An Improved MU-MIMO OFDM Scheme for the 3GPP LTE-Advanced Standard

Wang Yuechao, Wu Zhanji, Wu Guanghao

Beijing University of Post & Telecommunications

Beijing, China

E-mail: wang_yuechao@163.com, wuzhanji@bupt.edu.cn, 604503597@qq.com

Abstract—In order to achieve the required data rates of 1Gbit/s and more in 3GPP LTE-Advanced wireless systems, there is an increased need for high spectral efficiency. Multi-user multiple-input-multiple-output (MU-MIMO) is a promising way to improve frequency usage efficiency. In this paper, a new multi-user MIMO OFDM scheme, named Joint Coding Modulation Diversity (JCMD) is proposed for the 3rd generation partnership project (3GPP) long term evolution advanced (LTE-Advanced). This new scheme can take advantage of the modulation diversity of rotated constellation, the frequency diversity of orthogonal frequency-division multiplexing (OFDM) system and Turbo codes all together. The simulation results show that this proposed scheme can significantly outperform the current Bit Interleaved Coded Modulation (BICM) scheme in the 3GPP LTE-Advanced standard.

Keywords-coded modulation; rotated modulation; modulation diversity.

I. INTRODUCTION

In recent years, demands on high rate wireless communication have increased. Orthogonal Frequency-Division Multiplexing (OFDM) has been proven to be a viable technique, which is proposed for next generation wireless communication system. OFDM is used to transmit data over extremely hostile channel at a comparable low complexity with high data rates [1, 2]. Channel equalization is simplified because OFDM may be viewed as using many slowly modulated narrowband signals rather than one rapidly modulated wideband signal. Compared with the conventional FDM scheme, it can save about half the bandwidth. On the other hand, Multiple-Input Multiple-Output (MIMO) systems have also been one of the key techniques to achieve high rate and high reliability over wireless channels because it can allow modulated symbols to transmit on multiple antennas at the same time to obtain high spectral efficiency. Using sophisticated coding at the transmitter and substantial signal processing at the receiver, the MIMO channel can be provisioned for higher data rates and resistance to multipath fading. A Layered Space-Time (LST) architecture that allows processing of multidimensional signals in the space domain by one dimension processing steps was proposed by Foschini [3]. Bit-Interleaved Coded Modulation (BICM) [4,5,6] is used in most of the existing communication standards (cf. HSPA, IEEE 802.11a/g, IEEE 802.16, DVB-S2, etc.) [7,8,9,10]. In BICM, the channel encoder and the modulator are separated by a bit-level interleaver, which makes the design simple and flexible. Based on BICM-LST, which is the conventional spectral-efficient spatial multiplexing technology to deal with MIMO fading channel, MIMO-OFDM system can be used to overcome the various challenges that come with the high bit-rate, reliable and power efficient transmission demand in wide-band communication, and has become the foundation of advanced wireless transmission technologies in the current 3GPP LTE-Advanced standard.

J. Boutros et al. [11] studied the uncoded rotated multidimensional modulation schemes over Rayleigh fading channels. To distinguish from the other well-known diversity (time, frequency, code, space), the rotated modulation schemes have an essential modulation diversity and the modulation diversity order is the minimum number of distinct components between any two multidimensional components between any two multidimensional constellation points. The schemes are essentially uncoded and can achieve very high modulation diversity, and can approach Addition White Gaussian Noise (AWGN) error performance over independent flat fading channels without time-dispersion. However, it cannot be directly used for frequency selective fading channels with severe inter-symbol interference (ISI). To make full use of modulation diversity and frequency diversity over the time-dispersion fading channels with ISI, it should cooperate with OFDM. A rotated coding modulation OFDM system was proposed in [12], which extends the modulation diversity in coded OFDM systems. Through the combination of rotating the multiple phase shift keying or multiple quadrature amplitude modulation (MPSK/MQAM) constellation and interleaving the symbol components in space-time-frequency domain, the performance of wireless communications systems can be significantly improved.

A MIMO-OFDM system based on the two-dimensional modulation diversity [12] is proposed for the 3GPP LTE-advanced MU-MIMO schemes. Through jointly optimizing Turbo coded MIMO-OFDM and the rotated coding modulation for each user, the proposed scheme can take full advantage of the coding-gain of Turbo codes, the time and frequency diversity of OFDM system and the spatial diversity of MIMO all together. It is named the joint coding and modulation diversity (JCMD) MU-MIMO OFDM scheme. The rest of this paper is organized as follows.

General multi-user MIMO system model is introduced in Section II. In Section III, the JCMD MU-MIMO

OFDM scheme is proposed. Rotational modulation, the component interleaver and the detection algorithm are introduced in Sections IV, V, and VI, respectively. Section VII describes the procedure of selection of the codebook and the user, which makes the system throughput maximized and Section VIII gives the simulation results. Section IX concludes the paper.

II. MU-MIMO SYSTEM MODEL

In the multi-user MIMO system, which is showed in Figure 1, K users are assumed. We consider a MU-MIMO downlink channel, where M_T transmit antennas are located at the base station and M_{R_i} receive antennas are located at the ith user terminal (UT), $i = 1, 2, ..., K$. The total number of receive antennas is as follows

$$M_R = \sum_{i=1}^{K} M_{R_i}.$$

Let the notation $\{M_{R_1}, ..., M_{R_K}\} \times M_T$ denote the antenna configuration of the system. Let the MIMO channel of user i be denoted as $\mathbf{H}_i \in \mathbb{C}^{M_{R_i} \times M_T}$. Then, the combined channel matrix is given by

$$\mathbf{H} = [\mathbf{H}_1^T, \mathbf{H}_2^T, ..., \mathbf{H}_K^T]^T \in \mathbb{C}^{M_R \times M_T} \qquad (1)$$

The data vectors $\mathbf{x}_k \in \mathbb{C}^{b_k \times 1}, k = 1, ..., K$, for the K users are stacked in the vector $\mathbf{x} = [\mathbf{x}_1^T, ..., \mathbf{x}_K^T]^T \in \mathbb{C}^{b \times 1}$. The received vector is given by

$$\mathbf{y} = \mathbf{G}(\mathbf{Hwx} + \mathbf{n}) \qquad (2)$$

where $\mathbf{y} = [\mathbf{y}_1^T, ..., \mathbf{y}_K^T]^T \in \mathbb{C}^{b \times 1}$ is the received data vector, $\mathbf{n} = [\mathbf{n}_1^T, ..., \mathbf{n}_K^T]^T \in \mathbb{C}^{M_R \times 1}$ is the stacked vector of the zero mean additive white Gaussian noise at the input of the receive antennas. The joint precoding matrix and decoding matrices are denoted by $\mathbf{w}$ and $\mathbf{G}$, respectively. Define the joint precoder matrix as

$$\mathbf{w} = [\mathbf{w}_1, \mathbf{w}_2, ..., \mathbf{w}_K] \in \mathbb{C}^{M_T \times b} \qquad (3)$$

where $\mathbf{w}_k \in \mathbb{C}^{M_T \times b_k}$ is the kth user's precoder matrix. Moreover, b_k is the number of data stream sequences transmitted to the kth user, and $b = \sum_{k=1}^{K} b_k \leq rank(\mathbf{H}) \leq \min(M_R, M_T)$ is the total number of the transmitter data streams. The decoding matrix G can be written as

$$\mathbf{G} = \begin{pmatrix} \mathbf{G}_1 & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \mathbf{G}_K \end{pmatrix} \in \mathbb{C}^{b \times M_R} \qquad (4)$$

where $\mathbf{G}_k \in \mathbb{C}^{b_k \times M_{R_k}}$ is the kth user's decoding matrix.

The received symbol at the kth user can be expressed as $\mathbf{r}_k = \mathbf{H}_k \mathbf{wx} + \mathbf{n}_k \in \mathbb{C}^{M_{R_k} \times 1}$, and the received data vector is given by $\mathbf{y}_k = \mathbf{G}_k \mathbf{r}_k = \mathbf{G}_k (\mathbf{H}_k \mathbf{wx} + \mathbf{n}_k) \in \mathbb{C}^{b_k \times 1}$. For the sake of simplicity, we consider a system with

$b_k = N_L, k = 1, ..., K$ and each generic user with J antennas, $M_{R_k} = J, k = 1, ..., K$ in the rest of this paper.

III. JOINT CODING AND MODULATION DIVERSITY MU-MIMO OFDM SCHEME

In [13], the principle of MU-MIMO with per user unitary and rate control (PU2RC) is described. The feedback does not indicate the channel itself, but a preferred precoder from a set of predefined precoders. Based on the conventional MU-MIMO scheme with PU2RC, an improved MU-MIMO OFDM scheme based on modulation diversity is proposed, which is illustrated in Figure 2. The blocks drawn in dotted line are the additional proposed processing based on the conventional MU-MIMO OFDM system in the 3GPP LTE standard.

At the transmitter, the base station firstly selects a group of users in the set of users to be scheduled according to the procedure of PU2RC-MIMO with partial feedback, which will be described in Section VI. Then, the information bits of selected group users are firstly encoded by 3GPP LTE Turbo code [14]. The conventional quadrature amplitude modulation (QAM) is carried out to produce complex-valued symbols by constellation mapper. The conventional modulated symbols are converted into new complex-valued rotated symbols by a rotational modulation mapper. The rotated symbols contain In-phase (I) components and Quadrature (Q) components and subsequently are mapped onto N_L transmission layers. Afterwards, the spatial interleaving and spatial Q-interleaving that will be described in Section V are carried out over each user's N_L transmission layers. The interleaved symbols are then mapped onto different time-frequency resource elements (REs). For the Q-components, an additional time-frequency Q-interleaver is performed. After the codebook-based precoding [15], N_{FFT}-point Inverse Fast Fourier Transform (IFFT) and inserting cyclic prefix (CP) are implemented on each transmit antenna.

At the receiver of each UT, removing CP and N_{FFT} point FFT are carried out first. The minimum mean squared-error (MMSE) equalization in the frequency domain is performed. Afterwards, the received symbols of each spatial stream are obtained after the time-frequency Q-component deinterleaving, time-frequency resource element demapping, spatial Q-deinterleaving, spatial deinterleaving, and layer demapping. For each symbol, the maximum likelihood (ML) rotational demodulation is used to produce the log-likelihood-ratios (LLRs) of encoded bits. Finally, the Turbo decoder utilized the
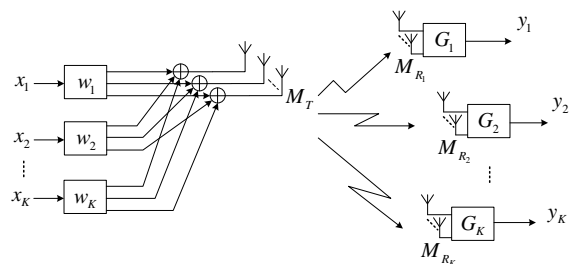

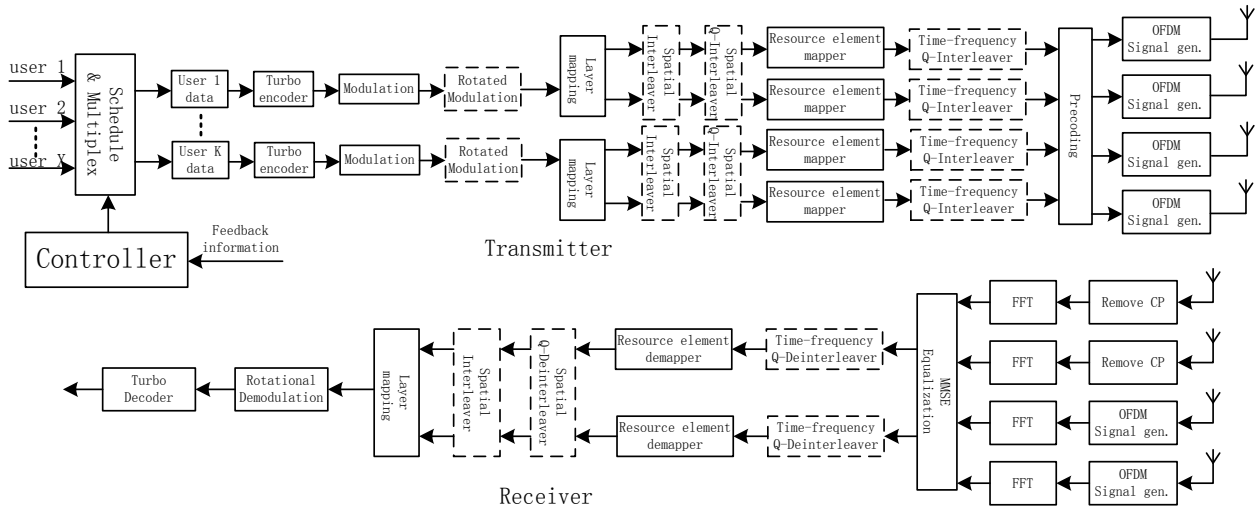Figure 1.   Block diagram of a multi-user MIMO downlink system

Figure 2.   A Transmitter and Receiver Diagram of the JCMD MU-MIMO OFDM scheme

LLRs to decode the information bits.

IV.   ROTATIONAL MODULATION

Compared with the usual QAM, rotated constellation can obtain the modulation diversity by rotating some angle [11]. Combining rotating the conventional constellation and interleaving the components, the performance of wireless communications systems can be improved greatly in fading channels.

At the transmitter, coded bits $\mathbf{K} = (k_1, k_2, ..., k_{N_{bit}})$ are converted into a sequence of complex-valued modulation symbols $\mathbf{D} = (d_1, d_2, ..., d_{N_{Sym}})$ by Gray-mapping [11].

The rotational complex-valued symbol $d_i' = X_i + jY_i$ can be obtained by multiplying the conventional modulated complex-valued symbol $d_i = I_i + jQ_i$ with a rotational matrix $\mathbf{R}$, which is shown as follows.

$$\begin{pmatrix} X_i \\ Y_i \end{pmatrix} = \mathbf{R} \begin{pmatrix} I_i \\ Q_i \end{pmatrix} = \begin{pmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{pmatrix} \begin{pmatrix} I_i \\ Q_i \end{pmatrix} \qquad (5)$$

This processing can also be illustrated in Figure 3. In the conventional square QPSK constellation, the I-component and Q-component of one complex-valued modulated symbol just carry one bit, respectively. After the operation of constellation rotation, both I-component and Q-component of each point in the new constellation carry the information of two bits, which means that the information of two bits exists across the I-component and Q-component. After appropriate interleaving of the Q-components, the fading coefficients of the I-components and Q-components are uncorrelated. The modulation diversity order L of a multidimensional signal set is the minimum number of distinct components between any two constellation points. Assuming independent Rayleigh fading channel and the maximum likelihood (ML) detection, multidimensional QAM constellation becomes insensitive to fading channel when the diversity L is large [16].

By adjusting the rotation angle, the optimum modulation diversity can be obtained to minimize bit error rate (BER). J. Boutros et al. [11] studied the optimum criteria of the uncoded rotation modulation over independent Rayleigh channel, and proposed that the diversity order L and product distance should be maximized [11]. Simulation results show that the best angle of rotation depends on modulation order and coding rate. Also, there is weak correlation between the code rate and the optimum rotation angle, especially from a high coding rate of 3/4 and 8/9. However, it is strongly relevant to the modulation order. For the same code rate, the optimal rotation angle values are different for QPSK, 16QAM and 64QAM, and it decreases as the modulation order increases. Based on the above theoretical analysis and computer simulation, we obtain the optimal rotation angles that are suitable for the improved MU-MIMO OFDM systems based on modulation diversity, which are shown in Table I.

V.   Q-COMPONENT INTERLEAVER

A. Spatial interleaver and Spatial Q-component interleaver

The spatial interleaving is the conventional spiral layer interleaving among each user's transmit layers. Let $\mathbf{x}_t^i$ and $\bar{\mathbf{x}}_t^i$ denote the input-symbol and the output-symbol of the ith transmit layer at the t instant. The interleaving is defined as follows.

$$\bar{\mathbf{x}}_t^k = \mathbf{x}_t^i, k = (i+t) \bmod N_L \qquad (6)$$

where $k, i \in [0, N_L - 1]$. The spatial Q-component interleaver is carried out after the spatial interleaving to ensure the independence of the I-components and Q-components. Let Q_t^i and $\tilde{Q}_t^i$ denote the input Q-component and the output Q-component of the Q-component interleaver on the ith transmit layer at the t instant. The Q-component interleaving is defined as follows.

TABLE I.   THE OPTIMAL ROTATION ANGLES FOR DIFFERENT CODE RATE

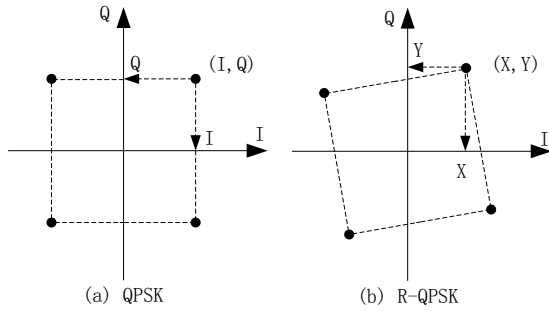| Code rate	QPSK	16QAM	64QAM
3/4	26.6°	18.4°	14.0°
5/6	26.6°	18.4°	14.0°
8/9	26.6°	18.4°	14.0°

Figure 3.   Conventional QPSK constellation (left) and Rotated QPSK constellation

$$\tilde{Q}_t^k = Q_t^i, k = N_L - i - 1 \qquad (7)$$

where $k, i \in [0, N_L - 1]$.

B. Time-frequency Q-component interleaver

The time-frequency Q-component interleaving is aimed to make the I-component and the Q-component in one modulated symbol as uncorrelated as possible in the time and frequency domain. We assume the OFDM system has L subcarriers in frequency domain for each user and N_{sym} OFDM symbols in time domain. One Q-component of a complex-valued modulated symbol is mapped onto one resource element Q (f, t) at the fth subcarrier in the tth OFDM symbol. After time- frequency Q-component interleaving, the output is Q (f, t). The Q-component interleaving is defined as follows.

$$f' = (f + L / 2) \bmod L$$
$$t' = (t + N_{sym} / 2) \bmod N_{sym} \qquad (8)$$

where $f \in [0, L-1], t \in [0, N_{sym}-1]$. Thus, the time interval between I-component and Q-component is $N_{sym} / 2$ OFDM symbol duration. In frequency domain, the frequency interval is half the bandwidth for each user. So, the time-frequency Q-component interleaver can make full use of the frequency diversity and the time diversity of OFDM system, and it can make the I-components and Q-components as uncorrelated as possible. Based on the use of spatial interleaver, the achievable rate of the proposed scheme can be increased, which is demonstrated as follows.

In (3), $\mathbf{w}_k$ denotes the kth user's precoder matrix, and the equivalent channel matrix between the base station and k^{th} user is expressed as $\mathbf{H}_E = \mathbf{H}_k \mathbf{w}_k \in \mathbb{C}^{J \times N_L}$. Let $q = \min(J, N_L)$. Assuming the transmission power is P and the bandwidth is B, and the q singular values are

$\sqrt{\lambda_1}, ..., \sqrt{\lambda_q}$ ($\sqrt{\lambda_1} \geq \sqrt{\lambda_2} \geq, ..., \geq \sqrt{\lambda_q}$). The total rate each user can achieve is the sum of partial achievable rate on each singular-value-decomposition (SVD) spatial layer of the MIMO channel [17].

For the system without spatial Q-component interleaver, the fading coefficients of I-component and Q-component in each symbol are identical. The achievable rate can be calculated as follows.

$$C_1 = B \cdot \log_2 \{ (1 + \frac{\lambda_1 P}{2\sigma^2})(1 + \frac{\lambda_2 P}{2\sigma^2}) \cdots (1 + \frac{\lambda_q P}{2\sigma^2}) \} \qquad (9)$$

For the proposed system, thanks to the spatial Q-component interleaver, the fading coefficients of I-component and Q-component in each symbol are $\sqrt{\lambda_i}$ and $\sqrt{\lambda_{(i+1) \bmod q}}$, $i = 1, ..., q$, respectively. Thus, the achievable rate of the proposed system can be calculate as follows.

$$C_2 = B \cdot \log_2 \{ (1 + \frac{\lambda_1 + \lambda_2}{4\sigma^2} P)(1 + \frac{\lambda_2 + \lambda_3}{4\sigma^2} P) \cdots (1 + \frac{\lambda_q + \lambda_1}{4\sigma^2} P) \} \qquad (10)$$

When $q = 2$, we can easily get the following equations,

$$\begin{aligned} C_1 &= B \cdot \log_2 \{ (1 + \frac{\lambda_1 P}{2\sigma^2})(1 + \frac{\lambda_2 P}{2\sigma^2}) \} \\ &= B \cdot \log_2 \{ 1 + \frac{\lambda_1 + \lambda_2}{2\sigma^2} P + \frac{\lambda_1 \lambda_2}{4\sigma^4} P^2 \} \end{aligned} \qquad (11)$$

and

$$\begin{aligned} C_2 &= B \cdot \log_2 \{ (1 + \frac{\lambda_1 + \lambda_2}{4\sigma^2} P)(1 + \frac{\lambda_2 + \lambda_1}{4\sigma^2} P) \} \\ &= B \cdot \log_2 (1 + \frac{\lambda_1 + \lambda_2}{2\sigma^2} P + \frac{(\lambda_1 + \lambda_2)^2}{16\sigma^4} P^2) \end{aligned} \qquad (12)$$

It can easily obtain $C_1 \leq C_2$, because $(\lambda_1 + \lambda_2)^2 \geq 4\lambda_1 \lambda_2$. Likewise, for $q > 2$, we can get the same result.

VI.   DETECTION AND DEMODULATION

The received symbol at the kth user can be expressed as

$$\mathbf{r}_k = \mathbf{H}_k \mathbf{w}_k \mathbf{x}_k + \sum_{i=1, i \neq k}^{K} \mathbf{H}_k \mathbf{w}_i \mathbf{x}_i + \mathbf{n}_k \qquad (13)$$

where the first term on the right hand side is the useful data for k^{th} user. The middle term represents multi-user interference (MUI) and the last term is the noise vector, which contains i.i.d. complex Gaussian random numbers with variance σ^2. The minimum mean-squared error (MMSE) equalization is used for each antenna of user after the FFT operation. Assuming the perfect channel state information (CSI), the MIMO channel matrix between the base station and user terminal is known to the receiver. Let $\mathbf{H}_k = [\mathbf{h}_1^T, ..., \mathbf{h}_J^T]^T$ denote a $J \times M_T$ channel for that UT. We use $\widetilde{\mathbf{H}}_k = \mathbf{H}_k \mathbf{w}$ to denote the product matrix between the UT's channel and the precoder, where w denotes the precoding matrix. Based on the MMSE criterion, the receiver combining coefficients for the kth user can be defined as follows.

$$\mathbf{G}_k = \tilde{\mathbf{h}}_k^H \left(\widetilde{\mathbf{H}}_k \widetilde{\mathbf{H}}_k^H + 2K\sigma^2 \mathbf{I}_{M_T} \right)^{-1} \qquad (14)$$

$$\tilde{\mathbf{h}}_k = \mathbf{H}_k\mathbf{w}_k = [\widetilde{\mathbf{H}}_k]_k = [\mathbf{H}_k\mathbf{w}]_k \qquad (15)$$

where $\mathbf{w}_k$ denotes the kth user's precoder matrix. $[\bullet]_k$ denotes the column vectors that correspond to k^{th} user. Then, the received data vector is given by

$$\mathbf{y}_k = \mathbf{G}_k\mathbf{r}_k = \mathbf{G}_k\mathbf{H}_k\mathbf{w}_k\mathbf{x}_k + \sum_{i=1,i\neq k}^{K} \mathbf{G}_k\mathbf{H}_k\mathbf{w}_i\mathbf{x}_i + \mathbf{G}_k\mathbf{n}_k$$
$$\qquad\qquad (16)$$
$$= \mathbf{A}_k\mathbf{x}_k + \sum_{i=1,i\neq k}^{K} \widetilde{\mathbf{A}}_i\mathbf{x}_i + \mathbf{G}_k\mathbf{n}_k$$

where $\mathbf{A}_k \in \mathbb{C}^{N_L\times N_L}$, $\widetilde{\mathbf{A}}_i \in \mathbb{C}^{N_L\times N_L}, i=1, i\neq k$. Let $\mathbf{A}_k^{(m,n)}$ denotes the element on the mth row and the nth column. Likewise, $\widetilde{\mathbf{A}}_i^{(m,n)}$ denotes the element $\widetilde{\mathbf{A}}_i(m,n)$, and $\mathbf{G}_k^{(m,n)}$ denotes the element $\mathbf{G}_k(m,n)$. For k^{th} user, the symbol on the receive layer m can be expressed as

$$y_k^{(m)} = \{\mathbf{A}_k^{(m,1)}\mathbf{x}_k^{(1)} + \cdots + \mathbf{A}_k^{(m,m)}\mathbf{x}_k^{(m)} \cdots + \mathbf{A}_k^{(m,N_L)}\mathbf{x}_k^{(N_L)}\}$$
$$+ \{\sum_{i=1,i\neq k}^{K} (\widetilde{\mathbf{A}}_i^{(m,1)}\mathbf{x}_i^{(1)} + \cdots + \widetilde{\mathbf{A}}_i^{(m,N_L)}\mathbf{x}_i^{(N_L)})\} \qquad (17)$$
$$+ \{\mathbf{G}_k^{(m,1)}\mathbf{n}_k^{(1)} + \cdots + \mathbf{G}_k^{(m,J)}\mathbf{n}_k^{(J)}\}$$

Assuming the normalized transmit signal power $\sigma_s^2 = 1$, the signal to interference plus noise ratio (SINR) for the symbol on the receive layer m is estimated as follows.

$$SINR_k^m = \frac{\left|\mathbf{A}_k^{(m,m)}\right|^2}{\sum_{i=1,i\neq m}^{N_L}\left|\mathbf{A}_k^{(m,i)}\right|^2 + \sum_{i=1,i\neq k}^{K}(\left|\widetilde{\mathbf{A}}_i^{(m,1)}\right|^2 + \cdots + \left|\widetilde{\mathbf{A}}_i^{(m,N_L)}\right|^2) + 2\sigma^2\sum_{i=1}^{J}\left|\mathbf{G}_k^{(m,i)}\right|^2}$$
$$\qquad\qquad (18)$$

We assume the rotated constellation point in the transmitter is $S = (S_I, S_Q)$, and the corresponding received constellation point after the MMSE detection is $R = (R_I, R_Q)$, the SINR computed in equation (18) is the corresponding SINR for R_I and R_Q, denoted by $SINR_I$ and $SINR_Q$, respectively. Thanks to the above Q-component interleaving, the SINRs of I-component and Q-component in each symbol are different. The rotational demodulation produces the LLRs of encoded bits $\mathbf{K} = [k_1, k_2, ..., k_l]^T$ as follows.

$$LLR(k_i) = \ln\frac{P(k_i = 0)}{P(k_i = 1)} \qquad (19)$$

where k_i denotes the ith bit of one symbol, $l = \log_2 M$. Assuming equal *aprior* probabilities, P is calculated as follows.

$$P = \exp[-(R_I - S_I)^2 \bullet SINR_I] \bullet \exp[-(R_Q - S_Q)^2 \bullet SINR_Q] \qquad (20)$$

VII. PRECODING MATRIX AND USER SELECTION ALGORITH

According to the equation (18), for the kth user, the value of $SINR_k^m$ is not only dependent on its channel matrix $\mathbf{H}_k$, but also determined by its corresponding precoding vector $\mathbf{w}_k$ and the other user's precoding vector $\mathbf{w}_i$. A precoding matrix $\mathbf{w}^p$ will correspond to $K \times N_L$ SINRs. According to the known precoding codebook $\mathbf{W} = \{\mathbf{w}^1, \mathbf{w}^2, ..., \mathbf{w}^L\}$, then the total number of SINRs for each user is $K \times N_L \times L$. Each user will feedback these SINRs to the base station to perform scheduling.

The scheduling method of maximizing system capacity will be used in this paper. The transmitting side will select K users from the X users to be scheduled, and select the corresponding precoding matrix to maximize system throughput. For the multi-user MIMO system, system capacity can be expressed as

$$C_{MU}(SINR) = \sum_{k=1}^{K}\sum_{m=1}^{N_L}\log_2(1 + SINR_k^m) \qquad (21)$$

The method of choosing optimal precoding matrix that maximizes system capacity can be defined as

$$\mathbf{w} = \arg\max_{\mathbf{w}\in W} C_{MU}(SINR) \qquad (22)$$

where the constraint is that K users correspond to the different precoding vectors in the same precoding matrix, as follows

$$\mathbf{w}_k = [\mathbf{w}]_k \qquad (23)$$

The detailed steps are as follows.
1) Each user computes $K \times N_L \times L$ SINRs according to their channel matrix and the precoding codebook, and then sends these SINRs to the base station.
2) The base station collects all the feedback information. Select K users from the X users to be scheduled and corresponding precoding matrix according to the principle of maximizing system capacity. Each user in the selected users group corresponds to the precoding vectors $\mathbf{w}_k$ in the precoding matrix $\mathbf{w}$.
3) Apply precoding scheme corresponding to the selected group.

VIII. SIMULATION RESULTS

Assuming that each user has two data streams $N_L = 2$, computer simulations are carried out to compare the proposed scheme with the conventional MU-MIMO OFDM scheme in the current 3GPP LTE-Advanced standard. The frame structure, physical resource elements of OFDM system and Turbo code are given in Table II.

Figure 4 depicts the JCMD-vs-BICM frame error rate (FER) performance comparison for 3/4 and 5/6. In the figure, the proposed scheme JCMD MU-MIMO OFDM scheme can obtain obvious SNR gain over the conventional BICM MU-MIMO OFDM scheme in current 3GPP LTE standard.

For QPSK modulation, SNR gains are about 3.52dB and 4.55dB at FER=0.01 for 3/4 and 5/6 code rate, respectively. For 16QAM modulation, SNR gains are about 3.33dB and 3.83dB at FER=0.01 for 3/4 and 5/6 code rate, respectively. Thus, for high code rate, the SNR gain becomes more significant.

TABLE II.   SIMULATION PARAMETERS

Parameter	Value
Number of transmit antennas	4
Number of receive antennas	4
Type of transmit precoding	Unitary
Number of UEs	9
Transmission bandwidth	5MHz
Centre frequency	2GHz
DFT Size	512
Sample Period	$1/(15000 \times 512)$s
Channel Model	ITU-R IMT-Advanced(M.2135)
Mobile Speed	3km/h
Turbo Code Rates	3/4,5/6
Modulation Type	QPSK, 16QAM

Figure 4.   FER performance of JCMD vs BICM MU-MIMO OFDM

for 3/4 and 5/6 code rate, QPSK, 16QAM IMT-Advanced

IX.   CONCLUSION

In this paper, an improved joint coding and modulation diversity (JCMD) MU-MIMO scheme is proposed. Compared with the current solution BICM, the JCMD scheme only introduces limited complexity, which was brought by the several interleavers. A low-complexity time-frequency component interleaver is proposed to maximize the frequency diversity and the time diversity of OFDM system. The proposed scheme makes full use of Turbo codes coding gain, time and frequency diversity of OFDM system, and the spatial diversity of MIMO.

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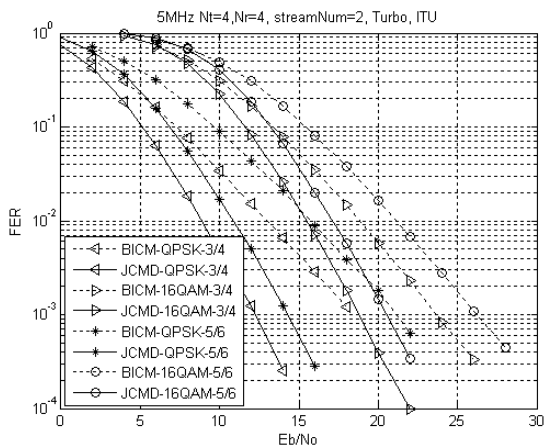[17] A. Goldsmith, "Wireless communications," Cambridge University Press, Aug. 2005, pp. 299-309.