Performance of MMSE Interference Rejection followed by Joint MLD for DAN

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SUMMARY  This paper applies minimum mean square error (MMSE) interference rejection followed by joint maximum likelihood detection (MLD) to a receiver in a distributed antenna network (DAN). DAN receivers capture not only the desired signals, but also the interference signals from nearby uncoordinated antennas. For the overloaded signal situation, non-linear detection schemes such as joint MLD can be applied to the received signals. However, the amount of metric calculations in joint MLD increases exponentially with the number of signal streams. Therefore, MMSE interference rejection followed by MLD detection is proposed. The proposed scheme reduces the complexity by a factor of $1/2^M(N_T-1)$ where $N_T$ is the number of interference signals with $2^M\text{QAM}$ modulation. The effect of residual interference after the MMSE interference rejection is evaluated. Numerical results obtained through computer simulation and experiment show that the performance of the proposed scheme is about 4.0 dB worse at a bit error rate (BER) of $10^{-3}$ than that of the joint MLD while its complexity is four times lower for QPSK signal streams. The BER performance degradation can be suppressed to about 2.5 dB by adjusting the value of the coefficient in the MMSE matrix.

key words: 5G, DAN, interference rejection, Joint MLD

1. Introduction

Because of the wide use of smart phones and mobile terminals, a higher data rate is an inevitable demand in wireless communications. The 5th generation (5G) mobile communication systems, which are expected in 2020, have been investigated to accommodate explosive traffic growth [1]. One solution to improve spectrum efficiency is the distributed antenna network (DAN) [2]. The DAN deploys a number of distributed antennas in each cell and they are connected to a baseband unit via optical links. Each mobile terminal is served by surrounding multiple distributed antennas [3]. It has been shown that the DAN allows frequency reuse and improves the spectrum efficiency as compared with that of a conventional cellular network [4]. For mobile wireless systems with full frequency reuse, co-channel interference at antenna coverage boundaries has a significant impact on signal reception performance [5]. Therefore, a detection scheme that is robust to co-channel interference is required. Our studies apply joint maximum likelihood detection (MLD) to desired and interference signals on the same channel [6]. If one of surrounding base stations (BSs) transmits signals on the same channel, the terminal receives not only its own desired signals, but also the interference signals transmitted for another terminal. The joint MLD treats the received signal as the superposition of the desired and interference signals and regards the results as a signal with a larger number of constellation points. Therefore, the amount of metric calculation increases exponentially with the number of signal streams.

To reduce detection complexity, interference rejection combining (IRC) through a minimum mean square error (MMSE) criteria is often used [7]- [9]. However, it has not been applied as preprocessing for reducing the amount of calculation in the MLD stage. In this research, the joint MLD combined with MMSE interference rejection is presented. Interference rejection is carried out with an MMSE algorithm so that the amount of metric calculation in the joint MLD can be reduced. Furthermore, the amount of residual interference is suppressed by changing the value of the coefficient in an MMSE weight matrix.

This paper is organized as follows. Section 2 explains the system model and the experimental setup. In Section 3, numerical results obtained through computer simulation and experience are presented. Finally, conclusions are presented in Section 4.

2. System Model

2.1 Signal Model

![Distributed antenna system ($N_T = N_R = 2$).](image)

In this paper, two adjacent cells in a distributed...
antenna system as shown in Fig. 1 are assumed. Those two small BSs are connected to the same signal processing center and work as antennas in the DAN. Those BSs transmit signals to user terminals without precoding for interference coordination between the adjacent cells. The number of transmit antennas at each BS is \(N_T\) and the number of receive antennas at each user terminal is \(N_R\). It is assumed that the terminal of interest receives its desired signals as well as the signals for the other terminal from the antennas in the adjacent cell as interference. Thus, the receiver of the terminal demodulates the desired signals as well as the interference signals by combing the MMSE interference rejection and the joint MLD. It is also assumed here that the terminals receive the pilot signals embedded in the desired signals and in the interference signals, and channel state information is available at the receiver side. The pilot signals are assumed to consist of orthogonal sequences and can be demodulated without interference.

Transmitter and receiver models are illustrated in Fig. 2. In the transmitter, the turbo encoder generates a coded bit sequence and the interleaver rearranges the bit sequence. The code rate at the transmitter is controlled by puncturing. After the puncturing, every \(M\) coded bits are assigned to a \(2^M\)-QAM symbol and multiplexed by orthogonal frequency division multiplexing (OFDM) in a frequency domain. Suppose that, \(S_p^D[l]\) is the transmit symbol on the \(l\)th subcarrier from the \(p\)th antenna of the base station and \(S_p^I[l]\) is the transmit symbol on the \(l\)th subcarrier from the \(p\)th antenna of the base station in the adjacent cell. In the following, the superscript letter "\(D\)" implies the desired signal and "\(I\)" implies the interference signal. The OFDM signals transmitted from the \(p\)th antenna of both base stations are given as follows:

\[
u^D_p[n] = \sum_{l=0}^{N-1} S_p^D[l] \exp \left( j \frac{2\pi nl}{N} \right), \tag{1}\]

where \(n(n = 0, 1, ..., N - 1)\) is the time index and \(N\) is the size of an inverse discrete Fourier transform (IDFT). The last part of the OFDM signal is appended to the beginning of the OFDM signal as a guard interval (GI). The OFDM signals, \(u^D_p[n]\) and \(u^I_p[n]\), pass through the transmit filter. The baseband signals are given as follows:

\[
v^D_p(t) = \sum_{n=-N_GI}^{N-1} u^D_p[n] p_{tp}(t - nT_s),
\]

\[
v^I_p(t) = \sum_{n=-N_GI}^{N-1} u^I_p[n] p_{tp}(t - nT_s), \tag{2}\]

where \(p_{tp}(t)\) is the impulse response of the transmit filter, \(T_s\) is the sampling interval of the OFDM signal, and \(N_GI\) is the length of the GI. At the receiver, the signal received by the \(q\)th receive antenna is given as

\[
y_q(t) = \sum_{p=1}^{N_T} y^D_q(t) + \sum_{p=1}^{N_T} y^I_q(t) + n_q(t) \tag{3}\]

where \(n_q(t)\) is the noise on the \(q\)th receive antenna and \(y^D_q(t)\) and \(y^I_q(t)\) are the received signals from the \(p\)th transmit antenna to the \(q\)th receive antenna of the terminal of interest. \(y^D_q(t)\) and \(y^I_q(t)\) are then given as follows:

\[
y^D_q(t) = \sum_{n=-N_GI}^{N-1} u^D_p[n] h_{qp}(t - nT_s),
\]

\[
y^I_q(t) = \sum_{n=-N_GI}^{N-1} u^I_p[n] g_{qp}(t - nT_s), \tag{4}\]

where \(h_{qp}(t)\) and \(g_{qp}(t)\) are the channel impulse responses including the transmit and receive filters. The following expressions can be derived as
The received signal passes through the filter, terminal of interest, and physical channels between the communicating and interfering base stations to the qth antenna of the terminal of interest, respectively, where \( p_r(t) \) is the impulse response of the receive filter in the terminal of interest, and, \( \Omega \) means convolution. The received signal passes through the filter, \( p_r(t) \), and is converted to a digital form with the sampling interval of \( T_s \). The received signal is sampled as follows:

\[
y_q[n] = y_q(n T_s). \tag{6}
\]

The received signal on the lth subcarrier that is transformed by a discrete Fourier transform (DFT) after the removal of the GI and it is given as

\[
Y_q[l] = \sum_{n=0}^{N_s-1} y_q[n] \exp \left( -\frac{2\pi i nl}{N} \right) = \sum_{p=0}^{N_s} H_{qp}[l] S_p[l] + \sum_{p=0}^{N_s} G_{qp}[l] S_p[l] + N_q[l], \tag{7}
\]

where \( H_{qp}[l] \) and \( G_{qp}[l] \) are the channel responses on the lth subcarrier. The received signal represented by a vector form is

\[
\]

where

\[
Y[l] = [Y_1[l], Y_2[l], \ldots, Y_{N_s}[l]]^T, \tag{9}
\]

\[
H[l] = \begin{bmatrix}
H_{11}[l] & \cdots & H_{1N_s}[l] \\
\vdots & \ddots & \vdots \\
H_{N_s1}[l] & \cdots & H_{N_sN_s}[l]
\end{bmatrix}, \tag{10}
\]

\[
G[l] = \begin{bmatrix}
G_{11}[l] & \cdots & G_{1N_s}[l] \\
\vdots & \ddots & \vdots \\
G_{N_s1}[l] & \cdots & G_{N_sN_s}[l]
\end{bmatrix},
\]

\[
S[l] = [S_1[l], S_2[l], \ldots, S_{N_s}[l]]^T, \tag{11}
\]

\[
S'[l] = [S_1[l], S_2[l], \ldots, S_{N_s}[l]]^T, \tag{12}
\]

\[
N[l] = [N_1[l], N_2[l], \ldots, N_{N_s}[l]]^T.
\]

The outputs of the DFT are demodulated to calculate soft information as the inputs of a turbo decoder.

### 2.2 Likelihood Calculation in MLD

For the MLD of the desired signals, a log-likelihood ratio (LLR) is calculated for each bit. In the conventional MLD, the interference signals are regarded as noise. However, in the joint MLD, the constellations of interference signals are also taken into account [6].

\[
P_{pm}^1 = \sum_{S_p[l] \in S_p^D} \exp \left( \frac{-\|Y[l] - \tilde{H}[l] S_p[l] - \tilde{D}[l] S'[l] \|^2}{\sigma^2(1 + \lambda\|S'[l]\|^2)} \right), \tag{13}
\]

\[
P_{pm}^0 = \sum_{S_p[l] \in \tilde{S}_p^D} \exp \left( \frac{-\|Y[l] - \tilde{H}[l] S_p[l] - \tilde{D}[l] S'[l] \|^2}{\sigma^2(1 + \lambda\|S'[l]\|^2)} \right), \tag{14}
\]

where \( \tilde{H} \) and \( \tilde{D} \) are the estimated channel vector, \( \sigma^2 \) is the noise variance, \( S_p^D \) and \( \tilde{S}_p^D \) are the vectors of the candidate \( N_T \) symbols of which the mth coded bit of a \( 2^M \) QAM symbol from the pth transmit antenna is "1" and "0", \( S[l] = [S_1[l], S_2[l], \ldots, S_{N_s}[l]]^T \) and \( S'[l] = [S_1[l], S_2[l], \ldots, S_{N_s}[l]]^T \) are the candidate symbol vectors for the desired signals and the interference signals, \( \{S[l]\}_{b_m=1}^N \) and \( \{S[l]\}_{b_m=0}^N \) are the set of the symbol vectors of which the mth coded bit of the pth \( 2^M \) QAM symbol, \( S[l] = [1^1, 1^0, \ldots, 0^1, 0^0] \) and \( D[l]_{b_m=1}^N \) and \( D[l]_{b_m=0}^N \) are the sums of the likelihood values for the mth coded bit of "1" and "0" in the pth symbol, respectively. \( \lambda \) is the ratio of the channel estimation error to the noise variance [10]. The LLR is calculated as

\[
L(b_m|Y[l]) = \log \frac{P_{pm}^1}{P_{pm}^0} \tag{15}
\]

where \( L(b_m|Y[l]) \) is the LLR of the mth bit of the symbol transmitted from the pth antenna on the lth subcarrier. The calculated LLR is put into the turbo decoder.

### 2.3 Interference Rejection with MMSE

In this paper, interference rejection with an MMSE algorithm is applied to reduce computational complexity in the following MLD stage. Before the MLD, MMSE interference rejection is applied to separate the interference signals. The MMSE matrix is defined as follows:

\[
W[l] = \tilde{G}[l] \left( \tilde{G}[l]^H \tilde{G}[l] + \sigma^2 \mathbf{I} \right)^{-1} \tag{16}
\]

where \( \mathbf{I} \) is the \( N_R \times N_R \) unit matrix and the superscript \( H \) indicates an Hermitian matrix.

The received signal is multiplied by the MMSE matrix. The received signal after interference rejection is given by

\[
Y'[l] = W[l] Y[l] \approx \tilde{H} S[l] + \tilde{S}'[l] + N[l], \tag{17}
\]

where

\[
\tilde{S}' = W[l] S'[l]. \tag{18}
\]
with interference rejection, it is possible to reduce the com-
inations are also included. In the equations, \( \gamma \) is the unit vector with a size of \( N_R \). With interference rejection, it is possible to reduce the computational complexity of the MLD by replacing \( \hat{G} \hat{S}^T \) with \( \gamma \hat{S}^T \).

The numbers of real-valued multiplication operations required in the joint MLD and the proposed scheme are presented in Table 1. The numerical examples of the numbers of real-valued multiplication operations are also included. In the equations, \( \hat{S}^{D_{pm}} \) implies either \( \hat{S}^{D_0} \) or \( \hat{S}^{D_1} \) for the calculation of \( D_{pm}^{0} \) or \( D_{pm}^{1} \) and \( O(*) \) means the order of * multiplication operations. Both schemes require to calculate the multiplication between channel responses and candidate symbols as soon as channel estimation is conducted. On the other hand, the MMSE algorithm requires the matrix inversion operation that costs \( 4 \times O(N_R^{-3}) \) real-valued multiplication operations. As for the LLR calculations, the number of real-valued multiplication operations increases exponentially with the number of the desired interference signals as shown in the table. However, the proposed scheme requires \( 3 \times 2^M(2N_T - N_R + 1) \) multiplication operations that are \( 4^N R^{-1} \) times smaller than that in the joint MLD.

### Table 1 Number of Real-Valued Multiplication Operations

<table>
<thead>
<tr>
<th>Equation</th>
<th>Number of Operations</th>
<th>Number of Operations</th>
</tr>
</thead>
<tbody>
<tr>
<td>Joint MLD (for each channel estimation vector) ( \hat{H}[l] \hat{S}^{D_{pm}} \hat{G}[l] \hat{S}^T ) (for each received signal vector) ( |Y[l] - \hat{H}[l] \hat{S}^{D_{pm}} - \hat{G}[l] \hat{S}^T|^2 \times \sigma^2 (1 + \lambda |S^T|^2) )</td>
<td>( 4 \times 2^{2M N_T N_R} )</td>
<td>128</td>
</tr>
<tr>
<td>MMSE+MLD (for each channel estimation vector) ( \hat{G}^H[l] (\hat{G}[l] \hat{G}^H[l] + \sigma^2 I)^{-1} ) ( \hat{W}^H[l] \hat{H}[l] \hat{S}^{D_{pm}} \hat{W}^H[l] \hat{G}^S[l], ) (for each received signal vector) ( | Y[l] - \hat{W}^H[l] \hat{H}[l] \hat{S}^{D_{pm}} - \gamma \hat{S}^T|^2 \times \sigma^2 (1 + \lambda |S^T|^2) )</td>
<td>( 4 \times (O(N_R^3) + N_R^2) )</td>
<td>( 4 \times (O(8) + 4) )</td>
</tr>
</tbody>
</table>

2.4 Coefficient for Noise Term in MMSE Matrix

The MMSE interference rejection eliminates the interference signal while avoiding noise enhancement by including the covariance matrix of the noise term. Thus, the part of the interference signal remains after the rejection and the amount of the residual interference should be adjusted.

In order to evaluate the amount of the interference, the coefficient, \( \zeta \), is included in Eq. (16) as \( \hat{W}^H[l] = \hat{G}^H[l] (\hat{G}[l] \hat{G}^H[l] + \zeta \sigma^2 I)^{-1} \)

and the ratio of \( \hat{S}_I^T / J_q (q = 1, \ldots, N_R - 1) \) in the following equation in accordance with \( \xi \) is evaluated,

\[
\begin{bmatrix}
\hat{S}_I^T \\
J_1 \\
\vdots \\
J_{N_R-1}
\end{bmatrix} = \hat{W}^H[l] \hat{G} \begin{bmatrix}
1 \\
0 \\
\vdots \\
0
\end{bmatrix}
\]

where \( \hat{S}_I^T \) corresponds to the remaining interference signal from the 1st antenna of the BS in the adjacent cell and \( J_q \) is the residual component of the qth received interference after the MMSE interference rejection, respectively. If the value of \( \xi \) increases, the power of the residual component of the interference included in \( \hat{S}^T[l] \) in Eq. (17) grows. On the other hand, if the value of \( \xi \) approaches to 0, noise enhancement occurs after the
MMSE interference rejection and the power of the noise term, \( N_\text{f} \), in Eq. (17) increases. There is a trade-off according to the value of \( \xi \) and it is evaluated numerically in Sec. 3. The MMSE coefficients are updated every after channel estimation and it is less frequent than the symbol rate.

2.5 Experimental Setup

The block diagram of an experimental setup is presented in Fig. 3. This setup assumes that one BS transmits to a user terminal on the downlink while the other BS causes interference to the same user terminal. Table 2 shows the specifications of equipments used in the experiment. The transmitter and the receiver are connected by a cable. As the transmitter the signal generator transmits an OFDM signal with carrier frequency of 2.422 GHz. At the receiver, a Sora-MIMO radio control board (RCB) receives the OFDM signal, down-converts it to baseband, and converts it to digital samples [11].

At the transmitter, information bits are generated randomly and put into a turbo encoder. The coded bits are then mapped to QPSK symbols and multiplexed with OFDM. The OFDM signal passes through a root-raised-cosine filter. Fading channels are also generated in the signal generator and the OFDM signal is convolved with those impulse responses. They are renewed for each codeword.

At the receiver, the received signal is first put into a root-raised-cosine filter and digitized by analog-to-digital (A/D) converters in the Sora board. Digital samples are resampled to the sample rate of the transmit signal because the sampling frequency at the receiver is fixed to 20 MHz. After that, the DC offset canceler calculates the average of each 512 samples, and remove the average value from the received samples. Frame synchronization and frequency offset compensation are then carried out [12]. The received samples are then converted to the frequency domain, the channel responses are estimated on all the subcarriers, and turbo decoding is carried out after the MLD.

The specifications of the OFDM signal is shown in Table 3. At the transmitter, the OFDM signal is generated with the sample rate of 3.84 MHz.

The frame format of the transmit signal in the experiments are presented in Fig. 4. In the data signal part, the number of subcarriers is 256 and the number of data subcarriers is 151. The short preamble is the same sequence as the one specified in the IEEE802.11ac standard [13] and it is also used for timing synchronization. The VHT-LTF 80 MHz sequence specified in the IEEE802.11ac standard is used for the long preamble sequences. The two of the six sequences are used for frequency offset compensation. The others are used for channel estimation. When four signal sequences are transmitted, these four sequences are multiplied with orthogonal codes in order to obtain the channel responses for each signal stream.

The PN sequence is used to recognize the frame index. After generating the PN sequence with length of 131071, every 64 chips of the sequence are inserted into each frame. The receiver knows the relationship between the PN sequence and the frame index. The position of the PN sequence is detected by a matched filter and the frame index is calculated with the outputs from the matched filter.

In the experiment, the short preamble and the PN sequence are assumed to be subject to no fading for ideal timing synchronization. On the other hand, long preambles for channel estimation passes through the fading channel and they are received with 20 dB larger signal power than that of the data signal in order to evaluate performance under accurate channel estimation.
Table 4  Simulation conditions.

<table>
<thead>
<tr>
<th>Modulation scheme for interference signal</th>
<th>QPSK, 16QAM/OFDM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Modulation scheme for desired signal</td>
<td>QPSK/OFDM</td>
</tr>
<tr>
<td>Signal to interference ratio</td>
<td>SIR=6.0 or 1.8 dB</td>
</tr>
<tr>
<td>Interleaver size</td>
<td>4800</td>
</tr>
<tr>
<td>Coding scheme</td>
<td>Turbo code</td>
</tr>
<tr>
<td>Decoding scheme</td>
<td>Log-MAP</td>
</tr>
<tr>
<td>Number of decoding iteration</td>
<td>8</td>
</tr>
<tr>
<td>Code rate</td>
<td>1/3 (CoMP:2/3)</td>
</tr>
<tr>
<td>DFT size</td>
<td>256</td>
</tr>
<tr>
<td>Number of data subcarriers</td>
<td>151</td>
</tr>
<tr>
<td>GI size (1st symbol)</td>
<td>5.12 µs</td>
</tr>
<tr>
<td>GI size (2-7th symbol)</td>
<td>4.69 µs</td>
</tr>
<tr>
<td>Number of transmit antennas (each BS)</td>
<td>2</td>
</tr>
<tr>
<td>Number of receive antennas (each user terminal)</td>
<td>2 (CoMP:1)</td>
</tr>
<tr>
<td>Channel model</td>
<td>Rician (K = 10)</td>
</tr>
<tr>
<td>Channel estimation</td>
<td>Ideal or Long preamble</td>
</tr>
<tr>
<td>Channel estimation error coefficient (λ)</td>
<td>0.0 (Ideal), 0.25 (Long preamble)</td>
</tr>
</tbody>
</table>

3. Numerical Results

3.1 Simulation and Experiment Conditions

The performance curves of CoMP (coordinated scheduling/coordinated beamforming) transmission, a conventional zero forcing (ZF) scheme, the joint MLD scheme, and the proposed scheme (ξ = 1.0, 0.25) are obtained through computer simulation. As for the CoMP transmission, null steering at the BSs is assumed since it eliminates the interference to the adjacent cell. In this case only one receive antenna at each user terminal is assumed since only one null can be formed with two antenna elements in each base station. Furthermore, the same signal with precoding coefficients for beam nulling is transmitted from those antenna elements. To normalize the throughput, the code rate is doubled and MLD is applied at each user terminal. As for the conventional ZF scheme, the interference signals from the adjacent cell is ignored.

The simulation and experiment conditions are shown in Table 4. The number of the transmit antennas at each BS is $N_T = 2$ while the number of the receive antennas is $N_R = 2$ (1 for CoMP) at each UE. The signal streams are modulated with QPSK or 16QAM and multiplexed through OFDM. The average signal-to-interference ratio (SIR) is set to SIR=6.0 or 1.8 dB. A turbo code with 8-state memory is applied as the error-correction code and the code rate is set to 1/3 (2/3 for CoMP). The roll-off factor of the root-raised-cosine filter is 0.1. A Rician model is assumed as a channel model and the $K$ factor is 10. The channel response is renewed for each OFDM symbol. At the receiver, the channel response is assumed to be ideal or estimated by demodulating long preamble symbols. In the latter case, $λ$ is set to 0.25 since four long preamble symbols are used for channel estimation [10]. In the CoMP transmission, CSI is assumed to be ideal.

3.2 QPSK Modulation

3.2.1 Effect of Noise Term

Firstly, the effect of the coefficient, $ξ$, in Eq. (25) is evaluated. It is assumed that SIR=3.7 dB and $E_b/N_0 = 10$ dB for the desired signal. The number of transmit and receive antennas are $N_T = 2$ and $N_R = 2$, while the modulation schemes for the interference signal and the desired signal are both QPSK. Three ratio curves versus $ξ$ are shown in Fig. 5.
(1) S/N ratio:

The S/N ratio is given as $|\overline{S}_D^j|^2/|\overline{N}_1|^2$, where $\overline{S}_D^j$ and $\overline{N}_1$ are the 1st elements of $\overline{S}^j$ and $\overline{N}$, respectively, and $\overline{S}^j = \textbf{H} \textbf{s}^j$. This is the ratio between the desired signal power and the noise power after the MMSE interference rejection. It can be seen that $|\overline{S}_D^j|^2/|\overline{N}_1|^2$ does not substantially change in accordance with $\xi$.

(2) I/N ratio:

The I/N ratio is given as $|\overline{S}_I^j|^2/|\overline{N}_1|^2$. At the stage of the joint MLD, the power of the remaining interference signal should be larger for detection. The MMSE matrix with $\xi = 1$ maximizes the remaining interference-to-noise ratio. When $\xi = 0.0$, it is equivalent to a zero forcing (ZF) method and noise enhancement occurs. It can be observed that $|\overline{S}_I^j|^2/|\overline{N}_1|^2$ increases as $\xi$ grows.

(3) I/J ratio:

The I/J ratio is given as $|\overline{S}_I^j|^2/|\overline{J}_1|^2$. This ratio represents the interference rejection capability. It can be seen that $|\overline{S}_I^j|^2/|\overline{J}_1|^2$ decreases as $\xi$ grows. From Fig. 5, there is a trade-off between the I/N ratio and the I/J ratio. In the joint MLD, it is preferable that $|\overline{S}_D^j|^2$ and $|\overline{S}_I^j|^2$ are larger while $|\overline{N}_1|^2$ and $|\overline{J}_1|^2$ are smaller.

As $\xi$ increases from 0 to 0.25, the I/N ratio increases significantly though the I/J ratio reduces sharply. The increase rate of the I/N ratio becomes smaller if $\xi$ grows more than 0.25 and the I/J ratio is still more than 10 dB at $\xi = 0.25$. Thus, $\xi = 0.25$ is selected in the following sections and the performance is compared to that of the typical MMSE scheme, i.e. $\xi = 1.0$.

3.2.2 BER Performance

In the following figures, open dots indicate the numerical results obtained through computer simulation and closed dots indicate the results of the experiment. The red line indicates the performance of the proposed scheme and the green line indicates the performance of the full joint MLD scheme. The solid line represents for $\xi = 1.0$ and the dash line is for $\xi = 0.25$. The blue line indicates the performance of the CoMP transmission with a coding rate of 2/3.

The BER versus $E_b/N_0$ is shown in Fig. 6. The modulation schemes for the interference signal and desired signal are both QPSK, and channel estimation is assumed to be ideal. The performance is worse than that of the proposed scheme even though the complexity is smaller. The performance obtained through the experiment deteriorates by about 1.0 dB from that obtained through computer simulation. In the case of the proposed scheme ($\xi = 1.0$), the performance obtained through the experiment realizes the better performance through that of the computer simulation. The reason is that the channel estimation error is ignored at the calculation of the MMSE weight matrix in the experiment. The channel estimation error effectively increases the amount of
Conclusions

This paper proposed, for DAN systems, MMSE interference rejection followed by the joint MLD. The proposed scheme reduces the computational complexity of the joint MLD. Furthermore, the amount of the residual interference after the MMSE rejection is suppressed by adjusting the coefficient in the MMSE matrix.

The BER performance is evaluated by computer simulations and experiments. When the SIR is 6.0 dB, under the ideal channel estimation assumption, the BER performance of the proposed scheme for $\xi = 1$ deteriorates by about 4.0 dB at a BER of $10^{-3}$ from that of the joint MLD. By setting $\xi = 0.25$, the performance improves by about 2.5 dB. The performance obtained through the experiment has shown the same tendency though it deteriorates by about 1.0 dB from that obtained through computer simulation.

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References


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