Open-Loop Correlation Reduction Precoding in Overloaded MIMO-OFDM Systems

Hikari MATSUOKA\(a)\), Yoshihito DOI\(b)\), Tatsuro YABE\(c)\), Student Members, and Yukitoshi SANADA\(d)\), Senior Member

SUMMARY This paper proposes an open-loop correlation reduction precoding scheme for overloaded multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) systems. In overloaded MIMO-OFDM systems, frequency diversity through joint maximum likelihood (ML) decoding suppresses performance degradation owing to spatial signal multiplexing. However, on a line-of-sight (LOS) channel, a channel matrix may have a large correlation between coded symbols transmitted on separate subcarriers. The correlation reduces the frequency diversity gain and deteriorates the signal separation capability. Thus, in the proposed scheme, open-loop precoding is employed at the transmitter of an overloaded MIMO system in order to reduce the correlation between codewords transmitted on different signal streams. The proposed precoding scheme changes the amplitude as well as the phase of the coded symbols transmitted on different subcarriers. Numerical results obtained through computer simulation show that the proposed scheme improves the bit error rate performance on Rician channels. It is also shown that the proposed scheme greatly suppresses the performance degradation on an independent Rayleigh fading channel even though the amplitude of the coded symbols varies.

key words: open-loop precoding, overloaded MIMO, rician fading channel

1. Introduction

With the popularity of mobile phones and mobile terminals, the enlargement of capacity and the improvement of throughput are expected in wireless communications. A multiple-input multiple-output (MIMO) transmission scheme has been extensively studied to resolve these demands \[1\], \[2\]. MIMO increases the communication capacity by using multiple transmit and receive antenna elements and improves the reliability of the communication links through diversity \[3\], \[4\].

However, due to the requirements of a form factor, the number of antenna elements that can be implemented in a mobile terminal is limited. In \[5\]–\[7\], overloaded MIMO in which the number of receive antennas is less than that of transmit antennas has been investigated. Although there are several signal separation schemes for MIMO, the maximum likelihood (ML) detection that searches the round-robin candidate points of the signal shows the best performance in theory. Nevertheless, each additional transmission signal stream deteriorates the performance in the overloaded MIMO. It has been shown in \[8\] that it is possible to suppress the amount of performance degradation through diversity in a receiver. Therefore, by using an error correction code instead of increasing the number of receive antennas, spatially multiplexed signal streams can be separated without a large amount of bit error rate (BER) degradation.

The joint decoding scheme for overloaded MIMO systems with error correction codes has been investigated \[9\]–\[11\]. If fading statistics among coded symbols are independent, the error rate approaches to a lower bound through frequency diversity on multipath channels. Nevertheless, on Rician channels which include a line-of-sight (LOS) component, the performance of joint decoding deteriorates because of correlations among the elements of a channel matrix \[12\], \[13\].

Precoding schemes for MIMO systems in order to reduce the correlation of received signal streams have been proposed. A precoding scheme in \[13\], \[14\] uses channel state information (CSI). When the precoder uses the CSI, the signal covariance is optimized and the channel capacity is maximized. However, the receiver needs the feedback of the CSI before transmission. Open-loop precoding MIMO has also been proposed in \[15\]. Nevertheless, this precoding scheme can be applied only to MIMO systems in which the numbers of transmit and receiver antennas are the same. Hence, this scheme is not applicable to overloaded MIMO systems. In \[16\], a precoding scheme for joint decoding in the overloaded MIMO system has been investigated. This scheme needs to detect the random precoding coefficients with preamble sequences.

Several literature have also investigated phase rotation between two signal streams \[17\]–\[19\]. The assumed system in these references transmits multiple signal streams from one antenna. Thus, the optimum amount of phase rotation between the signal streams can be found. On the other hand, each signal stream is transmitted from different antenna element in the proposed scheme. Thus, the relative phase between the signal streams cannot be fixed at the receiver.

This paper proposes an open-loop precoding scheme with predetermined coefficients for overloaded MIMO systems with repetition codes. This precoding scheme changes the phases of coded symbols with the predetermined coefficients. It then reduces the correlation between coded symbols transmitted on different signal streams. In the proposed scheme the number of transmit antennas is limited by the...
correlation between the signal streams instead of the design of the precoding coefficients since the orthogonality between the coefficient sequences for different signal streams is not assumed. Furthermore, pilot-based channel estimation as well as preamble-based channel estimation can be applied with this scheme. However, the ill-condition of the channel matrix that leads to the overlap of signal constellation points cannot be avoided if only the phase of the coded symbols is rotated [20]. Thus, amplitude variation in addition to the phase rotation is applied through multiplying windowing coefficients.

This paper is organized as follows. Section 2 explains the system model and the joint decoding scheme. Proposed precoding scheme is described in Sect. 3. In Sect. 4, numerical results obtained through computer simulation are presented. Finally, conclusions are presented in Sect. 5.

2. System Model

2.1 Overloaded MIMO-OFDM System

An MIMO-OFDM system with \( N_T \) transmit antennas and \( N_R \) receive antennas is shown in Fig. 1. The overloaded MIMO is the case when \( N_T > N_R \). Information bits are demultiplexed to \( N_T \) transmit antennas in a bit-by-bit fashion. On each antenna branch, a transmit 2\(^M\) QAM symbol is generated with \( M \) information bits through Gray coding. With a rate 1/\( L \) repetition code, the \( i \)th transmit symbol, \( C_p^{(i)} \), is replicated and assigned to \( L \) subcarriers. That is, on the \( p \)th branch, the data subcarrier index assigned to the \( l \)th coded symbol of the \( i \)th transmit symbol is given as

\[
k_l = (l - 1)N_D/L + (i - 1) + (1 + \Delta)
\]

(1)

where \( N_D \) is the number of data subcarriers in OFDM modulation and \( \Delta \) is the number of null subcarriers on channel edges and \( N = N_D + 2\Delta \). Thus, the transmit symbol on the \( k \)th subcarrier of the \( p \)th branch is given as \( S_p[k] = C_p^{(i)} \) for \( l = 1...L \) where \( k_l \) is calculated with \( i \) and \( l \) by Eq. (1). From \( N_T \) antenna branches, \( N_T \) OFDM signal streams are transmitted. The received signal model of the overloaded MIMO system using a repetition code is presented in Fig. 2.

The OFDM signal of the \( p \)th branch is given as

\[
u_{p}[n] = \sum_{k=0}^{N-1} P_p[k] S_p[k] \exp\left(\frac{j2\pi n k}{N}\right)
\]

(2)

where \( P_p[k] \) is the precoding coefficient on the \( k \)th subcarrier, \( S_p[k] \) is the coded symbol on the \( k \)th subcarrier, \( n(n = 0, 1, ..., N - 1) \) is the time index, and \( N \) is the size of the inverse discrete Fourier transform (IDFT). The last part of the OFDM symbol is replicated and inserted at the beginning as a guard interval (GI). The generated OFDM signal passes through a transmit filter. Therefore, the transmit signal in the baseband is given as

\[
x_{p}(t) = \sum_{n=-N_{GI}}^{N-1} u_{p}[n] p_{s}(t - nT_s)
\]

(3)

where \( x_{p}(t) \) is the OFDM symbol in the \( p \)th signal stream, \( p_{s}(t) \) is the impulse response of the transmit filter, \( T_s \) is the sampling interval, and \( N_{GI} \) is the GI length. The received signal goes through a receive filter and the sum of the received signals in the \( q \)th antenna is given by

\[
y_q(t) = \sum_{p=1}^{N_T} y_{qp}(t) + w_q(t)
\]

(4)

where \( y_{qp}(t) \) is the received signal from the \( p \)th transmit antenna to the \( q \)th receive antenna, which is calculated as

\[
y_{qp}(t) = \sum_{n=-N_{GI}}^{N-1} u_{p}[n] h_{qp}(t - nT_s)
\]

(5)

and \( w_q(t) \) is the noise in the \( q \)th receive antenna, \( h_{qp}(t) \) is the impulse response of the composite channel including the transmit and received filters. In an A/D converter, the received signal is digitized to discrete samples as

\[
y_q[n] = y_q(nT_s).
\]

(6)

The signal on the \( q \)th subcarrier after removing the GI and taking the DFT of \( N \) samples is given as

\[
Z_q[k] = \sum_{n=0}^{N-1} y_q[n] \exp\left(-j\frac{2\pi nk}{N}\right)
\]
expressed in the vector form as Eq. (8) where \( H_{qp} \) symbol and is given in Eq. (1), the \( i \) signal streams. The joint ML decoding for the \( i \)th codeword is expressed as \[ \hat{C}_{i} = \arg \min_{\mathbf{C}_{i}} \sum_{l=1}^{L} \sum_{q=1}^{N_{T}} |Z_{q}[k]| \]

where \( k \) is given in Eq. (1) with \( i \) and \( l \), \( \Omega \) is the set of candidate codeword vectors, and

\[ \hat{C}_{p} = [\hat{C}_{p} \hat{C}_{p} \cdots \hat{C}_{p} \cdots \hat{C}_{p}]^{T} \]  

In the conventional scheme no precoding is assume, i.e., \( P_{p}[k_{i}] = 1 \) for any \( p \) and \( k_{i} \).
the channel response on different subcarriers in Eq. (8) increases. It then increases the correlation between the code-words transmitted from different antennas. The correlation reduces frequency diversity gain and deteriorates the separation capability of the signal streams [13], [14]. Thus, the open-loop correlation reduction precoding scheme is proposed in this paper.

In the proposed scheme, the precoding coefficients are multiplexed to the coded symbols of the codeword. The coefficients are selected from a DFT matrix. The precoding coefficient for the \( k \)th coded symbol of the \( i \)th codeword in the \( p \)th signal stream, \( P_{p}[k_i] \), is given as

\[
P_{p}[k_i] = \exp \left(-j \frac{2\pi (l-1)(p-1)}{N_T}\right)
\]

where \( k \) is given by Eq. (1). Because of the length of the coefficient sequence, that is expressed as \([P_{p}[k_1] \ P_{p}[k_2] \cdots P_{p}[k_L]]\), the coefficient sequences are not orthogonal each other. Even though the coefficient sequences are not orthogonal, it can still reduce the correlation between the codewords transmitted on different streams. Since the precoding coefficients are derived from the parameters of \( i, l, \) and \( p \), it is also known to the receiver. From Eq. (13), the metric for the joint ML decoding is given as

\[
\begin{align*}
\sum_{l=1}^{L} \sum_{q=1}^{N_q} |Z_q[k_l]| - \sum_{q=1}^{N_q} & P_{p}[k] H_{q,p}[k] \hat{C}_p \sum_{l=1}^{L} \sum_{p=1}^{N_T} |P_{p}[k_l] H_{q,p}[k_l] \hat{C}_p^*| \\
= & \sum_{q=1}^{N_q} \sum_{l=1}^{L} \left( \sum_{p=1}^{N_T} |P_{p}[k_l] H_{q,p}[k_l] C_p - \sum_{p' \neq p}^{N_T} P_{p'}[k_l] H_{q,p}[k_l] C_{p'}| \right) \\
\times & \left( \sum_{l=1}^{L} \sum_{p=1}^{N_T} |P_{p}[k_l] H_{q,p}[k_l] (C_p - \hat{C}_p)| \right)^2 + I_C \\
= & \sum_{q=1}^{N_q} \sum_{l=1}^{L} \sum_{p=1}^{N_T} \sum_{p' \neq p}^{N_T} |P_{p'}[k_l] P_{p}[k_l]^* H_{q,p}[k_l] H_{q,p'}[k_l]| \\
\times & (C_p - \hat{C}_p^*) (C_{p'} - \hat{C}_{p'})^* 
\end{align*}
\]

where * indicates complex conjugate, \( I_C \) is the sum of the cross correlation term among the signal streams and is given as

\[
I_C = \sum_{q=1}^{N_q} \sum_{l=1}^{L} \sum_{p=1}^{N_T} \sum_{p' \neq p}^{N_T} |P_{p'}[k_l] P_{p}[k_l]^* H_{q,p}[k_l] H_{q,p'}[k_l]| \\
\times (C_p - \hat{C}_p^*) (C_{p'} - \hat{C}_{p'})^* 
\]

The approximation in Eq. (18) is because the noise term, \( W_q[k_l] \), in Eq. (8) is neglected for simplicity. If \( H_{q,p}[k_l] H_{q,p'}[k_l]^* \) is almost constant over all the subcarriers due to the constant channel component, the cross correlation term may increases when \( C_p \neq \hat{C}_p \). Thus, the correlation between the precoding coefficient sequences, \( \sum_{l=1}^{L} P_{p}[k_l] P_{p}[k_l]^* \), should be small to reduce the total amount of interference by the cross correlation.

3.2 Precoding for Amplitude Variation

Even though the pulse of each symbol is rotated, the ill-condition of the channel matrix cannot be avoided [20]. This is because the channel response includes the constant channel component whose phase is random owing to propagation conditions. Thus, the amplitude of each symbol should also be varied with the coefficient. In order to further reduce the correlation, window coefficients based on an optimized better than raised-cosine (OBTRC) pulse is applied [22].

From Eq. (17), if the codeword length, \( L \), is the same as the number of signal streams, the correlation of the code-words on the \( p \)th and \( p' \)th \((p \neq p')\) streams diminishes with orthogonal precoding coefficient sequences. However, if \( l \) reaches only up to \( L < N_T \), the precoding coefficient sequences for the \( p \)th and the \( p' \)th streams are not orthogonal. The cross correlation of the precoding coefficient sequences is equivalent to the amount of inter-carrier interference (ICI) of OFDM subcarriers with the frequency offset. This cross correlation is given as

\[
\sum_{l=1}^{L} P_{p}[k_l] P_{p'}^*[k_l] = \sum_{l=1}^{L} \exp \left(-j \frac{2\pi (l-1)(p-p')}{N_T}\right).
\]

The OBTRC pulse has been proposed to reduce the ICI of the OFDM subcarriers with the frequency offset.

The coefficients for odd-numbered streams are given as

\[
P_{odd}[k_l] = \begin{cases} 1, & l < \alpha \\ \exp \left(-\ln 2 \left(\frac{\alpha - l}{\beta}\right)^\nu\right), & \alpha \leq l \leq \alpha + \beta \\ \frac{1}{2}, & \alpha + \beta < l \end{cases}
\]

and those for even-numbered streams is

\[
P_{even}[k_l] = P_{odd}[k_l+1].
\]

The coefficient for the \( k \)th coded symbol in the \( p \)th signal stream is then given as follows;

\[
P_{p}[k_l] = \begin{cases} P_{odd}[k_l] \exp \left(-j \frac{2\pi (l-1)(p-1)}{N_T}\right), & p \text{ is odd} \\ P_{even}[k_l] \exp \left(-j \frac{2\pi (l-1)(p-1)}{N_T}\right), & p \text{ is even} \end{cases}
\]

4. Numerical Results

4.1 Simulation Conditions

Simulation conditions are presented in Table 1. Some of the specifications such as the bandwidth or the number of subcarriers follow the IEEE 802.11a/g standards. A coded symbol is modulated with QPSK on each subcarrier and multiplexed with OFDM. The number of data subcarriers is 48
Table 1  Simulation conditions.

<table>
<thead>
<tr>
<th>Modulate scheme</th>
<th>QPSK/OFDM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of subcarriers</td>
<td>64</td>
</tr>
<tr>
<td>Number of date subcarriers</td>
<td>48 (Δ=8)</td>
</tr>
<tr>
<td>Bandwidth of subcarrier</td>
<td>312.5 [kHz]</td>
</tr>
<tr>
<td>Guard interval length</td>
<td>16 samples</td>
</tr>
<tr>
<td>Number of transmit antennas</td>
<td>4, 6, 8</td>
</tr>
<tr>
<td>Number of receive antennas</td>
<td>1</td>
</tr>
<tr>
<td>Parameters of window coefficients</td>
<td>α = 1, β = 3</td>
</tr>
<tr>
<td>Block coding</td>
<td>Repetition code</td>
</tr>
<tr>
<td>Coding rate</td>
<td>1/4</td>
</tr>
<tr>
<td>Interleaver</td>
<td>Block interleaver (8x6)</td>
</tr>
<tr>
<td>Channel model</td>
<td>Rician fading</td>
</tr>
<tr>
<td>Channel estimation</td>
<td>Ideal</td>
</tr>
</tbody>
</table>

and the DFT size is 64. Thus, the number of null subcarriers, Δ, is 16. The guard interval length is set to 16 samples. The bandwidth of each subcarrier is 312.5 kHz. The number of transmit antennas is 4, 6, and 8 while the number of receive antennas is set to 1. Thus, the number of signal streams is 4, 6, and 8. The coding rate of the repetition code is 1/4. A block interleaver with the size of (8 x 6) is employed to spread the coded symbols over the subcarriers. In order to show the validity of the proposed scheme, a Rician fading channel, a 1 path Rayleigh fading channel, and an independent Rayleigh fading channel are assumed. These channels corresponds to a flat fading channel in LOS environment, that in NLOS environment, and a frequency-selective fading channel in NLOS environment within the OFDM bandwidth, respectively. On the independent Rayleigh fading channel model each coded symbol is subject to independent Rayleigh fading distribution. Perfect channel estimation in the receiver is assumed. In the performance figures, $E_b$ is the bit energy and $N_0$ is the noise spectrum density.

4.2 Numerical Results

4.2.1 Design of Precoding for Amplitude Variation

The parameters of window coefficients for amplitude variation are set as $n = 1$, $\alpha = 1$, and $\beta = 3$, in this paper. With these parameters the window coefficients are given as Fig.4. The correlation between coefficient sequences for the $p$th and $p'$th streams are given in Table 2. Here, the number of signal streams is assumed to be $N_T = 8$. The correlation is calculated with the following equation:

$$\rho_{pp'} = \frac{\sum_{i=1}^{L} P_p[k_i]P_{p'}[k_i]}{\sqrt{\sum_{i=1}^{L} |P_p[k_i]|^2} \sqrt{\sum_{i=1}^{L} |P_{p'}[k_i]|^2}}$$  \hspace{1cm} (24)

and the DFT size is 64. Thus, the number of null subcarriers, $\Delta$, is 16. The guard interval length is set to 16 samples. The bandwidth of each subcarrier is 312.5 kHz. The number of transmit antennas is 4, 6, and 8 while the number of receive antennas is set to 1. Thus, the number of signal streams is 4, 6, and 8. The coding rate of the repetition code is 1/4. A block interleaver with the size of $(8 \times 6)$ is employed to spread the coded symbols over the subcarriers. In order to show the validity of the proposed scheme, a Rician fading channel, a 1 path Rayleigh fading channel, and an independent Rayleigh fading channel are assumed. These channels corresponds to a flat fading channel in LOS environment, that in NLOS environment, and a frequency-selective fading channel in NLOS environment within the OFDM bandwidth, respectively. On the independent Rayleigh fading channel model each coded symbol is subject to independent Rayleigh fading distribution. Perfect channel estimation in the receiver is assumed. In the performance figures, $E_b$ is the bit energy and $N_0$ is the noise spectrum density.

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### Table 2  Example of cross correlation of window coefficient sequences ($N_T = 8$).

<table>
<thead>
<tr>
<th>p'</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.000</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>2</td>
<td>0.653</td>
<td>1.000</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>3</td>
<td>0.300</td>
<td>0.653</td>
<td>1.000</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>4</td>
<td>0.271</td>
<td>0.000</td>
<td>0.653</td>
<td>1.000</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>5</td>
<td>0.000</td>
<td>0.271</td>
<td>0.000</td>
<td>0.653</td>
<td>1.000</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>6</td>
<td>0.653</td>
<td>0.000</td>
<td>0.271</td>
<td>0.000</td>
<td>0.653</td>
<td>1.000</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>7</td>
<td>0.000</td>
<td>0.271</td>
<td>0.000</td>
<td>0.271</td>
<td>0.000</td>
<td>0.653</td>
<td>1.000</td>
<td>-</td>
</tr>
<tr>
<td>8</td>
<td>0.653</td>
<td>0.000</td>
<td>0.271</td>
<td>0.000</td>
<td>0.271</td>
<td>0.000</td>
<td>0.653</td>
<td>1.000</td>
</tr>
</tbody>
</table>

(a) W/O amplitude variation

(b) With amplitude variation
The maximum correlation value reduces to 0.653 or 0.574 with or without amplitude variation while it is 1.0 in the conventional overloaded MIMO.

The correlation is smaller if the window coefficient for each stream have larger variation. However, larger variation in the coefficients means larger variation in the amplitude of the coded symbols. As explained in the following section, the variation in the amplitude reduces diversity gain on the independent Rayleigh fading channel [23], [24]. Thus, it is a trade-off between the correlation and the diversity gain. As it has been shown in [24] that the symbol error rate degradation is less significant, the maximum difference in the amplitude of the coded symbols is set to 1/2.

4.2.2 BER Performance

The BER performance curves on a Rician fading channel ($K=10$) and a 1 path Rayleigh fading channel with the proposed precoding scheme are shown in Figs. 5 and 6. The number of transmit antennas is 8. While the receiver can decode almost no codewords in the absence of precoding, the proposed scheme achieves the BER of $10^{-3}$ at $E_b/N_0$ of 13 dB or 26 dB. On the other hand, the BER performance show almost the same with and without the coefficients for amplitude variation. This is because the Rayleigh fading component changes the amplitude of the coded symbols. The BER versus the number of signal streams on the Rician fading channel ($K=10$) is shown in Fig. 7. $E_b/N_0$ is set to 13 dB. As it is shown that the BERs grow as the number of signal streams increases. However, the BER degradation is suppressed with the proposed precoding schemes.

The BER performance on the Rician fading channel ($K=\infty$) with the proposed precoding scheme is shown in Fig. 8. The number of transmit antennas is 8. When the transmitter does not use the precoding, the BER is close to 0.5. In contrast, by using the proposed precoding for phase rotation, the BER improves to $2 \times 10^{-2}$. However, an error floor appears. Since the coded symbols in all streams have the same amplitude, the candidate signal constellation points overlap with a certain probability [16]. With the pre-
coding for amplitude variation, the BER performance reduces to $10^{-4}$ at $E_b/N_0$ of 15 dB. The BER versus the number of signal streams on the Rician fading channel ($K = \infty$) is shown in Fig. 9. $E_b/N_0$ is set to 10 dB. It is also clear that the precoding scheme with phase rotation and amplitude variation can suppress the BER degradation created by signal multiplexing.

The BER performance on the independent Rayleigh fading channel with the proposed precoding scheme is shown in Fig. 10. The number of transmit antennas is 8. The BERs of these three schemes are almost the same. In [24], it has been shown that the BER performance deteriorates if the signal is received by multiple branches (antenna branches in [24]) while $L$ subcarriers in the proposed scheme with unequal powers. Since the proposed scheme varies the amplitude of the coded symbols over the $L$ subcarriers with the window coefficients, it increases the BER especially when the subcarriers are subject to independent fading [23], [24]. It can be observed through the BER curves with and without amplitude variation in Fig. 10 that the power difference among the $L$ symbols does not lead to the significant BER degradation.

4.2.3 BER Performance with Higher Order Modulation

The BER performance with 16QAM signals is presented in Figs. 11 and 12. The Rician fading channels with $K = 10$ and $K = \infty$ are assumed in Figs. 11 and 12. The difference from Fig. 5, the performance of the proposed precoding scheme for phase rotation shows the error floor on the Rician channel with $K = 10$. This is also because the candidate signal constellation points overlaps even though the precoding scheme for phase rotation is applied. Since the number of the candidate signal constellation points increases with 16QAM symbols and the fading channel varies the amplitude of the received signal, more overlaps among the candidate signal constellation points occur and the level of the error floor increases. On the other hand, the error floor on the channel with $K = \infty$ is lower with the proposed precod-
ing for phase rotation. On the channel with a fixed amplitude response, the different signal amplitudes in multi-level modulation reduces the probability of the overlaps among the candidate signal constellation points. On both channels, the proposed precoding scheme for amplitude variation and phase rotation reduces the BER level of the error floor effectively.

5. Conclusions

This paper has presented a correlation reduction precoding scheme that uses predetermined coefficients for phase rotation and amplitude variation over the codeword. The phase rotation coefficients are derived by the DFT coefficients while the amplitude variation coefficients are calculated based on the OBTRC pulse. With the use of the precoding for phase rotation, the receiver is able to separate the signal streams even those received on the Rician channels. Furthermore, with the use of the precoding for amplitude variation, the proposed scheme eliminates the error floor of the BER performance curves that appear on the channel with only the constant component. Even though the proposed scheme leads to the gain imbalance in the channel responses on different subcarriers, the performance on the independent Rayleigh fading channel shows almost the same BER as the conventional scheme.

Acknowledgments

This work is supported in part by Ministry of Internal Affairs and Communications in Japan under the project name of Strategic Information and Communications R&D Promotion Programme (SCOPE 141303004).

References

Hikari Matsuoka was born in Tokyo, Japan in 1992. He received his B.E. degree in electronics engineering from Keio University, Japan in 2014. Since April 2013, he has been a graduate student in School of Integrated Design Engineering, Graduate School of Science and Technology, Keio University. His research interests are mainly concentrates on MIMO-OFDM system.

Yoshihito Doi was born in Wakayama, Japan in 1989. He received his B.E. degree in electronics engineering from Keio University, Japan in 2013. Since April 2013, he has been a graduate student in School of Integrated Design Engineering, Graduate School of Science and Technology, Keio University. His research interests are mainly concentrated on MIMO-OFDM system.

Tatsuro Yabe was born in Shizuoka, Japan, in 1990. He received his B.E. degree in electronics engineering from Keio University, Japan, in 2013. Since April 2013, he has been a graduate student in the School of Integrated Design Engineering, Graduate School of Science and Technology, Keio University. His research interests are in MIMO systems.

Yukitoshi Sanada was born in Tokyo in 1969. He received his B.E. degree in electrical engineering from Keio University, Yokohama, Japan, his M.A.Sc. degree in electrical engineering from the University of Victoria, B.C., Canada, and his Ph.D. degree in electrical engineering from Keio University, Yokohama, Japan, in 1992, 1995, and 1997, respectively. In 1997 he joined the Faculty of Engineering, Tokyo Institute of Technology as a Research Associate. In 2000, he joined Advanced Telecommunication Laboratory, Sony Computer Science Laboratories, Inc, as an associate researcher. In 2001, he joined the Faculty of Science and Engineering, Keio University, where he is now a professor. He received the Young Engineer Award from IEICE Japan in 1997. His current research interests are in software defined radio, cognitive radio, and OFDM systems.