

Channel Ranking based Joint Symbols Detection for MQRD-PCM/MIMO-OFDM

Chang-Jun AHN
Hiroshima City University

Incheol JEONG
Sungkonghoe University

Abstract MIMO-OFDM is considered a key technology in emerging high-data rate systems. With MIMO techniques, the transmission quality deteriorates due to co-channel interference (CCI). Several signal detection schemes have been proposed to mitigate this problem. However, it is impractical to use the conventional methods without reducing their computational complexity. Previously, we have proposed a parallel detection algorithm using multiple QR decompositions with permuted channel matrix (MQRD-PCM) for MIMO-OFDM to reduce the system complexity. This method achieves a good BER performance with low system complexity. However, since MQRD-PCM is a kind of parallel detection method, the wrong detection probability is increased due to the bad channel SINR of the transmitted signals. To overcome this problem, in this paper, we propose the channel ranking based joint symbols detection for MQRD-PCM/MIMO-OFDM.

1 Introduction

Orthogonal frequency division multiplexing (OFDM) is one of the most promising physical layer technologies for high data rate wireless communications due to its robustness to frequency selective fading, high spectral efficiency, and low computational complexity [1]. OFDM can be used in conjunction with a multiple-input multiple-output (MIMO) transceiver to increase the diversity gain and the system capacity by exploiting spatial domain. Because the OFDM system effectively provides numerous parallel narrowband channels, MIMO-OFDM is considered a key technology in emerging high-data rate systems such as 4G, IEEE 802.16, IEEE 802.11n, and the next generation broadcasting systems. [2],[3].

However, with MIMO techniques, the transmission quality deteriorates due to co-channel interference (CCI). Several signal detection schemes have been proposed to mitigate this problem, such as minimum mean square error (MMSE), parallel or serial interference canceller (PIC or SIC), and maximum likelihood detection (MLD). MLD, which is the best performing scheme, gen-

erates replicas of the received signals from candidates of the desired and CCI signals using pilot symbols [4]. Be that as it may, it is impractical to use a full MLD without reducing its computational complexity, since it would be prohibitively large for implementation. Regarding the problem of implementation, a promising approach that applies QR decomposition in association with an M-algorithm to MLD has been proposed as QRD-M [5]. However, the QRD-M performance depends on the number of surviving symbol replica candidates. When QRD-M is used with a small number of surviving symbol replica candidates, the performance declines. Meanwhile, when there is a large number of surviving symbol replica candidates and the transmitter antenna branches, QRD-M requires a large memory to maintain their branch metrics, and long latency time is also required [6].

To reduce the above-mentioned QRD-M problems, we have proposed a parallel detection algorithm using multiple QR decompositions with permuted channel matrix (MQRD-PCM) for MIMO-OFDM [7]. Since MQRD-PCM/MIMO-OFDM is a kind of parallel detection method, the wrong detection probability is increased due to the bad channel SINR of the transmitted signals. As a result, the average BER performance of MQRD-PCM/MIMO-OFDM is influenced by the wrong detection probability of the bad channel SINR. To overcome this problem, in this paper, we propose the channel ranking based joint symbols detection for MQRD-PCM/MIMO-OFDM system. This paper is organized as follows. The system model is described in Section 2. The configuration of the proposed system is described in Section 3. In Section 4, we show the simulation results. Finally, the conclusion is given in Section 5.

2 System Model

2.1 Channel Model

We assume that the propagation channel consists of L discrete paths with different time delays. The impulse response between the m th transmitter and the n th re-

ceiver antenna $h_{m,n}(\tau, t)$ is represented as

$$h_{m,n}(\tau, t) = \sum_{l=0}^{L-1} h_{m,n,l}(t) \delta(\tau - \tau_{m,n,l}), \quad (1)$$

where $h_{m,n,l}$, $\tau_{m,n,l}$ are the complex channel gain and the time delay of the l th propagation path, respectively. The channel transfer function $h_{m,n}(f, t)$ is the Fourier transform of $h_{m,n}(\tau, t)$ and is given by

$$\begin{aligned} h_{m,n}(f, t) &= \int_0^{\infty} h_{m,n}(\tau, t) \exp(-j2\pi f\tau) d\tau \\ &= \sum_{l=0}^{L-1} h_{m,n,l}(t) \exp(-j2\pi f\tau_{m,n,l}). \end{aligned} \quad (2)$$

2.2 MIMO/OFDM Transmitter

The transmitter block diagram of MIMO/OFDM system is shown in Fig. 1(a). Firstly, the coded binary data sequence is modulated, and N_p pilot symbols are appended at the beginning of the sequence. The MIMO/OFDM transmit signal for the m th transmit antenna can be expressed in its equivalent baseband representation as

$$\begin{aligned} s_m(t) &= \sum_{i=0}^{N_p+N_d-1} g(t-iT) \cdot \left\{ \sqrt{\frac{2S}{N_c}} \sum_{k=0}^{N_c-1} u_m(k, i) \right. \\ &\quad \left. \cdot \exp[j2\pi(t-iT)k/T_s] \right\}, \end{aligned} \quad (3)$$

where N_c is the number of carriers, T_s is the FFT time length, N_d and N_p are the number of data and pilot symbols, S is the average transmitting power, T is the OFDM symbol length, respectively. The frequency separation between adjacent orthogonal subcarriers is $1/T_s$ and can be expressed, by using the k th subcarrier of the i th modulated symbol $d(k, i)$ with $|d(k, i)| = 1$, as

$$u_m(k, i) = c_{PN}(k) \cdot d_m(k, i), \quad (4)$$

where c_{PN} is a long pseudo-noise (PN) sequence as a scrambling code to reduce the peak average power ratio (PAPR). For example, if we consider “1” and “-1” chips, the phase rotation of the signal is 0 deg and 180 deg, respectively. Therefore, we can reduce the PAPR with considering the randomized phase of the signal. In this case, the chip rate of c_{PN} is the same as the sampling rate of the OFDM signal. The guard interval T_g is inserted in order to eliminate the inter-symbol interference (ISI) due to multipath fading, and hence, we have

$$T = T_s + T_g. \quad (5)$$

In Eq. (3), $g(t)$ is the transmission pulse given by

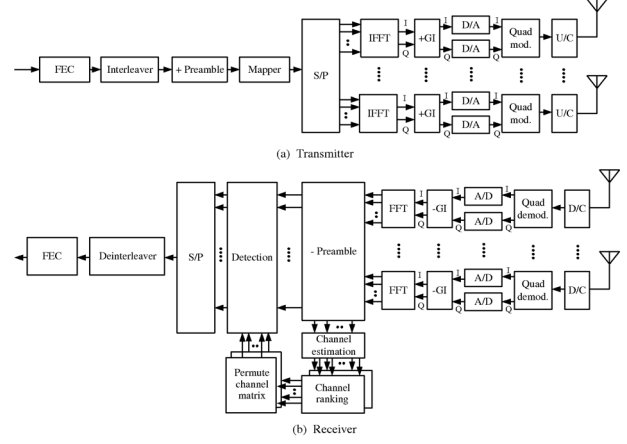


Fig. 1: Proposed MIMO/OFDM system.

$$g(t) = \begin{cases} 1 & \text{for } -T_g \leq t \leq T_s \\ 0 & \text{otherwise.} \end{cases} \quad (6)$$

For $0 \leq i \leq N_p - 1$, the transmitted pilot signal of the k th subcarrier of the m th transmit antenna element is given by

$$d_m(k, i) = \Phi_m(i) \quad (7)$$

where Φ is the orthogonal spreading code. For 4×4 MIMO systems, the orthogonal spreading code Φ as $\Phi_1 = \{1, 1, 1, 1\}$, $\Phi_2 = \{1, -1, 1, -1\}$, $\Phi_3 = \{1, 1, -1, -1\}$, and $\Phi_4 = \{1, -1, -1, 1\}$ are used as pilot signals for the different transmit antennas.

2.3 MIMO/OFDM Receiver

The receiver structure is illustrated in Fig. 1(b). By applying the FFT operation, the received signal $\mathfrak{R}(t)$ is resolved into N_c subcarriers. The received signal for the n th received antenna $\mathfrak{R}_n(t)$ in the equivalent baseband representation can be expressed as

$$\mathfrak{R}_n(t) = \sum_{m=0}^{M-1} \int_{-\infty}^{\infty} h_{m,n}(\tau, t) s_m(t - \tau) d\tau + \zeta_n(t), \quad (8)$$

where $\zeta_n(t)$ is additive white Gaussian noise (AWGN) with a single sided power spectral density of N_0 for the n th received antenna. The k th subcarrier $\tilde{\mathfrak{R}}_n(k, i)$ is given by

$$\begin{aligned} \tilde{\mathfrak{R}}_n(k, i) &= \frac{1}{T_s} \int_{iT}^{iT+T_s} \mathfrak{R}_n(t) \\ &\quad \cdot \exp[-j2\pi(t-iT)k/T_s] dt \\ &= \sqrt{\frac{2S}{N_c}} \sum_{m=0}^{M-1} \sum_{e=0}^{N_c-1} u_m(e, i) \\ &\quad \cdot \frac{1}{T_s} \int_0^{T_s} \exp[j2\pi \cdot (e-k)t/T_s] \end{aligned}$$

$$\cdot \left\{ \int_{-\infty}^{\infty} h_{m,n}(\tau, t + iT)g(t - \tau) \cdot \exp(-2\pi e\tau/T_s)d\tau \right\} dt + n_n(k, i), \quad (9)$$

where $n_n(k, i)$ is AWGN noise with zero-mean and a variance of $2N_0/T_s$. After abbreviation, Eq. (9) can be rewritten as

$$\begin{aligned} \tilde{\mathfrak{R}}_n(k, i) &\approx \frac{1}{T_s} \sqrt{\frac{2S}{N_c}} \sum_{m=0}^{M-1} \sum_{e=0}^{N_c-1} u_m(e, i) \\ &\cdot \int_0^{T_s} \exp[j2\pi \cdot (e - k)t/T_s] \\ &\cdot \left\{ \int_{-\infty}^{\infty} h_{m,n}(\tau, t + iT)g(t - \tau) \cdot \exp(-2\pi e\tau/T_s)d\tau \right\} dt + n_n(k, i) \\ &= \sqrt{\frac{2S}{N_c}} \sum_{m=0}^{M-1} h_{m,n}(k, i)u_m(k, i) + n_n(k, i). \quad (10) \end{aligned}$$

After descrambling, the output signal $y_n(k, i)$ of the n -th receive antenna element is given by

$$\begin{aligned} y_n(k, i) &= \frac{c_{PN}^*(k)}{|c_{PN}(k)|^2} \{ \tilde{\mathfrak{R}}_n(k, i) \}, \\ &= \sqrt{\frac{2S}{N_c}} \sum_{m=0}^{M-1} h_{m,n}(k, i)d_m(k, i) + n_n(k, i), \quad (11) \end{aligned}$$

where $\frac{c_{PN}^*(k)}{|c_{PN}(k)|^2}$ is the descrambling operation. Since descrambling operation is to rotate the phase of each subcarrier by using PN codes as shown in Eq. (4), we can use the same notation $n_n(k, i)$ of Eqs. (10) and (11). In the evaluation, we estimated the channel state information (CSI) using the orthogonal pilot signals as shown in Eq. (7). The estimated CSI of the k th subcarrier between the m th transmitter and the n th receiver $h_{m,n}(k)$ is calculated by

$$h_{m,n}(k) = \frac{1}{N_p} \sum_{i=0}^{N_p-1} y_n(k, i) \otimes \Phi_m(i), \quad (12)$$

where \otimes denotes convolution. Alternately, (11) can be written in matrix form as

$$Y(k) = H(k)D(k) + N(k) \quad k = 1, \dots, K \quad (13)$$

where $Y(k)$ is the $N \times 1$ received signal matrix, $D(k)$ is the $M \times 1$ transmitted signal matrix, $H(k)$ is the $N \times M$ channel matrix, and $N(k)$ is the $N \times 1$ noise matrix, respectively.

2.4 MQRD-PCM

From Eq. (13), we can denote the signal model as

$$\begin{bmatrix} y_1(k) \\ y_2(k) \\ \vdots \\ y_N(k) \end{bmatrix} = \begin{bmatrix} h_{1,1}(k) & \dots & h_{1,M}(k) \\ h_{2,1}(k) & \dots & h_{2,M}(k) \\ \vdots & \ddots & \vdots \\ h_{N,1}(k) & \dots & h_{N,M}(k) \end{bmatrix} \cdot \begin{bmatrix} d_1(k) \\ d_2(k) \\ \vdots \\ d_M(k) \end{bmatrix} + \begin{bmatrix} n_1(k) \\ n_2(k) \\ \vdots \\ n_N(k) \end{bmatrix}. \quad (14)$$

Suppose a matrix is decomposed as $H(k) = Q(k) \cdot R(k)$. The QR decomposition is a coordinate rotation that left multiplies the vector $Y(k)$ in Eq. (13) by $Q(k)^H$ to produce a sufficient statistic

$$\begin{aligned} Z(k) &= [z_1(k), z_2(k), \dots, z_M(k)]^T \\ &= Q(k)^H \cdot Y(k) = R(k) \cdot D(k) + \hat{N}(k), \quad (15) \end{aligned}$$

where $\hat{N}(k) = Q(k)^H \cdot N(k)$ and $(\cdot)^T$ is the transpose matrix. Since $R(k)$ is an upper triangular matrix, $d_M(k)$ can be easily detected as $\Theta\left(\frac{z_M(k)}{r_{M,M}(k)}\right)$ where $\Theta\left(\frac{z}{r}\right) = \arg \min (|\eta - r \cdot s|)_{s \in \sum_{c=1}^C d_c}$. If we change the order of $D(k)$, the column order of $H(k)$ is also changed. For example, if we exchange the positions of $d_1(k)$ and $d_M(k)$, we can rewrite Eq.(14) as

$$\begin{aligned} \begin{bmatrix} y_1(k) \\ y_2(k) \\ \vdots \\ y_N(k) \end{bmatrix} &= \begin{bmatrix} h_{1,M}(k) & \dots & h_{1,1}(k) \\ h_{2,M}(k) & \dots & h_{2,1}(k) \\ \vdots & \ddots & \vdots \\ h_{N,M}(k) & \dots & h_{N,1}(k) \end{bmatrix} \\ &\cdot \begin{bmatrix} d_M(k) \\ d_2(k) \\ \vdots \\ d_1(k) \end{bmatrix} + \tilde{N}(k) \\ &= H_1(k)D_1(k) + \tilde{N}(k), \quad (16) \end{aligned}$$

where $H_1(k)$ is the channel matrix that exchanged the first and M th columns, and $D_1(k)$ is the transmitted signal matrix that exchanged the first and M th rows, respectively. The QR decomposition of $H_1(k)$ is given by

$$H_1(k) = Q_1(k) \cdot R_1(k), \quad (17)$$

where $R_1(k)$ is given by

$$R_1(k) = \begin{bmatrix} r_{1,1,1}(k) & r_{1,2,1}(k) & \dots & r_{1,M,1}(k) \\ 0 & r_{2,2,1}(k) & \dots & r_{2,M,1}(k) \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & r_{M,M,1}(k) \end{bmatrix}. \quad (18)$$

From Eqs. (16) and (17), we can obtain the different matrix compared with Eq. (15) as follows,

$$Z_1(k) = Q_1(k)^H \cdot Y(k) = R_1(k) \cdot D_1(k) + \tilde{N}(k). \quad (19)$$

From Eqs. (16), (17), and (18), we also easily calculate $d_1(k)$ as

$$d_1(k) = \Theta \left(\frac{Q_1(k)_M^H \cdot Y(k)}{r_{M,M,1}(k)} \right). \quad (20)$$

where $Q(k)_M^H$ is the M th row of $Q(k)^H$. Moreover, if we exchange the positions of $d_2(k)$ and $d_M(k)$, we also obtain the channel matrix $H_2(k)$ and calculate $d_2(k)$. In the same manner, we can calculate $Q_3(k), \dots, Q_{M-1}(k)$ and $R_3(k), \dots, R_{M-1}(k)$ with permuted channel matrix. The main point of this method is to reorder $D(k)$ such that the coordinate to be detected is in the last row, the order of other signals is unrestricted. As a result, the detected signals are given by

$$\begin{aligned} d_2(k) &= \Theta \left(\frac{Q_2(k)_M^H \cdot Y(k)}{r_{M,M,2}(k)} \right), \\ d_3(k) &= \Theta \left(\frac{Q_3(k)_M^H \cdot Y(k)}{r_{M,M,3}(k)} \right), \\ &\vdots \\ d_{M-1}(k) &= \Theta \left(\frac{Q_{M-1}(k)_M^H \cdot Y(k)}{r_{M,M,M-1}(k)} \right). \end{aligned} \quad (21)$$

Observing Eqs. (20) and (21), we can see that the detection accuracy is depended on the elements $r_{M,M,1}(k), r_{M,M,2}(k), \dots, r_{M,M,M-1}(k)$. The elements mean the channel SINR of the transmitted signals. Therefore, the wrong detection probability is increased due to the bad channel SINR of the transmitted signals.

3 Proposed method

In order to decrease the wrong detection probability of the bad channel SINR, the transmitted signals are ranked according to the estimated received SINR. The estimated channel matrix $\tilde{H}(k)$ at the k th subcarrier after ranking is expressed as

$$\tilde{H}(k) = \begin{bmatrix} \tilde{H}_{1,rank(M)}(k) & \dots & \tilde{H}_{1,rank(1)}(k) \\ \tilde{H}_{2,rank(M)}(k) & \dots & \tilde{H}_{2,rank(1)}(k) \\ \vdots & \ddots & \vdots \\ \tilde{H}_{N,rank(M)}(k) & \dots & \tilde{H}_{N,rank(1)}(k) \end{bmatrix}, \quad (22)$$

where $rank(v)$ is the ranked transmit antenna index at the v th order. From Eq. (22), the left first column means the worst channel SINR. Therefore, $d_{rank(M)}(k)$ shows the worst detection. To decrease the wrong detection probability of the worst channel SINR, we propose the channel ranking based joint symbol detection. Firstly, we permute the channel matrix as

$$\tilde{H}_M(k) = \begin{bmatrix} \dots & \tilde{H}_{1,rank(M)}(k) & \tilde{H}_{1,rank(1)}(k) \\ \dots & \tilde{H}_{2,rank(M)}(k) & \tilde{H}_{2,rank(1)}(k) \\ \vdots & \vdots & \vdots \\ \dots & \tilde{H}_{N,rank(M)}(k) & \tilde{H}_{N,rank(1)}(k) \end{bmatrix}, \quad (23)$$

where $\tilde{H}_M(k)$ is the permuted channel matrix with adjacent assignment of the first and M th ranked channel columns. By using Eq. (23), we can rewrite Eq. (14) as

$$\begin{bmatrix} y_1(k) \\ y_2(k) \\ \vdots \\ y_N(k) \end{bmatrix} = \begin{bmatrix} \dots & \tilde{H}_{1,rank(M)}(k) & \tilde{H}_{1,rank(1)}(k) \\ \dots & \tilde{H}_{2,rank(M)}(k) & \tilde{H}_{2,rank(1)}(k) \\ \vdots & \vdots & \vdots \\ \dots & \tilde{H}_{N,rank(M)}(k) & \tilde{H}_{N,rank(1)}(k) \end{bmatrix} \cdot \begin{bmatrix} d_{rank(2)}(k) \\ \vdots \\ d_{rank(M)}(k) \\ d_{rank(1)}(k) \end{bmatrix} + \tilde{N}(k). \quad (24)$$

The QR decomposition of $\tilde{H}_M(k)$ is given by

$$\tilde{H}_M(k) = \tilde{Q}_M(k) \cdot \tilde{R}_M(k), \quad (25)$$

where $\tilde{R}_M(k)$ is given by

$$\tilde{R}_M(k) = \begin{bmatrix} \tilde{r}_{1,1}(k) & \dots & \tilde{r}_{1,M}(k) \\ 0 & \dots & \tilde{r}_{2,M}(k) \\ \vdots & \ddots & \vdots \\ 0 & \dots & \tilde{r}_{M,M}(k) \end{bmatrix}. \quad (26)$$

From Eqs. (24) and (25), we can obtain the following matrix form,

$$\tilde{Z}(k) = \tilde{Q}_M(k)^H \cdot Y(k) = \tilde{R}_M(k) \cdot \tilde{D}_M(k) + \tilde{N}(k), \quad (27)$$

and we can detect the transmitted signals as,

$$\begin{aligned} \tilde{d}_{rank(1)}(k) &= \Theta \left(\frac{\tilde{Q}_M(k)_M^H \cdot Y(k)}{\tilde{r}_{M,M}(k)} \right) = \Theta \left(\frac{\tilde{z}_M(k)}{\tilde{r}_{M,M}(k)} \right) \\ \tilde{d}_{rank(M)}(k) &= \Theta \left(\frac{\tilde{Q}_M(k)_M^H \cdot Y(k) - \tilde{r}_{M-1,M}(k) \cdot \tilde{d}_{rank(1)}(k)}{\tilde{r}_{M-1,M-1}(k)} \right) \\ &= \Theta \left(\frac{\tilde{z}_M(k) - \tilde{r}_{M-1,M}(k) \cdot \tilde{d}_{rank(1)}(k)}{\tilde{r}_{M-1,M-1}(k)} \right). \end{aligned} \quad (28)$$

Furthermore, we permute the channel matrix with adjacent assignment of the second and $M-1$ th ranked channel

columns as

$$\tilde{\mathbf{H}}_{M-1}(k) = \begin{bmatrix} \dots & \tilde{H}_{1,rank(M-1)}(k) & \tilde{H}_{1,rank(2)}(k) \\ \dots & \tilde{H}_{2,rank(M-1)}(k) & \tilde{H}_{2,rank(2)}(k) \\ \vdots & \vdots & \vdots \\ \dots & \tilde{H}_{N,rank(M-1)}(k) & \tilde{H}_{N,rank(2)}(k) \end{bmatrix}. \quad (29)$$

After QR decomposition of $\tilde{\mathbf{H}}_{M-1}(k)$, we can detect the transmitted signals as

$$\begin{aligned} \tilde{a}_{rank(2)}(k) &= \Theta \left(\frac{\tilde{Q}(k)_{M-1}^H \cdot Y(k)}{\tilde{r}_{M,M,M-1}(k)} \right) = \Theta \left(\frac{z_{M-1}(k)}{\tilde{r}_{M,M,M-1}(k)} \right) \\ \tilde{a}_{rank(M-1)}(k) &= \Theta \left(\frac{\tilde{Q}(k)_{M-1}^H \cdot Y(k) - \tilde{r}_{M-1,M,M-1}(k) \cdot \tilde{a}_{rank(2)}(k)}{\tilde{r}_{M-1,M-1,M-1}(k)} \right) \\ &= \Theta \left(\frac{z_{M-1}(k) - \tilde{r}_{M-1,M,M-1}(k) \cdot \tilde{a}_{rank(2)}(k)}{\tilde{r}_{M-1,M-1,M-1}(k)} \right) \end{aligned} \quad (30)$$

where $\tilde{r}_{M,M,M-1}(k)$ is the M th column and low element of $\tilde{\mathbf{R}}_{M-1}(k)$. As the same manner, we obtain the channel matrix $\tilde{\mathbf{H}}_{M-2}(k)$, $\tilde{\mathbf{H}}_{M-3}(k)$, \dots , $\tilde{\mathbf{H}}_{\frac{M}{2}+1}(k)$, and detect the all transmitted signals. Observing Eqs (28) and (30), we can decrease the wrong detection probability of the bad channel SINR by using the channel ranking based joint symbol detection with higher and worse channel SINR of the transmitted signals.

4 Computer Simulated Results

Figure 1 shows a simulation model of the proposed MIMO/OFDM with $N_c = 64$ subcarriers. On the transmitter side, the data stream is firstly encoded. Here, convolutional codes (rate $\mathcal{R}=1/2$, constraint length $\xi=7$) with bit interleaving are used. The coded bits are QPSK modulated and then serial-to-parallel transformed. The OFDM time signals are generated by an IFFT and transmitted from each transmit antenna branch over the frequency selective and time variant radio channel after cyclic extensions have been inserted. The transmitted signals are subject to broadband channel propagation. In this model, $L = 15$ path Rayleigh fading have exponential shapes with path separation $T_{path} = 140nsec$ and the RMS delay spread is $\tau_{rms} = 0.65\mu s$. This situation causes severe frequency selective fading. The maximum Doppler frequency is assumed to be 10Hz. The packets consist of 64 subcarriers and 24 OFDM symbols (number of pilot signals: $N_p = 4$, number of data signals: $N_d = 20$) with an effective data rate of 20 Msymbol/s for each transmit antenna branch.

Figure 2 shows the BER of the full MLD-based MIMO, ZF-based MIMO, QRD-M-based MIMO ($S_1, S_2, S_3 = 4, 4, 4$), MQRD-PCM based MIMO, and the proposed MIMO system with $M=4$, $N=4$ and a 10Hz Doppler

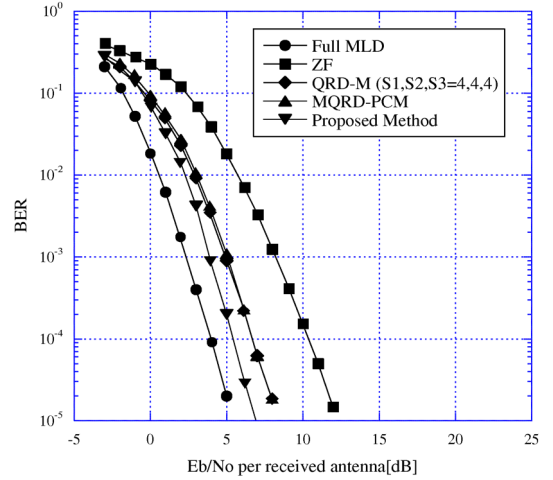


Fig. 2: BER of the full MLD-based MIMO, ZF-based MIMO, QRD-M-based MIMO ($S_1, S_2, S_3 = 4, 4, 4$), MQRD-PCM based MIMO, and the proposed MIMO system with $M=4$, $N=4$ and a 10Hz Doppler frequency.

frequency. In the ZF-based MIMO, the detected signals include enhanced noise components. The performance of the QRD-M depends on the number of surviving symbol replica candidates. With a small number of surviving candidates in the QRD-M, it is difficult to include the desired signal in the surviving symbol replica candidates every time. Thus, the MQRD-PCM based MIMO obtains the approximately same BER at BER of 10^{-4} compared with QRD-M ($S_1, S_2, S_3 = 4, 4, 4$) where S_m is the surviving symbol replica candidates for the m -th stage of M-algorithm corresponding to transmitted signal from the m -th transmit antenna. In MQRD-PCM based MIMO, the wrong detection probability is increased due to the bad channel SINR of the transmitted signals. As a result, the average BER performance of MQRD-PCM/MIMO-OFDM is influenced by the wrong detection probability of the bad channel SINR of the transmitted signals. Meanwhile, the proposed MIMO systems consider the channel ranking based joint symbols detection. Therefore, the proposed MIMO can achieve better BER than those of MQRD-PCM and QRD-M in high E_b/N_0 . The full MLD produces the best BER performance. This is because the full MLD calculates all possible transmitted signals.

Figure 3 shows the BER of QRD-M-based MIMO ($S_1, S_2, S_3 = 4, 4, 4$), QRD-M-based MIMO ($S_1, S_2, S_3 = 4, 16, 16$), and the proposed MIMO with $M=4$, $N=4$ and a 10Hz Doppler frequency. The performance of the QRD-M depends on the number of

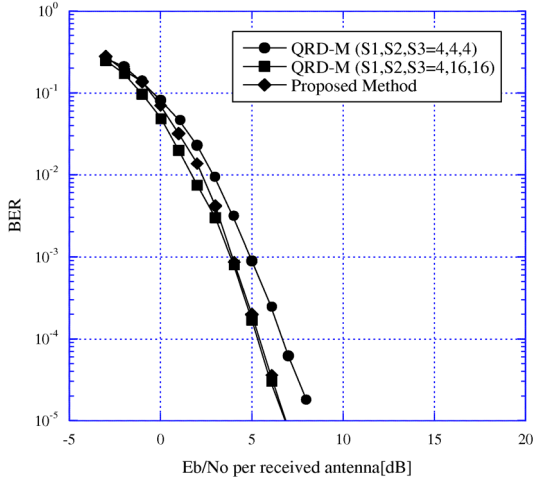


Fig. 3: BER of QRD-M-based MIMO ($S_1, S_2, S_3 = 4, 4, 4$), QRD-M-based MIMO ($S_1, S_2, S_3 = 4, 16, 16$), the proposed MIMO system with $M=4$, $N=4$ and a 10Hz Doppler frequency.

surviving symbol replica candidates. With a small number of surviving candidates in the QRD-M, it is difficult to include the desired signal in the surviving symbol replica candidates every time. Meanwhile, with a large number of surviving candidates in the QRD-M, it is easy to include the desired signal in the surviving symbol replica candidates. Thus, the proposed scheme obtains the approximately same BER and 1.5dB gain at BER of 10^{-4} compared with QRD-M ($S_1, S_2, S_3 = 4, 16, 16$) and QRD-M ($S_1, S_2, S_3 = 4, 4, 4$), respectively.

Table 1 shows the required multiplication per packet. C is the constellation size (QPSK), M is the number of transmit antenna branches, K is the number of carriers, N_d is the number of data symbols, S is the surviving symbol replica candidates for the proposed method, respectively. From Table 2, the total required multiplica-

表 1: Required multiplication per packet.

Detection scheme	Required multiplication
Full MLD	Gen. of replica candidates ($4M^2CK$) Squared Euclidian dist. ($2MC^MKN_d$)
QRD-M	QR decom. of H ($4M^3K$) Q^H ($4M^2KN_d$) Squared Euclidian dist. ($(1 + \sum_{m=1}^{M-1} S_m)2CKN_d$)
MQRD-PCM	QR decom. of H ($4M^4K$) Q^H ($4M^2KN_d$) Squared Euclidian dist. ($2MCKN_d$)
Proposed method	QR decom. of H ($2M^4K$) Q^H ($2M^3KN_d$) Squared Euclidian dist. ($(1 + S)MCKN_d$)

tions per packet of the full MLD, QRD-M ($S_1, S_2, S_3 = 4, 16, 16$), QRD-M ($S_1, S_2, S_3 = 4, 4, 4$), MQRD-PCM based MIMO, the proposed scheme are 2637824, 477184, 231426, 188416, and 299008, respectively. Therefore, the required multiplication per packet of the proposed scheme is reduced about 0.62 and 0.11 times when compared with QRD-M ($S_1, S_2, S_3 = 4, 16, 16$) and the full MLD, respectively.

5 Conclusion

In this paper, we have proposed the channel ranking based joint symbols detection for MQRD-PCM/MIMO-OFDM system. The proposed scheme achieves a good BER performance with low system complexity. The required multiplication per packet of the proposed scheme is reduced about 0.62 and 0.11 times when compared with QRD-M ($S_1, S_2, S_3 = 4, 16, 16$) and the full MLD, respectively.

参考文献

- [1] L. Cimini, "Analysis and simulation of a digital mobile channel using OFDM," *IEEE Trans. on Comm.*, vol. 33, pp. 665-675, July 1985.
- [2] A. V. Zelst, R. V. Nee, and G. Awater, "Space division multiplexing (SDM) for OFDM systems," *Proc. of VTC*, pp. 15-18, 2000.
- [3] A. Nallanathan and Chang Lih Yun "Eigenbeam-space division multiplexing for OFDM systems with optimum resource allocation," *Proc. of GLOBECOM'04*, vol. 4, pp. 2366-2370, 2004
- [4] P. Vandenameele, L. V. Perre, M. G. E. Engels, B. Gyselinck, and H. D. Man, "A combined OFDM/SDMA approach," *IEEE Journal of sel. Area in Commun.*, vol. 18, no. 11, pp. 2312-2321, 2000.
- [5] K. J. Kim, Y. Yue, R. A. Iltis, and J. D. Gibson, "A QRD-M/Kalman Filter-Based Detection and Channel Estimation Algorithm for MIMO-OFDM Systems," *IEEE Trans. on Wireless Commun.*, vol.4, pp710-pp721, 2005.
- [6] H. Kawai, K. Higuchi, N. Maeda, M. Sawahashi, T. Ito, Y. Kakura, A. Ushirokawa, and H. Seki, "Likelihood function for QRM-MLD suitable for soft-decision turbo decoding and its performance for OFCDM MIMO multiplexing in multipath fading channel," *IEICE Trans. on Communication*, vol. E88-B, no. 1, pp. 47-57, 2005.
- [7] C. Ahn, "Parallel Detection Algorithm Using Multiple QR Decompositions with Permuted Channel Matrix for SDM/OFDM," *IEEE Trans. on Vehicular Technology*, vol.57, no.4, pp.2578-2582, July 2008.