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Non-Linear Limited-Feedback Precoding for ICI Reduction in Closed-Loop Multiple-Antenna OFDM Systems

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Abstract— **We consider precoding for intercarrier interference (ICI) reduction due to frequency offsets in closed-loop orthogonal frequency-division multiplexing (OFDM) systems. Since the complete channel state information (CSI) is difficult to obtain at the transmitter, we develop a novel non-linear limited-feedback Tomlinson-Harashima (LFB-TH) precoder to suppress ICI in closed-loop multiple-antenna OFDM. Our precoder applies a pre-designed codebook of matrices, which is available at both the transmitter and the receiver. The receiver selects the feedback matrix from the codebook according to a specific selection criterion and transfers the index of the chosen matrix to the the transmitter. Hence the CSI is not required at the transmitter in the proposed precoder. The codebook design criterion and feedback matrix selection criteria are analyzed. Our LFB-TH precoder significantly reduces the bit error rate (BER) degradation due to frequency offset; the degradation is virtually eliminated for normalized frequency offsets as high as** 10%.

Index Terms— **Orthogonal frequency-division multiplexing (OFDM), multiple-input multiple-output (MIMO), frequency offset, Tomlinson-Harashima (TH) precoding, limited feedback**

I. Introduction

Orthogonal frequency-division multiplexing (OFDM) is an attractive technique for high-throughput transmission over wireless radio channels. Spectral efficiency and the ability to mitigate the dispersive effects of wideband channels have made OFDM a popular choice for several applications including the IEEE 802.11a and 802.16 and a strong candidate for fourth-generation (4G) cellular communication systems. A wireless communication system employing closed-loop multiple-input multiple-output (MIMO) links assures capacity and performance gain, and enables precoding to avoid interference with complete channel state information (CSI) at the transmitter. A combination of OFDM and closed-loop MIMO techniques is hence very promising. However, an essential feature of OFDM is subcarrier orthogonality, which is lost when a carrier frequency offset is present. The frequency offset may be due to a Doppler shift or oscillator mismatch, and results in intercarrier interference (ICI) and an error floor. The performance degradation caused by ICI becomes significant as the carrier frequency, block size, and vehicle velocity increase [1].

For a single-antenna system, ICI may be reduced by time-domain widowing [2], frequency-domain equalization [3] and ICI self-cancellation coding [4]. These methods are for **open-loop** systems; i.e., complete CSI, including frequency offset and channel impulse response, is only available at the receiver. Some precoding schemes using complete or partial CSI at the transmitter are also proposed for ICI reduction in **closed-loop** OFDM. In [5], the performance of successive interference cancellation in linear-precoded OFDM is analyzed. For this scheme, the transmitter needs the complete CSI, which is not feasible for practical implementation. To overcome this problem and to improve the precoder performance, we proposed a non-linear Tomlinson-Harashima (TH) precoder [6] - [8] to suppress ICI in [9], where it requires only channel response, but not frequency offset, at the transmitter.

If accurate CSI is available, closed-loop systems offer a remarkable performance gain over their open-loop counterparts. The CSI can be provided at the transmitter using one of two approaches: first, for time-division duplex (TDD) systems, the transmitter directly estimates the CSI exploiting the radio channel's reciprocity as the forward and reverse links have the same channel response; second, CSI can be estimated by the receiver and fed back to the transmitter. However, in many wireless systems, the complete CSI may not be available at the transmitter because the forward and reverse channels are not reciprocal or capacity of the feedback links is limited. consequently, limited-feedback signal design and linear precoders have been explored in [10] - [12] for MIMO systems. The basic idea is to design the precoding matrix at the receiver rather than at the transmitter. Nevertheless, this approach has not yet been applied in MIMO OFDM for ICI suppression.

In this paper, we propose a non-linear limited-feedback TH (LFB-TH) precoder for ICI reduction in closed-loop MIMO OFDM systems. Our proposed LFB-TH precoder consists of a feedforward filter and a modulo arithmetic device at the receiver, and a modulo arithmetic feedback structure at the transmitter (see Fig. 2). The novel feature of this precoder is that the feedback matrix is determined at the receiver, i.e., the CSI is not required at the transmitter. The receiver chooses the feedback matrix from a finite codebook based on a selection criterion and conveys the chosen matrix to the transmitter using a limited number of bits. The codebook of feedback matrices is known a priori to both the transmitter and the receiver. The codebook design algorithms and different matrix selection criteria are analyzed. Our LFB-TH precoder significantly reduces the bit error rate (BER) degradation of OFDM due to frequency offset; the degradation is virtually eliminated for normalized frequency offsets as high as 10%.

II. System Model

This section will introduce a MIMO OFDM system model employing N subcarriers with M_T transmit antennas and M_R receive antennas in the presence of frequency offset.

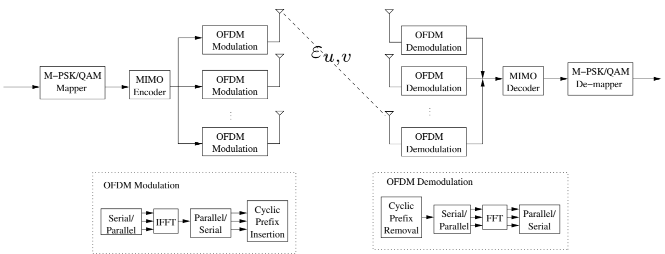

Fig. 1. Block diagram of a MIMO OFDM link.

The structure of a MIMO OFDM link is shown in Fig. 1. The k-th sample of the inverse discrete Fourier transform (IDFT) output of the u-th transmit antenna is

$$x_u(k) = \sum_{n=0}^{N-1} X_u[n] e^{\jmath \frac{2\pi}{N} nk}, \qquad (1)$$

where $X_u[n]$ is an M-ary quadrature amplitude modulated (QAM) symbol sent by the u-th transmit antenna. The input OFDM symbol can then be written as $\mathbf{X}_u = [X_u[0] \ldots X_u[N-1]]^T$. Assuming that the mulitpath fading channel consists of L resolvable paths, the time-domain received signal of the v-th receive antenna can be written as

$$y_v(n) = \sum_{u=1}^{M_T} \sum_{l=0}^{L-1} h_{u,v}(l,n) x_u(n-l) e^{\jmath \frac{2\pi}{N} n\varepsilon_{u,v}} + w_{u,v}(n), \qquad (2)$$

where $h_{u,v}(l,n)$ represents the complex path gain for the l-th path at time n. The usual assumption for precoding is that the multipath channel is sufficiently slowly fading. The $h_{u,v}(l,n)$ thus does not change at least over an OFDM symbol interval T_s, i.e, $h_{u,v}(l,n) = h_{u,v}(l)$. The $w_{u,v}(n)$ is a discrete-time additive white Gaussian noise (AWGN) sample. The $\varepsilon_{u,v}$ is the normalized frequency offset and assumed constant for one or more symbol periods. A cyclic prefix, which is longer than the expected maximum excess delay, is customarily inserted at the beginning of each time-domain OFDM symbol to prevent intersymbol interference (ISI). Discarding the cyclic prefix, the demodulated signal in the frequency domain is obtained by performing the FFT of $\mathbf{y}_v$ as

$$\begin{aligned} Y_v[k] &= \underbrace{\sum_{u=1}^{M_T} H_{u,v}[0] X_u[k]}_{\text{desired signal}} \\ &+ \underbrace{\sum_{u=1}^{M_T} \sum_{m=0, m\neq k}^{N-1} H_{u,v}[k-m] X_u[m]}_{\text{ICI component}} + \sum_{u=1}^{M_T} W_{u,v}[k], \end{aligned} \qquad (3)$$

where

$$H_{u,v}[k-m] = \frac{1}{N} \sum_{l=0}^{L-1} h_{u,v}(l) e^{-\jmath \frac{2\pi}{N} ml} \sum_{n=0}^{N-1} e^{-\jmath \frac{2\pi}{N} n(k-m+\varepsilon_{u,v})}, \qquad (4)$$

and $W_{u,v}[k]$ is an AWGN sample with zero mean and variance $\sigma^2_{W_{u,v}}$; $W_{u,v}[k]$ for different subcarriers k are assumed independent and identically distributed (i.i.d).

For a MIMO OFDM system, the signal on the v-th receive antenna is $\mathbf{Y}_v = [Y_v[0] \ldots Y_v[N-1]]^T$, $v = 1, \ldots, M_R$. The NM_R-dimensional received signal vector $\mathbf{Y} = [\mathbf{Y}_1^T, \ldots, \mathbf{Y}_{M_R}^T]^T$ can therefore be represented in as

$$\mathbf{Y} = \mathbf{G}\mathbf{X} + \mathbf{W}, \qquad (5)$$

where the $NM_R \times NM_T$ MIMO channel matrix is

$$\mathbf{G} = \begin{pmatrix} \mathbf{G}_{1,1} & \cdots & \mathbf{G}_{M_T,1} \\ \vdots & \ddots & \vdots \\ \mathbf{G}_{1,M_R} & \cdots & \mathbf{G}_{M_T,M_R} \end{pmatrix}, \qquad (6)$$

and the $N \times N$ sub-matrix $\mathbf{G}_{u,v}$ in (6) has the $\{m,k\}$th entry of $G_{u,v}(m,k) = H_{u,v}[k-m]$. The transmitted vector $\mathbf{X} = \left[\mathbf{X}_1^T \ldots \mathbf{X}_{M_T}^T\right]^T$. The noise matrix $\mathbf{W} = \left[\mathbf{W}_1^T \ldots \mathbf{W}_{M_R}^T\right]^T$ and $\mathbf{W}_v = \sum_{u=1}^{M_T} \mathbf{W}_{u,v}$ is the noise on the v-th receive antenna.

III. Limited-Feedback Tomlinson-Harashima Precoding

A. LFB-TH Precoding

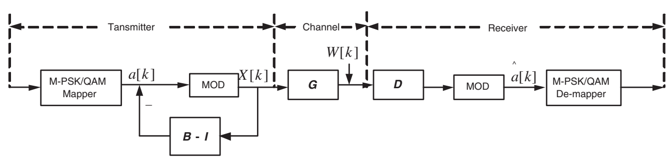

Fig. 2. Limited-feedback Tomlinson-Harashima precoder for MIMO OFDM.

Previous work in limited-feedback precoding has focused on linear precoding design. This section develops our non-linear LFB-TH precoder for ICI mitigation in uncoded MIMO OFDM systems. The non-linear property of the TH precoder eliminates noise enhancement in linear precoding, and the precoding structure avoids error propagation common in decision feedback equalization (DFE), which is used in vertical Bell Lab's layered space time (V-BLAST) [13]. We construct a finite codebook and choose the feedback matrix from the codebook at the receiver. The codebook is known at both the transmitter and the receiver such that only the index of the selected matrix needs to be sent to the transmitter. The codebook design algorithms and matrix selection criteria are analyzed. For convenience of signal detection, we assume $M_R \geq M_T$.

The structure of the proposed precoder is illustrated in Fig. 2. The receiver side consists of a feedforward filter $\mathbf{D}$ and a modulo arithmetic device. The transmitter side includes a modulo arithmetic feedback structure employing the matrix $\mathbf{B}$, by which the transmitted symbols $X[k]$ are calculated for the data carrying symbols $a[k]$ drawn from a signal constellation.

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We build up a codebook $\mathcal{B} = \{\mathbf{B}_1, \mathbf{B}_2, ..., \mathbf{B}_K\}$, which is used at both the transmitter and the receiver. Each complex matrix $\mathbf{B}_i$ in the codebook is a unitary $NM_T \times NM_T$ matrix with orthonormal columns, i.e, $\mathcal{B} \in \Theta(NM_T, NM_T)$, where $\Theta(m,n)$ is a set of $m \times n$ matrices with orthonormal columns. The feedback matrix $\mathbf{B}$ is chosen from the codebook $\mathcal{B}$ at the receiver according to the current or average channel conditions. We assume that $\mathbf{B} = \mathcal{P}(\mathbf{G})$, where $\mathcal{P}(\mathbf{G})$ is a mapping from the channel matrix $\mathbf{G}$ to $\mathcal{B}$, which optimizes the feedback filter according to some performance criterion. The limited feedback only represents an index of the selected feedback matrix with a limited number of bits. When we use a codebook with K feedback matrices, only $\lceil \log_2 K \rceil$ bits need to be sent to the transmitter. With the zero-forcing (ZF) criterion, the $NM_T \times NM_R$ feedforward matrix can be immediately obtained as $\mathbf{D} = \left[\mathbf{G}\mathbf{B}^H\right]^\dagger$, where † denotes Moore-Penrose pseudo-inverse.

With an M-ary QAM square signal constellation

$$\mathcal{A} = \left\{ a_I + \jmath a_Q \mid a_I,\ a_Q \ \in\ \pm 1, \pm 3, ..., \pm(\sqrt{M}-1) \right\}, \tag{7}$$

the transmitted symbols after the modulo-$2\sqrt{M}$ operation can be written as

$$\begin{aligned} X[k] &= \text{MOD}_{2\sqrt{M}} \left\{ a[k] - \sum_{q=0}^{N_B-1} B(k,q) X[q] \right\} \\ &= a[k] + c[k] - \sum_{q=0}^{N_B-1} B(k,q) X[q], \end{aligned} \tag{8}$$

where N_B is the number of columns of feedback matrix $\mathbf{B}$ and $c[k]$ is the precoding symbol $c[k] \in \mathcal{C} = \left\{ 2\sqrt{M}(c_I + \jmath c_Q) \mid c_I, c_Q \in \mathbb{I} \right\}$ which restricts $X[k]$ to the interval $\left(-\sqrt{M}, \sqrt{M}\right]$; $\mathbb{I}$ is the set of integers. The congruent signal points are obtained by extending the signal set $\mathcal{A}$ to the set $\mathcal{J} = \{a + c | a \in \mathcal{A},\ c \in \mathcal{C}\}$. From the set $\mathcal{J}$ the current effective data symbol is selected such that $X[k]$ lies in the boundary region of $\mathcal{A}$. At the receiver, a slicer yields estimates $\hat{c}[k]$, from which the unique estimated data symbols are produced by the same non-linear modulo reduction into the initial constellation $\mathcal{A}$. Consequently, after discarding the modulo congruence, the proposed LFB-TH precoder cancels the fading channel distortion and ICI. Our precoder produces a near i.i.d transmit sequence $X[k]$, uniformly distributed over the initial constellation $\mathcal{A}$. Moreover, $X[k]$ can be assumed mutually uncorrelated with variance $E_s = \mathrm{E}\left[|X[k]|^2\right], \forall k$. After linear pre-distortion via $\mathbf{B}^{-1}$ and the cascade $\mathbf{GD}$, the transmitted symbols $X[k]$ are corrupted by an additive noise and become $Y[k] = X[k] + W'[k]$. Note that $W'[k]$, the entry of the filtered noise vector $\mathbf{W}' = \mathbf{DW}$, is the k-th sample with individual variance $\sigma^2_{W'_k}$.

In our proposed LFB-TH precoder, the transmitter does not require any channel information and only a limited number of bits are delivered to it. When a better performance is required, a larger codebook can be constructed, i.e., more bits can be fed back. Furthermore, due to the non-linear property, THP avoids power efficiency loss as in linear precoding; and because the feedback part is moved to the transmitter, error propagation which is typically associated with the DFE, can be avoided. Therefore, low BER can be expected for LFB-THP.

B. Feedback Matrix Selection Criteria

In this subsection, we analyze the performance criteria used for selecting the feedback matrix.

Given the channel matrix $\mathbf{G}$ at the receiver, a singular value decomposition (SVD) of the channel matrix generates

$$\mathbf{G} = \mathbf{U}\mathbf{\Gamma}\mathbf{V}^H, \tag{9}$$

where $\mathbf{U}$ and $\mathbf{V}$ are unitary matrices with size of $NM_R \times NM_R$ and $NM_T \times NM_T$, respectively; $\mathbf{\Gamma}$ is a $NM_R \times NM_T$ diagonal matrix with real, non-negative entries $\gamma[k]$, $k = 1, ..., NM_T$, in descending order $\gamma_1 \geq \gamma_2 \geq \cdots \geq \gamma_{NM_T} \geq 0$; the $\gamma[k]$ are singular values of the channel matrix $\mathbf{G}$.

1) Minimum Mean Squared Error (MMSE) Criterion: A specific design target is considered in [14] by minimizing the symbol MSE to improve the overall system performance. The MSE of the feedback filter can be expressed as

$$\overline{\text{MSE}}(\mathbf{B}) = E_s \left(\mathbf{I} + \mathbf{B}\mathbf{G}^H \mathbf{R}^{-1}_{W'W'} \mathbf{G}\mathbf{B}^H \right)^{-1}, \tag{10}$$

where $\mathbf{R}_{W'W'} = E\left[\mathbf{W}'\mathbf{W}'^H\right]$ is the noise covariance matrix. We use (10) to select $\mathbf{B}$ from $\mathcal{P}(\mathbf{G})$ according to

$$\mathcal{P}(\mathbf{G}) = \arg \min_{\mathbf{B}_i \in \mathcal{B}} \text{tr}\left[\overline{\text{MSE}}(\mathbf{B}_i)\right], \tag{11}$$

where $\text{tr}(\mathbf{A})$ is trace of the matrix $\mathbf{A}$. As given in [12] and [14], the optimal precoding matrix $\mathbf{B}_{\text{opt}} = \mathbf{V}^H$, and the corresponding $\mathbf{D} = \mathbf{\Gamma}^\dagger \mathbf{U}^H$.

2) Maximum Singular Value (MSV) Criterion: Using a selection criterion based on the minimum signal-to-noise ratio (SNR) is difficult to implement since it requires the computation of the SNR every OFDM symbol interval. In [14], the criterion with a simple closed form of SNR is designed

$$\text{SNR}(\mathbf{B}, \mathbf{D}) = E_s \mathbf{B}\mathbf{G}^H \mathbf{D}^H \left(\mathbf{D}\mathbf{R}_{W'W'}\mathbf{D}^H \right)^{-1} \mathbf{D}\mathbf{G}\mathbf{B}^H. \tag{12}$$

The minimum substream SNR can be approximately achieved by maximizing the minimum singular value of the effective channel $\mathbf{G}\mathbf{B}_i^H$. The solution of the optimization problem is therefore

$$\mathcal{P}(\mathbf{G}) = \arg \max_{\mathbf{B}_i \in \mathcal{B}} \gamma_{\min}(\mathbf{G}\mathbf{B}_i^H), \tag{13}$$

i.e., the minimum singular value of (12) should be as large as possible. As shown in [12], the optimal feedback filter is $\mathbf{B}_{\text{opt}} = \mathbf{V}^H$.

C. Codebook Design Criterion

In this subsection, we analyze a codebook design criterion using the codebook selection function $\mathcal{P}(\mathbf{G})$. As in [15], we need to minimize

$$\rho = \max_{1<p<q<K} \|\mathbf{B}_p^H \mathbf{B}_q\|_F, \tag{14}$$

where $\|\mathbf{A}\|_F$ is Frobenius norm of the matrix $\mathbf{A}$. The set of all possible column spaces of the matrices in $\Theta(NM_T, NM_T)$ is a complex Grassmannian $\Xi(NM_T, NM_T)$, in which Grassmannian $\Xi(m,n)$ is the set of n-dimensional subspaces in an

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m-dimensional vector space. The minimum chordal distance of any two arbitrary subspaces is defined as

$$\mathrm{d}_{\min} = \sqrt{M_T(1-\rho^2)}. \tag{15}$$

We thus choose the set of K subspaces in $\Xi(NM_T, NM_T)$ such that $\mathrm{d}_{\min}$ is as large as possible.

A scheme for designing the desired codebook with large minimum distance has been presented in [15]. Each precoding matrix in the set $\mathcal{B}$ can be given as

$$\mathbf{B}_i = \mathbf{\Phi}^{i-1}\mathbf{B}_1, \tag{16}$$

where $\mathbf{B}_1$ is a $NM_T \times NM_T$ unitary matrix; $\mathbf{\Phi}$ is a $NM_T \times NM_T$ unitary matrix and $\mathbf{\Phi}^K = \mathbf{I}$, i.e., $\mathbf{\Phi}$ is a K-th root of unity. The diagonal entries of $\mathbf{\Phi}$ are $e^{\jmath\frac{2\pi}{K}\varphi_1}$, ..., $e^{\jmath\frac{2\pi}{K}\varphi_{NM_T}}$, and $0 \le \varphi_1$, ..., $\varphi_{NM_T} \le K-1$, and $NM_T < K$. To get the lowest correlations, we choose the parameters φ_1, ..., φ_{NM_T} to achieve

$$\min_{0\le\varphi_1,\ldots,\varphi_{NM_T}\le K-1} \rho = \min_{0\le\varphi_1,\ldots,\varphi_{NM_T}\le K-1} \max_{i=2,\ldots,K} \left\|\mathbf{B}_1^H\mathbf{B}_i\right\|_F . \tag{17}$$

Geometrically, this construction is to rotate an initial NM_T-dimensional subspace using a K-th root of unity to form K different NM_T-dimensional subspaces.

Once the codebook is designed, the feedback matrix $\mathbf{B} \in \mathcal{B}$ using the performance criteria in (11) or (13) is chosen at the receiver. The selected matrix is then delivered to the transmitter using only $\lceil \log_2 K \rceil$ bits.

IV. Simulation Results

We perform simulations to verify the BER performance of our LFB-TH precoder. The codebook is designed as described in Section III, and known a priori at both the transmitter and the receiver. A MIMO QPSK-OFDM system with 64 subcarriers on a 6-tap Rayleigh fading channel is considered. The vehicular B channel specified by ITU-R M. 1225 [16] is used where the channel taps are zero-mean complex Gaussian random processes with normalized variances 0.3226, 0.5737, 0.0302, 0.0574, 0.0017, and 0.0144. The spatial channels between different transmit-receive antenna pairs are uncorrelated. We assume that the receiver has perfect knowledge of CSI, including frequency offset and channel response. For many wireless systems, the multipath channels fade slowly. As an example, for Mode-I digital audio broadcasting (DAB) systems [17], the Doppler shift is 41.67 Hz at a mobile speed of 120 km/h. The corresponding normalized frequency offset is roughly 0.04, i.e., the fading coefficients can be assumed constant over several OFDM symbol intervals. Here, we consider the worst case, where the fading gains $h(l)$, $l = 0, ..., L-1$ change from one OFDM symbol to another.

Fig. 3 shows the BER of single-input single-output (SISO) OFDM for two groups. First, the codebook consists of $K = 256$ precoding matrices, i.e., 8 bits are transferred to the transmitter; while 128 matrices are included in the second group. The BER curve of an OFDM system with zero-frequency offset is shown as a reference. The MMSE selection criterion is used to select the feedback matrix. The LFB-THP reduces ICI notably: even with a normalized frequency offset of 10%, OFDM with our precoder performs as well as the reference, i.e., the ICI has been eliminated completely. The system performs better when the number of feedback bits is increased. In Fig. 4, for an SNR of 20 dB, the BER of LFB-THP OFDM with a 30% normalized frequency offset approaches the performance of OFDM with zero-frequency offset when the size of the codebook is larger than 250.

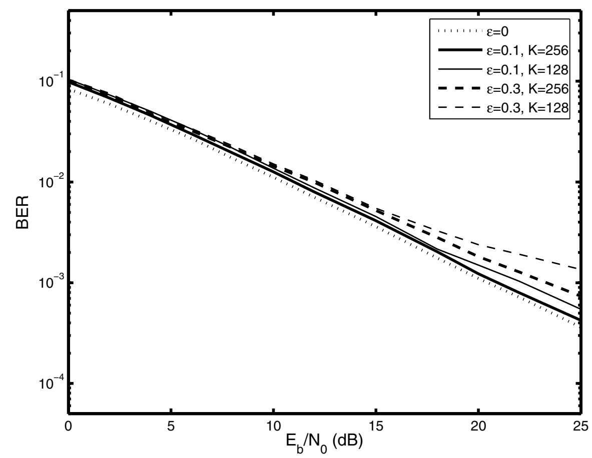

Fig. 3. BER of LFB-THP as a function of the SNR for different values of the normalized frequency offset and 64-subcarrier SISO QPSK-OFDM. MMSE selection criterion is used.

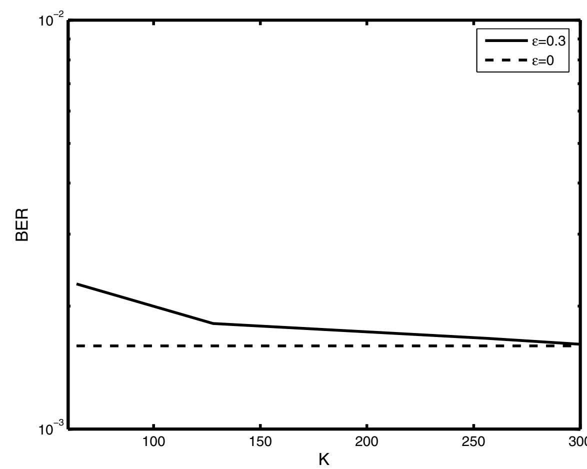

Fig. 4. BER of LFB-THP as a function of the codebook size in 64-subcarrier SISO QPSK-OFDM at SNR=20 dB. MMSE selection criterion is used.

The BER of MIMO OFDM systems with our precoder is shown in Fig. 5 and Fig. 6. The codebook size K is set as 512. The BER curve of OFDM with $\varepsilon = 0$ is shown as a reference. In Fig. 5, the MMSE and MSV selection criteria are used in a 2×2 OFDM system. As before, our precoder reduces ICI significantly; for a 10% normalized frequency offset, the ICI has been suppressed completely. Furthermore, the system with the MMSE criterion outperforms the system with the MSV criterion. Fig. 6 provides the BER of 2×4 OFDM

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with the MMSE criterion and shows that our precoder can be used in orthogonal space-time coded systems. We consider the Alamouti code [18] in this figure. The BER degradation due to ICI is dramatically reduced by our precoder, and the Alamouti-coded OFDM achieves a 3 dB gain over the uncoded system at BER=10^{-4}.

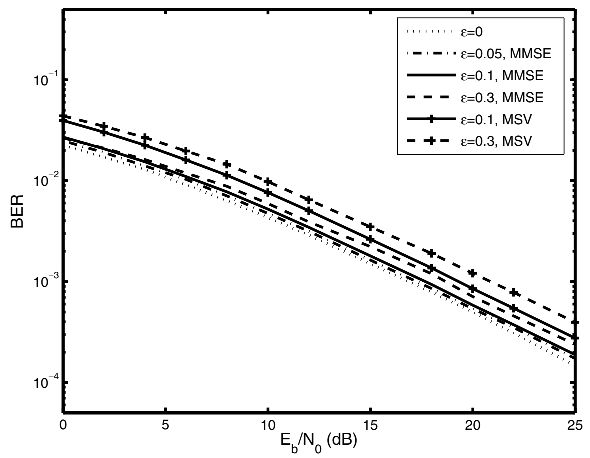

Fig. 5. BER of LFB-THP as a function of the SNR for different values of the normalized frequency offset and 64-subcarrier 2×2 QPSK-OFDM. MMSE and MSV selection criteria are compared.

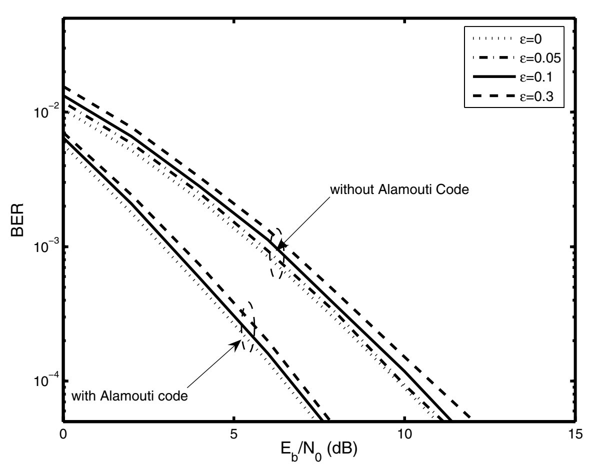

Fig. 6. BER of LFB-THP as a function of the SNR for different values of the normalized frequency offset and 64-subcarrier 2×4 QPSK-OFDM and Alamouti-coded QPSK-OFDM. MMSE selection criterion is used.

V. Conclusion

We have developed a non-linear LFB-TH precoder to suppress ICI in MIMO OFDM. The feedback matrix of our precoder is chosen at the receiver from a finite pre-designed codebook of matrices according to a specific selection criterion. The codebook is known at both the transmitter and the receiver such that only the index of the chosen matrix needs to be fed back. The CSI is hence not required at the transmitter. The codebook design criterion and feedback matrix selection criteria have been analyzed. Simulation results have demonstrated that our proposed LFB-TH precoder significantly reduces the BER degradation due to frequency offset; the degradation is virtually eliminated for normalized frequency offsets as high as 10%.

VI. Acknowledgement

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