities. A signal generator, a spectrum analyzer and an amplifier are used for the measurement. Shielding effectiveness is defined as the received power normalized to that without shielding sheet. As shown in the Fig. 1(b), the shielded loop antennas and the metal cavity are modeled for the simulation that 3D electromagnetic simulator HFSS [9] is used. The ends of each shielded loop antenna are terminated with 50Ω.

![Figure 1](image1.png)

**Fig. 1.** KEC method to evaluate shielding effectiveness.

### 2.2 Evaluation results

First, a shielding effectiveness of a laminated sheet is simulated in order to clarify the optimum thickness ratio of the copper sheet to the laminated sheet. The laminated sheet, of which the total thickness is 53 µm, is composed of copper (conductivity $5.8 \times 10^7$ S/m) and magnetic sheets. Here, real part $\mu'_r$ and imaginary part $\mu''_r$ of the relative permeability of the magnetic sheet are assumed to be 1000 and 0, respectively. Fig. 2(a) shows the shielding effectiveness against the thickness.
ratio of the copper sheet to the laminated sheet. The thickness ratio at maximum shielding effectiveness is also shown with large symbols in the Fig. 2(a). The shielding effectiveness decreases with decreasing frequency; therefore, improving the shielding effectiveness in the lower frequency is required. The thickness ratio at the maximum shielding effectiveness is 0.5–0.7 in the frequency from 0.1 MHz to 10 MHz, and the shielding effectiveness at 0.1 MHz is 7 dB higher than that of copper sheet. Therefore, the optimum ratio of the copper sheet to laminated sheet is 0.5–0.7 for obtaining wideband shielding effectiveness.

Secondly, the relative permeability of the magnetic sheet is treated. The thickness of the copper sheet is set to 35 µm (total thickness of 53 µm) to satisfy the optimum thickness condition. Fig. 2(b) shows the shielding effectiveness of the laminated sheet against $\mu'$ at $\mu'' = 0$. The shielding effectiveness of copper sheet with 53 µm thickness is also shown for comparison. The shielding effectiveness increases with increasing $\mu'$. This result means the reflection loss contributes to the shielding effectiveness. The color values in Fig. 2(b) indicate $\mu'$ at the intersection points of the shielding effectiveness of the laminated sheet and that of the copper sheet. The values are minimum conditions of the magnetic sheet for obtaining higher shielding effectiveness than that of the copper sheet. Fig. 2(c) shows the shielding effectiveness of laminated sheet against $\tan\delta = \mu''/\mu'$. The shielding...
effectiveness increases drastically as \( \tan \delta \) of the magnetic sheet exceeds 1. In order to interpret the simulated results, the relationship between \( \tan \delta \) of the magnetic sheet and the reflection loss of the laminated sheet is examined. Characteristic impedance \( Z_S \) of a material is expressed by Eq. (1).

\[
Z_S = \sqrt{\frac{j \omega \mu}{\sigma + j \omega \varepsilon}}
\]

(1)

where, \( \sigma, \mu \) and \( \varepsilon \) are the conductivity, permeability and permittivity of the material and \( \omega \) is the angular frequency. On the assumption that \( \sigma = 0 \), and \( \varepsilon \) is constant, the real part of \( Z_S \) is given by Eq. (2). Here, \( Z_0 \) is impedance of free spaces.

\[
Re(Z_S) = Z_0 \sqrt{\frac{\mu'}{2} \left( 1 + \sqrt{1 + \tan^2 \delta} \right)}
\]

(2)

Eq. (2) exhibits the drastic increase in \( Z_S \) with \( \tan \delta \) above 1. The high shielding effectiveness is due to increasing the reflection loss by the impedance mismatch between the low-impedance copper sheet and high-impedance magnetic sheet. Therefore, the magnetic sheet with \( \tan \delta \) over 1 should be selected for increasing the reflection loss of the laminated sheet.

3 Measurements

Shielding effectiveness of the laminated sheet is measured to validate the above discussions. From the conditions obtained in previous section, the magnetic sheet of amorphous ribbon containing iron as a main component is selected. Thickness of the magnetic sheet is 18 \( \mu \text{m} \). Fig. 3 shows (a) \( \mu' \) and (b) \( \tan \delta \) of the magnetic sheet. The conditions of relative permeability of the magnetic sheet are also shown with dotted lines. The \( \mu' \) is smaller than the conditions over 4 MHz, but the laminated sheet would have higher shielding effectiveness than the copper sheet because its \( \tan \delta \) meet the conditions in the frequency from 0.1 MHz to 10 MHz. Fig. 3(c) shows the measured and simulated shielding effectiveness of the laminated sheet. The shielding effectiveness of the copper sheet with 53 \( \mu \text{m} \) thickness is also shown for comparison. The measurement results are consistent with the simulation within 5 dB errors. The shielding effectiveness of the laminated sheet at 0.1 MHz is 26 dB higher than that of the copper sheet. Additionally, the shielding effectiveness of the laminated sheet is higher than that of the copper sheet in the frequency from 0.1 MHz to 10 MHz. The conditions are validated by the measurement results.
4 Conclusions

This paper presents conditions for obtaining wideband shielding effectiveness of the laminated sheet below 10 MHz by using electromagnetic simulation. As a result, the following conditions were clarified: (1) the optimum thickness ratio of the copper sheet to the laminated sheet is 0.5–0.7; (2) tanδ of the relative permeability of the magnetic sheet is above 1 for increasing reflection losses of the laminated sheet. To validate the above discussions, the shielding effectiveness of the laminated sheet satisfying these conditions was measured. As a result, the shielding effectiveness of the laminated sheet at 0.1 MHz was 26 dB higher than that of the copper sheet. Moreover, the shielding effectiveness of the laminated sheet is higher than that of the copper sheet in the frequency from 0.1 MHz to 10 MHz. Therefore, the conditions were validated by the measurement results.

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Numerical study on a rotationally-symmetrical dipole array antenna position for a MIMO full-duplex system

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Abstract: In this letter, a rotationally-symmetrical array (RSA) arrangement of a receiving dipole array antenna is optimized to maximize the performance of the null-beamforming method for the suppression of self-interference. The optimum RSA for a receiving dipole array with four elements is clarified. Through numerical simulations, we found that the optimized RSA arrangement suppressed the self-interference significantly.

Keywords: MIMO full-duplex, self-interference, antenna arrangement, null-beamforming

Classification: Antennas and Propagation

References


1 Introduction

Full-duplex system and multiple-input multiple-output (MIMO) technology are promising technologies that can improve communication capacity without using additional frequency bands [1, 2]. The full-duplex system enables simultaneous two-way wireless communication in the same frequency band while the MIMO technology improves frequency utilization efficiency using multiple antennas. The combination of the MIMO technology and the full-duplex system is called the MIMO full-duplex system. However, it is well-known that the full-duplex system suffers from the so-called self-interference [3]. The self-interference problem leads to the saturation of the radio frequency (RF) front-end because the level of the transmitting signal is considerably higher than that of the receiving signal. Consequently, the RF front-end suffers from undesired nonlinearity and can be damaged.

In addition, the eigen-beamforming (EBF) method is a technique to improve the communication quality by controlling the radiation pattern at the transmitter side using transmission weight vector [4]. The control of the radiation pattern at the transmitter side can be also effective in circumventing the self-interference because the interference cannot reach the receiving antennas ideally. The null-beamforming (NBF) method is defined as the EBF method used to suppress self-interference using transmission weight vectors obtained by singular value decomposition (SVD) of the self-interference channel [5]. The performance of the NBF method is strongly affected by the nature of self-interference channel, in terms of spatial correlation. In general, high spatial correlation leads to higher performance of the NBF method.

Previously, an array antenna arrangement such as rotationally-symmetrical array (RSA) has been proposed to increase the spatial correlation and improve the performance of the NBF method [6]. In RSA, the NBF method can considerably suppress the self-interference if all antennas are point wave sources. However, the actual antenna has mutual coupling from other antennas.

Due to the highly symmetric arrangement of the RSA, the self-interference can be suppressed but the influence of asymmetry of mutual coupling between the array antenna elements can negatively impact its performance. The mutual coupling between the array antenna depends on the position of the array antenna.

In this letter, the receiving antenna arrangement for the RSA is optimized from the viewpoint of self-interference channel between the transmitting and receiving array antenna elements. Both transmitting and receiving antennas are array antenna consisting of four-element dipole antennas.

2 Antenna arrangement suitable for NBF method

Fig. 1 shows a system model of our proposed MIMO full-duplex system. \( N \) and \( M \) are the number of transmitting antenna (Tx) and receiving antenna (Rx), respectively. \( \mathbf{H} \) is the self-interference channel, \( \lambda_i \) is the \( i \)-th singular value of \( \mathbf{H} \) in
descending order, and \(v_i\) is the \(i\)-th transmission weight vector obtained by SVD of \(H\). By arranging the array antenna so that the rank of \(H\) becomes 1, the singular values other than the first singular value degenerate. The NBF method uses the transmission weight vectors excluding the first singular value. The magnitude of the self-interference channel norm using NBF method is given by

\[
\|H\tilde{V}_1\|^2_F = \sum_{i=2}^{\min(M,N)} \lambda_i^2,
\]

where \(\tilde{V}_1\) is the transmission weight matrix excluding the first singular value, \(\tilde{V}_1 = [v_2, \cdots, v_N]\). This means that Tx weighted by \(\tilde{V}_1\) forms a null in Rx direction to suppress \(\lambda_1\) using the NBF method.

Fig. 2 shows the antenna arrangement to gain high spatial correlation. The Tx linear array is located on the \(z\)-axis. The Rx circular array is located on the circumference around the \(z\)-axis. \(D\) is the distance from the end of the Tx array to the origin of the Rx array, \(d_e\) is the element spacing of the Tx array, and \(r_e\) is the radius in the Rx array. We assume that all antennas are point wave sources and there is no mutual coupling.

In RSA, Line-of-sight paths from the arbitrary transmitting element to all of Rx elements are the same strength and phase because the distance between an arbitrary transmitting point and all of receiving point wave sources are equal. Therefore, all elements in each column vector of the self-interference channel matrix, \(H\), are identical, and the rank of self-interference channel \(H\) equals 1.

The self-interference power \(P_{w/NBF}\) and \(P_{w/oNBF}\) are defined as
\[ P_w = \frac{N}{C_0} \quad \text{for} \quad \min(M,N) \]

where \( P \) is the sum of the transmission power. The singular value of the self-interference channel becomes 0 except for first singular value if the rank of \( H \) is 1. Therefore, the self-interference power with and without NBF method are \( P_{w/NBF} = 0 \) and \( P_{w/\text{oNBF}} = P \mu^2_i / N \). Therefore, the NBF method can suppress the self-interference by excluding only the first singular value.

However, the gain of each Rx antenna varies if the mutual coupling of the Rx array antenna is asymmetric due to the mutual coupling among Rx. In this letter, the symmetry of the mutual coupling of the RSA antenna is evaluated from the magnitude of the mutual coupling of each Rx antenna.

### 3 Numerical analysis results

#### 3.1 Simulation setup

The antenna arrangement used for numerical analysis is shown in Fig. 2. The antenna is a half-wavelength dipole antenna. The distance from the end of the Tx array to the origin of the Rx array, the element spacing of the Tx array and the mutual coupling of the Rx array antenna is evaluated from the magnitude of the mutual coupling of each Rx antenna.
radius of the Rx array are set to $D = 10\lambda$, $d_e = 0.5\lambda$, and $r_e = 0.5\lambda$, respectively ($\lambda$: wavelength in a vacuum). $\theta_e$ is the rotation angle of the receiving antennas moving in the direction from $x$-axis to $y$-axis. In this numerical analysis, we changed the rotation angle $\theta_e$ from $0^\circ$ to $90^\circ$.

### 3.2 Simulation results

Fig. 3(a) shows the singular value characteristics. The first singular value was almost the same at all rotation angles, the second singular value become smaller as it approaches $\theta_e = 45^\circ$. We found that the second singular value at $\theta_e = 45^\circ$ was, at maximum, 48 dB lower than that at $\theta_e = 0^\circ$. From the results, we found that the spatial correlation become higher as approach at $\theta_e = 45^\circ$.

Fig. 3(b) shows the mutual coupling of the Rx antenna versus rotation angle. The mutual coupling of $i$-th Rx antenna $C_i$ is given as

$$C_i = \sum_{k=1}^{N} |S_{i,k}|^2 \quad (i \neq k), \quad (4)$$

where $S_{i,k}$ is S-parameter from $k$-th Rx antenna to $i$-th Rx antenna. The mutual coupling difference of each Rx antenna become smaller as it approaches $\theta_e = 45^\circ$. In particular, all mutual coupling at $\theta_e = 45^\circ$ were almost equal. The sum of mutual
coupling from the adjacent antenna $(\theta_e \pm 90^\circ)$ is equal if the directivity of the RSA antenna is point-symmetric. Therefore, $C_i$ is affected by the mutual coupling of the diagonal RSA antenna $(\theta_e + 180^\circ)$. The dipole antenna is omnidirectional on the $xz$-plane perpendicular to the element axis ($y$-axis) and forms a null at the end of element on the $xy$-plane. Thus, the mutual coupling $C_1$ and $C_3$ were large, and $C_2$ and $C_4$ were small at $\theta_e = 0^\circ$. However, mutual coupling between Rx antennas at $\theta_e = 45^\circ$ were almost equal because the mutual coupling of the dipole antennas on the diagonal was equal. Therefore, the mutual coupling at $\theta_e = 45^\circ$ was symmetrical.

Fig. 3(c) shows the self-interference power versus rotation angle. The self-interference power without NBF method ($P_{w/oNBF}$) was $-37$ dB at all rotation angles. In other hand, the self-interference power with NBF method ($P_{w/NBF}$) was suppressed by more than 95 dB at all rotation angles. In particular, the self-interference power was suppressed by 143 dB at $\theta_e = 45^\circ$. This shows that the NBF method at $\theta_e = 45^\circ$ suppresses the self-interference significantly because the self-interference channel including the influence of mutual coupling was symmetrical.

4 Conclusions

Numerical analysis confirmed that the self-interference power with NBF method varies with the rotation angle of the receiving dipole array antenna of RSA. The self-interference power was changed according to the rotation angles because the mutual coupling of the receiving dipole array antennas alters the symmetry. The self-interference power with NBF method ($P_{w/NBF}$) was suppressed by 143 dB at $\theta_e = 45^\circ$. Therefore, we found that the most efficient arrangement of the dipole array antenna in RSA is at $\theta_e = 45^\circ$.

Acknowledgments

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Errata

The following editorial correction has been found in Vol. 6, No. 9, and should be corrected as follows.

Wrong

p. 517 (in Fig. 2)  
time interval = 800 µs

Correct

time interval = 4 ms