A Novel MIMO-OFDM Scheme Based on Modulation Diversity for IEEE 802.11ac Standard

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***Abstract*—Wireless communications always strive for higher throughput and higher performance. The Multiple Inputs Multiple Outputs (MIMO)- Orthogonal Frequency Division Multiplexing (OFDM) scheme is the key Physical (PHY) layer feature of the next generation wireless local area networks (WLAN) IEEE 802.11ac standards. In this paper, we propose a novel MIMO-OFDM scheme based on modulation diversity for IEEE 802.11ac. This proposed scheme jointly optimizes the MIMO-OFDM and modulation diversity together, which makes full use of time diversity, frequency diversity and space diversity. It exhibits high spectral efficiency and low error rate in fading channels. The simulation results show that the proposed scheme outperforms the current MIMO-OFDM scheme based on Bit-Interleaved Coded Modulation (BICM) in the 802.11ac standard, which is up to 7dB SNR gain.**

Keywords-WLAN; 802.11ac; modulation diversity; MIMO; OFDM

I. INTRODUCTION

In recent years, the applications of wireless local area networks (WLAN) have grown very fast in hot spots all over the world. Through the combination of Multiple Inputs Multiple Outputs (MIMO) and Orthogonal Frequency Division Multiplexing (OFDM), IEEE 802.11n significantly increases the maximum data rate up to 600 Mbit/s [1]. Although existing 802.11n can support data rate up to 600 Mbit/s, it cannot yet meet the growing demands for higher throughput and higher performance. Since 2008, IEEE 802.11ac Task Group (TG) has developed the next generation wireless local area networks (WLAN) IEEE 802.11ac standard in order to achieve 1 Gbit/s data throughput below 6 GHz to meet the fast-growing market demands [2]. It extends the air interface technologies in IEEE 802.11n [3]. The combination of OFDM and MIMO is still a key feature for high-throughput transmission in the IEEE 802.11ac standard.

OFDM is an orthogonal multi-carrier frequency-division multiplexing (FDM) modulation scheme. OFDM allocates its transmitted symbols into narrow-band sub-carriers and maintain its orthogonality between the sub-carriers, so that OFDM can avoid inter symbol interference (ISI) due to a frequency selective fading channel. MIMO systems have attracted much more attention in wireless communications, because it can obtain very high spectral efficiency. Thanks to these properties, MIMO-OFDM has become the foundation of advanced wireless transmission technologies in current WLAN 802.11standards.

The uncoded modulation diversity schemes based on multidimensional rotated constellations were proposed in [4][5][6] for the conventional single-input single-output (SISO) scenario. The schemes are essentially uncoded and can achieve very high modulation diversity, which can approach to the additive white Gaussian noise (AWGN) error performance over fading channels. A rotated coding modulation OFDM system was put forward in [7], which extends the two-dimensional modulation diversity in coded OFDM systems. Through the combination of rotating the MPSK/MQAM constellation and interleaving the symbol-components, the performance of wireless communications systems can be significantly improved in flat fading channels without time-dispersion. Hence, a MIMO-OFDM system based on two-dimensional modulation diversity for IEEE 802.11ac is proposed, which jointly optimizes MIMO-OFDM and the rotated coding modulation. This scheme can take full advantage of the coding-gain of channel codes, the time and frequency diversity of OFDM system and the spatial diversity of MIMO all together. It is more suitable for space-time-frequency selective fading channels and exhibits the much better performance in wireless communications.

The rest of this paper is organized as follows. An improved MIMO-OFDM system based on modulation diversity for IEEE 802.11ac is proposed in Section II. Rotational modulation, the component interleaver and the detection algorithm in receiver are introduced in Sections III, IV and V, respectively. Simulation results are given in Section VI. Conclusions are reached in Section VII.

II. MIMO-OFDM SYSTEMS BASED ON MODULATION DIVERSITY FOR 802.11AC

An improved MIMO-OFDM scheme based on modulation diversity is proposed, which is shown in Figure 1. The blocks drawn in dotted line are the additional proposed processing based on the conventional MIMO-OFDM system introduced in [8]. N_T transmit and N_R receive antennas are assumed.

In the transmitter, the number of encoders is determined by rate-dependent parameters [8]. Information bits are firstly divided between the encoders by sending bits to different encoders in a round robin manner, and then coded by binary convolutional codes (BCC) or low density parity codes (LDPC).

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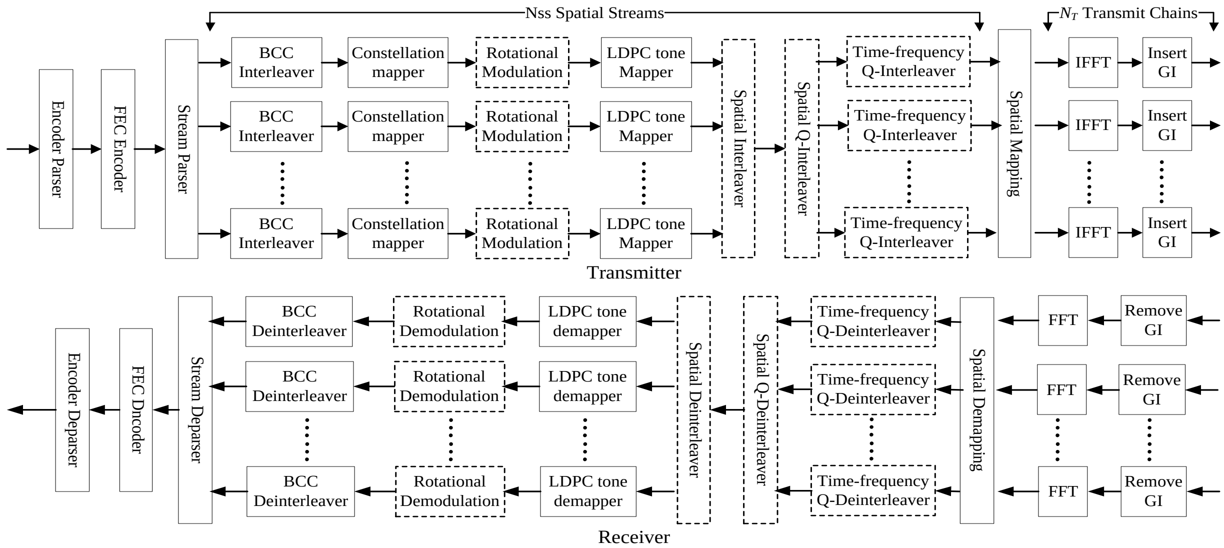

Fig. 1. Transmitter and receiver diagram of MIMO-OFDM system based on modulation diversity for 802.11ac

Afterwards, the coded bit stream is parsed into N_{SS} spatial streams.

If BCC encoding is used, BCC interleaving is performed within each spatial stream according to a rule in [8]. For each of the encoded, parsed, and interleaved spatial stream, convert the bit group into complex-valued symbols according to the modulation encoding tables. And then, the rotational modulation mapper converts the conventional modulated symbols into new complex-valued rotated symbols which contain in-phase (I) components and Quadrature (Q) components. LDPC tone mapping is skipped in this case.

If LDPC is used, after the stream parsing, BCC interleaver is not used, and the coded bits on N_{SS} spatial streams are converted into complex-valued symbols according to the modulation encoding tables. Based on the conventional modulated constellations, rotational modulation is carried out, and then the LDPC tone mapping shall be performed on all LDPC coded streams.

Afterwards, the spatial interleaving and spatial Q-interleaving are carried out over N_{SS} spatial streams, which will be described in section V. The interleaved symbols are then mapped onto different time-frequency resource elements (REs) on each stream. For the Q-components, an additional time-frequency Q-interleaver is performed on each stream. In this paper, we consider that STBC is not used, so the number of space-time streams is the same as the number of spatial streams. After the spatial mapping, N_{FFT} -point Inverse FFT (IFFT) and inserting guard interval (GI) are implemented for each group of subcarriers on each transmit antenna.

In the receiver, removing GI and N_{FFT} -point FFT are carried out firstly. Afterwards, the spatial demapping is performed. And then, the received symbols of each spatial stream are obtained after time-frequency Q-component deinterleaving, spatial Q-deinterleaving and spatial deinterleaving.

If LDPC encoding is used, LDPC tone demapping is carried out after the above operations. For each symbol, the maximum likelihood (ML) rotational demodulation is used to produce the log-likelihood-ratios (LLRs) of encoded bits.

If BCC encoding is used, LDPC tone demapping is skipped. After the ML rotational demodulation, BCC deinterleaving is performed.

After the above operations, stream deparser is carried out, and then the channel decoder (BCC or LDPC) utilizes the LLRs to decode the information bits.

III. ROTATIONAL MODULATION

Modulation diversity can be achieved via the rotational modulation, which provides an additional modulation diversity gain. Through the combination of rotating the conventional constellation and interleaving the components, the performance of wireless communications systems can be improved greatly in fading channels.

In the transmitter, coded bits $\mathbf{K} = (k_1, k_2, k_3 ... k_{N_{bit}})$ are modulated according to Gray-mapping constellation described in [8] to produce a sequence of complex-valued modulation symbols $\mathbf{D} = (d_1, d_2, d_3 ... d_{N_{Sym}})$.

Rotating the conventional constellation can be expressed as a complex-valued modulation symbol d_i multiplied by a rotational matrix $\boldsymbol{R}$. The relationship between conventional modulated complex-valued symbol $d_i = I_i + jQ_i$ and the rotational complex-valued symbol $d_i' = X_i + jY_i$ is shown as follows.

$$\begin{pmatrix} X_i \\ Y_i \end{pmatrix} = \mathbf{R} \begin{pmatrix} I_i \\ Q_i \end{pmatrix} = \begin{pmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{pmatrix} \begin{pmatrix} I_i \\ Q_i \end{pmatrix} \quad (1)$$

This processing can be illustrated in Figure 2. In the conventional square QPSK constellation, the I- component and Q-component of one complex-valued modulated symbol just carry one bit, respectively. In the rotational QPSK constellation,

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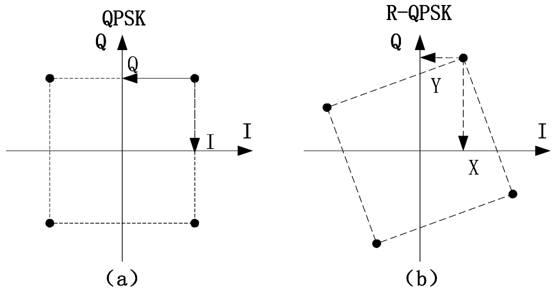

Fig. 2. conventional QPSK constellation (left) and Rotated QPSK constellation (right)

TABLE I. THE OPTIMAL ROTATION ANGLES FOR 802.11AC

MCSs	Modulation	Code rate	Angle
MCS 2	QPSK	3/4	arctan1/2
MCS 4	16QAM	3/4	arctan1/3
MCS 7	64QAM	5/6	arctan1/4
MCS 8	256QAM	3/4	arctan1/7

two bit are carried by both components at the same time, which means the information of two bits is spread over 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 [4].

The key of rotational modulation technique is finding an optimum rotational angle to minimize bit error rate (BER). J.Boutros studied the optimum criteria of the uncoded rotation modulation over the independent Rayleigh channel, and proposed that the diversity order L and product distance should be maximized [4]. Simulation results indicate that the optimum rotation angle depends on the modulation order and code rate for coded modulation. The correlation between the code rate and optimum rotation angle is weak. Especially for the high code rates from 3/4 to 8/9, the optimal rotation angles keep almost the same value. However, it is strongly correlated 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 IEEE 802.11ac, which are shown in Table I.

IV. Q-COMPONENT INTERLEAVER

A. Time-frequency Q-component interleaver

The 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 $Q(f,t)$, which at the f^{th} subcarrier in the t^{th} OFDM symbol. After time-frequency Q-component interleaving, the output is $Q(f',t')$. The Q-component interleaving is defined as follows.

$$\begin{aligned} f' &= (f + L/2)\%L \\ t' &= (t + N_{sym}/2)\%N_{sym} \end{aligned} \tag{2}$$

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 of 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.

B. Spatial interleaver and spatial Q-component interleaver

The spatial interleaving is the conventional spiral layer interleaving among all layers. Let x_t^i and $\overline{\mathrm{x}}_t^i$ denote the input-symbol and the output-symbol of the spatial interleaver on the i^{th} spatial stream at the t instant. The interleaving is defined as follows.

$$\overline{\mathrm{x}}_t^k = \mathrm{x}_t^i, k = (i+t) \bmod N_{SS} \tag{3}$$

where $k, i \in [0, N_{SS}-1]$ .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 $\overline{\mathrm{Q}}_t^i$ denote the input Q-component and the output Q-component of the Q-component interleaver on the i^{th} spatial stream at the t instant. The Q-component interleaving is defined as follows.

$$\overline{\mathrm{Q}}_t^k = \mathrm{Q}_t^i, k = N_{SS} - i - 1 \tag{4}$$

where $k, i \in [0, N_{SS}-1]$ .The achievable rate of the proposed scheme can increase via the spatial Q-component interleaver. For the sake of simplicity, a system with two transmit and two receive antennas is firstly considered, and the symbols are mapped to two spatial streams. That is to say, $N_T = N_R = N_{SS} = 2$. We assume the total transmission power is P and the bandwidth is W. Channel matrix H_{2x2} is known at the transmitter, and the two singular values are $\sqrt{\lambda_1}$ and $\sqrt{\lambda_2}$ ($\sqrt{\lambda_1} \geq \sqrt{\lambda_2}$). The total achievable rate is the sum of partial achievable rate on each singular-value-decomposition (SVD) spatial layer of the MIMO channel [11].

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

$$C_1 = W \cdot \log_2\left(1+\frac{\lambda_1 P}{2\sigma^2}\right)\left(1+\frac{\lambda_2 P}{2\sigma^2}\right) = W \cdot \log_2\left(1+\frac{\lambda_1+\lambda_2}{2\sigma^2}P + \frac{\lambda_1\lambda_2}{4\sigma^4}P^2\right) \tag{5}$$

For the proposed system, thanks to the spatial Q-component interleaver, the fading coefficients of I- and Q-component in each symbol are $\sqrt{\lambda_1}$ and $\sqrt{\lambda_2}$, respectively. Thus, the

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achievable rate of the proposed system can be calculate as follows.

$$C_2 = W\cdot\log_2\left(1+\frac{(\lambda_1+\lambda_2)}{4\sigma^2}P\right)^2 = W\cdot\log_2\left(1+\frac{\lambda_1+\lambda_2}{2\sigma^2}P+\frac{(\lambda_1+\lambda_2)^2}{16\sigma^4}P^2\right) \quad (6)$$

It is easy to obtain $C_1 \le C_2$, because $(\lambda_1+\lambda_2)^2 \ge 4\lambda_1\lambda_2$.

V. DETECTION AND DEMODULATION

Assuming perfect channel state information (CSI), the MIMO channel with $N_T \times N_R$ matrix $\mathbf{H}$ is well known to have the singular value decomposition (SVD) as

$$\mathbf{H} = \mathbf{UDV}^H \quad (7)$$

where the $N_R \times N_R$ matrix $\mathbf{U}$ and the $N_T \times N_T$ matrix $\mathbf{V}$ are unitary matrices and $\mathbf{D}$ is an $N_R \times N_T$ diagonal matrix of singular values $\{\sigma_i\}$ of $\mathbf{H}$. For the i^{th} eigenvalue λ_i of $\mathbf{HH}^H$, $\sigma_i = \sqrt{\lambda_i}$. By using SVD, MIMO channel is divided into parallel independent spatial subchannels.

In the transmitter, the spatial mapping takes a vector $\mathbf{x}^p = \left[x^p_{(1)}, x^p_{(2)} \ldots x^p_{(N_{SS})}\right]^T$ on N_{SS} space-time streams as input and generates a vectors $\mathbf{y}^p = \left[y^p_{(1)}, y^p_{(2)} \ldots y^p_{(N_T)}\right]^T$ to be mapped onto N_T transmit chains at p symbol-instant, where $x^p_{(i)}$ represents the symbol p on space-time stream i, $y^p_{(i)}$ represents the symbol p on transmit chain i, where i=1, 2 for $N_T = N_{SS} = 2$. The input to the transmit chain $\mathbf{y}^p$ is the output of a linear transformation on input vector $\mathbf{x}^p$ as $\mathbf{y}^p = \mathbf{V}\cdot\mathbf{x}^p$.

In the receiver, let $\mathbf{r}^p = \left[r^p_{(1)}, r^p_{(2)} \ldots r^p_{(N_R)}\right]^T$ denote the received signal vector on N_R antenna ports, which can be expressed for $N_T = N_R = N_{SS} = 2$ as follows.

$$\mathbf{r}^p = \mathbf{H}\cdot\mathbf{y}^p + \mathbf{n} \quad (8)$$

where, $\mathbf{n}^p = \left[n^p_{(1)}, n^p_{(2)} \ldots n^p_{(N_R)}\right]^T$ denote a white noise vector on N_R antenna ports. The white noise variance is σ^2.

Receiver shaping performs a similar operation at the receiver by multiplying the channel output $\mathbf{r}^p$ with $\mathbf{U}^H$, as shown in (9).

$$\begin{aligned}\tilde{\mathbf{y}}^p &= \mathbf{U}^H\cdot\mathbf{r}^p \\ &= \mathbf{U}^H\cdot\mathbf{H}\cdot\mathbf{y}^p + \mathbf{U}^H\mathbf{n} \\ &= \mathbf{U}^H\mathbf{UDV}^H\mathbf{V}\cdot\mathbf{x}^p + \mathbf{U}^H\mathbf{n} \\ &= \mathbf{D}\cdot\mathbf{x}^p + \mathbf{n}'\end{aligned} \quad (9)$$

where $\mathbf{n}' = \mathbf{U}^H\mathbf{n}$. Note that multiplication by a unitary matrix does not change the distribution of the noise, so $\mathbf{n}'$ and $\mathbf{n}$ are identical Gaussian distribution. Thus, the transmit precoding and receiver shaping transform the MIMO channel into parallel independent channels. The received symbol after detection on each stream is

$$\tilde{\mathbf{y}}^p_{(i)} = \sigma_i\cdot\mathbf{x}^p_{(i)} + \mathbf{n}'^p_{(i)} \quad (10)$$

After the spatial demapping and space-time-frequency deinterleaving, the rotational demodulation is carried out on each antenna. We assume the i^{th} rotated constellation point in the transmitter is $S^i = (S_I^i, S_Q^i)$, and the corresponding received constellation point after the above operations is $R^i = (R_I^i, R_Q^i)$. Thanks to the above Q-component interleaving, the fading coefficient of I-component H_I and that of Q-component H_Q in each symbol are different, as shown in Figure 3. For the i^{th} symbol, the relationship between S^i and R^i is shown as follow,

$$(R_I^i, R_Q^i) = (H_I\cdot S_I^i + n_I, H_Q\cdot S_Q^i + n_Q) \quad (11)$$

Where n_I and n_Q are the additive Gaussian noise of I-component and that of Q-component, respectively. Rotational demodulation produces the LLRs of encoded bits $\mathbf{K}^i = (k_1^i, k_2^i, ..., k_m^i)$ as follows.

$$LLR(k_x^i) = \ln\frac{P(k_x^i = 0)}{P(k_x^i = 1)} \quad (12)$$

where k_x^i denotes the x^{th} bit of symbol i, $m = \log_2 M$. Assuming equal *aprio*r probabilities, P is calculated as follows.

$$P = \frac{1}{\sqrt{2\pi}\sigma}\exp\left(-\frac{(R_I^i - H_I^iS_I^i)^2}{2\sigma^2}\right)\cdot\exp\left(-\frac{(R_Q^i - H_Q^iS_Q^i)^2}{2\sigma^2}\right) \quad (13)$$

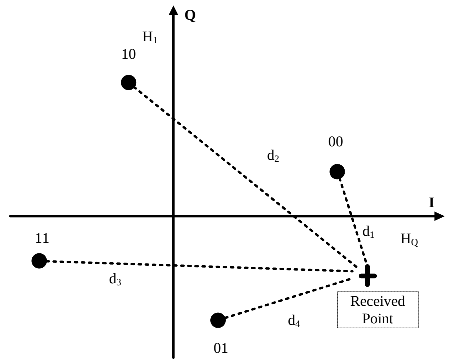

Fig. 3. Received constellation points for R-QPSK

VI. SIMULATION RESULTS

To validate the advantages of the proposed scheme, an IEEE 802.11ac link level simulation platform is programmed by using Microsoft Visual Studio 2005 (VS2005). Computer simulations are carried out to compare the proposed scheme with the conventional MIMO-OFMD scheme in current 802.11ac standard draft. The frame structure, physical resource elements of OFDM system, channel model and coding are all based on 802.11ac standard draft [8][9], which are given in Table II.

Figure 4 and Figure 5 depict frame error rate (FER) performance comparison in NLOS channel for BCC coding and LDPC coding, respectively. The solid line shows the FER performance of our proposed scheme, while the dotted line is that of the conventional MIMO-OFMD scheme in current 802.11ac standard draft. From these figures, the proposed scheme can obtain obvious SNR gain as compared with the current 802.11ac MIMO-OFDM scheme. In NLOS channel, for BCC coding, SNR gains are about 7 dB and 4.8 dB at FER=0.01 for QPSK and 16QAM modulation, respectively. For LDPC coding, SNR gains are about 6.8 dB and 4.5 dB at FER=0.01 for QPSK and 16QAM modulation, respectively. Therefore, the proposed scheme obviously outperforms the current MIMO-OFDM system in 802.11ac. For higher code rate or lower modulation order, the SNR gain becomes more significant.

VII. CONCLUSION

An improved MIMO-OFDM scheme based on modulation diversity is proposed. A simple and effective space-time-frequency component interleaver is designed to make the best use of space-time-frequency diversities. An efficient demodulation method is also put forward for it. This scheme can take full advantage of the coding-gain of channel codes (BCC and LDPC), the time and frequency diversity of OFDM system and the spatial diversity of MIMO all together. Simulation results turn out that it can obtain obvious SNR gain as compared with the current MIMO-OFDM scheme in IEEE 802.11ac standard. In the future, more theoretical analysis and simulations are still required to research for the impact of channel estimation to our proposed scheme. In a word, our proposed scheme is simple, efficient and robust for the next generation WLAN standard.

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TABLE II SIMULATIONS PARAMETERS

Parameters	Value
N_T	2
N_R	2
Carrier frequency	5 GHz
Bandwidth	20.0 MHz
Subcarrier Number	52
Useful Subcarrier Number	56
FFT Size	64
Coding	BCC, LDPC
MCSs	MCS2, MCS4, MCS7, MCS8
sub-carrier bandwidth	312.5 kHz
Channel Models	TGac Channel Model case E
Mobile Speed	3km/h
Coherent Time	1/0.4326=2.427s
Coherent Bandwidth	B_{CLOS}=1.7 MHz, B_{CNLOS}=1.6 MHz
Channel estimation	perfect CSI

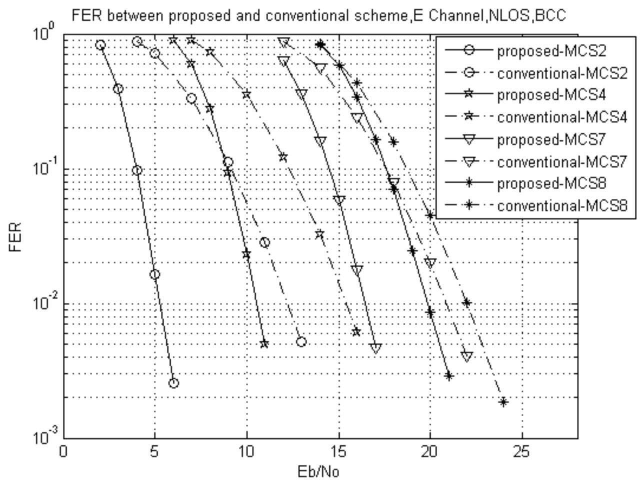

Fig. 4. FER performance of proposed scheme vs conventional MIMO-OFDM for different MCSs, BCC coding, AC case E Channel, NLOS

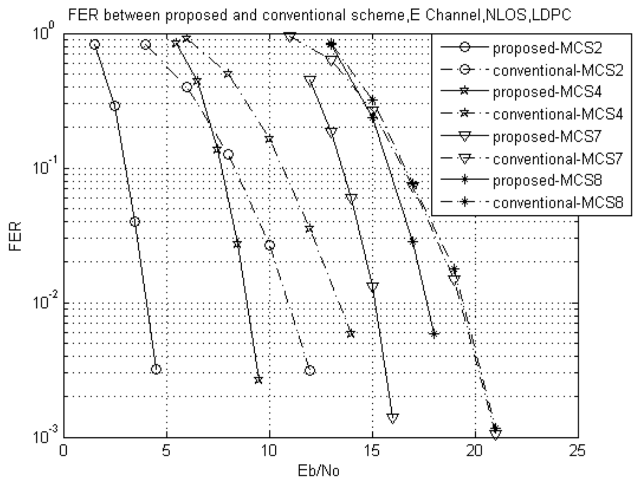

Fig. 5. FER performance of proposed scheme vs conventional MIMO-OFDM for different MCSs, LDPC coding, AC case E Channel, NLOS

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