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A Low Complexity Detection Scheme of Space Frequency Codes for MIMO OFDM Systems with Minimum Euclidean Distance Metric Calculation

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Abstract

In this paper a design criteria of full diversity space frequency codes (SFC) is proposed for MIMO-OFDM systems with a low complexity detection scheme. It can achieve full diversity. This system utilizes space distributive properties of SFCs to make correct decisions. In the proposed scheme detection of SFCs is done using a novel minimum Euclidean distance metric calculation method followed by decision making. It can achieve the performance of partial interference cancellation methods. Partial interference cancellation (PIC) group decoding is an attractive decoding alternative for MIMO wireless communications. It can well deal with the tradeoff among rate, diversity and decoding complexity of SFC codes. Simulation results of the newly proposed codes demonstrate the theory.

***Index Terms*:** MIMO-OFDM, SFC, ML decoding, PIC group decoding.

INTRODUCTION

The idea of using multiple transmit and receive antennas in wireless communication systems to accommodate high data rates has attracted considerable attention recently.The multiple input multiple output(MIMO) systems offer considerable performance improvement over single antenna systems.by combining the orthogonal frequency division multiplexing (OFDM) modulation with MIMO systems,space frequency (SF) codes have been proposed to exploit the spatial and frequency diversity present in frequency selective MIMO channels.The strategy of SF coding is to distribute the channel symbols over different transmit antennas and OFDM tones with one OFDM block. Partial interference cancellation group decoding is an attractive decoding alternative for multiple input multiple output(MIMO) wireless communications.It can well deal with the trade off among rate,diversity and decoding complexity of space time block codes.

In this paper, an SF coded MIMO-OFDM system is first formulated to a general MIMO system. Then, a design criterion of SFC is proposed to obtain full diversity with the PIC group decoding in MIMO-OFDM systems similar to that in for MIMO systems. With the proposed new criterion, a systematic design of SFC is constructed for any number of transmit antennas *Mt* and any number of delay paths *L* for a particular grouping scheme. It is proved that the proposed SFC can obtain full diversity including multipath diversity gain *L* when the PIC group decoding is applied. Furthermore, the PIC group decoding leads to a joint decoding of *Mt*L* complex symbols that has much lower decoding complexity than the ML decoding. This is the first attempt of the design criterion and full-diversity SFC design in MIMO OFDM systems with the PIC group decoding.

Therefore, compared with the existing SFC designs aforementioned, challenges and contributions in this work are highlighted as follows: 1) SFC designs based on the ML or ZF receiver suffer from either the high decoding complexity or the low code rate. To address the tradeoff between code rate and decoding complexity, PIC group decoding is one of the promising solutions. How to explore both spatial and multipath diversity with the PIC group decoding is a challenging problem in space-frequency coded MIMO-OFDM systems. 2) In this paper, a design criterion of SFC is proposed for MIMO-OFDM systems with the PIC group decoding. Based on the criterion, a systematic design of full-diversity SFC is proposed that can achieve full diversity including spatial and multipath diversity under the PIC group decoding. The code rate of our proposed SFC design approaches *J* (number of layers in the codeword) with a large number of transmit antennas, while the decoding complexity is significantly reduced compared to the ML decoding.

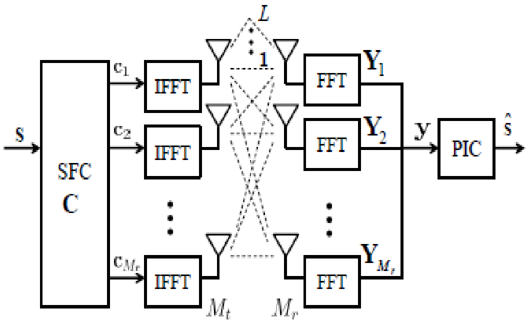

Fig.1 MIMO–OFDM System with the PIC Group Decoding.

MIMO - OFDM Overview

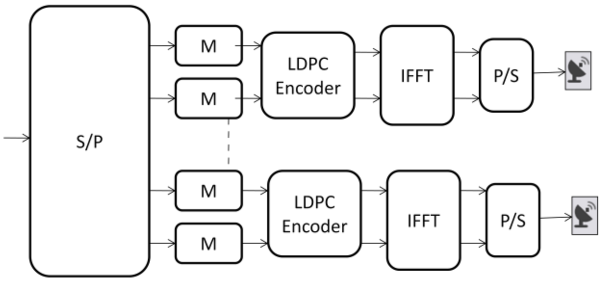

Fig. 2 Transmitter

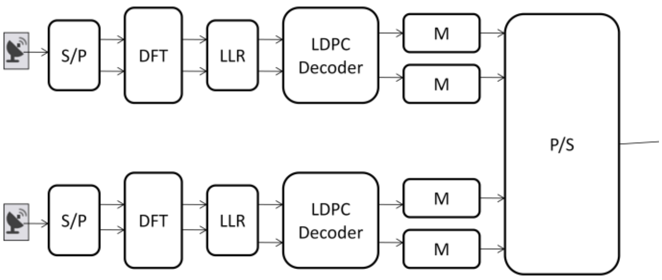

Fig. 3 Receiver

SF CODED MIMO-OFDM SYSTEM

Fig. 1 shows that a MIMO-OFDM system is equipped with *Mt* transmit antennas, M_r receive antennas, and *N*-tone OFDM. The channel between each pair of transmit and receive antennas is assumed to have *L* independent channel taps and the same power delay profile. We assume that $N \geq M_t L$. The MIMO channel keeps constant over each OFDM block period and varies independently from one block to another. The channel impulse response from transmit antenna *i* to receive antenna *j* is modeled as

$$h_{i,j}(\tau) = \sum_{l=1}^{L} h_{i,j}(\tau)\ \delta(\tau - \tau l),$$

where τl is the delay of the *l*th path, and the *l*th path gain $h_{i,j}(l)$ is a complex Gaussian random variable distributed with zero mean and variance σ_l^2. The powers of the *L* paths are normalized such that $\sum_{l=1}^{L} \sigma_l^2 = 1$. Define the channel vector between transmit antenna *i* and receive antenna *j* as

$h_{i,j} = [h_{i,j}(1)\ h_{i,j}(2) \ldots\ldots h_{i,j}(l)]T$

and the channel vector for the MIMO system is denoted by

h = $[h_{1,1}^T \ldots\ldots h_{M_t,1}^T \ldots\ldots h_{1,M_r}^T \ldots\ldots h_{M_t,M_r}^T]^T$

The frequency response of the channel vector in (2) can be,represented as

$$h_{i,j} = [H_{i,j}(1)\ H_{i,j}(2)\ \ldots\ldots\ H_{i,j}(N)]^T$$
$$= W h_{i,j}$$

where $H_{i,j}(k)$ for $k = 1, 2, \cdots, N$ denotes the *k*th subcarrier in $H_{i,j}$, and $\mathbf{W} \in C^{NxL}$ is given by

$$W = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ \xi^{T_1} & \xi^{T_2} & \cdots & \xi^{T_L} \\ \vdots & \vdots & \ddots & \vdots \\ \xi^{(N-1)T_1} & \xi^{(N-1)T_2} & \cdots & \xi^{(N-1)T_L} \end{bmatrix}$$

Here, we have $\xi = e^{-j2\pi/Ts}$ and *Ts* is the OFDM symbol period. In addition, we suppose that the channel state information **H***i,j* is available at the receiver only.

We consider an information symbol vector

s = $[s1, s2, \cdots, s_{Ns}]^T$ selected from a finite constellation *A*, such as, a normalized quadrature amplitude modulation (QAM). The symbol vector $\mathbf{s} \in A^{N_s}$ is firstly parsed into blocks and mapped onto an SF codeword $\mathbf{C} \in C^{N \times Mt}$

$\mathbf{C} = [\boldsymbol{c}_1\ \boldsymbol{c}_2 \cdots \boldsymbol{c}_{Mt}]$,

where the *i*th column $\boldsymbol{c}_i$ for $i = 1, 2, \cdots, Mt$ is sent from the *i*th transmit antenna. After IFFT modulation and the insertion of cyclic prefix, OFDM symbols are transmitted simultaneously from all transmit antennas.

The received signals are assumed to have perfect timing and synchronization. After cyclic prefix removal and FFT demodulation, the received signal at the *j*th receive antenna is given by

$$Y_j = \sqrt{\frac{\rho}{\mu}}\ \sum_{i=1}^{Mt} diag(c_i)\, H_{i,j} + Z_j,$$

where all elements in the noise vector $\mathbf{Z}j \in C^{N \times 1}$are assumed to be i.i.d. complex Gaussian with zero mean and unit variance, ρ denotes the average signal-to-noise ratio (SNR) per receive antenna, and the transmitted power is normalized by a factor μ such that the average energy of the coded symbols transmitting from all antennas during one subcarrier is one. Let $\Phi_l = diag([1\ \xi^{T_1} \ldots .\ \xi^{(N-1)T_l}]^T)$, and $h_j^D(l) = [h_{1,j}(l) h_{2,j}(l) \ldots\ h_{Mt,j}(l)]^T$ for l= 1,2,….L .Then, the received signals from all the receive antennas is written as [6].

$$Y = \sqrt{\frac{\rho}{\mu}}\ \wedge h^D + Z^D,$$

Where

$$\wedge = \boldsymbol{I}_{\boldsymbol{Mr}} \otimes [\boldsymbol{\Phi}1\mathbf{C}\ \ \boldsymbol{\Phi}2\mathbf{C} \cdots\ \boldsymbol{\Phi}L\mathbf{C}],$$
$$h^D = [(h_1^D)^T\ (h_1^D)^T \ldots (h_{Mr}^D)^T]^T,$$
$$h_j^{\ D} = [(h_j^D(1))^T\ (h_j^D(2))^T \ldots (h_j^D(L))^T]^T,$$

For simplicity, the entries in the first column of **W** in (5) will be replaced by all ones in the following since we always have $\tau 1 = 0$ (i.e., $\boldsymbol{\Phi}1 = \mathbf{I}N$). It is seen that h^D and Z^D are obtained from **h** in (3) and **Z***j* in (7) after some row permutations, respectively.

In order to decode the independent symbols in **s**, the model in (8) can be rewritten as an equivalent signal model

$$y = \sqrt{\frac{\rho}{\mu}}\ G(h)s + z,$$

where $\mathbf{G(h)} \in C^{NMr \times N_s}$ is an equivalent channel matrix with **h** in (3). For convenience, we use **G** instead of **G(h)** in what follows. Here, we have to mention that the transformation from (8) to (12) is held if the SFC **C** in (6) is linear.

In the following, the basic idea of the PIC group decoding will be described. The PIC group decoding in MIMO systems is firstly specified in [10]. Define an index set $I = \{1, 2, \cdots, Ns\}$. I is then partitioned into K groups: $I1, I2, \cdots, IK$ with $I_k = \{I_{k,1}, I_{k,2}, , \cdots, I_{k,l_k}\}$ for $k = 1, 2, \cdots, K$, where I_k is the cardinality of the group I_k (i.e., $I_k = |I_k|$). $\{I_1, I_2, \cdots, I_k\}$ is defined as a grouping scheme. For such a grouping scheme, we have $I = \bigcup_{k=1}^{K} I_k$ and $\sum_{k=1}^{K} I_k = N_s$

Let the column vectors of the equivalent channel matrix **G** be **g**1, **g**2, $\cdots$, $\boldsymbol{g_{N_s}}$. Define

$s_{I_k} = [s_{I_{k,1}}\ s_{I_{k,2}} \dots s_{I_{k,l_k}}]^T$

$G_{I_k} = \left[g_{I_{k,1}}\ g_{I_{k,2}} \dots g_{I_{k,l_k}}\right]$. With these notations, (12) is further rewritten as

$$y = \sqrt{\frac{\rho}{\mu}}\ G_{I_k} s_{I_k} + z$$

Suppose that we want to decode the symbols in group s_{I_k} . Linear interference cancellation is firstly implemented in the PIC group decoding by using a cancellation matrix $\boldsymbol{Q_{I_k}} \in C^{NMr \times NMr}$, which can completely eliminate the interferences from other groups [10], i.e., $\boldsymbol{Q_{I_k}}\ \boldsymbol{G_{I_q}} = \mathbf{0}$, $\forall q = k$ and $q = 1, 2, \cdots, K$. The interference cancellation matrix $\boldsymbol{Q_{I_k}}$ is chosen as follows [9],

$$\boldsymbol{Q_{I_k}} = \boldsymbol{I} - \boldsymbol{G^c_{I_k}}((\boldsymbol{G^c_{I_k}})^{\boldsymbol{H}}\ \boldsymbol{G^c_{I_k}})^{-1}\ (\boldsymbol{G^c_{I_k}})^{\boldsymbol{H}}\ , \qquad (14)$$

where

$\boldsymbol{G^c_{I_k}} = [\boldsymbol{G_1} \dots \boldsymbol{G_{I_{k-1}}}\ \boldsymbol{G_{I_{k+1}}} \dots \boldsymbol{G_{I_k}}]$. In fact, $\boldsymbol{Q_{I_k}}$ is the projection matrix that projects a vector into the orthogonal complementary subspace of $\boldsymbol{G^c_{I_k}}$. Then, we have

$$y_{I_k} = Q_{I_k}\, y$$

$$= \sqrt{\frac{\rho}{\mu}}\ \boldsymbol{Q_{I_k}} G_{I_k} s_{I_k} + \boldsymbol{Q_{I_k}} z \qquad (15)$$

The $\boldsymbol{Q_{I_k}}$ in (14) is one possible solution to realize the interference cancellation and any matrix $\boldsymbol{Q_{I_k}}$ with the property that $\boldsymbol{Q_{I_k}} G_{I_k} = \mathbf{0}$, $\forall q \neq k$ can be used.

Afterwards, the symbols in the group $\boldsymbol{s_p}$ are decoded with the ML group decoding algorithm based on the signal model (15) as follows,

$$\hat{s}_{I_k} = \arg\min \left\| y_{I_k} - \sqrt{\frac{\rho}{\mu}}\ \boldsymbol{Q_{I_k}} G_{I_k} s_{I_k} \right\|^2, \qquad (16)$$

for $k = 1, 2, \cdots, K$. The ML metric can be used in (16) since the new noise term $\boldsymbol{Q_{I_k}}\mathbf{z}$ is proved to be a degenerated Gaussian white noise .

In [10], successive interference cancellation (SIC) aided PIC group decoding (i.e., PIC-SIC group decoding) was proposed, which can obtain a better performance by following a proper decoding order than the PIC group decoding. Suppose that the grouping scheme of the PIC-SIC decoder is $I_1, I_2, \cdots, I_k$. The algorithm is briefly described as follows.

Initially, we let $k = 1$ and **y**1 = **y** where **y** is given in (13).

1) The symbol vector **s***Ik* is firstly decoded to be $\hat{s}_{I_k}$ by using PIC grouping decoding (16);

2) $\hat{s}_{I_k}$ is removed from **y***k* : $\boldsymbol{y_{k+1}} = \boldsymbol{y_k} - \sqrt{\frac{\rho}{\mu}} G_{I_k}\ \hat{s}_{I_k}$

Then, we let $k = k + 1$.

3) If $k \leq K$, then go to 1). Else, stop the algorithm.

In the transmitter section the signal is parallelly converted ,mapped and fed into the LDPC encoder.Then signals are embedded with cyclic prefixing. Then this is fed into IFFT block.The output is converted back into the serial signal and fed into the antenna for transmission.Figure 2 depicts this.

Figure 3 shows the receiver section of MIMO OFDM systems.The signals are picked up by the antenna.These signals are parallelly converted.Then cyclic prefix is removed.Then FFT is taken.These signals are decoded using LDPC codes and is fed into LLR block.The signals are serially converted and the output is taken.

SFC DESIGN WITH THE PIC GROUP DECODING

In this section, we propose a systematic design of fulldiversity SFC for MIMO-OFDM system with the PIC group decoding. Then, some code design examples are presented.

SFC Design

Consider the MIMO-OFDM system with $N \geq MtL$. A systematic SFC design for any Mt and L is constructed as

$$C_{M_t,J,L,u} = [B_1^T\ B_2^T\ \dots\ B_p^T\ 0_{M_t X(N-TP)}]^T$$

where $P = [N/T]$, T is the length of block **B***p*, and J represents the number of diagonal layers embedded in **B***p*. Additionally, **B***p* for $p = 1, 2, \cdots, P$ is given in (21), which is shown on top of the next page, where $T = L(Mt+J-1)+(J-1)u$ and $0 \leq u \leq L$. From the design in (21), one may see that each **B***p* has layered structure with multiple diagonal layers of **X***j*,1,**X***j*,2, . . . ,**X***j*,*Mt* for $j = pJ - J + 1, \cdots, pJ$. The parameter u in this design is defined as the number of zeros padded to suppress the interference between any two diagonal layers in (21). Note that u does not have to be a nonzero number, in other words, zero padding between diagonal layers is not necessary in the above design [21].

For each diagonal layer, we have

$$\boldsymbol{X_j} = [\boldsymbol{X^T_{j,1}}\ \boldsymbol{X^T_{j,2}}\ \dots\ \boldsymbol{X^T_{j,Mt}}]^{\boldsymbol{T}}$$

$$= \boldsymbol{\Theta}\mathbf{s}j, \qquad j = pJ - J + 1, \cdots, pJ. \ (22)$$

where **s***j* is performed by a linear transformation matrix **Θ** and

$$\mathbf{X}j,i = [X_{j,(i-1)L+1}\ X_{j,(i-1)L+2} \dots X_{j,iL}]^T \qquad , (23)$$

for $i = 1, 2, \cdots, Mt$.

In this paper, the optimal cyclotomic lattices design proposed in [20] is used. For $M = MtL$, from [20, Table I] we can get a set of integers (m, n) and let $U = mn$. Then, the optimal lattice **Θ** of size $M \times M$ is given by [20, Eq. (16)]

$$\mathbf{\Theta} = \sqrt{\square/\square}\,[\;\begin{matrix} \square_{\square} & \square^{\square}_{\square} & \cdots & \square^{\square}_{\square} \\ \square^{\square+\square\square\square}_{\square} & \square^{\square(\square+\square\square\square)}_{\square} & \cdots & \square^{\square(\square+\square_{\square}\square)}_{\square} \\ \vdots & \vdots & \ddots & \vdots \\ \square^{\square+\square_{\square}\square}_{\square} & \square^{\square(\square+\square_{\square}\square)}_{\square} & \cdots & \square^{\square(\square+\square_{\square}\square)}_{\square} \end{matrix}\;]$$

Where $\square_{\square}$ = exp(**j**2π/U) and $n2$, $n3$, · · · , nM are distinct integers such that 1 + $\square_{\square}m$ and U are co-prime for any $2 \le i \le M$. $\theta i,j = \square^{\square(\square+\square_{\square}\square)}_{\square}$ U denotes the (i, j)th entry in the matrix $\mathbf{\Theta}$.

In [20], $\Gamma\zeta m$ denotes a 2-dimensional real lattice with the generator matrix

$$\begin{bmatrix} 1 & \cos 2\square/\square \\ 0 & \sin 2\square/\square \end{bmatrix}$$

Obviously, $\Gamma\zeta m$ is a subset of Z[ζm] that is the number ring generated by the integer ring Z and ζm, i.e., $\Gamma\zeta m \subset$ Z[ζm]. Note that $\Gamma\zeta m$ is located on square lattice when m = 4, i.e., a QAM constellation. Hence, the signal constellation A can be $\Gamma\zeta m$ or a subset of $\Gamma\zeta m$.

According to the codeword matrix, the total number of information symbols in (20) transmitted during N subcarriers is $Ns = JMtLP$. Thus, the code rate of the proposed SFC is

$$R = \square_{\square}/\square \; = JMtLP/N$$

When N is equal to TP, the code rate is $R = JMtLP$

$TP = JMtL/\ L(Mt+J-1)+(J-1)u$. For a large value of Mt, the rate approaches J. On the other hand, when zeros are padded between diagonal layers in (21), i.e., $u > 0$, the code rate is reduced.

SIMULATION RESULTS

In the simulations, we consider a MIMO-OFDM system equipped with 2 transmit antennas, 2 receive antennas and N = 64 OFDM tones. The MIMO frequency-selective blockfading channel is simulated as an L ray with equal power delay profile. The bandwidth is supposed to be 20 MHz and the length of the cyclic prefix is 16.

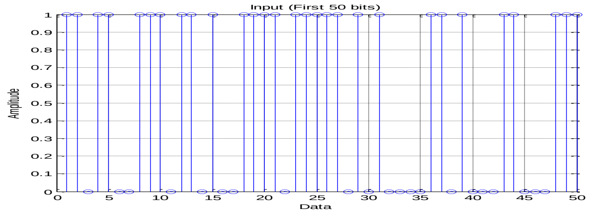

Fig. 4 Input- First 50 bits

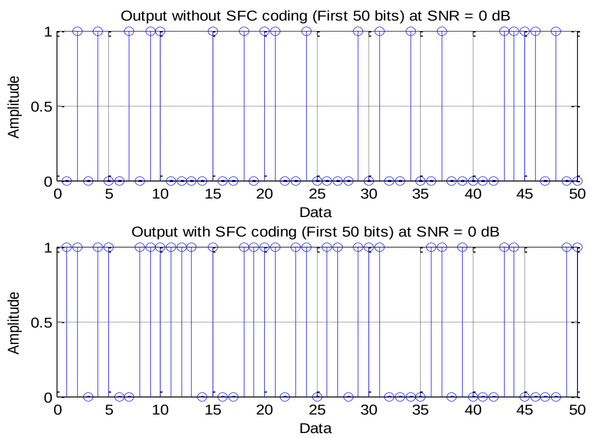

Fig.5 Output without and with SFC coding at SNR=0dB

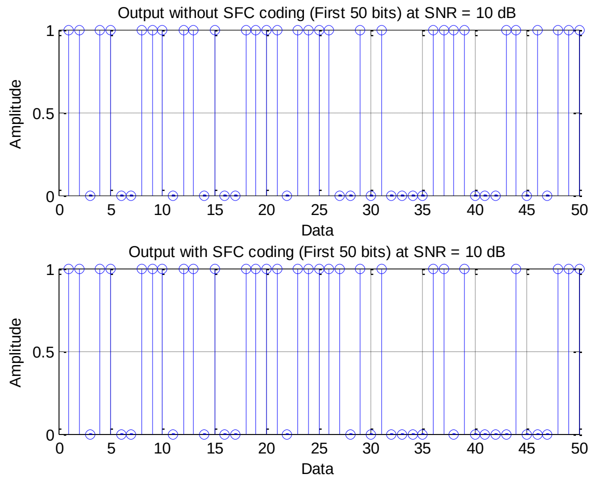

Fig.6 Output without and with SFC coding at SNR=10dB

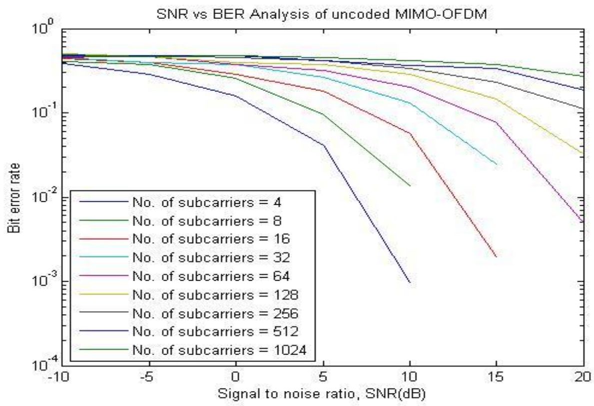

Fig.7 SNR Vs BER Analysis of Uncoded MIMO-OFDM

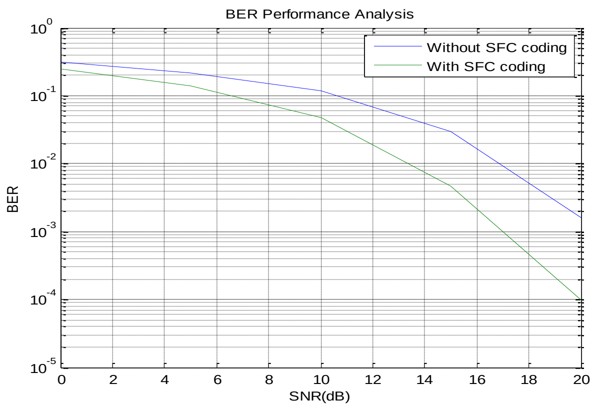

Fig. 8 BER Performance Analysis

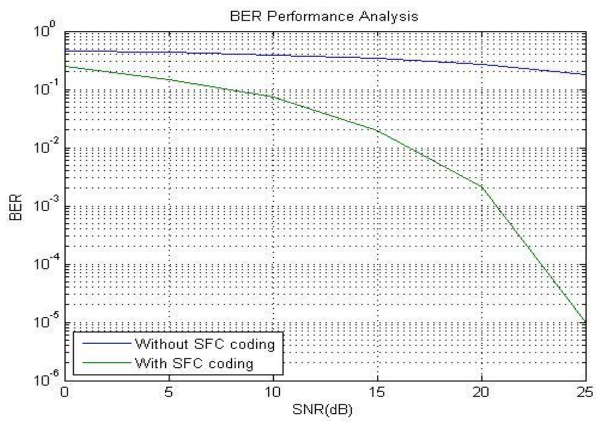

Fig. 9 BER Performance Analysis

Conclusion

In this paper, a design criterion was proposed for an SFC to achieve full diversity in MIMO-OFDM systems with the PIC group decoding. Based on the criterion, a design of SFC was proposed that can obtain the full diversity including spatial diversity and multipath diversity under the PIC group decoding. The complexity of the PIC group decoding is equivalent to a joint decoding of *MtL* complex symbols for *Mt* transmit antennas and *L* delay paths. Simulation results were presented to demonstrate that the proposed SFC can obtain full diversity and outperform the existing SFC in MIMO-OFDM systems with the PIC group decoding when the bandwidth efficiency is not too small.

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