Bit Interleaved Space-Frequency Coded Modulation for OFDM Systems

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***Abstract*— We present an orthogonal frequency-division multiplexing (OFDM) system with bit-interleaved space-frequency coded modulation for frequency selective fading channels employing multiple transmit and receive antennas. OFDM is used to transform a frequency selective fading channel into multiple flat fading channels, and space-frequency trellis coding combined with bit-interleaving is used to exploit space and frequency diversity. The performance of such an approach is evaluated by analytical bounds and simulation. With the use of space-frequency coding and BICM, we can provide increased rates and also improved performance by means of increased diversity.**

I. Introduction

Achieving reliable communications is challenging in wireless channels due to signal fading caused by multipath propagation. Space-time (ST) coding is an effective method to combat the severe effects of fading in wireless channels, by improving the diversity gain and the spectral efficiency with the use of multiple transmit and receive antennas. There has been a great interest in literature on space-time trellis codes [2], [3], [4] and space-time block codes [5], [6], [7].

The design of space-time codes in frequency-selective fading channel is complicated because of the existence of intersymbol interference (ISI). Orthogonal frequency division multiplexing (OFDM) is an efficient method to combat the ISI problem. OFDM effectively converts a frequency-selective channel into parallel flat fading channels. On the other hand, channels with ISI provide frequency diversity that can be exploited by a careful design. Space-time trellis codes [8], [9], [10] and space-time block codes [11], [12], [13], [14] for OFDM systems have been proposed in literature. A hybrid method, which combined both space-time trellis coding and space-time block coding to achieve full frequency and space diversity is proposed in [11]. It is well known that, to exploit full frequency diversity, the effective length of the space-time code must be at least equal to the frequency-selectivity order of the channel, where *effective length* is defined as the minimum number of instances for which any pair of space-time codewords differ [9], [14]. Therefore, space-time codes with high trellis complexities are needed for channels with large frequency-selectivity orders.

In this paper, we consider bit-interleaved space-frequency coded modulation (BI-SFCM) for OFDM systems. The advantage of BI-SFCM is that, with bit-interleaving, the diversity order can be increased to the minimum Hamming distance of the code, which can be larger than the effective length with the same trellis complexity. In addition, bit interleaving separates the design of the error correcting code and that of the modulation. Hence standard and well-known convolutional codes can be used directly.

Space-time coding with BICM was proposed before in [15], which used the space-time block codes of [5]. This idea was then applied to OFDM systems [14]. In the bit interleaved space-time code (BI-STC) approach suggested in [14], coded bits are interleaved across the subcarriers of OFDM symbols in the BICM fashion and then space-time block coding is performed on the same subcarriers of multiple OFDM symbols. Our method is based on interleaving the coded bits across different subcarriers of multiple OFDM symbols without the space-time block coding step. Since space-time trellis coding is applied across the subcarriers of OFDM symbols, *space-frequency* coding actually results. One major advantage of our method is that BI-SFCM gives higher data rates than that of the BI-STC approach suggested in [14] for the same number of transmit antennas. In particular, we show that the information rate of the proposed BI-SFCM scheme increases linearly with the number of transmit antennas, while the diversity order increases linearly with the number of receive antennas.

The rest of paper is organized as follows. In Section II, the channel model and the BI-SFCM OFDM system is described. We present analytical bounds for the pairwise error probabilities in Section III. In Section IV, we introduce a block precoding technique to increase the diversity order that can be provided by BI-SFCM. Section V includes numerical results from simulations, and Section VI concludes the paper.

II. Bit-Interleaved Space-Frequency Coded Modulation

The proposed BI-SFCM method will be discussed in this section. We assume a multiple-antenna system consisting of t transmit and r receive antennas. A block diagram depicting the system is given in Fig.1 for the case of two transmit and two receive antennas. The components of the transmitter are the convolutional encoder (*ENC*), bit-level interleaver (*ILV*), modulator (μ) that maps tm bits into a vector consisting of t complex signals from the constellation set $\mathcal{X}$ and the usual IFFT (of size K) and cyclic prefix (CP) insertion

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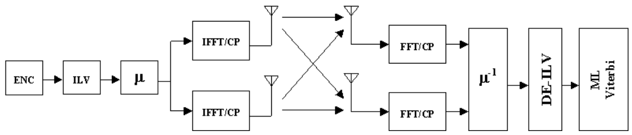

Fig. 1. Model for the proposed BI-SFCM OFDM system for two transmit and two receive antennas.

steps of OFDM. At the receiver side, the received signals are OFDM-demodulated with FFT and the CP is removed. Then, the demodulator (μ^{-1}) produces bit metrics, which are deinterlaved (*DE-ILV*) and fed to the maximum likelihood (ML) Viterbi decoder.

Encoding at the transmitter is done as follows. All the bits of a packet output from the convolutional encoder are interleaved on bit-level and then grouped together in blocks of tKm bits. Each group of tm bits from a block is mapped into a vector of t complex signals. These t complex signals are separated into t substreams. In each substream, K complex signals are grouped together and input to the IFFT. Then, a CP is added to produce one OFDM symbol at each substream. Finally, the complex OFDM signals from the t substreams are transmitted through the t transmit antennas simultaneously. The overall data rate is $R_c tmK$ bits per OFDM symbol, where R_c is the convolutional code rate.

We assume a frequency-selective channel model with proper cyclic prefix insertion and perfect synchronization. In frequency domain, for the kth subcarrier of the nth OFDM symbol, we have the received signal vector $\mathbf{Y}(k,n)$ for the transmitted signal vector $\mathbf{X}(k,n)$ given by

$$\mathbf{Y}(k,n) = \mathbf{H}(k,n)\mathbf{X}(k,n) + \mathbf{N}(k,n), \tag{1}$$

for $k = 1, \ldots, K$. Here $\mathbf{N}(k,n)$ is an $r \times 1$ additive Gaussian noise vector, $\mathbf{X}(k,n)$ is an $t \times 1$ code symbol, $\mathbf{Y}(k,n)$ is an $r \times 1$ received signal vector after the FFT, and $\mathbf{H}(k,n)$ is the channel matrix for the kth subcarrier of the nth symbol. The (i,j)th entry of $\mathbf{H}(k,n)$, $H_{ij}(k,n)$, is the gain of the kth subcarrier channel from the ith transmit antenna to the jth receive antenna. In addition, with ideal interleaving, we assume that the $\mathbf{H}(k,n)$'s are independent and identically distributed (iid) for different values of k and n. For each subchannel, we will assume independent Rayleigh fading between the transmit receive antenna pairs. Thus elements of the channel matrix $\mathbf{H}(k,n)$ will be iid zero-mean circularly symmetric complex Gaussian random variables. The noise term $\mathbf{N}(k,n)$ is assumed to be an independent circularly symmetric complex Gaussian random vector. To simplify our notation, we will serialize the subcarrier symbols of the OFDM system. For instance, we will write $\mathbf{X}(k+Kn)$ for $\mathbf{X}(k,n)$, where K is the FFT size.

We consider maximum-likelihood decoding of the transmitted information bit sequence. As in the case of BICM for single-antenna systems, the Viterbi decoder is used together with the bit metrics calculated based on the BICM approach [1]. The difference is that the decoder calculates the bit metric based on the sequence of observations from the r receive antennas.

Following the notation in [1], the bit-interleaving step is defined by the mapping $\pi : k \to (k^{'}, i)$, where k is the ordering of the coded bits c_k, $k^{'}$ denotes the ordering of the t dimensional complex vector $\mathbf{X}(k^{'}) = [X_1(k^{'}), \ldots X_t(k^{'})]^T$, and i indicates the position of the bit c_k in the label $\mathbf{X}(k^{'})$. We use $\mathcal{X}_b^i$ to denote the subset of all the signal points in the constellation, which have the bit value b at the ith bit position. Equivalently, $\mathcal{X}_b^i = \{\mathbf{X} \in \mathcal{X} \times \ldots \times \mathcal{X} : l^i(\mathbf{X}) = b\}$, where $l^i(\mathbf{X})$ stands for the bit in the ith position of the signal vector $\mathbf{X}$.

In order to find the bit metric for a particular bit k, we first find $(k^{'}, i)$ from the interleaver mapping $\pi : k \to (k^{'}, i)$. For the bit sent at ith bit position of the signal vector $\mathbf{X}(k^{'})$, we can write the conditional density function of the corresponding received signal vector $\mathbf{Y}(k^{'})$, given the bit position and value, and the channel state $\mathbf{H}(k^{'})$ as

$$\begin{aligned} & p\left(\mathbf{Y}(k^{'}) \middle| l^i(\mathbf{X}(k^{'})) = b, \mathbf{H}(k^{'})\right) \\ = & \sum_{\mathbf{X} \in \mathcal{X} \times \ldots \times \mathcal{X}} p\left(\mathbf{Y}(k^{'}) \middle| \mathbf{X}, \mathbf{H}(k^{'})\right) P\left(\mathbf{X}(k^{'}) \middle| l^i((\mathbf{X}(k^{'})) = b\right) \\ = & \frac{1}{2^{tm-1}} \sum_{\mathbf{X} \in \mathcal{X}_b^i} p\left(\mathbf{Y}(k^{'}) \middle| \mathbf{X}, \mathbf{H}(k^{'})\right). \end{aligned} \tag{2}$$

Thus the bit metric corresponding to the hypothesis that the kth coded bit has the value of b is

$$\lambda^i(\mathbf{Y}(k^{'}), b) = \log \sum_{\mathbf{X} \in \mathcal{X}_b^i} p\left(\mathbf{Y}(k^{'}) \middle| \mathbf{X}, \mathbf{H}(k^{'})\right). \tag{3}$$

These bit metrics are then used in the Viterbi decoder to make the final decision on the transmitted information bit sequence. The following simplified method, where $\log \sum$ is approximated by $\max \log$, can also be used.

$$\lambda^i(\mathbf{Y}(k^{'}), b) \approx \max_{\mathbf{X} \in \mathcal{X}_b^i} \log p\left(\mathbf{Y}(k^{'}) \middle| \mathbf{X}, \mathbf{H}(k^{'})\right). \tag{4}$$

III. Error-Probability Analysis

In this section, we use the technique derived in [1] to find an upper bound for the bit error probability, P_b, of the proposed BI-SFCM approach. Consider two distinct sequences of coded bits $\underline{\mathbf{c}}$ and $\underline{\hat{\mathbf{c}}}$, which diverge from the same state and remerge after a number of trellis steps corresponding to a Hamming distance of d. Assume that the sequence $\underline{\mathbf{c}}$ is transmitted and define the pairwise error probability (PEP), $P(\underline{\mathbf{c}} \to \underline{\hat{\mathbf{c}}})$, as the probability of decoding in favor of the sequence $\underline{\hat{\mathbf{c}}}$ erroneously, when $\underline{\mathbf{c}}$ and $\underline{\hat{\mathbf{c}}}$ are the only two outcomes of the decoding process. With ideal interleaving and symmetrized channel [1], the PEP is a function of the distance d, labeling map μ and constellation set $\mathcal{X}$, write $P(\underline{\mathbf{c}} \to \underline{\hat{\mathbf{c}}}) = f(d, \mu, \mathcal{X})$. Then, the bit error probability can be upper bounded by

$$P_b \leq \frac{1}{k_c} \sum_{d=d_{free}}^{\infty} W_I(d) f(d, \mu, \mathcal{X}), \tag{5}$$

where $W_I(d)$ is the total input weight of error events at Hamming distance d and k_c is the number of information bits per unit time of the convolutional code used [16]. We will try to find exressions for the PEP $f(d, \mu, \mathcal{X})$ assuming perfect knowledge of CSI at the receiver and use of the simplified bit metrics given in (4).

The total metric difference Δ for the paths of $\underline{\mathbf{c}}$ and $\underline{\hat{\mathbf{c}}}$ is

$$\Delta(\underline{\mathbf{X}}, \underline{\mathbf{Z}}) = \sum_{k=1}^{d} \Delta(\mathbf{X}_k, \mathbf{Z}_k), \tag{6}$$

where $\underline{\mathbf{X}}$ and $\underline{\mathbf{Z}}$ are the sequences of signal vectors corresponding to the paths $\underline{\mathbf{c}}$ and $\underline{\hat{\mathbf{c}}}$, respectively. Each term in the summation is the metric difference for the signal vector components $\mathbf{X}$ and $\mathbf{Z}$ of the two sequences $\underline{\mathbf{X}}$ and $\underline{\mathbf{Z}}$, given by,

$$\Delta(\mathbf{X}, \mathbf{Z}) = \log p(\mathbf{Y}|\mathbf{X}, \mathbf{H}) - \log p(\mathbf{Y}|\mathbf{Z}, \mathbf{H}). \tag{7}$$

Using the Laplace transform approach [1], we can write

$$f(d, \mu, \mathcal{X}) \leq \frac{1}{2\pi j} \int_{\alpha - j\infty}^{\alpha + j\infty} [\psi(s)]^d \frac{ds}{s}, \tag{8}$$

where $\psi(s)$ is defined as

$$\psi(s) = \frac{1}{mt2^{mt}} \sum_{i=1}^{mt} \sum_{b=0}^{1} \sum_{\mathbf{X} \in \mathcal{X}_b^i} \sum_{\mathbf{Z} \in \mathcal{X}_{\bar{b}}^i} \Phi_{\Delta(\mathbf{X}, \mathbf{Z})}(s) \tag{9}$$

and $\Phi_{\Delta(\mathbf{X}, \mathbf{Z})}(s)$ is the Laplace transform of the pdf of the metric difference $\Delta(\mathbf{X}, \mathbf{Z})$. The integration in (8) can easily be evaluated numerically by the method developed in [17].

One can also use the Chernoff bound to obtain a looser but computationally simpler bound as

$$f(d, \mu, \mathcal{X}) \leq \min_{0 < \alpha < \alpha_2} [\psi(\alpha)]^d. \tag{10}$$

Here, $\Phi_{\Delta(\mathbf{X}, \mathbf{Z})}(s)$ has the convergence region defined by the vertical strip $\alpha_1 < Re(s) < \alpha_2$ and α is taken from the intersection of the positive real line with the convergence regions of $\Phi_{\Delta(\mathbf{X}, \mathbf{Z})}(s)$'s for all possible $(\mathbf{X}, \mathbf{Z})$ pairs.

Now, we can obtain closed-form expressions for the bounds using the signal set and the channel model. Conditioning on the random channel matrix $\mathbf{H}$, we can write the Laplace transform as

$$\Phi_{\Delta(\mathbf{X}, \mathbf{Z})|\mathbf{H}}(s) = \exp\left(\frac{-(2s - s^2) |\mathbf{H}(\mathbf{x} - \mathbf{z})|^2}{4\sigma^2} \right). \tag{11}$$

To obtain the bounds for $f(d, \mu, \mathcal{X})$, we need to take the expectation of the conditioned transform in (11) with respect to the elements of the random channel matrix $\mathbf{H}$. We can write the quadratic term as [18],

$$|\mathbf{H}(\mathbf{X} - \mathbf{Z})|^2 = \mathbf{h}^H \mathbf{Q}(\mathbf{X}, \mathbf{Z}) \mathbf{h}, \tag{12}$$

where $\mathbf{h} = vec(\mathbf{H})$ is the $tr \times 1$ column vector consisting of the elements of $t \times r$ matrix $\mathbf{H}$, and $\mathbf{Q} = (\mathbf{X} - \mathbf{Z})(\mathbf{X} - \mathbf{Z})^H \otimes \mathbf{I}_r$, with $\otimes$ denoting the Kronecker product between two matrices. Then, we can write [19]

$$\Phi_{\Delta(\mathbf{X}, \mathbf{Z})}(s) = \left| \mathbf{I} + \frac{(2s - s^2) \mathbf{Q}(\mathbf{X}, \mathbf{Z})}{4\sigma^2} \right|^{-1}, \tag{13}$$

where $|\cdot|$ denotes the determinant of a matrix. This expression can then be used to evaluate the bound of $f(d, \mu, \mathcal{X})$ in (8) numerically or to obtain the Chernoff bound in (10). For the Chernoff bound, after some simple manipulations, we have

$$f(d, \mu, \mathcal{X}) \leq \left(\frac{1}{mt2^{mt}} \sum_{i=1}^{mt} \sum_{b=0}^{1} \sum_{\mathbf{X} \in \mathcal{X}_b^i} \sum_{\mathbf{Z} \in \mathcal{X}_{\bar{b}}^i} \left| \mathbf{I} + \frac{\mathbf{Q}(\mathbf{X}, \mathbf{Z})}{4\sigma^2} \right|^{-1} \right)^d. \tag{14}$$

We note that the bounds given in (8) and (14) are often loose. To obtain tighter bounds, one can employ the approach described in [1] to expurgate irrelevant error events. In this way, approximate but tighter bounds similar to the ones in (8) and (14) can be obtained by replacing the last summation in (9) and (14) for each $\mathbf{X} \in \mathcal{X}_b^i$ by the closest $\mathbf{Z} \in \mathcal{X}_{\bar{b}}^i$.

From (14), we see that the spatial diversity order is determined by the rank of the $tr \times tr$ matrix $\mathbf{Q}(\mathbf{X}, \mathbf{Z})$. Defining $\mathbf{d} = \mathbf{X} - \mathbf{Z}$, we have

$$rank(\mathbf{Q}(\mathbf{X}, \mathbf{Z})) = rank(\mathbf{d}\mathbf{d}^H \otimes \mathbf{I}_r) = rank(\mathbf{d}\mathbf{d}^H) \times r = r, \tag{15}$$

because $rank(\mathbf{d}\mathbf{d}^H) = 1$. Hence the spacial diversity order is equal to the number of receive antennas. In particular, it can be shown that

$$\left| \mathbf{I} + \frac{\mathbf{Q}(\mathbf{X}, \mathbf{Z})}{4\sigma^2} \right| = \left(1 + \frac{|\mathbf{X} - \mathbf{Z}|^2}{4\sigma^2} \right)^r. \tag{16}$$

Together with BICM, the total diversity order is rd. In summary, the diversity order of BI-SFCM and the information rate increase linearly with the number of receive antennas and the number of transmit antennas, respectively.

IV. Increasing Diversity with Block Precoding

In this section, we explore a method to increase the diversity order of the system by precoding blocks of signal vectors at the transmitter properly and decoding the corresponding blocks of received vectors together. The idea is similar to space-time block codes, however, with the distinction that the data rate is not reduced.

Encoding at the transmitter is modified as follows. Convolutionally encoded and interleaved bits of a packet are grouped together in blocks of $ntmK$ bits. Each group of ntm bits from a block is mapped into n vectors, each of which contains t complex signals. These n vectors are chosen from n subcarriers having independent channel responses. We concatenate these n vectors into one block signal vector $\mathbf{X}$ of size $tn \times 1$, which is then precoded by multiplication with a precoding matrix $\mathbf{A}$. The precoded sequence of tn complex signals are then separated into t substreams. Again, in each substream, K-point IFFT is performed on a group of K complex signals and CP is inserted to produce one

OFDM symbol. Finally, we transmit t complex signals through the t transmit antennas simultaneously. This precoded scheme has the same overall data rate with the BI-SFCM without precoding described in Section II.

The operations at the receiver needs to be modified accordingly. For each subcarrier, we obtain a $t \times 1$ received signal vector by taking one complex signal from the output of each FFT block in the t substreams. We take n such vectors from n subcarriers in accordance with the grouping order employed at the transmitter and concatenate them into a single received signal vector of size $tn \times 1$. In this case, the $rn \times 1$ received vector $\mathbf{Y}$ after the FFT can be written as

$$\mathbf{Y} = \mathbf{HAX} + \mathbf{N}. \tag{17}$$

Here, $\mathbf{X}$ is the $tn \times 1$ transmission signal, $\mathbf{A}$ is the $tn \times tn$ precoding matrix and $\mathbf{H} = diag\left(\mathbf{H}_1, \ldots, \mathbf{H}_n\right)$ is the $rn \times tn$ channel matrix, where each $\mathbf{H}_i$ is the $r \times t$ channel matrix for the corresponding subcarrier. Decoding can be performed in the same way, by using (3) or (4) together with (2). However, it is necessary to substitue the modified channel matrix $\tilde{\mathbf{H}} = \mathbf{HA}$ instead of $\mathbf{H}$.

Analysis similar to the case of BI-SFCM without precoding can be applied to the precoded system by using $\tilde{\mathbf{X}} = \mathbf{AX}$ instead of $\mathbf{X}$. Again, the crucial point is the $ntr \times ntr$ matrix $\mathbf{Q}$, whose rank multiplied with d gives the total diversity order of the system. Note that we can write

$$\mathbf{Q}(\tilde{\mathbf{X}}, \tilde{\mathbf{Z}}) = diag\left(\mathbf{Q}_1, \ldots, \mathbf{Q}_n\right). \tag{18}$$

Then, letting $\mathbf{d}_i = \left(\tilde{\mathbf{X}} - \tilde{\mathbf{Z}}\right)_i$, we have

$$rank(\mathbf{Q}) = \sum_{i=1}^{n} rank(\mathbf{Q}_i) = \sum_{i=1}^{n} rank(\mathbf{d}_i \mathbf{d}_i^H \otimes \mathbf{I}_r). \tag{19}$$

Eqn. (19) implies that it is possible to achieve a diversity order of nrd, with the use of a precoding matrix $\mathbf{A}$ such that $\mathbf{d}_i \neq \mathbf{0}$ for all $i = 1, \ldots, n$ and all pairs of signal vectors $\tilde{\mathbf{X}}, \tilde{\mathbf{Z}}$. Similar to the method in [20], it is possible in many cases to employ the following form of precoding to achieve this:

$$\mathbf{A} = \mathbf{F\Theta}, \tag{20}$$

where $\mathbf{F}$ is an FFT matrix and the $\mathbf{\Theta}$ is a diagonal matrix with diagonal elements on the unit circle.

V. Numerical results

In this section, we provide numerical results to evaluate the performance of the proposed BI-SFCM schemes and compare them with other approaches. We employ the channel model with ideal interleaving, that is, for each subcarrier, we have independent Rayleigh fading between all the transmit and receive antenna pairs. An OFDM system with 64 subcarriers is considered for the case of 2 transmit and 2 receive antennas. *ENC* is the optimum rate-$1/2$ binary convolutional code with minimum Hamming distance of $d = 5$. Signals are taken from a Gray coded QPSK constellation. The rate of the BI-SFCM formed in this way is $R = 2K$ bits per OFDM symbol and the overall spectral efficiency is $2K/(K + K_c)$ b/s/Hz, where K_c is the duration of the cyclic prefix. For simplicity of notation, we will refer the code rate of this scheme as $R = 2$. For BI-SFCM with precoding, we take $n = 2$ and use the precoding matrix [20]

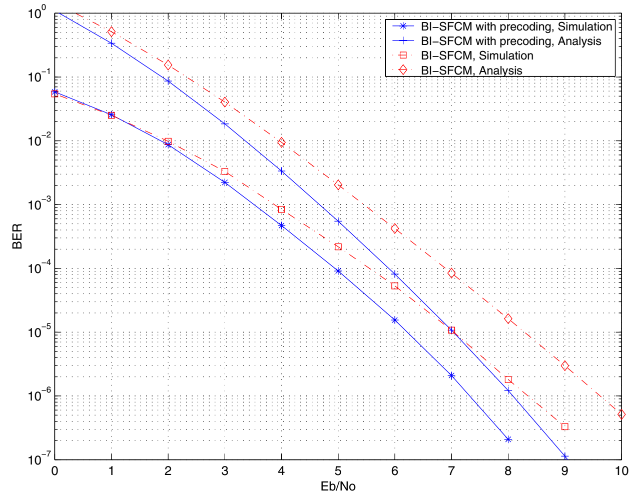

Fig. 2. Performance of the BI-SFCM OFDM system with a 2x2 array: analysis and simulation.

$$\mathbf{A} = \mathbf{F\Theta}, \tag{21}$$

where $\mathbf{F}$ is the 4×4 FFT matrix and the $\mathbf{\Theta} = diag\left(e^{j\theta_0}, e^{j\theta_1}, e^{j\theta_2}, e^{j\theta_3}\right)$, with $\theta_i = \pi i/8$ for $i = 0, \ldots, 3$. It can be shown that with this precoding matrix, (19) gives the value of $nr = 4$ for the rank of $\mathbf{Q}$ and the overall diversity order is $nrd = 20$.

In Fig. 2, we compare the results from simulations with the results from the analytical bounds obtained in Section III, for the cases with and without precoding. We can see that the bounds are about 1dB away from simulation at the high SNR region. These bounds provide tools for system design issues such as precoding. For instance, the increase in the diversity order with precoding can be observed directly from the bounds. In addition, we observe that we do not suffer from a loss in coding gain by employing precoding in this case.

In Fig. 3, we compare the performance of our methods with the BI-STC scheme proposed in [14]. Using our notation, the code rate of the BI-STC with the QPSK constellation is $R = 1$. For a fair comparison, we also consider the BI-STC with the 16-QAM constellation. The resulting code rate is $R = 2$. It is seen that BI-STC and BI-SFCM with precoding achieves the same order of diversity. At the BER of 10^{-6}, the performance of BI-SFCM without precoding is about 2.5 dB worse than the rate-1 BI-STC, but is about 1 dB better than the rate-2 BI-STC. On the other hand, BI-SFCM with precoding achives the same order of diversity with BI-STC and performs about 2 dB better than the rate-2 BI-STC. We note that the advantage of BI-SFCM will be more significant for the cases of higher spectral efficiencies.

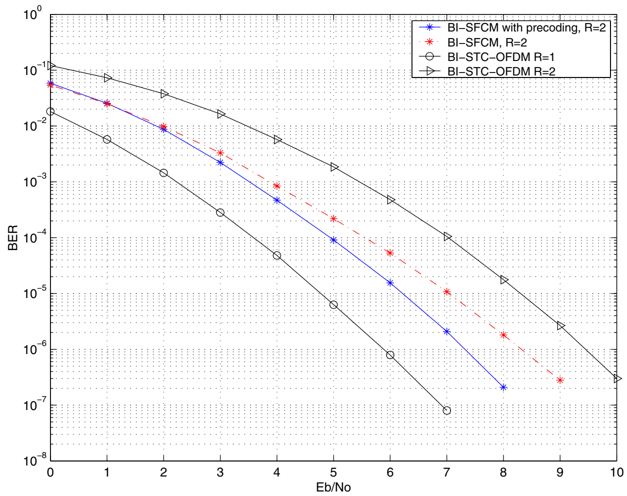

Fig. 3. Performance comparison of the proposed BI-SFCM OFDM system and BI-STC OFDM system with a 2x2 array.

VI. CONCLUSION

Bit-interleaved space-frequency coded modulation (BI-SFCM), based on bit-interleaved coded modulation and space-time trellis coding, was proposed in this paper. In addition, a precoding method was developed to increase the diversity order of the proposed BI-SFCM scheme. It was also shown that the proposed method provides better performance compared to a similar architecture, based on BICM and space-time block coding, that has the same spectral efficiency. Therefore, the proposed BI-SFCM approach is more suitable for systems with high spectral efficiencies.

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