Low complexity precoding matrix selection method for Long-Term Evolution systems

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Abstract: In a Long-Term Evolution system, a transmitter can appropriately transmit layered data streams via multiple antennas with the help of a precoding matrix. The precoding matrix can be selected using the feedback precoding matrix indicator information from the receiver. Different precoding matrixes will affect the system capacity and bit error rate performance dramatically. In this paper, we propose a low complexity precoding matrix selection method based on trace calculation of the MIMO channels’ eigenvalue. The method will have almost the same capacity as the maximum mutual information (MI) method, as well as the low computational complexity advantage. Simulation results show that the proposed method requires less than 20% calculations needed by the maximum MI method.

Keywords: MIMO, LTE, Precoding Matrix Indicator (PMI)

Classification: Wireless Communication Technologies

References


1 Introduction

Long-Term Evolution Advanced (LTE-A), which is defined by the 3rd Generation Partnership Project (3GPP), has been deployed widely as the fourth-generation (4G) wireless communication system owing to its high spectral utilization and throughput performance. The LTE-A system adopts the orthogonal frequency division multiple access (OFDMA) and multiple input multiple output (MIMO) techniques to achieve high-speed service for the served user equipment (UE).

To cope with time-varying fading variations of the wireless channels, the transmitter (Tx) usually uses feedback channels to report current channel state information (CSI) and adjusts transmit parameters such as modulation, code rate, spatial layer, and adopted precoding matrix. In the LTE system, CSI can be acquired at the Tx in terms of three primary parameters, i.e., rank indicator (RI), precoding matrix indicator (PMI), and channel quality indicator (CQI). These parameters can help the Tx to dynamically select the appropriate parameters for the signals to achieve better and reliable transmission. Therefore, deciding on appropriate and efficient selection of feedback information is an important issue for LTE systems.

Among the various pieces of feedback information, the precoding matrix serves to appropriately distribute the layered data streams transmitted by the multiple Tx antennas. Different precoding matrix selections will dramatically influence system capacity and system bit error rate (BER) performance. For the 3GPP LTE system, the two-transmit antennas, four-transmit antennas, and eight-transmit antennas precoding matrices are well-defined [1].

Besides, many literatures have provided methods for determining the appropriate PMI for LTE systems. In [2], the authors described a precoding matrix selection method based on mutual information (MI) maximization. This method has very good capacity performance; however, the associated computational complexity is high, especially because of the intensive signal to noise ratio (SNR) and MI calculations. In [3], the authors reported a low complexity precoding matrix selection method based on the maximization of the equivalent channel gains. In this paper, we propose a low complexity precoding matrix selection method involving maximization of the trace of the eigenvalue of the MIMO channels. The proposed method will have almost the same capacity ability as the maximum MI method [2], but it will be characterized by lower computational complexity compared with the literate methods. The proposed method and the compared literature methods are described in Section 2. Performance analysis of these methods, including system capacity and computational complexity, is reported in Section 3. The conclusions of this paper are summarized in Section 4.

2 Method descriptions

The signal model considered in this paper is shown in Fig. 1. The $N_x \times 1$ received signal vectors $y_k$ at the $k$th subcarriers of the OFDM signal can be expressed as follows:

$$y_k = H_k W_k x_k + n_k$$ (1)
where $H_k$ is the $N_r \times N_t$ MIMO channel matrix; $N_t$ is the number of transmit antennas; $N_r$ is the number of receiving antennas; $W_i$ is the $N_t \times 1$ precoding matrix to be determined; $x_k$ is the $l \times 1$ transmitted signal vector, where $l$ is the spatial layer during transmission; and $n_k$ is the $N_r \times 1$ additive white Gaussian noise (AWGN). Assume that the spatial layer has been determined before the PMI selection. After equalization, the output signal is

$$r_k = G_k y_k = G_k H_k W_i x_k + G_k n_k$$  \hspace{1cm} (2)$$

where $G_k$ is the equalizer compensation, which can be expressed by Eqs. (3) and (4) for the minimum mean square error (MMSE) and zero forcing (ZF) criteria, respectively [4].

$$G_{ZF} = [(H_k W_i)^H (H_k W_i)]^{-1} (H_k W_i)^H$$  \hspace{1cm} (3)$$

$$G_{MMSE} = [(H_k W_i)^H (H_k W_i) + \left( \frac{N_r N_0}{E_s} \right) I]^{-1} (H_k W_i)^H$$  \hspace{1cm} (4)$$

2.1 Maximum mutual information selection [2]

In [2], the authors reported a precoding matrix selection method based on maximizing the Shannon capacity criteria of the transmission. The signal to interference and noise ratio (SINR) after the equalizer processing can be expressed as follows:

$$SINR_{l,k} = \frac{|K_k(l, l)|^2}{\sum_{i \neq l} |K_k(l, i)|^2 + \sigma_n^2 \sum_l G_k(l, i)}$$ \hspace{1cm} (5)$$

$$K_k = G_k H_k W_i$$ \hspace{1cm} (6)$$

where $k$ is the subcarrier index, and $l$ and $i$ are the indexes for the row and column, respectively. It is well-known that the channel capacity is the maximum of the MI between the transmitted signal and the equalized output. The maximum MI of a fading channel can be given by the Shannon theorem as follows:

$$I_k = \sum_{l=1}^{L} \log_2 (1 + SINR_{l,k})$$ \hspace{1cm} (7)$$

where $L$ is the number of spatial layers of the LTE system. The precoding matrix can be selected using Eq. (8).

$$W_i = \arg \max_{W \in W} \sum_{k=1}^{K} I_k(W_i)$$ \hspace{1cm} (8)$$

Clearly, many calculation steps are required to determine the SINR from $K_k$ and $G_k$ obtained using Eq. (5). Therefore, the same paper described a low complexity
selection version as well by replacing Eq. (7) with Eq. (9) to avoid the SINR calculations and reduce computational complexity.

\[ I_k(W_i) = \log_2 \det \left( I + \frac{1}{\sigma_n^2} W_i^H H_k H_k^H W_i \right) \] (9)

### 2.2 SVD-based selection [3]

In [3], the authors adopted singular value decomposition (SVD) for selecting the appropriate PMI. Let the SVD of the channel matrix be

\[ H = UDV^H. \] (10)

where \( U \) and \( V \) are the unitary matrices with dimensions \( N_r \times N_t \) and \( N_t \times N_t \), respectively. \( D \) is a diagonal matrix. Let the characteristic of the equalizer at Rx be \( G_k = U^H \), and let the precoding matrix at Tx be \( W_i = V \). Then, the received signal can be expressed as follows:

\[ Y = U^H W_i X = U^H UDV^H V X = DX \] (11)

\( X \) is the transmitted signal. Define an equivalent channel (EC) as follows:

\[ P(W_i) = U^H W_i = D \text{tr}^H W_i \] (12)

Then, the suitable PMI for this method can be determined using Eq. (13):

\[ W_j = \arg \max_{W_i \in W} \left| \sum_{j=1}^{L} P(W_j)(j, j) \right| \] (13)

\( L \) is the number of spatial layers. Basically, the method tries to find a precoding matrix that can achieve the best equivalent channel gain during the ongoing MIMO transmission.

### 2.3 Proposed PMI selection method

In the proposed method, the channel capacity of the MIMO system can be expressed as follows:

\[ C(H) = \log_2 \det \left( I_{N_r} + \frac{\rho}{N_t} H W_i^H H^H \right) \]

\[ = \sum_{i=1}^{N_t} \log_2 \left( 1 + \frac{\rho}{N_t} \lambda_i \right) \] (14)

where \( \lambda_i \) is the eigenvalue of matrix \( HH^H \). It can be easily shown that the eigenvalues share the following relation with the channel matrix:

\[ \text{tr}(H W_i^H H^H) = \sum_{i=1}^{N_t} \lambda_i \] (15)

where \( \text{tr}(\cdot) \) is the trace operator. The proposed method selects the precoding matrix according to Eq. (16) to ensure the maximum channel capacity for the MIMO transmission.

\[ W_j = \arg \max_{W_i \in W} \{ \text{tr}(H W_i^H H^H) \} \] (16)

Theoretically, \( \lambda_i \) can be viewed as the energy gain of the \( i \)th channel. Therefore, the proposed method finds the precoding matrix with the best channel gain.
summation of the MIMO channels (Eqs. 15 and 16). It is shown in Section 3 that the proposed method has almost the same capacity performance as the maximum MI method, but with computational complexity lower than the other methods in the literature.

3 Computer simulations

In this section, the channel capacity performance and computational complexity of the proposed method are evaluated and compared with the corresponding values of other methods in the literature. In addition, it is assumed that the channel estimation can be acquired perfectly at the Rx. The MIMO channels are assumed to be identical, independent distributed complex Gaussian random variables. The antenna numbers at the Tx and Rx are 4 and 3, respectively. The MMSE equalizer is used in this simulation.

Fig. 2 shows the system capacity performance of different methods. From the results, the SVD-based method has the worst capacity performance, whereas the maximum MI method, its simplified version, and the proposed method have almost identical capacity performance. In Eq. (14), it is clear that the channel capacity depends mainly on the eigenvalues of the MIMO channel matrix. The proposed method uses the trace of the $HW_iW_i^H$ which is directly related to the eigenvalues in Eq. (15) to select the precoding matrix. For the SVD method, it has a worse result than the other methods since that it adopts the equivalent channel to conduct the precoding matrix selection rather than the eigenvalues of the channel.

![Fig. 2. Capacity performance.](image)

However, the maximum MI method is plagued with high computational complexity, even the simplified version, which is further analyzed in the following text. The major contribution of the proposed method is the large reduction in computational complexity during PMI selection. Let $l = \min(N_t, N_r)$. Table I lists the computational complexity values of the methods compared herein. If we insert
the simulation parameters in the calculations and set the calculations required by
the maximum MI method to 100%, the proposed method needs only 19.5% of the
calculations used by the maximum MI method. The SVD method has almost the
same low complexity advantage. However, the proposed method has better capacity
performance than the SVD method, as shown in Fig. 2. Therefore, when consid-
ering both the capacity performance and the computational complexity in a
practical implementation, the proposed method will be the best choice among the
methods discussed herein.

### 4 Conclusions

In this paper, we proposed a low complexity precoding matrix selection method for
LTE systems. Without loss of any channel capacity, the proposed method offers
the advantage of avoiding intensive PMI calculations at the Rx. Based on the
simulation results, the proposed method requires less than 20% of the calculations
used by the maximum MI method. Consequently, Rx can more speedily inform the
Tx via the feedback channels, thus reducing the impact of the time delay due to the
time-varying fading variations of the MIMO channels.

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<th>Example</th>
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<td>[ \sum_{l=1}^{L} { N_l l N_r l (N_r - 1) + \hat{P}_l l N_r l + \hat{P}_l (N_r - 1) + \hat{P}_l + l ] + [ \hat{P}_l N_r l + \hat{P}_l (l - 1) + \hat{P}_l N_r l + \hat{P}_l (N_r - 1) + l + 2(l - 1) ]</td>
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<tr>
<td>Low complexity maximum mutual information selection</td>
<td>2096</td>
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<tr>
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<tr>
<td>SVD-based selection</td>
<td>1776</td>
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<tr>
<td>[ { N_l l N_r l (N_r - 1) + N_l^2 l + N_l l (N_r - 1) + N_r } ]</td>
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<tr>
<td>Proposed selection</td>
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<tr>
<td>[ { N_l l N_r l (N_r - 1) + \hat{P}_l N_r l + \hat{P}_l (N_r - 1) + N_r } ]</td>
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Scaling law of distributed estimation in sensor networks with semi-orthogonal MAC

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Abstract: This letter is concerned with distributed estimation in a Gaussian sensor network operating under a semi-orthogonal multiple-access channel (MAC). In such a semi-orthogonal MAC, all sensors are divided into groups and signals from the sensors in each group are directly combined (opposed to being coherently combined) and then transmitted on one orthogonal channel to a fusion center (FC). The behavior of the average distortion in terms of the number of sensors, namely the scaling law, is analyzed. The analytical derivation and simulation results show that the semi-orthogonal MAC with adaptive sensor grouping can achieve the optimal scaling law while requiring only a small amount of feedback from the FC to the sensors.

Keywords: sensor networks, distributed estimation, multiple-access channel, scaling law

Classification: Network

References

1 Introduction

This letter considers the problem of distributed estimation in a Gaussian sensor network in which a scalar Gaussian random variable is observed in a memoryless fashion by $K$ sensors and each observation is subject to white Gaussian noise. The sensors are connected to a fusion center (FC) via wireless channels. Using data collected from the sensors, the FC estimates the underlying source signal to within the smallest distortion possible. The distortion of such distributed estimation is affected by many factors, such as the number of sensors, the total transmit power and the communication bandwidth. Of particular interest in our study is the behavior of the average distortion as a function of the number of sensors, i.e., the scaling law as defined in [1]. In particular, [1, Theorem 1] tells us that the optimal scaling law achieved by analog transmission is $1/K$.

Regarding the wireless communications between the sensors and the FC, many different types of multiple access channel (MAC) can be used. The coherent MAC [1, 2, 3] is an example of analog transmission which can achieve the optimal scaling law of $1/K$. In the coherent MAC, signals from all sensors are transmitted on a single channel. To obtain coherent combination at the FC, the phases of the wireless channel responses need to be compensated at the sensors, which entails a large amount of channel state information (CSI) feedback from the FC to the sensors. In another analog transmission scheme, namely the orthogonal MAC [4], since orthogonal channels are used by different sensors, no channel phase compensation is required at the sensors. However, reference [4] shows that the orthogonal MAC does not achieve the optimal scaling law.

Recently, a novel analog transmission scheme, known as the semi-orthogonal MAC with sensor grouping, is studied in [5]. Such a scheme offers a better tradeoff among the bandwidth, feedback overhead, and estimation performance than both the coherent and orthogonal MACs. As a further development of [5], this letter establishes the important scaling law achieved by the semi-orthogonal scheme with both fixed and adaptive sensor groupings. In particular, it is proved that the semi-orthogonal MAC with adaptive sensor grouping can achieve the optimal scaling law of $1/K$, even with less feedback overhead as compared to the coherent MAC.

2 System model

In a distributed Gaussian sensor network proposed in [5], $K$ sensors are used to monitor a source signal $s$ and communicate their observations to a FC over $N$ orthogonal channels. All sensors are divided into disjoint groups and the sensors within each group are allocated to transmit on one orthogonal channel. Since the consumed bandwidth is proportional to the number of the orthogonal channels, $N$ shall be chosen to be much smaller than $K$ to conserve bandwidth.

The observation of the $i$th sensor can be expressed as $x_i = s + v_i$, $i = 1, 2, \ldots, K$, where the source signal $s$ and observation noise $v_i$ are modeled as Gaussian random variables with zero mean and variances $\sigma_s^2$ and $\sigma_v^2$, respectively. The observation signal-to-noise ratio (SNR) is defined as $\gamma_i = \sigma_s^2 / \sigma_v^2$.

With analog transmission, the $i$th sensor simply amplifies $x_i$ with a gain $a_i$ and transmits the result to the FC on one orthogonal channel. Let $h_i$, $i = 1, \ldots, K$,
represent the channel response from sensor $i$ to the FC. These channel responses are modeled as i.i.d. complex Gaussian random variables with zero mean and unit variance. At the FC, the received signal on the $n$th orthogonal channel is $y_n = \sum_{i \in \Omega_n} a_i(s + v)h_i + \omega_n$, $n = 1, 2, \ldots, N$, where $\Omega_n$ is the index set of the sensors that transmit on the $n$th orthogonal channel, and $\omega_n$'s are the i.i.d. (over $n$) complex AWGN components with zero mean and variance $\sigma^2_n$.

The received signal can be rewritten as $y_n = \tilde{h}_n s + \tilde{v}_n + \omega_n$, where $\tilde{h}_n = \sum_{i \in \Omega_n} a_i h_i$ and $\tilde{v}_n = \sum_{i \in \Omega_n} a_i v_i h_i$. Since $y_n$ is complex, while $s$ is real, the phase of $\tilde{h}_n$ is compensated to obtain $\tilde{y}_n = R\{\tilde{h}_n y_n/|\tilde{h}_n|\} = \tilde{h}_n s + \tilde{v}_n + \bar{\omega}_n$, where $\tilde{v}_n = |\sum_{i \in \Omega_n} a_i h_i|$, $\bar{\omega}_n = R\{\tilde{h}_n^* \omega_n/|\tilde{h}_n|\}$ and $\tilde{y}_n = R\{\tilde{h}_n^*(\sum_{i \in \Omega_n} a_i v_i h_i)/|\tilde{h}_n|\}$. The above phase compensation discards halves of observation noise and channel noise. Since it is performed at the FC, no phase information is needed at the sensors and feedback of CSI from the FC to the sensors is not required.

Let $\tilde{y} = [\tilde{y}_1, \tilde{y}_2, \ldots, \tilde{y}_N]$, $\tilde{h} = [\tilde{h}_1, \tilde{h}_2, \ldots, \tilde{h}_N]$, $\tilde{v} = [\tilde{v}_1, \tilde{v}_2, \ldots, \tilde{v}_N]$ and $\bar{\omega} = [\bar{\omega}_1, \bar{\omega}_2, \ldots, \bar{\omega}_N]$. Then one has $\tilde{y} = \tilde{h} s + \tilde{v} + \bar{\omega}$. The LMMSE estimator is adopted at the FC. Accordingly, the estimate of $s$ based on $\tilde{y}$ is [6]

$$\hat{s} = \sigma_s^{-2}\tilde{h}^T (\sigma_s^2 \tilde{h} \tilde{h}^T + \Sigma_v + \Sigma_\omega)^{-1}\tilde{y},$$

where

$$\Sigma_v = \mathbb{E}\{\tilde{v}\tilde{v}^T\} = \text{diag}\left(\sum_{i \in \Omega_n} a_i^2 \sigma^2_{r_{ni}}, n = 1, \ldots, N\right),$$

$$r_{ni} = (R\{h_i\} R\{\tilde{h}_n\} + I\{h_i\} I\{\tilde{h}_n\})/|\tilde{h}_n|,$$

$$\Sigma_\omega = \mathbb{E}\{\bar{\omega}\bar{\omega}^T\} = \text{diag}\left(\frac{\sigma^2_{\omega_1}}{2}, \frac{\sigma^2_{\omega_2}}{2}, \ldots, \frac{\sigma^2_{\omega_N}}{2}\right).$$

The corresponding MSE distortion is

$$[\sigma_s^{-2} + \tilde{h}^T (\Sigma_v + \Sigma_\omega)^{-1}\tilde{h}]^{-1} = \left[\sigma_s^{-2} + \sum_{n=1}^N \frac{|\sum_{i \in \Omega_n} a_i h_i|^2}{\sum_{i \in \Omega_n} a_i^2 r_{ni}} \right]^{-1}.$$  

(5)

### 3 Performance analysis

First, consider equal power allocation, i.e., $a_i = \bar{a} = \sqrt{P_{\text{tot}}/[K(\sigma_s^2 + \bar{\sigma}_r^2)]}$. This simplifies (5) to

$$\epsilon = \sigma_s^2 \left[1 + \sum_{n=1}^N \frac{|\sum_{i \in \Omega_n} h_i|^2}{\sum_{i \in \Omega_n} a_i^2 r_{ni}} \right]^{-1} \left[\frac{1}{\gamma_c} + \frac{1}{\gamma_c^*} (1 + \frac{1}{\gamma_c^*})\right]^{-1},$$

where $\gamma_c = P_{\text{tot}}/\bar{\sigma}_r^2$ is the channel SNR. In the following, for simplicity, assume $\sigma_s^2 = 1$ and define $\gamma_c^* = \gamma_c/(1 + \frac{1}{\gamma_c^*})$.

The main objective of this letter is to establish the scaling law achieved by the semi-orthogonal MAC. The scaling law, as defined in [1], is the decaying rate that the average MSE (AMSE) distortion achieves as $K$ increases. In the following, the AMSE of different cases of semi-orthogonal MAC is examined in more detail.

### 3.1 Semi-orthogonal MAC with fixed sensor grouping

With sensor grouping, $N$ is fixed to a small value to save bandwidth. For fixed sensor grouping (FSG), $K$ sensors are divided into $N$ disjoint groups such that

1. For complex quantities, $R\{\cdot\}$ denotes the real part and $I\{\cdot\}$ denotes the imaginary part.
2. For random variables, $\mathbb{E}\{\cdot\}$ denotes expectation.
\(\Omega_n = \{(n-1)(K/N) + 1, \ldots, n(K/N)\}\) and the sensors in each group transmit on one orthogonal channel. In this case, \(K\) is an integer multiple of \(N\). The strategy of FSG means that the orthogonal channel used by each sensor, once assigned, does not change during the communication phase. As a result, no feedback of channel information is required in this scheme. Based on (6), a lower bound on the AMSE of the semi-orthogonal MAC with FSG can be found as

\[
\mathcal{E}^\star \geq \mathcal{E}' \left( 1 + \sum_{n=1}^{\frac{N}{\xi}} \left| \frac{\sum_{i \in \Omega_n} h_i}{\sqrt{K}} \right|^2 \right)^{-1}
\]

(7)

Since \(\sum_{i \in \Omega_n} h_i/\sqrt{K}\) is a complex Gaussian random variable with zero mean and variance \(1/N\), the lower bound turns to

\[
\mathcal{E}^\star \geq \left( 1 + \sum_{n=1}^{\frac{N}{\xi}} \frac{1}{\frac{N}{\xi}} \right)^{-1} = (1 + 2\xi/C)^{-1}
\]

(8)

The above lower bound is independent of \(K\), which means that the optimal scaling law of \(1/K\) cannot be achieved by the semi-orthogonal MAC with FSG.

### 3.2 Semi-orthogonal MAC with adaptive sensor grouping

For the adaptive sensor grouping (ASG), the sensors are grouped based on the phases of their channel responses. To this end, the whole phase region of \(2\pi\) is partitioned into \(N\) equal sub-regions (each of length \(\frac{2\pi}{N}\)), and the sensors with channel phases in the same sub-region are assigned to transmit on the same orthogonal channel. In this case, feedback of orthogonal channel allocation from the FC to the sensors is required. For each sensor, \(\log_2 N\) bits are needed, which is a much smaller amount of feedback compared to the CSI required in the coherent MAC, especially when \(N\) is small. The scaling law achieved by the semi-orthogonal MAC with ASG is stated in the following theorem.

**Theorem 1.** The AMSE achieved by the semi-orthogonal MAC with ASG scales like \(1/K\) when \(K \to \infty\), i.e., \(\lim_{K \to \infty} KE^\star = c\), for some constant \(c > 0\).

**Proof.** First, Equation (6) is rewritten in the following form:

\[
\epsilon \geq \left( 1 + \sum_{n=1}^{\frac{N}{\xi}} \left( \frac{\sum_{i \in \Omega_n} \xi_i}{\sqrt{K}} \right)^2 \left( \frac{\sum_{i \in \Omega_n} \eta_i}{\sqrt{K}} \right)^2 \right)^{-1}
\]

(9)

where \(\xi_i\) and \(\eta_i\) are, respectively, the real and imaginary parts of \(h_i\), and \(K_n\) is the size of \(\Omega_n\). When \(K \to \infty\), \(K_n\) approaches to its mean value \(K/N\) (which is also infinity). Then it follows from the strong law of large numbers that when \(K_n \to \infty\) one has \(\sum_{i \in \Omega_n} \xi_i/K_n \xrightarrow{a.e.} \mathcal{E}(\xi_i)\), \(\sum_{i \in \Omega_n} \eta_i/K_n \xrightarrow{a.e.} \mathcal{E}(\eta_i)\) and \(\sum_{i \in \Omega_n} \xi_i^2/K_n \xrightarrow{a.e.} \mathcal{E}(\xi_i^2)\). It then follows that

\[
\epsilon \xrightarrow{a.e.} \left( 1 + \sum_{n=1}^{\frac{N}{\xi}} \left( \frac{\mathcal{E}(\xi_i^2) + \mathcal{E}(\eta_i^2)}{\frac{N}{\xi} \mathcal{E}(\xi_i^2) + \frac{N}{\xi} \mathcal{E}(\eta_i^2)} \right) \right)^{-1}
\]

(10)

As \(K \to \infty\), all sensor groups have identical distributions of channel responses. Therefore, (10) turns to
\[ \epsilon \xrightarrow{a.e.} \left[ 1 + \frac{N K}{\pi} \frac{\mathcal{E}^2(x_i) + \mathcal{E}^2(y_i)}{\mathcal{E}(r_{ni}^2) \frac{1}{\gamma_n} + \frac{1}{\pi N}} \right]^{-1} = \left[ 1 + \frac{K}{\pi} \frac{\mathcal{E}^2(x_i) + \mathcal{E}^2(y_i)}{\mathcal{E}(r_{ni}^2) \frac{1}{\gamma_n} + \frac{N}{\pi \gamma_n}} \right]^{-1}. \] (11)

For the ASG, the channel responses in sensor group \( n \) have i.i.d. distribution of
\[ \frac{N}{\pi} \exp[-(x_i^2 + y_i^2)], \]
where \( \theta_i \) is the phase of \( h_i \). Based on this distribution function, one has
\[ \mathcal{E}(x_i) = (N/2\sqrt{\pi}) \cos \alpha \sin \beta, \quad \mathcal{E}(y_i) = (N/2\sqrt{\pi}) \sin \alpha \sin \beta, \quad \mathcal{E}(x_i^2) = (N/4\pi) \cos 2\alpha \sin 2\beta + 1/2, \quad \mathcal{E}(y_i^2) = -(N/4\pi) \cos 2\alpha \sin 2\beta + 1/2 \quad \text{and} \quad \mathcal{E}(x_i y_i) = (N/4\pi) \sin 2\alpha \sin 2\beta, \]
where \( \alpha = (2n - 1)\pi/N \) and \( \beta = \pi/N \). In addition,
\[ \mathcal{E}(r_{ni}^2) = \mathcal{E}\left( (x_i \cos \varphi_n + y_i \sin \varphi_n)^2 \right) = \mathcal{E}(x_i^2) \cos^2 \varphi_n + \mathcal{E}(y_i^2) \sin^2 \varphi_n + 2\mathcal{E}(x_i y_i) \cos \varphi_n \sin \varphi_n, \]
where \( \varphi_n \) is the phase of \( \hat{h}_n \).

It is easy to prove that when \( K \to \infty \), \( \varphi_n \) approaches \( \mathcal{E}(\theta_i) = (2n - 1)\pi/N \) with probability 1. Thus,
\[ \mathcal{E}(r_{ni}^2) \xrightarrow{a.e.} \frac{1}{2} + \frac{N}{4\pi} \sin\left(\frac{2\pi}{N}\right). \]
Therefore, when \( K \to \infty \),
\[ \epsilon \xrightarrow{a.e.} \left[ 1 + K \frac{\frac{N^2}{\pi^2} \sin^2\left(\frac{2\pi}{N}\right)}{\frac{1}{2} + \frac{N}{4\pi} \sin\left(\frac{2\pi}{N}\right)} \right]^{-1}. \] (14)

Finally,
\[ \lim_{K \to \infty} K \mathcal{E}(\epsilon) = \left[ \frac{1}{2} + \frac{N}{4\pi} \sin\left(\frac{2\pi}{N}\right) \right] \frac{1}{\gamma_n} + \frac{N}{\pi \gamma_n}, \]
which is a constant. \( \square \)

The analysis of the scaling laws in this section is valid for equal power allocation. If the optimal power allocation\(^3\) can be found, whether the optimal scaling law of \( 1/K \) can be achieved by the semi-orthogonal MAC with FSG remains to be answered and deserves a further study. On the other hand, for the semi-orthogonal MAC with ASG, the optimal scaling law has already been achieved with the equal power allocation and surely it is also achieved with the optimal power allocation.

\[ ^3\text{The optimal power allocation divides the total power among sensors to obtain the smallest instantaneous MSE. Such a technique requires extra feedback from the FC to sensors.} \]

4 Simulation results

In all simulations, it is assumed that \( \gamma_0 = 20 \text{ dB} \) and \( \gamma_c = 25 \text{ dB} \). Fig. 1 plots the AMSEs of the orthogonal MAC and the semi-orthogonal MAC with FSG versus the number of sensors, \( K \). As \( K \) increases, the AMSEs of both MACs asymptotically converge to positive constants. For the semi-orthogonal MAC with FSG, the lower bound of \( (1 + 2\gamma_c)^{-1} \) obtained in (8) is quite loose for small values of \( K \) and \( N \), but it becomes tighter when \( K \) and \( N \) get larger.

The scaling law achieved by the semi-orthogonal MAC with ASG is illustrated in Fig. 2, where the AMSEs of the orthogonal and coherent MACs are also provided for comparison. Similar to the coherent MAC, the AMSE of the semi-orthogonal MAC with ASG appears as a straight line when \( K \) is large. Since the plots are in log-log fashion, a straight line means that the AMSE decays in an order of \( 1/K \) as \( K \) increases, showing that the optimal scaling law of the studied analog
Gaussian sensor networks is achieved. In addition, it can be shown that the constant $c$ of the ASG scheme (Eq. (15)) is larger than that of the coherent MAC. This means that the ASG scheme has a higher distortion compared to the coherent MAC with the same number of sensors. Therefore, while the semi-orthogonal MAC with ASG is as optimal as the coherent MAC in the scaling-law sense, it requires more sensors than the coherent MAC to achieve the same distortion.

5 Conclusions

The AMSE distortions of distributed estimation under two versions of the semi-orthogonal MAC, namely fixed sensor grouping and adaptive sensor grouping, are analyzed under equal power allocation. The semi-orthogonal MAC with FSG requires no feedback from the FC to sensors but fails to achieve the optimal scaling law. In contrast, for the semi-orthogonal MAC with ASG, the optimal scaling law of $1/K$ can be achieved. This means that the estimation distortion can be decreased to an arbitrary low level by employing more sensors. Compared to the coherent MAC, such optimal scaling law of the semi-orthogonal MAC with ASG can be achieved with a much smaller amount of feedback.
Beam scanning capability and suppression of endfire radiation of dipole array antennas coupled to two-wire parallel transmission line

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Abstract: In this letter, radiation pattern of a dipole array antenna coupled to a two-wire transmission line is numerically investigated. The effect of array spacing on the main beam direction of the array antenna is shown. In addition, it is shown that the endfire radiation of the array antenna can be suppressed when series inductors are loaded with the two-wire transmission line.

Keywords: array antenna, leaky wave antenna, beam scanning

Classification: Antennas and Propagation

References


1 Introduction

In recent years, antennas for the next-generation wireless communication system have received much attention. Beamwidth of such antennas is expected to be narrower than that for the existing wireless communication systems because high frequency band such as millimeter wave band will be used for the systems. In addition, propagation loss in high frequency band is much higher than microwave and main beam must be radiated directly from a transmitting antenna to a receiving antenna. Therefore, beam scanning capability is necessary for antennas of the next-generation wireless communication systems using the high frequency band.

Previously, various beam scanning antennas have been reported. A microstrip array antenna and transmit array with diodes have been reported as antennas with electronic beam scanning capability [1, 2]. On the other hand, antennas with mechanical beam scanning capability have received much attention [3]. One of the advantages of such antennas is low loss because its beam scanning capability is realized without using lossy microwave components such as diodes or phase shifters. Such low loss performance of the antennas with mechanical beam scanning capability can be a significant advantage when the antennas are used for the next generation wireless communication systems because the systems suffer from high propagation loss. The disadvantage of the antennas with mechanical beam scanning capability is slow-speed scanning. Therefore, the antennas with mechanical beam scanning capability can be useful for the next-generation indoor wireless communication systems in high frequency band because high-speed scanning is not necessary for tracking the movement of mobile terminals in an indoor environment. A dipole array antenna coupled electromagnetically to a two-wire transmission line is one of these antennas [4]. In previous researches, mutual coupling between a dipole antenna and two-wire transmission line has been investigated experimentally [5]. In addition, radiation pattern of the array antenna has been clarified experimentally [6] and analytically [7]. However, to the best of our knowledge, the array antenna with beam scanning capability has not been reported so far.

In this letter, a dipole array antenna coupled to a two-wire transmission line with beam scanning capability is proposed. Results of numerical simulation show that the beam direction of the proposed antenna can be scanned when its array spacing changes. It is shown that the endfire radiation of the antenna can be suppressed when series inductors are loaded with the two-wire transmission line.
Numerical simulation

A dipole array antenna coupled to a two-wire transmission line is shown in Fig. 1. The antenna operates at \( f = 20 \) GHz. Dipole elements have the same length \( l \) and are separated with equal spacing \( d_x \). Dimensions of the two-wire transmission line are designed so that characteristic impedance becomes 50 Ω using following formula.

\[
Z_0 = \frac{1}{\pi} \sqrt{\frac{\mu_0}{\varepsilon_0}} \ln \frac{W - a}{a}.
\]

(1)

A voltage source is applied to the two-wire transmission line terminated in load resistance \( Z_L = 50 \) Ω through a series resistance \( Z_s = 50 \) Ω. The Richmond’s method of moments (MoM) was used for numerical simulation [8]. Conductivity of antennas and two-wire transmission lines is \( \sigma = 5.8 \times 10^7 \) S/m.

Fig. 2 shows radiation patterns of the array antenna when array spacing \( d_x = 0.5, 0.6, 0.7, 0.8 \) ι. It is found that the main beam direction of the array antenna depends on its array spacing. As shown in [9], main beam direction of a leaky-wave antenna can be obtained as follows.

\[
\sin \theta_m \approx \frac{\beta_n}{k_0} \quad \text{where} \quad \beta_n = \beta_0 + n \frac{2\pi}{d_x},
\]

(2)

where \( \theta_m \) is main beam direction of the leaky-wave antenna and \( \beta_n \) is the propagation constant of the \( n \)th space harmonic. \( \beta_0 \) is the propagation constant in the two-wire transmission line and agrees with the wavenumber of free space \( k_0 \) because the TEM mode is dominant in the two-wire transmission line. Eq. (2) shows that endfire radiation shown in Fig. 2 corresponds to the dominant TEM mode and the scanned beam corresponds to \( n = -1 \)th space harmonic. The main beam is radiated in \( x \leq 0 \) because \( \beta_{-1} \leq \beta_0 \). \( \theta_m \) calculated by Eq. (2) is 42 deg. at
dx = 0.6λ, 25 deg. at dx = 0.7λ and 15 deg. at dx = 0.8λ, respectively. On the other hand, θm calculated by MoM is 41 deg. at dx = 0.6λ, 27 deg. at dx = 0.7λ and 14 deg. at dx = 0.8λ, respectively. It is found that the main beam directions obtained by Eq. (2) agree well with those obtained by MoM.

Beam scanning capability of the array antenna was clarified but endfire radiation which is comparable to its main beam level can be observed in Fig. 2. For practical applications, this undesired endfire radiation must be suppressed as much as possible. In order to suppress the undesired endfire radiation, a dipole array antenna coupled to a two-wire transmission line inserted inductors are proposed. Propagation constant of the electromagnetic wave propagating on the two-wire transmission line is expected to be modified when these inductors are loaded with the two-wire transmission line. All of these inductors are 0.01 nH and loaded with the two-wire transmission line as a lumped element in the middle of all adjacent two dipole elements. An example of radiation pattern of the array antenna is shown in Fig. 3 when array spacing dx = 0.6λ. As shown in this figure, it is found that the endfire radiation of the array antenna is suppressed when inductors are loaded with the two-wire transmission line. It is thought that the propagation constant β0 corresponding to the dominant TEM mode which contributes the endfire radiation is modified by inductors. As a result, main beam direction is tilted when inductors are loaded with the two-wire transmission line because β0 is the function of β0 as shown in Eq. (2). Even when the array spacing is changed mechanically, endfire radiation still can be suppressed without moving inductors because β0 is almost independent of the array spacing.
3 Conclusion

The dipole array antenna coupled to the two-wire transmission line with beam scanning capability was proposed and studied by numerical simulation. It was shown that array spacing between dipole elements affects beam direction of the array antenna. In addition, it was found that the main beam direction of the array antenna can be switched as a function of the array spacing. High level radiation at the endfire direction of the array antenna was suppressed when inductances are loaded with the two-wire transmission line.

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A reduced-complexity near-ML detection for broadband single-carrier transmission

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Abstract: In our recently proposed reduced-complexity near-maximum likelihood block signal detection employing QR decomposition and M-algorithm (QRM-MLBD) for broadband single-carrier (SC) block transmission, unreliable symbol candidates are removed from the symbol tree based on the hard-decision minimum mean square error based frequency-domain equalization (MMSE-FDE). In this paper, to further remove the unreliable symbol candidates and consequently reduce the computational complexity, we introduce a soft-decision MMSE-FDE based selection of reliable symbol candidates into reduced-complexity QRM-MLBD. It is shown that the soft-decision MMSE-FDE based selection significantly reduces the computational complexity while achieving a bit error rate (BER) performance similar to the original QRM-MLBD.

Keywords: single carrier, maximum likelihood detection, QR decomposition, M-algorithm, MMSE-FDE

Classification: Wireless Communication Technologies

References


1 Introduction

The BER performance of broadband SC transmission severely degrades due to strong inter-symbol interference (ISI). Although it has been shown that the QRM-MLBD can greatly improve the BER performance of SC block transmission [1, 2], it requires high computational complexity. To remedy this problem, we proposed the complexity-reduced QRM-MLBD [3]. However, its complexity reduction is limited.

In this paper, we introduce a soft-decision MMSE-FDE based selection of the reliable symbol candidates [4] into our reduced-complexity QRM-MLBD for further complexity reduction. We evaluate, by computer simulation, the average BER performance of SC transmissions achievable with the reduced-complexity QRM-MLBD utilizing the soft-decision MMSE-FDE and discuss its computational complexity.

2 Principle of reduced-complexity QRM-MLBD [3]

This section summarizes the principle of reduced-complexity QRM-MLBD proposed in Ref. [3]. Throughout the paper, the $T_s$ symbol-spaced discrete time representation is used. Assuming the single-input single-output cyclic prefix (CP) inserted SC block transmission same as Ref. [3], the frequency-domain received signal $\mathbf{Y} = [Y(0), \ldots, Y(k), \ldots, Y(N_c-1)]^T$, in which $N_c$ is the size of discrete Fourier transform (DFT), is given as

$$\mathbf{Y} = \sqrt{\frac{2E_s}{T_s}} \mathbf{H} \mathbf{d} + \mathbf{N},$$

where $E_s$ is the symbol energy, $\mathbf{H} = \text{diag}[H(0), \ldots, H(k), \ldots, H(N_c-1)]$ is the $N_c \times N_c$ frequency-domain channel matrix, $\mathbf{d} = [d(0), \ldots, d(t), \ldots, d(N_c-1)]^T$ is the data symbol block, $\mathbf{F}$ is the $N_c \times N_c$ DFT matrix whose $(i,j)$ component is $(1/\sqrt{N_c}) \exp(-j2\pi(i \times j)/N_c)$, and $\mathbf{N} = [N(0), \ldots, N(k), \ldots, N(N_c-1)]^T$ is the noise vector whose elements are the independent zero-mean Gaussian variables having the variance $2N_0/T_s$ with $N_0$ being the one-sided power spectrum density of the additive white Gaussian noise (AWGN).

In the reduced-complexity QRM-MLBD, MMSE-FDE is performed before QRM-MLBD. The MMSE-FDE output $\tilde{\mathbf{d}} = [\tilde{d}(0), \ldots, \tilde{d}(t), \ldots, \tilde{d}(N_c-1)]^T$ and its $t$-th component can be respectively represented as

$$\tilde{\mathbf{d}} = \mathbf{F}^H \mathbf{W} \mathbf{Y},$$

$$\tilde{d}(t) = \sqrt{\frac{2E_s}{T_s}} \left( \frac{1}{N_c} \sum_{k=0}^{N_c-1} \tilde{H}(k) \right) d(t) + \mu_{\text{ISI}}(t) + \mu_{\text{noise}}(t),$$

where $(\cdot)^H$ is the Hermitian transpose operation, $\tilde{H}(k) = W(k)H(k)$, and $\mathbf{W} = \text{diag}[W(0), \ldots, W(k), \ldots, W(N_c-1)]$ is the MMSE weight matrix given as [5]

$$\mathbf{W} = \text{diag} \left[ \begin{array}{ccc} H^*(0) & \vdots & H^*(N_c-1) \\ |H(0)|^2 + N_0/E_s & \cdots & |H(N_c-1)|^2 + N_0/E_s \end{array} \right]$$

In Eq. (3), the first, second, and third terms are the desired signal, residual ISI, and noise, respectively. $\tilde{d}(t)$ is normalized by $\sqrt{2E_s/T_s(1/N_c)} \sum_{k=0}^{N_c-1} \tilde{H}(k) = P_{\text{norm}}$ as
\[ \tilde{d}(t) = d(t) + \mu_{ISI}(t) + \mu_{\text{noise}}(t), \]  

where

\[ \left\{ \begin{array}{l}
\mu_{ISI}(t) = P_{\text{norm}}^{-1} \mu_{ISI} = P_{\text{norm}}^{-1} \sqrt{\frac{2E_s}{T_s}} \frac{1}{N_c} \sum_{k=0}^{N_s-1} \tilde{H}(k) \left\{ \frac{1}{N_c} \sum_{n=0, n \neq t}^{N_s-1} d(n) \exp \left( j2\pi k \frac{t-n}{N_c} \right) \right\} \\
\mu_{\text{noise}}(t) = P_{\text{norm}}^{-1} \mu_{\text{noise}} = P_{\text{norm}}^{-1} \frac{1}{N_c} \sum_{k=0}^{N_s-1} W(k)N(k) \exp \left( j2\pi k \frac{t}{N_c} \right)
\end{array} \right. \]  

Approximating \( \mu_{ISI}(t) \) as a zero-mean complex Gaussian variable from the central limit theorem, \( \mu_{ISI}(t) + \mu_{\text{noise}}(t) = \mu(t) \) can be treated as a new zero-mean complex Gaussian variable with the variance \( \sigma_{\mu}^2 \) being

\[ \sigma_{\mu}^2 = \frac{1}{2} E[|\mu(t)|^2] = \sigma_{ISI}^2 + \sigma_{\text{noise}}^2, \]  

where

\[ \left\{ \begin{array}{l}
\sigma_{ISI}^2 = P_{\text{norm}}^{-2} \frac{E_s}{T_s} \left\{ \frac{1}{N_c} \sum_{k=0}^{N_s-1} |\tilde{H}(k)|^2 \right\} + \left\{ \frac{1}{N_c} \sum_{k=0}^{N_s-1} \tilde{H}(k) \right\}^2 \\
\sigma_{\text{noise}}^2 = P_{\text{norm}}^{-2} \frac{1}{N_c} \frac{N_0}{T_s} \sum_{k=0}^{N_s-1} |W(k)|^2
\end{array} \right. \]  

Assuming that all symbols are transmitted with equal probability, the a posteriori probability (APP) of symbol candidate \( c_i \) (\( i = 0 \sim X - 1 \) with \( X \) being the modulation level) for the given \( \tilde{d}(t) \) can be expressed using Bayes’ theorem as

\[ P(c_i|\tilde{d}(t)) = \frac{1}{\sum_{n=0}^{X-1} p(\tilde{d}(t)|c_n)} \frac{1}{2\pi \sigma_{\mu}^2} \exp(-|\tilde{d}(t) - c_i|^2/2\sigma_{\mu}^2) \]

\[ = \frac{1}{\sum_{n=0}^{X-1} p(\tilde{d}(t)|c_n)} \sum_{n=0}^{X-1} \left\{ \frac{1}{2\pi \sigma_{\mu}^2} \exp(-|\tilde{d}(t) - c_n|^2/2\sigma_{\mu}^2) \right\}, \]

where \( p(\tilde{d}(t)|c_n) \) is the conditional probability of \( \tilde{d}(t) \). In the reduced-complexity QRM-MLBD, the selection of reliable symbol candidate based on the APP distribution is performed. Removing the unreliable symbol candidates from the symbol tree before QRM-MLBD, the complexity for tree search reduces.

In Ref. [3], the APP distribution after hard-decision MMSE-FDE is used. However, the symbol candidates which are located at the same distance from hard-decision MMSE-FDE results have the same APP and therefore, the selection of reliable symbol candidates is fixed according to hard-decision MMSE-FDE results and the complexity reduction is limited.

3 Soft-decision MMSE-FDE based selection of reliable symbol candidates for reduced-complexity QRM-MLBD

In this paper, the soft-decision MMSE-FDE based selection of reliable symbol candidates [4] is used in the reduced-complexity QRM-MLBD. Differently from the hard-decision MMSE-FDE based selection, the soft-decision MMSE-FDE based selection can variably select the reliable symbol candidates by adapting to
the APP distribution after soft-decision MMSE-FDE. As shown in Fig. 1(a), the symbol candidates are selected in descending order of APP until the accumulated AAP (AAPP) exceeds the threshold $\alpha (0 \leq \alpha \leq 1)$. The number of remaining symbol candidates associated with the $t$-th symbol $d(t)$ is denoted by $N_{\text{cand}}(t)$. In Sect. 4, we show the effect of $\alpha$ on the symbol error rate (SER).

After the reliable symbol candidate selection, only one symbol candidate may remain for $d(t)$, i.e., $N_{\text{cand}}(t) = 1$. In this case, $d(t)$ having $N_{\text{cand}}(t) = 1$ is detected by hard decision after MMSE-FDE. Such symbols can be removed from the symbol tree and the corresponding column components are removed from an $N_c \times N_c$ equivalent channel matrix $\mathbf{H}_F = \mathbf{H}$ as shown in Fig. 1(b-2). Denoting $C$ as the number of $d(t)$ having $N_{\text{cand}}(t) = 1$ in $d$, the size of equivalent channel matrix reduces from $N_c \times N_c$ to $N_c \times (N_c - C)$, thereby the computational complexity of QR decomposition reduces. The depth of the symbol tree also reduces from $N_c$ to $N_c - C$. Note that in Ref. [3], $N_{\text{cand}}(t)$ is always more than 1 and the computational complexity of QR decomposition cannot be reduced.

In each stage of the M-algorithm, $M$ paths having the smallest accumulated path metrics (i.e., accumulated squared Euclidean distance between the received signal and the symbol candidates) survive [1, 2]. When $M$ is small, the BER performance of QRM-MLBD degrades. In the QRM-MLBD, the equivalent channel is transformed to an upper triangular matrix. The lower-right elements of the upper triangular matrix, which are associated with the symbols closer to the end of symbol block, get weaker and the accuracy of path metrics in earlier stages lowers [2]. Consequently, when $M$ is small, pruning the correct path in earlier stages occurs with a high probability. To remedy this problem, the $N_{\text{cand}}(t)$-based ordering [4] is performed. The symbols are reordered in descending order of $N_{\text{cand}}(t)$ and the symbol tree is reconstructed as shown in Fig. 1(b-3). Since the symbols having smaller $N_{\text{cand}}(t)$ are gathered in earlier stages, the number of paths in earlier stages becomes smaller. Hence, the BER performance improves even if small $M$ is used.

**Fig. 1.** Complexity reduction operations in reduced-complexity QRM-MLBD using the soft-decision MMSE-FDE based selection of reliable symbol candidates.
Next, QRM-MLBD is performed. After removing \(d(t)\) having \(N_{\text{cand}}(t) = 1\) from the symbol tree and performing the \(N_{\text{cand}}(t)\)-based ordering, the frequency-domain received signal is expressed as

\[
\hat{Y} = \sqrt{\frac{2E_s}{T_s}} \hat{H} \hat{d} + N_t, \tag{10}
\]

where \(\hat{H}\) is the \(N_c \times (N_c - C)\) transformed equivalent channel matrix and \(\hat{d}\) is the \((N_c - C) \times 1\) symbol vector. QR decomposition is applied to \(\hat{H}\) as

\[
\hat{H} = \hat{Q} \hat{R}, \tag{11}
\]

where \(\hat{Q}\) is the \(N_c \times (N_c - C)\) matrix satisfying \(\hat{Q}^H \hat{Q} = I_{N_c - C}\) with \(I_{N_c - C}\) being the \((N_c - C) \times (N_c - C)\) identity matrix and \(\hat{R}\) is the \((N_c - C) \times (N_c - C)\) upper triangular matrix. Then, the transformed frequency-domain received signal \(\hat{Z} = [\hat{Z}(0), \ldots, \hat{Z}(k), \ldots, \hat{Z}(N_c - 1 - C)]^T\) is obtained as

\[
\hat{Z} = \hat{Q}^H \hat{Y} = \sqrt{\frac{2E_s}{T_s}} \hat{R} \hat{d} + \hat{Q}^H N_t, \tag{12}
\]

Finally, the tree search using the M-algorithm is performed. The tree search is significantly simplified by selecting the reliable symbol candidates only, removing \(d(t)\) having \(N_{\text{cand}}(t) = 1\) from the symbol tree, and performing the \(N_{\text{cand}}(t)\)-based ordering.

4 Computer simulation

We assume a SC block transmission with 16QAM, \(N_c = 64\), and the CP length is \(N_g = 16\). The channel is assumed to be a frequency-selective block Rayleigh fading channel having symbol-spaced \(L = 16\)-path uniform power delay profile (PDP). Ideal channel estimation is assumed.

First, we discuss the effect of \(\alpha\) on the error probability. The value of \(\alpha\) is an important design parameter which affects the trade-off between the error probability and the computational complexity. Fig. 2(a) shows the average SER of QRM-MLBDs with \(M = 256\) as a function of \(\alpha\) at the average received \(E_b/N_0 = 14\) dB (\(E_b/N_0 = (E_b/N_0)(1 + N_g/N_c)/4\) for 16QAM). The original QRM-MLBD requires \(M = 256\) for achieving the average BER = 10\(^{-5}\) with a less than 1 dB \(E_b/N_0\) degradation from the matched-filter (MF) bound [5] as seen from Fig. 2(c). Note that removal of \(d(t)\) having \(N_{\text{cand}}(t) = 1\) from the symbol tree and \(N_{\text{cand}}(t)\)-based ordering are not applied in Fig. 2(a) (our proposed QRM-MLBD consists of the soft-decision MMSE-FDE based selection of reliable symbol candidates, removal of \(d(t)\) having \(N_{\text{cand}}(t) = 1\) from the symbol tree and \(N_{\text{cand}}(t)\)-based ordering). In the following simulation, for each average received \(E_b/N_0\), \(\alpha\) is set to satisfy that the SER increase caused by the symbol candidate selection is less than 10\% from the original QRM-MLBD with \(M = 256\). For example, \(\alpha = 0.9992\) is set when the average received \(E_b/N_0 = 14\) dB.

Fig. 2(b) plots average \(N_{\text{cand}} = (1/N_c) \sum_{t=0}^{N_c-1} N_{\text{cand}}(t)\) and \(C\) as a function of the average received \(E_b/N_0\), obtained by 1 million simulation runs. The value of \(\alpha\) for each average received \(E_b/N_0\) are also shown. It is seen from Fig. 2(b) that average \(N_{\text{cand}}\) reduces from \(N_{\text{cand}} = 16\) for original QRM-MLBD. Fig. 2(b) also shows that many stages can be removed from the symbol tree and hence, the
complexity of QR decomposition and the tree search can be reduced. Finally, Fig. 2(c) shows the average BER performance comparison. It is seen from Fig. 2(c) that the proposed QRM-MLBD with $M = 64$ achieves almost the same BER performance as the original QRM-MLBD with $M = 256$. Note that in Ref. [3], the average $N_{\text{cand}} = 6.25$ and always $C = 0$, and $M = 256$ is required to achieve almost the same BER performance as the original QRM-MLBD.

In Table I, the average number of complex multiply operations is compared. The proposed QRM-MLBD requires computations of MMSE-FDE and the var-

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Fig. 2. Computer simulation results.
iance of residual ISI plus noise, selection of reliable symbol candidates, and removal of \( d(t) \) having \( N_{\text{cand}}(t) = 1 \) from the symbol tree, as well as QR decomposition. However, the complexity of QR decomposition and the path metric computation in the proposed QRM-MLBD can be reduced. Therefore, the computational complexity of proposed QRM-MLBD is greatly reduced compared to the original QRM-MLBD. For example, the proposed QRM-MLBD requires only 22% (26%) of computational complexity of original QRM-MLBD (complexity-reduced QRM-MLBD in Ref. [3]) when the average received \( E_b/N_0 = 14 \) dB.

5 Conclusion

The soft-decision MMSE-FDE based selection of reliable symbol candidate for reduced-complexity QRM-MLBD was proposed. Applying the soft-decision MMSE-FDE based symbol selection, the more number of unreliable symbol candidates can be removed from the symbol tree compared to Ref. [3]. Additionally applying removal of \( d(t) \) having \( N_{\text{cand}}(t) = 1 \) from the symbol tree and \( N_{\text{cand}}(t) \)-based ordering, our proposed reduced-complexity QRM-MLBD greatly reduces the computational complexity from original QRM-MLBD while achieving almost the same BER performance. In this paper, the parameter \( \alpha \) which affects the trade-off between the SER/BER and the computational complexity was predetermined for each average received \( E_b/N_0 \) by the preliminary computer simulation. The adaptive setting of \( \alpha \) is left as an important future study.
Countermeasure against fingerprinting attack in Tor by separated contents retrieval

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Abstract: Tor (The Onion Router) realizes that anonymous web surfing without revealing the user’s identity. However, A. Panachenko et al. reveals that an onion router that directly communicates with a user can infer which website a user accesses by leveraging site-specific traffic features, e.g., volume and time, and this attack is called the fingerprinting attack. In this paper, we propose a countermeasure against the fingerprinting attack by obfuscating site-specific traffic features. The idea is to establish two distinct Tor connections and to separately request text-based contents and image-based one through them. We show the effectiveness of our scheme with experiments.

Keywords: anonymous communication, Tor, fingerprinting attack

Classification: Internet

References

1 Introduction

Safe browsing on the Internet is crucial to preserve privacy for us. In order to realize it, anonymous communication techniques have been extensively studied [1, 2]. Anonymous communication realizes that a user retrieves web contents without revealing his/her identity to a recipient. Especially, Tor is one of the measure implementation of anonymous communication in the Internet [3]. Fig. 1 shows an example that a user communicates with a web server through Tor. A client, who wants to anonymously communicate with a web server, chooses multiple ORs (Onion Router), exchanges a symmetric key with each OR, and connects with a web server via each OR. By sequentially encrypting packets by each symmetric key, any OR cannot identify a website that a client accesses and it realizes anonymous communication.

However, A. Panachenko et al. reveals that an entry OR, that directly communicates with a user, can infer which website a user accesses with high accuracy. They collect site-specific traffic features, e.g., volume, time, and direction of the traffic, and use SVM (Support Vector Machine) to identify browsed websites. This attack is called the fingerprinting attack [4, 5, 6].

The conventional countermeasures try to request dummy requests to obfuscate the traffic features [4, 5]. However, they increase traffic volumes which are not desired from the network aspect. Therefore, it is necessary to propose a countermeasure against a fingerprinting attack without increasing unnecessary traffic.

In this paper, we propose a countermeasure against the fingerprinting attack by obfuscating site-specific traffic features. The idea is to establish two Tor connections and to separately request text-based contents and image-based contents through them. We show the effectiveness of our scheme with experiments.

2 Attacker model

We define the attacker model dealt in this paper. The aim of an attacker is to identify websites that a victim browses. The attacker’s procedure consists of two phases: (1) collecting fingerprints of websites and (2) identifying a website. In the following, we describe each phase in detail.
2.1 Fingerprints collection phase

An attacker accesses websites that a user likely to access to record traffic features by each website. Popular websites can be obtained with Alexa1, which discloses the top accessed websites. The collected traffic features are as follows:

- \( S^p_{\text{total}} \): the total amount of packets from \( p \) to \( q \) (Bytes)
- \( N^p_{\text{total}} \): the total number of packets from \( p \) to \( q \)
- \( S^p_{\text{avg}} \): the average packet size from \( p \) to \( q \) (Bytes)
- \( S^p_{\text{var}} \): the variance of packet size from \( p \) to \( q \) (Bytes)
- \( C^p_{\text{avg}} \): the average chunk size from \( p \) to \( q \) (Bytes)
- \( C^p_{\text{var}} \): the variance chunk size from \( p \) to \( q \) (Bytes)

\( p \) and \( q \) denote either a client \( c \) or a web server \( w \), respectively. Therefore, totally 12 (= 6 \times 2) features are used and we call a set of them \((S^c_{\text{total}}, S^c_{\text{total}}, \ldots, C^c_{\text{var}})\) as a fingerprint. The chunk size denotes cumulative amount of packets until a client (or a website) receives a packet from the other side.

2.2 Identification phase

After collecting fingerprints for popular websites, an attacker setups an OR, which we call a malicious OR. When the malicious OR is chosen as an entry OR of a client, the attacker tries to identify which site the client browses. In order to do that, the attacker records a fingerprint in the same way as the previous phase. Then, the attacker calculates the similarity between the recorded fingerprint and each fingerprint collected in the previous phase. Finally, the attacker judges the highest similar website as the website that the client browses.

3 Proposed scheme

Here, we propose a countermeasure against the fingerprinting attack by connecting with two distinct Tors and separately retrieving a text-based content (such as an HTML file) and other contents (such as an image file) with them. Fig. 2(a) shows a concept of our scheme. At first, a web browser only retrieves and shows text-based HTML contents through a Tor connection. Then a user clicks desired non-text contents, i.e., images and videos, and a web browser retrieves them through another Tor connection. By doing this, our scheme makes a malicious OR infeasible to identify an accessed website from collected traffic features, since it may only contain only text-based contents or non-text-based contents and timing to retrieve contents is also obfuscated. We argue that it is unlikely for a client to choose two different entry ORs controlled by an attacker.

3.1 Implementation

We explain a way to implement the proposal in a client system. In order to realize that a client can retrieve images on demand, when a text-based HTML is retrieved, let a web browser attach a HTML code to show a button to retrieve an image above the image location. Fig. 2(b) and 2(c) show an implementation of an image retrieval button.

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1http://www.alexa.com/topsites
In the current release of Tor\(^2\), a client cannot simultaneously boot up multiple Tor clients. Therefore, we boot up a virtual machine on a client and establish another Tor connection to retrieve image contents. Retrieved images are saved in a shared directory that is accessible by the host OS (Operating System). Finally, a web browser on a host OS displays a web page by combining text-based contents and images on the shared directory.

4 Evaluation

In order to show the effectiveness of our scheme, we compare attacker’s detection accuracy against the real website traffic information between (1) with our scheme and (2) with no countermeasure. We evaluate the attacker’s success rate \( r_i \) for a website \( i \) and the overall success rate \( R \) averaged over \( i \) and they are defined as follows:

\[
    r_i = \frac{N_{\text{success}}(i)}{N_{\text{trial}}(i)}, \quad R = \frac{1}{N_{\text{sites}}} \sum_{i} N_{\text{success}}(i)
\]

where \( N_{\text{success}}(i) \), \( N_{\text{trial}}(i) \), \( N_{\text{sites}} \) denote the number of websites that successfully identified by attackers, the number of trials that an attacker tries to identify, and the

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\(^2\)https://www.torproject.org/
number of websites, respectively. In this evaluation, we set $N_{\text{sites}} = 100$, $N_{\text{trial}}(i) = 10$ for $i \in [1, 100]$ if not stated otherwise.

For evaluation, a client accesses the top $N_{\text{sites}} = 100$ websites from Alexa by two minutes with and without the proposed system and collects fingerprints. We repeat the same procedures by $N_{\text{trial}}(i) = 10$ times for each. An attacker accesses the same 100 websites by ten times for each, collects fingerprints, and identifies browsed websites by the procedure described in Section 2. We assume that an attacker against the proposed scheme knows a client only retrieves text-based contents or image-based contents but cannot retrieve both of them because a client is unlikely to choose two entry ORs from ORs controlled by the same attacker. We evaluate it on a computer that operates Windows 7 Professional, Firefox 28.0 as a web browser, and Tor v0.2.3.25.

At first, we evaluate attacker’s success rate when a client only obtains HTML contents. Fig. 3 shows a CDF (Cumulative Density Function) versus the success rate $r_i$. CDF quickly approaches 1 against a better approach since it means that most of websites are incorrectly identified. As we can see from Fig. 3, our scheme clearly decreases success rate against without any countermeasure. The overall success rate $R$ is 57% for no countermeasure approach while 35% for our scheme (with HTML), respectively. Therefore, by separately retrieving contents, we can decrease the attacker’s success rate by 22% on average.

We pay attention to 20 websites whose $r_i \geq 0.9$ if no countermeasure is taken. That is, we focus on websites that an attacker can easily identify. Here we also compare $R$ of (1) no countermeasure, (2) our scheme with only HTML, and (3) our scheme with only images, respectively. Here, ‘our scheme with only images’ denotes that an attacker knows that a client retrieves only images through his/her OR. Assuming the worst case, a client retrieves the same images as the attacker chooses in the training phase but the order and timing of images retrieval may differ. In collecting image contents, an attacker uses the same web browser that can
separately retrieve text contents and images and only 50% of images for a website. Our scheme with HTML and our scheme with images both decrease $R$ from 93% to 31% and 53%, respectively. Again, our schemes much decrease the attacker’s success rate by 40%–62%. We can also see that $R$ is higher when an attacker observe fingerprints calculated from only images. This is because the size of images is much larger than that of texts and the fingerprints calculated from images can be more characteristic. In addition, we assume that an attacker can retrieve the same images that a client does and thus it is easier for an attacker to infer websites.

5 Conclusion

We have proposed a countermeasure against the fingerprinting attack by separately retrieving website contents with two Tor connections. The aim is to obfuscate site-specific traffic features measured by an attacker. We implement our scheme with a computer and show that our scheme effectively decreases the attacker’s overall detection accuracy by approximately 22%.
Effect of miniaturization of the unit cell of a meta-surface using double-layered FSS

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Abstract: This paper proposes a meta-surface for a miniaturized unit cell that consists of a double-layered Frequency Selective Surface (FSS) and the ground plane. Additional miniaturization of the unit cell is achieved by sliding one layer of the double-layered FSS in the polarization direction of an incident wave. Moreover, the unit cell becomes smaller than half the size of a conventional single-layered meta-surface by using the double-layered FSS.

Keywords: meta-surface, AMC, PMC, FSS, double-layered FSS

Classifications: Antennas and Propagation

References


1 Introduction

A meta-surface that realizes controlled distributions of a reflection phase has been studied [1]. A meta-surface that reflects an electromagnetic wave without phase rotation at a specific frequency is called an Artificial Magnetic Conductor (AMC). A low-profile and high-gain antenna is realized by using an AMC reflector [2, 3].

A meta-surface can be constructed easily by placing a frequency selective surface (FSS) on the ground plane [4]. The FSS is a planar structure that realizes filtering characteristics such as the low-pass and band-pass and is constructed by arranging metal elements periodically at short intervals [5]. The unit of the periodic structure is called a unit cell.

In the application of a reflector of antennas, the unit cell must be small to enable flexibility in placing the reflector, and preventing antenna characteristics from deteriorating [6].

We propose a meta-surface structure using a double-layered FSS to miniaturize a unit cell [7]. In addition, further miniaturization of the unit cell is achieved by sliding one layer of the double-layered FSS in the polarization direction of an incident wave. The dependence of the miniaturization effect on the polarization of an incident wave is also clarified. Moreover, the unit cell becomes smaller than half the size of the conventional single layered meta-surface by using a double-layered FSS.

2 Proposed structure of meta-surface using double-layered FSS

Fig. 1 shows the structure of the meta-surface using a double-layered patch-type FSS. The meta-surface consists of the double-layered patch-type FSS and ground plane. The structure is determined by four parameters: the unit cell size of the FSS $p$, the patch length $l$, the distance between the FSSs $t$, and the distance between the lower FSS and ground plane $h$. Here, the single layer of the FSS (FSS-2) is stationary, and the other FSS layer (FSS-1) can slide. The slide distances along the $x$ and $y$ directions are labeled as $d_x$ and $d_y$, respectively. The state at which each metal element of the FSS overlaps completely, as shown in Fig. 1(a), is defined as $d_x = 0$ and $d_y = 0$. The alternated structure, as shown Fig. 1(b), is similarly defined as $d_x/p = 0.5$ and $d_y/p = 0.5$. 
3 Effect of single layer sliding in double-layered FSS

Fig. 2(a) shows the reflection phase characteristics of the meta-surface. The solid and dotted lines denote the results of the meta-surface using the double-layered FSS and single-layered FSS, respectively. Here, $p = 0.10\lambda_0$, $l = 0.09\lambda_0$, $t = 0.01\lambda_0$, and $h = 0.15\lambda_0$.

The slide distances $d_x$ and $d_y$ are 0. The meta-surface using the single-layered FSS is constructed by removing the upper FSS, FSS-1, shown in Fig. 1. The frequency is normalized by $f_0$. In other words, the frequency with the reflection phase value equals 0° in the meta-surface when using the single-layered FSS. When the frequency with a reflection phase of 0° is focused, the solid line appears in a lower frequency. Reducing the frequency means that the unit cell of the meta-surface is miniaturized electrically.

A case involving relative sliding on the other surface of the FSS is also discussed. Here, this surface is assumed to slide along the $x$ direction only. The relationship between the slide distance $d_x$ and the frequency with a reflection phase value equal to 0° is shown in Fig. 2(b). The horizontal axis is used to denote the slide distance along the $x$ direction. Here, the slide distance in the $y$ direction is fixed at $d_y = 0.0$. The vertical axis is the frequency with a reflection phase of 0° $f_{PMC}$. The solid and dotted lines show corresponding results in cases where the
incident wave is polarized along the $x$ and $y$ axes, respectively. The miniaturized effect of the unit cell size varies based on the polarization direction. When the polarization direction corresponds to the slide direction of FSS-1, the reflection phase characteristics are affected by the slide distance. This is because the variation in the structure caused by sliding FSS-1 is observed by means of electricity. In addition, we demonstrate that the frequency with a reflection phase equal to 0° appears in the lowest frequency side, that is, at $d_y/p = 0.5$. However, when the polarization crosses to the slide direction of FSS-1, the reflection phase is unaffected because the variation of the structure is invisible.
Based on the aforementioned phenomenon, we demonstrate that the structure, as shown Fig. 2(b), can be used to obtain the miniaturized effect in both polarizations.

4 Miniaturized effect of unit cell by distance between FSSs

The frequency with a reflection phase of 0° appears at the lower frequency side. This means that the unit cell of the meta-surface is miniaturized electrically. We clarify the effect of the distance between FSS-1 and FSS-2 on the frequency with a reflection phase value of 0°. Here, the meta-surface uses the double-layered FSS, and the slide distances $d_x/p$ and $d_y/p$ are 0.5. Fig. 3 shows the effect of the unit cell size $p$ of the meta-surface and distance $t$ between the FSSs on the frequency with a reflection phase value of 0°. The vertical and horizontal axes denote the variables $p$ and $t$, respectively, and the color indicates $f_{PMC}$. When red, $f_{PMC}$ is the same frequency with when the meta-surface uses a single-layered FSS. Thus, we demonstrate that the unit cell becomes smaller than half the size of the unit cell when a single-layered FSS is used. Even if the unit cell size is changed, the reflection phase can be 0° at the same frequency by setting the appropriate distance between the FSSs. We show that the miniaturization of the meta-surface is designed by setting the distance between the FSSs as short as possible.

Fig. 3. Relationship between unit cell size and distance between FSSs on $f_{PMC}$

5 Conclusion

A meta-surface using a double-layered FSS was proposed. We showed that an electrical miniaturization of the unit cell size of the meta-surface was possible by using double-layered FSS and by sliding one side in relation to the other. In addition, we clarified that the miniaturization effect depends on the polarized
direction of the incident wave. The effect was obtained by sliding one side of the FSS in relation to the other in the same direction as the incident electric field. In addition, it was shown that the miniaturization effect of the unit cell size is affected by the distance between the FSSs. We clarified that the miniaturized unit cell is designed by setting the distance between the FSSs as short as possible. Consequently, the unit cell becomes smaller than half the size of the conventional single-layered meta-surface when using a double-layered FSS.
A meander branch antenna for MIMO antenna decoupling

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Abstract: MIMO technology has been widely used for several years for the purpose of improving communication speed and communication capacity. However, if MIMO antenna is mounted closely, radiation efficiency and throughput of MIMO are decreased due to strong mutual coupling. In this paper, a decoupling method using meander branch antenna is proposed. It is confirmed that the mutual coupling is reduced and the radiation efficiency is increased by utilizing this method. Moreover, further developing this method, it is confirmed that a decoupling bandwidth can be enhanced.

Keywords: MIMO, mutual coupling, branch antenna, radiation efficiency

Classification: Antennas and Propagation

1 Introduction
In recent years, MIMO technology \cite{1} has been introduced in various small wireless terminals for the purpose of an increase in communication speed and communication capacity. However, if MIMO antenna is mounted on limited space such as mobile phones, a strong mutual coupling occurs because a distance between antenna elements is close. The strong mutual coupling causes lower radiation efficiency and decreased throughput of MIMO. Therefore, a decoupling method of MIMO antenna is required.

Decoupling methods for $2 \times 2$ MIMO have been developed. Among them, the decoupling method using branch shape elements without utilizing circuit components \cite{2} could be applied to the small wireless terminals. This conventional
method utilized the following principle. When $|Y_{21}|$ of antennas is nearly zero at desired frequency, the mutual coupling of the antennas is reduced [2]. Here, the frequency at which the real and imaginary part of $Y_{21}$ changed is defined as “inflection point”. Regular monopole such as Fig. 1(a)-(1) has one inflection point. However, a branch shape antenna has two inflection points. As a result, frequency of $|Y_{21}| = 0$ was appeared between two inflection points. Then, adjusting the branch shape element length as the desired frequency is sandwiched between two inflection points, mutual coupling of the antennas was reduced because it obtained $|Y_{21}| = 0$ at desired frequency.

However, conventional report only discussed at higher frequency than inflection point of own antenna, it did not discuss at lower frequency than it. Especially, it has been demanded to reduce mutual coupling at lower frequency because small cellular antenna which has higher inflection point has been widely used for several years.

Then, this paper shows a branch antenna with meander structure. This meander branch antenna enables decoupling condition $|Y_{21}| = 0$ and small antenna size. Moreover, we report that optimizing its length allows decoupling bandwidth to spread.

The decoupling method and analysis models are shown in Section 2. The result of utilizing this method and decoupling bandwidth enhancement are shown in Section 3. Finally, the conclusion on this study is presented in Section 4.

2 Decoupling method and analysis models

2.1 Decoupling method

Analysis models accounting for the decoupling method are shown in Fig. 1(a). (1) 30 × 1 mm monopole, this is hereinafter referred to “primary element” in this paper, (2) monopole with a short element added to the primary element, (3) monopole with a long element added to the primary element. Fig. 1(b) is $|Y_{21}|$ of each antenna. First, the primary element generates “inflection point $\alpha$” corresponding to its electrical length. By contrast, the monopole added the short element generates another inflection point at higher frequency than inflection point $\alpha$. Similarly, the monopole added the long element generates another inflection point at lower frequency than inflection point $\alpha$.

![Fig. 1. Analysis models and $|Y_{21}|$ of them.](image)
Therefore, when a scope of decoupling is at lower frequency than inflection point $\alpha$, an additional long element generating inflection point at lower frequency is required. It is important to miniaturize whole antenna size including the additional element considering the practical use. Then, to miniaturize the whole antenna size, a branch antenna inserted inductors in antenna elements was reported [3]. However, large ohmic loss due to the inductors was concerned. In this paper, a proposed antenna is used meander structure instead of inductors.

2.2 Analysis models

Fig. 2(a) is 2-element monopole spaced 12 mm between elements for our analysis. The proposed antenna with a meander element is shown in Fig. 2(b). These antennas having symmetrical structure and can operate at 1.5 GHz by using matching circuits. Each antenna array is implemented on a 95 × 50 × 0.8 mm one-side 35 µm copper plate FR4 substrate, a ground plate is 95 × 50 mm. The proposed antenna is added meander element to the primary element to obtain $|Y_{21}| = 0$ at desired frequency, and an additional element used meander structure to reduce its size. Thereby the proposed antenna is disposed closely, the closest interval is 2 mm ($\lambda/100$ mm in the free space at desired frequency).

Fig. 2. 2-element monopole w/o decoupling and proposed antenna.

3 Decoupling using the meander branch antenna

Fig. 3. Comparison of the proposed antenna and the wideband model.
3.1 Design method of the proposed antenna

In this section, decoupling result of the proposed antenna is shown. Fig. 3(a) is $Y_{21}$ of the proposed antenna. In Fig. 1(a)-(1) the primary element, the inflection point $\alpha$ exists at 1.7 GHz. By adding the meander element to this primary element, another inflection point is generated at low frequency of 1.2 GHz. $(\text{Re})Y_{21}$ has a large negative value at an inflection point. On the other hand, $(\text{Im})Y_{21}$ is changed from negative value to positive value according to the increased frequency. Therefore, $(\text{Re})Y_{21}$ is almost 0 mS at the desired frequency of 1.5 GHz sandwiched between two inflection points. Moreover, it is possible that $(\text{Im})Y_{21}$ can cross 0 mS line by moving two inflection points. That is, it is able to satisfy the decoupling condition $|Y_{21}| = 0$ at the desired frequency by adjusting meander element length.

In Fig. 3(b), S-parameters of the proposed antenna is shown. The goal of this paper is keeping S-parameters of the matching and coupling under $-10$ dB. The proposed antenna obtained good impedance matching at 1.5 GHz because matching circuits are disposed. In addition, the mutual coupling is improved to $-13.7$ dB at 1.5 GHz. $|Y_{21}|$ is exactly 0 at this frequency. Moreover, a radiation efficiency of the proposed antenna is high, $-1.8$ dB, due to decoupling effect. Compared to 2-element monopole without decoupling, this result shows that the mutual coupling reduced by 11.3 dB and the radiation efficiency increased by 2.8 dB. However, a fractional bandwidth of $S_{11}$ and $S_{21}$ under $-10$ dB is 1.2% narrower in this antenna, bandwidth enhancement is required considering the practical use. The solution is elaborated in the following section.

3.2 Bandwidth enhancement by optimizing element length

In this section, it is shown that it is possible to spread the decoupling bandwidth by optimizing proposed antenna’s monopole and meander element length. Since $(\text{Re})Y_{21}$ of the proposed antenna value is almost kept 0 mS at the frequency between two inflection points as Fig. 3(a) shows, it can be considered that $(\text{Re})Y_{21}$ has a small influence of the mutual coupling. On the other hand, although $(\text{Im})Y_{21}$ is 0 mS at the desired frequency, it has a slope. It is assumed that by making its slope gentle, the bandwidth of $|Y_{21}|$ which is almost 0 mS and the decoupling bandwidth will be spread.

To make the slope of $(\text{Im})Y_{21}$ gentle, a wideband model (Fig. 3(b)) separated two inflection points is made. It is designed by shortening the monopole element by 2.0 mm and widening the meander element by 1.5 mm compared to the proposed antenna.

Fig. 3(c) is $Y_{21}$ of the wideband model. Then index $S$ about the slope of $(\text{Im})Y_{21}$ is introduced to evaluate the slope of $(\text{Im})Y_{21}$ in a quantitative way. The index $S$ is calculated by

$$S[\text{mS/MHz}] = \frac{|Y_{21}(\text{Im})_{\text{High}} - Y_{21}(\text{Im})_{\text{Low}}|[\text{mS}]}{f_{\text{High}} - f_{\text{Low}}[\text{MHz}]}$$

where $f_{\text{High}}$ is upper frequency of $S_{21}$ under $-10$ dB before matching circuits are disposed, $f_{\text{Low}}$ is its lower frequency, $(\text{Im})Y_{21_{\text{High}}}$ is value of $(\text{Im})Y_{21}$ at $f_{\text{High}}$ and $(\text{Im})Y_{21_{\text{Low}}}$ is value of $(\text{Im})Y_{21}$ at $f_{\text{Low}}$ before matching circuits are disposed. That is, low $S$ value represents gentle slope of $(\text{Im})Y_{21}$. Based on our calculation...
using (1), the proposed antenna has $S$ of 0.30 [mS/MHz], the wideband model has $S$ of 0.23 [mS/MHz]. From these results, it can be seen that the slope of the wideband model is gentle.

In Fig. 3(d), S-parameters of the wideband model disposed matching circuits is shown. The wideband model’s fractional bandwidth of $S_{11}$ and $S_{21}$ under $-10$ dB is 3.1%. It is possible to obtain more than twice the fractional bandwidth relative to the proposed antenna 1.2%.

4 Conclusion

In this paper, a decoupling method for lower frequency than primary element’s $Y_{21}$ inflection point by adding the meander element is proposed. The proposed antenna can generate $Y_{21}$ inflection point at lower frequency and can satisfy the decoupling condition $|Y_{21}| = 0$. Therefore, the proposed antenna has produced low mutual coupling and high radiation efficiency. Moreover, separating two $Y_{21}$ inflection points by optimizing the monopole and meander element length, it led to the spread of the bandwidth of $|Y_{21}| = 0$ and the decoupling bandwidth.

As for future studies, widening the decoupling bandwidth is required. Furthermore, supporting the proposed method to multiple antennas more than 2 elements is also future task.
Design of matching networks based on image impedances for near field MIMO

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Abstract: This paper proposes matching method based on image impedances for near field MIMO. In this letter, matching method, which can realize matching of all port at the same time by extending image impedances theory, is proposed. Numerical analysis and experimental results demonstrate the proposed matching method is effective in enhancing the channel capacity of the near field MIMO, where facing antennas strongly couple to each other.

Keywords: matching network, near field MIMO, conjugate image impedances

Classification: Antennas and Propagation

References


1 Introduction

Recently, noncontact short range communication attracts attentions. Also, with the rapid spread of compact terminal including smartphone, high speed data transmission in short range is required.

A short range MIMO (Multiple-Input Multiple-Output) [1, 2] has been proposed for extending the data-rate of the short range wireless communication, where the channel capacity is enhanced by increasing the number of antennas at the transmitting and receiving antenna. When the distance between the transmitting and receiving array antennas is sufficiently short, the channel has low spatial correlation characteristics even with line-of-sight (LOS) situation. As a result, the short range MIMO offers a higher channel capacity than conventional MIMO does. In many studies, the short range MIMO has been evaluated at a high frequency band such as a millimeter wave band since larger bandwidth is available. However, the propagation loss is serious in such high frequency band.

The near field MIMO offers less propagation loss and high energy efficiency due to near field electromagnetic coupling since the distance between transmitting and receiving antennas is shorter than the wavelength [3, 4]. However, the impedances of the antennas are seriously affected by other antennas as well as their feed networks in the near field situation. Therefore, near field MIMO cannot use the conventional matching method based on only one side impedance of antenna. The feed networks for near field MIMO need to be designed iteratively since it is necessary to consider the effect of feed network, which is connected to the facing antenna [5]. Image impedance theory [6, 7] offers matching condition for SISO (Single-Input Single-Output) considering effect of other antennas. However, further extension of the theory is needed for near field MIMO.

In this letter, matching method, which can achieve matching of all port at the same time by using image impedances theory, are proposed [8, 9]. Numerical analysis and experimental results demonstrate the proposed matching method is effective in enhancing the channel capacity of near field MIMO. Also, it is found that the mutual coupling among the non-facing antennas does not affect the channel capacity, which has not been well considered in [9].

2 Matching method based on image impedance for near field MIMO

Fig. 1 shows equivalent circuit of $2 \times 2$ near field MIMO array. $2 \times 2$ near field MIMO array antenna system is treated as four port network. The signals transmitted
from port 1 and 2 are received by port 3 and 4. In this figure, \( Z \) and \( Y \) represent impedance and admittance matrices of antenna system, respectively. \( Z_{i1} \) and \( Z_{i3} \) represent the image impedances for designing matching network (MN) of port 1 and 3, respectively. When we assume neighboring ports have a sufficiently low mutual coupling with the exception of two facing antenna ports, MNs can be designed using image impedance theory by focusing only on two facing antennas. For example, when port 2 and 4 are terminated by \( Z_0 \) represented reference for \( S \)-parameter, \( Z_{i1} \) and \( Z_{i3} \) are calculated by

\[
Z_{i1} = \frac{1}{G_{11}(\theta_g + j\theta_b) - jB_{11}}
\]

\[
Z_{i3} = \frac{1}{G_{33}(\theta_g + j\theta_b) - jB_{33}}
\]

where \( \theta_g, \theta_b \) are defined as

\[
\theta_g = \sqrt{ \left( 1 + \frac{G_{13}^2}{G_{11}G_{33}} \right) \left( 1 + \frac{B_{13}^2}{G_{11}G_{33}} \right) } \]

\[
\theta_b = \frac{G_{13}B_{13}}{G_{11}B_{33}}
\]

\[
G_{pq} = \text{Re}(Y_{pq}) \quad (p = 1, 2, 3, 4, \quad q = 1, 2, 3, 4)
\]

\[
B_{pq} = \text{Im}(Y_{pq}) \quad (p = 1, 2, 3, 4, \quad q = 1, 2, 3, 4)
\]

where \( Y_{pq} \) represents the element of \( Y \). When input and terminate impedances of port 1 and 3 are in the complex conjugate relation, matching can be achieved. Also the matching conditions for ports 2 and 4 can be determined in the same manner. Nevertheless, the MN design satisfying the above condition needs to be considered as follows.

As shown in Fig. 1, MNs for two facing coupled ports are designed by neglecting other two facing ports because the terminated non-facing ports little affect the matching condition due to small mutual coupling among them. When we focus on one certain antenna port, the observed reflection coefficient at the port is expressed as

\[
\Gamma = \frac{Z_i^* - Z_0}{Z_i^* + Z_0}
\]

where \( Z_i^* \) represent conjugate complex of the image impedance \( Z_i \). The facing ports are assumed to be terminated by the image impedance element, and all other ports are terminated by \( Z_0 \). When the MN is designed using \( \Gamma \), \( S \)-parameter of the MN is defined as
$$S_M = \begin{pmatrix} S_{M11} & S_{M12} \\ S_{M21} & S_{M22} \end{pmatrix}.$$  \hspace{1cm} (8)

The reflection coefficient through the MN becomes

$$S_{\text{ref}} = S_{M11} + S_{M12}(\Gamma^{-1} - S_{M22})^{-1}S_{M21} = 0.$$  \hspace{1cm} (9)

Also, the design condition of MN is given by

$$S_{M22} = \Gamma^*.$$  \hspace{1cm} (10)

Additionally, the lossless MN satisfies

$$S_M \cdot S_M^H = I$$  \hspace{1cm} (11)

where the operation, \(\{\}^H\), is the complex conjugate transpose. From (9)∼(11), \(S\)-parameter of MN is calculated by

$$S_{M22} = \Gamma^*$$  \hspace{1cm} (12)

$$S_{M12} = \sqrt{1 - |S_{M22}|^2} e^{i\theta}$$  \hspace{1cm} (13)

$$S_{M21} = S_{M12}$$  \hspace{1cm} (14)

$$S_{M11} = -S_{M12}(\Gamma^{-1} - S_{M22})^{-1}S_{M21}$$  \hspace{1cm} (15)

where \(\theta\) is arbitrary value.

3 Simulation

![Simulation Model](image1)

![S-parameters](image2)

![Channel Capacity](image3)

Fig. 2. Simulation results

Fig. 2(a) shows the antenna model in the simulation. An antenna system consists of four planar spiral antennas. This antenna element [10] is the square shaped spiral antenna printed on FR-4 substrate and the center frequency is set to 150 MHz. FR-4 substrate’s relative dielectric constant is 4.4. In this figure, element spacing \(d\) is 65 mm, and distance between transmitting and receiving antennas is \(D\). \(S\)-parameter of this model is analyzed by the method of moments. The signals transmitted from \#1 and \#2 are received by \#3 and \#4, respectively. Fig. 2(b) shows \(S\)-parameters of this model by the numerical analysis when \(D\) is 10 mm. In the figure, solid lines and broken lines represent the \(S\)-parameters with and without MNs respectively. The conjugate image impedances were

$$\begin{pmatrix} Z_{i1} & Z_{i2} \\ Z_{i3} & Z_{i4} \end{pmatrix} = \begin{pmatrix} 94.16 + j201.5 & 94.14 + j201.5 \\ 97.44 + j204.0 & 97.66 + j203.8 \end{pmatrix}.$$  \hspace{1cm} (16)
in this model. From this result, it can be seen that the proposed matching method improves $S_{11}$ significantly and transmitting power $S_{31}$ by 5 dB. Moreover, the mutual coupling, $S_{21}$, is even lower than transmission between facing antennas, $S_{31}$.

Fig. 2(c) plots the channel capacity versus distance between transmitting and receiving antennas. In this figure, solid line shows the channel capacity with the MN optimized for every distance, i.e. the MN is changed depending on $D$, dotted line shows the channel capacity with the MN designed only for $D = 10$ mm, and dashed line shows the channel capacity without MN. The transmitting power is set to 0 dBm, and noise power is $-90$ dBm. From this result, it can be seen the MN improves channel capacity by 12 bits/s/Hz at maximum since SNR (Signal to Noise Ratio) is improved thanks to the MNs at all port. Also, it is found that MN designed for 10 mm offers higher capacity than that without MN.

4 Experiment

Fig. 3(a) shows the photos of fabricated antennas and MN based on the proposed matching method. Configuration of the antenna system and antenna element for experiment is same as condition of simulation. MNs were fabricated on FR-4 substrate, and consist of chip capacitors and inductors. The conjugate image impedances of all ports are calculated by using the measured $S$-parameter of the fabricated antennas. The calculated image impedances were

\[
\begin{pmatrix}
  Z_{i1} \\
  Z_{i2} \\
  Z_{i3} \\
  Z_{i4}
\end{pmatrix} = \begin{pmatrix}
  68.58 + j123.6 \\
  86.16 + j173.0 \\
  57.44 + j128.3 \\
  50.22 + j133.1
\end{pmatrix}
\]

and the MNs were designed from this result.
Fig. 3(b) shows that the frequency characteristics of the measured reflection characteristic $S_{11}$ when $D$ is 8.4 mm. In this result, reflection characteristics $S_{11}$ are less than $-10\,\text{dB}$ at the center frequency when the MNs are used. Reflection characteristics of other ports are also improved.

Fig. 3(c) shows that the experimental result of transmission and coupling characteristics. This result indicates MN improves transmission, $S_{31}$. Moreover, the mutual coupling, $S_{21}$, is even lower than transmission, $S_{31}$, as with simulation results.

Fig. 3(d) plots the channel capacity versus frequency between transmitting and receiving antennas. The condition of calculating channel capacity is the same as that of the simulation. From this result, it is found the proposed MNs improve the channel capacity by 1.54 bits/s/Hz at the center frequency since reflection characteristics are improved.

5 Conclusion

This paper has proposed matching method based on image impedances suitable for near field MIMO. Simulations and experiments showed that the proposed matching method improves the channel capacity since reflection characteristics of all ports are improved at the same time. Both simulation and experiment results demonstrated that the proposed method is effective in enhancing channel capacity of near field MIMO.

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