Development of circuit model of AC/coaxial adapter using for calibration of AMN

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Abstract: Transmission characteristics of an AC/coaxial adapter that is used for calibration of an AMN had been evaluated up to 30 MHz, which is the upper limit of frequency in the use of conventional AMNs. However, the new type AMNs for using over 30 MHz has been developed, thus the transmission characteristics of the adapter beyond 30 MHz should be evaluated. In this paper, a circuit model of the AC/coaxial adapter is developed in order to obtain easily its transmission characteristics up to 1 GHz. The calculated characteristics using the developed circuit model agreed well with that of simulated by using a numerical electromagnetic analysis.

Keywords: AC/coaxial adapter, circuit model, artificial mains network

Classification: Electromagnetic compatibility (EMC)

References

1 Introduction

In recent years, energy-efficient equipment such as a power conditioner and LED bulbs have been rapidly spread by higher awareness for energy conservation. The energy is conserved by using a high-efficient power supply switching circuit in the equipment. However, the switching circuit often cause an electromagnetic disturbance over a wide frequency range from kHz to GHz [1]. Since this disturbance may interfere in the other electronics devices and broadcastings, the measurement of the disturbance is necessary.

An artificial mains network (AMN) [2] is used for measuring the conducted disturbance voltage below 30 MHz. Fig. 1(a) shows an example of conducted disturbance measurement using the AMN. As shown in the figure, an equipment under test (EUT) and a measuring receiver are connected to the AMN. The AMN supplies AC power to the EUT and transmits the disturbance occurring in the EUT to the measuring receiver. To obtain measurements result of the disturbance voltage, fundamental characteristics of the AMN such as the AMN impedance and the voltage division factor [2] need to be evaluated. Fig. 1(b) shows an example schema of AMN’s characteristics measurement. As shown in the figure, the AC/coaxial adapter [3] is used for connecting the AC outlet of the AMN to a vector network analyzer. Thus, the measured characteristics of the AMN include the characteristics of the adapter. The characteristics of the adapter should be evaluated for obtaining the characteristics of the AMN without the adapter.

The characteristics of the adapter had been evaluated up to 30 MHz which is the upper limit frequency of the conducted disturbance measurement using the conventional AMN [3, 4, 5]. However, since the conducted disturbance measurement of over 30 MHz is required by the above background, new type AMN for using up to 1 GHz has been developed [6]. Therefore, the characteristics should be also evaluated beyond 30 MHz.

In this paper, we develop a new circuit model of the adapter for obtaining the transmission characteristics of the adapter beyond 30 MHz. Although the characteristics of the adapter can be also evaluated by using a 3D electromagnetic simulation, it is difficult to clarify the phenomenon from the result in the simulator. On the other hand, the circuit model can reduce a calculation cost for obtaining the characteristics and clarify the phenomenon. The characteristics obtained by the developed circuit model are compared with those of an electromagnetic simulation to clarify the validity of the developed circuit model.
Developed circuit model of AC/coaxial adapter

2.1 Structure

Fig. 1(c) shows an AC/coaxial adapter. This adapter consists of two coaxial connectors, an AC plug, and a metal plate. The neutral and the phase lines connect directly each inner conductor of the connector. The protective earth (PE) line connects each outer conductor of the connector through the metal plate.

Fig. 2 shows the developed circuit model of the adapter. The AC plug can be treated as a transmission line because it consists of the conductors having uniform cross-section. There are three conductor lines #1, #2 and #3 above the infinite ground plane (GND) as shown in Fig. 2(a). The lines #1 and #2 represent the neutral and the phase lines, and the line #3 represents the PE line. Input ports of the lines #1, #2 and #3 are defined as port 1, 2, and 3, respectively. Output ports of the lines #1, #2 and #3 are defined as port 4, 5, and 6, respectively. The dimensions of the cross-section shown in Fig. 2(b) meet the JIS standard [7], and “h” is the height between the line #3 and GND. The two adapters which are connected as a back-to-back state will be measured because the ports of the adapter should be coaxial shape for connecting the vector network analyzer. Hence, a line length of the model is 34 mm which means the length of twice the AC plug. The equivalent circuit shown in Fig. 2(c) is lossless, and a telegrapher’s equation of the circuit is given by Eq. (1).

$$-\frac{d}{dx}\begin{bmatrix} V(x) \\ I(x) \end{bmatrix} = j\omega \begin{bmatrix} O & L \\ C & O \end{bmatrix} \begin{bmatrix} V(x) \\ I(x) \end{bmatrix}$$

(1)

Here, $O$ is a square zero matrix. $V(x)$ and $I(x)$ are vectors of line voltages and currents, respectively. $L$ and $C$ are inductance and capacitance matrices, respectively.
2.2 Evaluation method

The F matrix of the AC plug $F$ is derived from Eq. (1) using the state variable method [8]. Here, the values of $L$ and $C$ are analyzed by “ANSYS Q3D Extractor” which is the commercially-available code of parameter extraction. When each input port's voltages and currents are defined as $V_1$, $V_2$, $V_3$, $I_1$, $I_2$, and $I_3$, and each the output port's voltages and currents are defined as $V_4$, $V_5$, $V_6$, $I_4$, $I_5$, and $I_6$, the relation between the voltages and the currents are given by Eq. (2).

\[
\begin{bmatrix}
V_1 \\
V_2 \\
V_3 \\
I_1 \\
I_2 \\
I_3
\end{bmatrix}
= F
\begin{bmatrix}
V_4 \\
V_5 \\
V_6 \\
I_4 \\
I_5 \\
I_6
\end{bmatrix}
\]  

(2)

The currents which flow in the neutral and the phase lines return in the PE line. Therefore, the relation of the currents is shown Eq. (4)

\[
I_1 + I_2 = -I_3, \quad I_4 + I_5 = -I_6
\]  

(4)

---

**Fig. 2.** The developed circuit model of the adapter.
Then, a new $F$ matrix $F'$ is derived by using Eqs. (3) and (4) that express the voltages and the currents conditions of the adapter. The relation between the voltages and currents using $F'$ are given by Eq. (5).

$$
\begin{bmatrix}
V'_1 \\
V'_2 \\
I_1 \\
I_2
\end{bmatrix} = 
F'
\begin{bmatrix}
V'_3 \\
V'_4 \\
I_4 \\
I_5
\end{bmatrix}
$$

(5)

The transmission characteristics of the adapter are evaluated by $S$ matrix converted from $F'$.

### 3 Validation of developed circuit model

Fig. 3(a) shows an analysis model of the AC/coaxial adapters which are connected as a back-to-back state. All conductors of the analysis model are considered as perfect electric conductors. The dimensions of the coaxial connector are designed so that its characteristic impedance is 50 Ohm, and that of AC plugs are the same as the developed circuit model mentioned in Section 2. Also, the two inner conductors of the connectors are extended to avoid contact of the metal plate with the neutral/phase lines as shown in Fig. 3(a). The 4-port $S$-parameters of the analysis model are numerically simulated by using the 3D electromagnetic simulator “HFSS”, in a case of changing the height “$h$” between the PE line and the GND.

![Analysis model](image)

(a) An analysis model the adapters which are connected as a back-to-back state

![S-parameters](image)

(b) $S$-parameters at $h=10mm$  
(c) $S$-parameters at $h=50mm$

Fig. 3. The calculated $S$-parameters of the adapter using the developed circuit model.
Fig. 3(b) and (c) show the calculated S-parameters of the adapter using the developed circuit model in cases of $h = 10\text{mm}$ and $50\text{mm}$. As shown in the figures, the calculated characteristics using the model agree with the simulated one using the electromagnetic simulation regardless of $h$. This result shows the validity of the developed circuit model. The transmission characteristic $|S_{31}|$ of the adapter keeps 0 dB in less than 30 MHz. Thus, the adapter affects hardly the characteristics of the conventional AMN. However, the characteristic $|S_{31}|$ decreases in a case where the frequency exceeds 30 MHz by increasing the reflection and the coupling between the neutral/phase lines. It is found that the adapter affect the characteristics of the new type AMN for using over 30 MHz.

4 Conclusions

In this paper, the circuit model of the adapter was developed for obtaining easily its transmission characteristics. The calculated S-parameters of the adapter using the developed circuit model were compared with the simulated one using the electromagnetic simulator. As a result, both S-parameters agreed well regardless of the height between the adapter and the GND. Therefore, the validity of the developed circuit model was indicated. Furthermore, it was found that the adapter affects the characteristics of the new type AMN for using over 30 MHz because the transmission characteristics of the adapter decrease due to its reflection and coupling characteristics. In future works, a balance and an impedance of the adapter will be evaluated by using the developed circuit model, and an influence of the adapter for the AMN will be evaluated.

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Compensation of optical nonlinear waveform distortion using neural-network based digital signal processing

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Abstract: We studied a novel nonlinear compensation scheme using digital signal processing based on a neural network (NN). Distortion of 16QAM signals caused by self-phase modulation (SPM) was compensated for by using a three-layer NN without feedback loops. In the scheme, the input layer of the NN has feedforward tapped delay lines. Input signals of I and Q components of the 16QAM signals are fed into the delay lines. The compensated 16QAM signals of the I and Q components are outputted by two neurons in the output layer of the NN. BER and EVM performance was investigated by numerical simulations, and the EVM was improved by more than 20% by the compensation.

Keywords: digital signal processing, nonlinear distortion, SPM, neural network

Classification: Fiber-Optic Transmission for Communications

References


1 Introduction

Multi-level modulation schemes are key technologies to accommodate the increasing data traffic on communication networks. In particular, quadrature amplitude modulation (QAM) is an important candidate for attaining modulation at higher than four bits per symbol. On the other hand, the waveforms of QAM signals are distorted by self-phase modulation (SPM), because the signal power varies according to the transmitted symbols, resulting in a large peak-to-average power ratio (PAPR). Finite impulse response (FIR) filters have been used to compensate for linear distortion caused by, e.g., chromatic dispersion [1], but they cannot be used to compensate for nonlinear distortion. Some methods have been proposed for compensating for nonlinear effects, including optical phase conjugation (OPC) [2], digital back propagation (DBP) [3, 4], and the Volterra series transfer function (VSTF) [5]. However, these methods need an enormous amount of calculations or additional optical/electronic hardware components. Digital signal processing based on neural networks (NNs) has the merit that NNs can adaptively compensate for nonlinear distortion by using supervised learning algorithms. Some methods to compensate for nonlinear effects in wireless communication systems have been studied [6, 7]. In optical communication systems, nonlinear distortion compensation using NNs has been conventionally studied for Intensity Modulation-Direct Detection (IM-DD) transmission systems [8]. Recently, nonlinear equalization using NNs in frequency domain was investigated for coherent optical orthogonal
frequency division multiplexing (CO-OFDM) transmission systems, where many sub-NNs were used for subcarriers [9, 10]. We proposed a novel nonlinear compensation method using an NN to compensate optical multi-level signals distorted by SPM [11]. We showed that the NN can effectively compensate for distortion of 16QAM optical signals caused by SPM. In that study, however, we used an NN including complicated feedback loops, which possibly destabilized the system performance. In the present study, we investigated the performance of a simple three-layer NN without feedback to compensate for nonlinear distortion caused by SPM. Numerical simulation of 16QAM transmission showed that the NN can compensate for the nonlinear distortion as efficiently as the case with feedback loops. We evaluated the performance in terms of bit error rate (BER) and error vector magnitude (EVM).

2 Nonlinear compensation using an NN

Fig. 1 shows the construction of the three-layer NN that we used in the nonlinear compensation. The input layer of the NN has feedforward tapped delay lines. Input signals of in-phase (I) and quadrature (Q) components are fed into the delay lines. Neurons in the hidden layer have a sigmoidal output function. The neurons in the input and output layers have linear functions.

The neurons in the output layer output compensated signals described by

\[ y = f \left( \sum_{k=1}^{n} w_k x_k + b \right), \tag{1} \]

where \( x_k \) is the input from \( k \)-th neuron, \( w_k \) is the weight, \( b \) is the bias, and \( f \) is the output function. The values of the weight and bias are calculated by the Back Propagation (BP) algorithm to minimize the error, \( e \), which is defined as the difference between the output signals and supervised signals:

\[ e^2 = \sum_{k=1}^{n} (y_i - d_i)^2, \tag{2} \]

where \( d_i \) is the supervised signal. In this simulation, the numbers of input-layer neurons for both I and Q components were set to 12. The numbers of neurons in the hidden and output layers were set to 6 and 2, respectively.
3 System setup

Fig. 2 shows a 50-km 16QAM signal transmission system used in our simulations. 10-Gsymbol/s 16QAM optical signal was modulated by PRBS $2^{20}-1$ data and transmitted by a standard single mode fiber (SSMF) and a dispersion compensation fiber (DCF) having a total length of 50 km, cancelling the chromatic dispersion. The input power to the optical fibers was 10 dBm. The noise figure of EDFAs were 3 dB. After the transmission, the optical signal was received by optical homodyne detection. Here, we assumed that the local oscillator (LO) was ideally synchronized to the optical signal. The optical power of the LO was 3 dBm. In the simulation, electrical noise was taken into account only at PDs as thermal noise with power density of $1 \times 10^{-10}$ pW/Hz. Then, the distorted signal after the transmission was compensated using an NN in a digital signal processor (DSP). The sampled data were processed in 1-sample/symbol manner in the DSP consistently. The NN was trained using the Levenberg–Marquardt algorithm, which is a kind of BP [12]. We used random data sequence of about 50,000 symbols repeatedly for the training. The compensation performance was evaluated by BER and EVM.

4 Result and discussion

Fig. 3(a) shows the constellation of the received 16QAM signal in a back-to-back (BtB) configuration when the received power was $-10$ dBm. Fig. 3(b) shows the constellation after the transmission. Due to the large input power, the outer symbols of 16QAM signals were rotated in the clockwise direction by SPM, and Fig. 3(c) shows the constellation after the compensation using an NN. The distorted symbols were successfully compensated. EVM was improved by about 24%. Next, we investigated the compensation performance in the case where the received power is limited by the attenuator (ATT). Fig. 3(d) shows the constellation of the received 16QAM signal in the BtB configuration when the received optical power was $-32$ dBm, and Fig. 3(e) shows the constellation after the transmission. Fig. 3(f) shows the constellation after the compensation using the NN. The EVM was improved by about 20%. However, we could not recover the original constellation shown in Figs. 3(a) and (d) completely by the equalization. We calculated the EVM and BER versus received optical power, which was adjusted by the attenuator at the receiver side. Fig. 3(g) shows the EVM and BER characteristics calculated in the simulation. An EVM of less than 10% was achieved when the received power was

![Fig. 2. System setup of 16QAM transmission.](image-url)
higher than about 36 dBm, whereas the EVM without the compensation was about 30%. A BER of less than $10^{-6}$ was achieved by the compensation. In the figure, we also plotted EVM values in the case where an NN with feedback loops was employed as reported in [11]. However, no significant difference due to the feedback loops was observed. By removing the feedback loops, we can eliminate the possibility of unstable behavior and oscillation of the NN. In our simulations, the DSP was only used to calculate the equalization with the NN. In practical receivers, however, a DSP includes some functional blocks such as chromatic dispersion compensation, polarization demux, and carrier phase recovery. The nonlinear equalization can be used after all the linear processing and phase recovery. However, the best order in the processing sequence has to be investigated in the succeeding studies.

Fig. 3. Constellations and EVM/BER characteristics.
5 Conclusion

We investigated the compensation performance of a three-layer NN to compensate for nonlinear distortion in optical communication systems. Our numerical simulation of 16QAM transmission showed that the NN could efficiently compensate for the nonlinear distortion caused by SPM, and improved the performance in terms of BER and EVM.
Continuous beam scanning performance of dipole array antenna coupled to meander two-wire parallel transmission line

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Abstract: A mechanical beam scanning dipole array antenna coupled to a meander two-wire parallel transmission line is proposed. A meander transmission line is introduced to the proposed antenna in order to suppress its grating lobe without using lumped element inductors. Numerical simulation demonstrates that the proposed antenna is capable of continuous beam scanning ranging from 2°(@\(\phi = 0\)) to 42°(@\(\phi = 180°\)) without grating lobe due to the meander transmission line.

Keywords: array antenna, leaky wave antenna, beam scanning

Classification: Antennas and Propagation

References


1 Introduction

The next generation wireless communication system is expected to be operated in a high frequency band such as a millimeter wave frequency band [1]. In the millimeter wave frequency band, relatively wide bandwidth is available and an extremely high-speed wireless communication system can be developed. One of the most challenging problems for the millimeter wave wireless communication system is how to compensate a high propagation loss.

Electronic beam scanning antennas using various microwave components are promising techniques in order to compensate the high propagation loss in the millimeter wave frequency band [2, 3, 4]. The electronic beam scanning antennas are capable of scanning their main beam using active/passive microwave components such as solid state devices, phase shifters, or couplers. High signal-to-noise ratio is expected because the main beams of a transmitting and receiving antenna are directed to each other during the wireless communication. Quick response is another advantage of the electronic beam scanning antennas while their disadvantages are large insertion loss and nonlinearity.

A mechanical reconfigurable antenna is an alternative approach to develop a beam scanning antenna. Performance of the mechanical reconfigurable antenna is controlled using an actuator. The mechanical reconfigurable antenna needs no active microwave components and is free from large insertion loss and nonlinearity of them. Therefore, the mechanical reconfigurable antenna is a promising approach.
in order to compensate high propagation loss in the millimeter wave frequency band. A substrate integrate waveguide (SIW) slotted array antenna [5] and a traveling wave slot array have been proposed [6]. The SIW slot antenna is operating in 24 GHz band and its ±35° beam scanning range was demonstrated experimentally while the traveling wave antenna is operating in 13 GHz band and its ±25° beam scanning range was demonstrated experimentally. A 5 x 1 mechanical beam scanning reflectarray using a nylon gearing arrangement has been fabricated and its beam scanning performance was measured [7].

On the other hand, a mechanical beam scanning dipole array antenna coupled to two-wire parallel transmission line has been proposed and its beam scanning performance was demonstrated [8, 9]. The proposed antenna is a kind of a leaky-wave antenna and all array elements are excited via the near-field of the two-wire parallel transmission line. Beam scanning capability of the proposed antenna comes from variable array spacing which is available using an actuator. In [9], it was demonstrated that a couple of inductors loaded with the transmission line can suppress its grating lobe. The inductors are modeled as lumped elements in [9] while the size of inductors is comparable to wavelength in millimeter-wave band.

In this letter, a mechanical beam scanning dipole array antenna coupled to a meander two-wire parallel transmission line is proposed. A meander transmission line is introduced to the proposed antenna in order to suppress its grating lobe without using lumped element inductors. Numerical simulation demonstrates that the proposed antenna has a continuous beam scanning capability without the grating lobe.

2 Meander transmission line

According to [9], the proposed antenna has a grating lobe and inductors should be loaded with the transmission line in order to suppress the grating lobe. However, in the millimeter wave frequency band, it is difficult to deal with an inductor as a lumped element because its dimension is comparable to the wavelength. Therefore, meander structure was applied to the transmission line as the component of inductance. Fig. 1 shows the proposed antenna with the meander two-wire parallel transmission line and Table I shows its electrical parameters and dimensions. The two meander lines are parallel and the radiation from the lines is canceled because the current of the lines is the same amplitude but out of phase.

Transmission line theory is helpful in order to design the meander line because the meander line is composed of a couple of short stubs. According to the transmission line theory, a load impedance $Z_s$ of a lossless short stub is expressed as,

$$Z_s = jZ_0 \tan(\beta l)$$

(1)

where $\beta$ is phase constant of the transmission line, $Z_0$ is a characteristic impedance of the stub, and $l$ is the length of the stub. Once a specific value of the inductance is known, the length of the short stub to be loaded is immediately obtained from (1). In this letter, $l \approx 0.1 \lambda_0$ is obtained from (1) by substituting $Z_s = j60\pi$ and $Z_0 = 276$ which are obtained from dimensions and electrical parameters shown in Table I.
Numerical simulations were performed using the method of moments (MoM) in order to demonstrate a beam scanning performance of the proposed antenna with Fig. 1. Dipole array antenna coupled to meander two-wire parallel transmission line.

Table I. Electrical parameters and dimensions of antennas.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Design frequency (f) [GHz]</td>
<td>30</td>
</tr>
<tr>
<td>Conductivity of antenna elements and transmission lines (\sigma) [S/m] (Copper)</td>
<td>(5.8 \times 10^7)</td>
</tr>
<tr>
<td>Conductivity of ground plane (\sigma) [S/m] (Aluminium)</td>
<td>(3.5 \times 10^7)</td>
</tr>
<tr>
<td>Load impedance (Z_L) [(\Omega)]</td>
<td>300</td>
</tr>
<tr>
<td>Reactance of inductors (or short stubs) (L_H) [nH]</td>
<td>1</td>
</tr>
<tr>
<td>Width of short stub (l_{w}) [mm]</td>
<td>1</td>
</tr>
<tr>
<td>Length of short stub (l_l) [mm]</td>
<td>1</td>
</tr>
<tr>
<td>Spacing between two inductors (or short stubs) (l_{in}) [mm]</td>
<td>7</td>
</tr>
<tr>
<td>The number of inductors (or short stubs)</td>
<td>9</td>
</tr>
<tr>
<td>Radius of conductors (a) [mm]</td>
<td>0.1</td>
</tr>
<tr>
<td>Width of transmission line (W) [mm]</td>
<td>1.22</td>
</tr>
<tr>
<td>Length of transmission line (L) [mm]</td>
<td>120</td>
</tr>
<tr>
<td>Length of dipole element (l) [mm]</td>
<td>5</td>
</tr>
<tr>
<td>Height of dipole elements from transmission line (h) [mm]</td>
<td>0.5</td>
</tr>
<tr>
<td>Array spacing (d) [mm]</td>
<td>5~8</td>
</tr>
<tr>
<td>Spacing between source and the first element (x_0) [mm]</td>
<td>15</td>
</tr>
<tr>
<td>Length of ground plane in (x) direction (l_x) [mm]</td>
<td>120</td>
</tr>
<tr>
<td>Length of ground plane in (y) direction (l_y) [mm]</td>
<td>15</td>
</tr>
<tr>
<td>Height of transmission line from ground plane (h_p) [mm]</td>
<td>2.5</td>
</tr>
<tr>
<td>Number of dipole elements (N)</td>
<td>10</td>
</tr>
</tbody>
</table>

3 Numerical simulation

Numerical simulations were performed using the method of moments (MoM) in order to demonstrate a beam scanning performance of the proposed antenna with...
the meander line [10, 11]. Fig. 2(a) shows a beam scanning performance of the proposed antenna with lumped element inductors while Fig. 2(b) shows a beam scanning performance of the proposed antenna with the meander line. It is found that the beam scanning performance of these antennas is in good agreement.
Therefore, it can be said that the meander line is equivalent to the transmission line with lumped element inductors.

Fig. 2(c) shows a beam scanning range of the proposed antenna at 30 GHz. Here, the beam scanning range indicates an angle region where a drop of the directivity from the maximum one is not over 3 dB. It is found that continuous beam scanning performance is available as an array spacing \( d \) varies. The maximum directivity is 17.2 dBi at \( d = 0.89\lambda_0 \) and the beam scanning range is from \( 2^\circ(\phi = 0) \) to \( 42^\circ(\phi = 180^\circ) \).

4 Conclusion

In this letter, a performance of the mechanical beam scanning dipole array antenna coupled to a meander two-wire parallel transmission line has been demonstrated. It has been shown that the meander line is equivalent to the transmission line with lumped element inductors and is able to suppress a grating lobe. Numerical simulation has demonstrated that beam scanning range of the proposed antenna is ranging from \( 2^\circ(\phi = 0) \) to \( 42^\circ(\phi = 180^\circ) \). Impedance matching and fabrication technique are not discussed here and are remaining as problems to be challenged in future.

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A simple upper bound on the capacity of BI-AWGN channel

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Abstract: The capacity of binary input additive white Gaussian noise (BI-AWGN) channel has no closed-form solution due to the complicated numerical integrations involved. In this letter, a simple upper bound to evaluate the capacity of BI-AWGN channel is presented. In addition, through the moment generating function (MGF) of fading channel gain, the upper bounds of fading channels are derived by averaging the proposed bound with respect to the probability density function (pdf) of fading gain.

Keywords: BI-AWGN, channel capacity, MGF

Classification: Fundamental Theories for Communications

References


1 Introduction

BI-AWGN channel is often used as a practical model in many digital communication schemes, and particularly, it plays a crucial role in the research of channel coding. In fact, it is the starting point for many theoretical studies due to its simple form. Although the capacity of BI-AWGN channel had been studied intensively in [1, 2, 3], there are no closed-form solution but complicated series expansions and approximations. The significance of a simple and efficient formula for the capacity is not just only to evaluate the value of the capacity but also to provide a better approach and simple solution to some complex problems such as the transmit power allocation for OFDM and MIMO systems [4].

Several methods including Monte Carlo simulations [5] and Gauss-Hermite Quadrature integrals [6] had been proposed to calculate the capacity because they give high accuracy, yet they require multiple numerical integrals, and thus are not simple enough. The authors of [7] employed the use of Jensen’s inequality to estimate the channel capacity of AWGN and fading channels through upper and lower bounds. These bounds are quite useful as they simplified the expression of the channel capacity into a closed-form solution, however, they are somewhat loose for AWGN channel. In this work, we propose a novel upper bound for the capacity of BI-AWGN channel. This new bound is much tighter than that obtained by Baccarelli and Fasano [7]. In addition, the upper bounds of fading channels are derived through the moment generating function (MGF) of fading channel gain.

2 System model

Let $X \in \{\pm 1\}$ be the transmitted signals with elements are independent identically distributed (i.i.d) zero-mean binary symbols with equal probabilities. The received signal $Y$ is

$$Y = \sqrt{\Gamma}X + Z,$$

where $Z$ is the zero-mean AWGN with variance 1, i.e. $Z \sim \mathcal{N}(0, 1)$. $\Gamma$ is the channel power gain independent of $X$ and $Z$.

With CSI known at the receiver, the average capacity can be obtained [7] as

$$C^*(\bar{\gamma}) = \int_0^\infty C(\gamma) f_\Gamma(\gamma) d\gamma,$$  \hspace{1cm} (2)

where $f_\Gamma(\gamma)$ is the probability density function (pdf) of $\Gamma$, $\bar{\gamma} = \mathbb{E}[\gamma]$ is the average SNR, and

$$C(\gamma) = \ln 2 - \int_{-\infty}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{\gamma^2}{2}} \ln(1 + e^{-2\sqrt{\gamma}u - 2\gamma}) du$$

in nats per symbol, is the instantaneous capacity when the channel gain is fixed to $\Gamma = \gamma$ [7, eq. (2)].

The upper bound of $C(\gamma)$ obtained by Baccarelli and Fasano [7] using Jensen’s inequality is denoted by
Next, we will present a new upper bound which is 3 dB tighter than $C_{UB1}(\gamma)$ [7].

### 3 New upper bound for the capacity of BI-AWGN channel

The upper bounds for the capacity of BI-AWGN channel can be written as

$$C(\gamma) \leq C_{UB1}(\gamma) = \ln 2 - \ln(1 + e^{-\gamma}). \quad (4)$$

**Proof:** Let

$$g(\gamma) = C_{UB1}(\gamma) - C(\gamma). \quad (6)$$

It is easy to see that $g(0) = g(+\infty) = 0$ and the derivative of $g(\gamma)$ can be written as

$$g'(\gamma) = \frac{e^{-\gamma}}{1 + e^{-\gamma}} - C'(\gamma). \quad (7)$$

It is known that the derivative of $C(\gamma)$ with respect to $\gamma$ is equal to half the minimum mean-square error (MMSE) which is achievable by optimal estimation of the input given the output [8]. That’s to say, $C'(\gamma) = \frac{1}{2} \text{mmse}(\gamma)$. For BI-AWGN channel, the MMSE can be obtained as [8, eq. (17)]

$$\text{mmse}(\gamma) = 1 - \frac{1}{2\pi} \int_{-\infty}^{\infty} e^{-\gamma} \tanh(\gamma - \sqrt{2}\gamma) dy. \quad (8)$$

It can be verified that $\text{mmse}(0) = 1$ and $\text{mmse}(\infty) = 0$, further, $g'(0) = g'(+\infty) = 0$ and $g'(\gamma)$ has only one zero $\gamma_0$ when $\gamma \in (0, +\infty)$. $g(\gamma)$ is monotonically increasing when $\gamma \in (0, \gamma_0)$, and decreasing when $\gamma \in (\gamma_0, +\infty)$, hence $g(\gamma) \geq 0$. 

### 4 Upper bounds for fading channels

By using the Taylor series expansion

$$\ln(1 + e^{-\gamma}) = \sum_{n=1}^{\infty} \frac{-\gamma^{n+1}}{n} e^{-n\gamma}, \quad (9)$$

(5) can be further expressed as

$$C_{UB1}(\gamma) = \ln 2 - \sum_{n=1}^{\infty} \frac{(-1)^{n+1}}{n} e^{-n\gamma}. \quad (10)$$

Putting (10) back into (2), the upper bounds for fading channels can be expressed as

$$C^*(\tilde{\gamma}) \leq \int_{0}^{\infty} \left( \ln 2 - \sum_{n=1}^{\infty} \frac{(-1)^{n+1}}{n} e^{-n\gamma} \right) f_\Gamma(\gamma) d\gamma$$

$$= \ln 2 - \sum_{n=1}^{\infty} \frac{(-1)^{n+1}}{n} \int_{0}^{\infty} e^{-n\gamma} f_\Gamma(\gamma) d\gamma$$

$$= \ln 2 - \sum_{n=1}^{\infty} \frac{(-1)^{n+1}}{n} \mathcal{M}_\gamma(-n), \quad (11)$$

where $\mathcal{M}_\gamma(x)$ is the moment generating function (MGF) of random variable $\gamma$ with distribution $f_\Gamma(\gamma)$, $\gamma \geq 0$. MGF can be obtained by using Laplace transform. The
MGFs of Nakagami-$m$ and Rician fading channels are given in (12) and (13), respectively.

$$\mathcal{M}_m(-n) = \left(1 + \frac{n\gamma}{m}\right)^{-m}$$  \hspace{1cm} (12)

$$\mathcal{M}_n(-n) = \frac{1 + K}{1 + K + n\gamma} e^{-\frac{n\gamma}{K - n\gamma}},$$  \hspace{1cm} (13)

where $m$ is the fading parameter of Nakagami-$m$ distribution and $K$ is the Rician factor. Putting the result of (12) and (13) back into (11), the result obtain is

$$C_{\text{Nakagami}}(\gamma) \leq \ln 2 - \sum_{n=1}^{\infty} \frac{(-1)^{n+1}}{n} \left(1 + \frac{n\gamma}{m}\right)^{-m}$$  \hspace{1cm} (14)

$$C_{\text{Rician}}(\gamma) \leq \ln 2 - \sum_{n=1}^{\infty} \frac{(-1)^{n+1}}{n} \frac{1 + K}{1 + K + n\gamma} e^{-\frac{n\gamma}{K - n\gamma}}.$$  \hspace{1cm} (15)

Note that $m = 1$ in (14) or $K = 0$ in (15) represent the upper bound for Rayleigh fading channel.

**5 Numerical results**

It is easy to see that (5) is 3 dB tighter than (4) obtained in [7] by using Jensen’s inequality. The result of (4) and (5) (in bits/symbol) for AWGN channel are shown in Fig. 1, also with the curves employed by Monte Carlo integration for comparison. We observe from the figure that the proposed upper bound (5) is asymptotically exact both for low and high SNR’s.

Fig. 1. Asymptotically upper bounds for AWGN channel with BPSK.

Fig. 2 shows the upper bounds (in bits/symbol) both for Nakagami-$m$ fading with $m = 2$ and Rician fading with $K = 10$. The bounds obtained by using Taylor series and MGF are quite close to the real capacity of (2) estimated by using Monte Carlo integration over all SNR ranges.
6 Conclusion

In this letter, we develop a simple tight upper bounds for the capacity of BI-AWGN channel. The proposed bound is 3 dB tighter compared to the bound obtained by Baccarelli and et al. Further, the upper bounds for the capacity of fading channels can be obtained by averaging the capacity of AWGN channel with respect to the pdf of fading gain.

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Classification waveform optimization for MIMO radar

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Abstract: The waveform algorithm for active target classification or identification based on spectral variance has been studied in various pieces of literature. The algorithm was extended for widely-separated MIMO Radar system to take advantage of spatial diversity gain by Bae et al. However, the MIMO waveform can be improved further by considering multiple objective functions from the multiple target paths. In this letter, we optimize the MIMO waveform for target identification system by maximizing the multiple objective functions. We show simulation results to compare the proposed algorithm to other MIMO waveform design methods.

Keywords: radar waveform, MIMO, target classification, mutual information, extended target

Classification: Sensing

References


1 Introduction

Goodman et al. derived a classification waveform algorithm for enhanced target classification by defining mutual information based on energy spectral variance (MIESV) across the transfer functions of the various target hypotheses in [1, 2, 3]. Energy spectral variance (ESV) quantifies the statistical variance over a set of finite-duration target transfer functions. The details on the use of MIESV for classification task are described in [2, 3]. Bae et al. extended the MIESV algorithm for multiple-input, multiple-output (MIMO) radar by adding the spectral variance of the monostatic target impulse response and the spectral variances of the bistatic target impulse responses in [4] and showed the diversity gain in MIMO radar simulations.

In a MIMO radar system, multiple waveforms from widely separated radars are transmitted and reflected by a target. Then, the reflected waveforms are captured and combined by multiple radar receivers. Each single waveform is simultaneously captured by multiple radar receivers. For single-input, multiple-output (SIMO) waveform optimization, the single waveform affects the observations of the multiple radar receivers as shown in Fig. 1, and the waveform optimization can be performed by maximizing multiple objective functions based on the multiple observations. In addition, MIESV has the form of mutual information, and a MIMO waveform can be designed from the sum of the multiple mutual information measures. This information measure for each MIMO path is a function of the waveform parameters. Thus, we propose a MIMO classification waveform method based on multi-objective optimization (MO).

2 Monostatic radar model

We consider a stochastic extended target model for a particular target $j$, according to [2, 3]

$$y(t) = w(t) \ast h_j(t) + n(t)$$

where $*$ denotes convolution, the target hypothesis index $j = 1, 2, \ldots, \mathcal{H}$, $y(t)$ is the received observation with duration $T_y$, $w(t)$ is a finite-energy waveform with duration $T_w$, $n(t)$ is zero-mean receiver noise process with power spectral density (PSD) $P_n(f)$, and the random target $h_j(t)$ is a wide-sense stationary process with PSD $S_h(f)$. For a finite-duration stochastic target model, we adopt a finite target model $g_j(t) = a(t)h_j(t)$ where $a(t)$ is a rectangular window function of duration $T_g$ [3]. The frequency-domain system model that results from the Fourier transform is

$$Y(f) = W(f)G(f) + N(f)$$

where $G(f)$ is a finite-energy process with zero-mean [3]. The classification waveform optimization based on MIESV is expressed by

max MIESV
subject to \( \int_B \left| w(f) \right|^2 df \leq E_{\text{max}} \)  \( \text{(2)} \)

where MIESV = \( T_x \int_B \ln \left[ 1 + \frac{\left| w(f) \right|^2 S_{kl}(f) f \right] df \), \( E_{\text{max}} \) is a given waveform energy constraint for \( w(f) \), the ESV for a random target model is \( S_{R}(f) = \sum_{j=1}^{H} Pr(H_j) \sigma_k^2(f) - \left| \sum_{j=1}^{H} Pr(H_j) \sqrt{\sigma_k^2(f)} \right| ^2 \), \( Pr(H_j) \) is the probability of the \( j^{th} \) target hypothesis, and \( \sigma_k^2(f) \) is the spectral energy of the \( j^{th} \) target. Now, let us define a MIMO radar signal model for use in waveform design.

3 MIMO radar model

For MIMO radar model, the observation of the \( k^{th} \) radar receiver due to the waveform from the \( l^{th} \) radar transmitter is defined as

\[ y_{kl}(f) = w_l(f)g_{jk}(f) + n_{kl}(f) \]

where the radar receiver index is \( k = 1, 2, \ldots, N_{\text{Rx}} \), the radar transmitter index is \( l = 1, 2, \ldots, N_{\text{Tx}} \), and \( g_{jk}(f) \) is a finite-energy process with zero-mean [3]. Here, we have \( N_{\text{Rx}} \times N_{\text{Tx}} \) number of observations. Each observation includes one of \( H \) target impulse responses. The MIMO waveform is composed of \( N_{\text{Tx}} \) SIMO waveforms since the multiple waveforms from the different transmit radars are separable in the radar receivers. First, let us consider the SIMO waveform for the 1st radar. Fig. 1 shows that the first radar transmits a waveform, and its reflections are captured by the \( N_{\text{Rx}} \) radar receivers. From this result, an optimization method based on \( N_{\text{Rx}} \) objective functions is derived.

Each of the objective functions is individually defined by an MIESV metric. Thus, the SIMO waveform optimization from the \( l^{th} \) radar is described by multiple-objective optimization problem as

\[ \max \sum_{k=1}^{N_{\text{Rx}}} C_k T_j \int_B \ln \left[ 1 + \frac{\left| w_l(f) \right|^2 S_{kl}(f) \right] df \]

subject to \( \int_B \left| w_l(f) \right|^2 df \leq E_{\text{max}} \)  \( \text{(3)} \)

where \( S_{kl}(f) = \sum_{j=1}^{H} Pr(H_j) \sigma_k^2(g_{jk}(f)) - \left| \sum_{j=1}^{H} Pr(H_j) \sqrt{\sigma_k^2(g_{jk}(f)) \right| ^2 \), \( C_k \) defines the weights of the combination. We can define the weight factors depending on the

Fig. 1. SIMO radar waveform transmission and reception: The 1st radar transmits a waveform, and all radars receive a reflection.
degree of importance of each MIESV_{(k,l)} for a given l. However, in this letter, we set the weighting factors to unity for simplicity.

To enable computer simulation, we now define a discrete optimization problem from (3) as

$$\max \sum_{k=1}^{N_k} T_y \sum_{m=1}^{M} \ln \left[1 + \frac{\Omega_l(m) S_{kl}(m)}{T_y P_n(m)}\right] \Delta f$$

subject to \(\sum_{m=1}^{M} \Omega_l(m) \leq E_w\) (4)

where \(m\) is the index of a discrete frequency bin after the transmit bandwidth has been divided into intervals, \(\Delta f\) is the width of a single frequency bin, \(|w_l(m)|^2 = \Omega_l(m)\), and \(E_w\) is a given waveform energy constraint for \(w_l(m)\). The scalar sum of the multiple objective functions is a concave function since each of the objective functions is \(\ln |\cdot|\) which is concave. The concavity property of the scalar sum of \(\ln |\cdot|\) was also discussed in [5]. Thus, we define a Lagrangian problem and derive the optimization procedure as

$$L(\lambda, \Omega_l(m)) = \sum_{k=1}^{N_k} T_y \sum_{m=1}^{M} \ln \left[1 + \frac{\Omega_l(m) S_{kl}(m)}{T_y P_n(m)}\right] \Delta f + \lambda \left(\sum_{m=1}^{M} \Omega_l(m) - E_w\right).$$

By applying the first derivative to \(L(\lambda, \Omega_l(m))\) with respect to \(\Omega_l(m)\), we obtain

$$\frac{\partial L(\lambda, \Omega_l(m))}{\partial \Omega_l(m)} = \sum_{k=1}^{N_k} T_y \frac{S_{kl}(m)}{1 + \Omega_l(m)} \frac{T_y P_n(m)}{A_k(m)} \Delta f + \lambda$$

$$= \sum_{k=1}^{N_k} T_y \frac{A_k(m)}{1 + \Omega_l(m) A_k(m)} \Delta f + \lambda$$

where \(S_{kl}(f) = \sum_{j=1}^{N} Pr(H_j) \sigma^2_{g_{kl}}(f) - \left|\sum_{j=1}^{N} Pr(H_j) \sqrt{\sigma^2_{\gamma_{kl}}(f)}\right|^2\) and \(A_k(m) = \frac{S_{kl}(m)}{T_y P_n(m)}\). For the maximum value of the linear sum of multiple objective functions, let us apply \(\frac{\partial L(\lambda, \Omega_l(m))}{\partial \Omega_l(m)} = 0\) to get

$$\frac{\partial L(\lambda, \Omega_l(m))}{\partial \Omega_l(m)} = 0 \Rightarrow \sum_{k=1}^{N_k} \frac{A_k(m)}{1 + \Omega_l(m) A_k(m)} = \frac{\lambda}{T_y \cdot \Delta f}.$$

The optimal \(\Omega_l(m)\) can be numerically calculated according to the equations

$$\sum_{m=1}^{M} \Omega_l(m) = E_w \text{ and } \sum_{k=1}^{N_k} \frac{A_k(m)}{1 + \Omega_l(m) A_k(m)} = \lambda$$

(5)

where \(\lambda = \frac{\lambda}{T_y \cdot \Delta f}\), \(\Omega_l(m)\), and \(\lambda\) are inversely proportional. This optimization procedure is an iterative water-filling algorithm [5]. Finally, we can obtain the energy spectrum of the \(l^{th}\) optimal SIMO waveform, \(W_l = [w_l(1), w_l(2), \ldots, w_l(M)]^T\) by \(|w_l(m)|^2 = \Omega_l(m)\). For the MIMO waveform matrix \(W\), we calculate multiple individual SIMO waveforms as \(W = [W_1, W_2, \ldots, W_{Tx}]\).

4 Simulation result

In this section, we present simulation results showing the performance of the multi-objective optimization algorithm (MO). For the computer simulations, the random
target signatures are generated in two different ways. The first model sets have ESV’s generated from colored spectra. The second model sets have ESV’s generated from flat (white) spectra. Also, we use the following system parameters. The number of target hypotheses \( H \) is 4, the discrete spectral waveform dimension \( M \) is 40, the measurement noise power \( \sigma^2 \) is normalized to 1, and the waveform energy allocation varies from \( 10^{-4} \) to \( 10^1 \) energy units. The probability of detection is calculated over 20,000 Monte Carlo trials. Based on the given system parameters, we evaluate the probability of detection in determining the true target transfer function for three different waveform algorithms and compare their performances in Figs. 2 and 3. The first waveform is a wideband waveform having a flat energy distribution across the transmission band, the second waveform is the waveform based on the sum of spectral variance (SSV) of [4], and the last waveform is the MO waveform that we propose in this letter. From the results, MO algorithm shows the best performance among the three waveform algorithms both in the cases of colored and non-colored target models. The results of Fig. 2 shows wider performance gain compared to that of Fig. 3 due to the more structured target spectra.

Fig. 2. Performance comparison in 4x4 MIMO radar system with colored target models

Fig. 3. Performance comparison in 4x4 MIMO radar system with non-colored target models
5 Conclusion

We derived a MIMO waveform for classification via mutual information based on energy spectral variance. The waveform was optimized by the multi-objective optimization for the widely-separated MIMO radar model. The proposed method showed the best results in computer simulations, including simulations based on both flat and colored spectral models. As another advantage of the proposed algorithm, the optimization procedure is performed by an iterative water-filling algorithm whose computational load is very light. In this letter, we set the weighting factors $C_k$ of (3) to unity for simplicity. However, it is also an interesting problem to find the best combination of the weighting factors since the factors are depending on the degree of importance of each MIESV metric and are related with the MIMO radar channel. This will be our future research topic.

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