Limited-Feedback Precoding for Closed-Loop Multiuser MIMO OFDM Systems with Frequency Offsets

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***Abstract*—Frequency offsets negatively impact the performance of closed-loop multiple-input multiple-output (MIMO) orthogonal frequency-division multiplexing (OFDM) systems. Particularly, when multiple users are active, the impact can be high. Linear precoding and non-linear Tomlinson-Harashima precoding (THP) are thus developed for spatially-multiplexed multiuser OFDM and orthogonal space-time block-coded (OSTBC) OFDM. The proposed precoders employ a limited feedback structure, which is implemented with a shared codebook of precoding matrices, and only the index of the selected optimal matrix is fed back to the transmitter. The conventional limited-feedback design criterion for flat-fading MIMO channels is only applicable to single-user OFDM without frequency offsets. We show that the ICI matrix due to frequency offset does not impact users' precoding individually, and precoding on a per-subcarrier basis is possible. Exploiting this property, the conventional design is generalized to multiuser OFDM with frequency offsets. Non-linear precoding uses a modulo arithmetic precoding matrix (which reduces the power efficiency loss inherent in linear precoding and leads to a lower error rate) and outperforms linear precoding. Our precoders not only offer significant bit error rate (BER) improvement for spatially-multiplexed multiuser MIMO OFDM with frequency offsets, but are equally effective for both OSTBC MIMO OFDM and spatially correlated channels.**

***Index Terms*—Limited feedback precoding, multiuser multiple-input multiple-output (MIMO), orthogonal frequency-division multiplexing (OFDM), frequency offset.**

I. Introduction

CLOSED-LOOP multiple-input multiple-output (MIMO) orthogonal frequency-division multiplexing (OFDM) techniques exploit spatial diversity, provide higher link capacity and throughput, and improve system performance [1]. Typically, either spatial multiplexing (SM) or space-time coding is used to exploit the benefits offered by MIMO links. The special case of space-time diversity coding known as orthogonal space-time block coding (OSTBC)[1] has been adopted by several wireless standards due to its simplicity of encoding and optimal decoding. In an open-loop SM system, the number of receive antennas must exceed the number of transmit antennas. This requirement may not be practical for several applications. On the other hand, OSTBC exists only for certain numbers of transmit antennas, which limits the scope of its potential applications. Moreover, open-loop OSTBC suffers a significant array gain penalty over its closed-loop counterpart. Closed-loop techniques have therefore been discussed in several third-generation (3G) standards [2], [3].

Closed-loop techniques, such as transmitter precoding, can overcome ill-conditioning of the channel matrix and improve the system performance. For multiuser systems, each user's mobile station (MS) knows the frequency offset and channel response affecting its own receiver only. Complete required processing at user level usually is not possible. Transmitter precoding techniques are therefore necessary. They usually require that complete CSI be available at the transmitter. The CSI may be gathered by the transmitter itself exploiting channel reciprocity in time-division duplex (TDD) systems, or by the receiver's feedback to the transmitter [4]. Nevertheless, in many wireless systems, full CSI is difficult to obtain at the transmitter. When the reciprocity of wireless channels does not hold, such as in frequency-division duplex (FDD) systems, perfect frequency offset and channel gains at the transmitter require a high-rate feedback link. Thus, limited-feedback signal design and linear precoders have been explored for flat-fading MIMO channels for feedback volume reduction [4]-[7], [14]. The basic idea of this approach is to use a pre-designed set of matrices between the transmitter and the receiver. The receiver selects the best precoding matrix using full CSI and only the index of the selected matrix is fed back to the base station (BS) transmitter.

Although a combination of OFDM and closed-loop MIMO realizes large capacity gains on frequency-selective fading channels, subcarrier orthogonality, an essential feature of OFDM, is lost if there are carrier frequency offsets, caused by a Doppler shift or oscillator mismatch, resulting in intercarrier interference (ICI) and an error floor. This performance loss increases as the carrier frequency, OFDM symbol size, and vehicle velocity increase [8]. Some precoders for closed-loop OFDM using complete or partial CSI at the transmitter

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[1]OSTBC stands for orthogonal space-time block-coded or orthogonal space-time block coding, depending on the context.

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have also been proposed for interference reduction. In [9], a Bezout precoder is developed for ICI mitigation in MIMO OFDM, when the transmitter has the complete CSI. We have proposed non-linear Tomlinson-Harashima precoding (THP) [10], [11] to suppress ICI [12], [13], using only partial CSI, not including the knowledge of frequency offsets, at the transmitter. Limited-feedback precoding [4]-[7], [14] not only can reduce the number of feedback bits, but also can minimize the system error rate and maximize capacity. **Nevertheless, this approach so far has only been considered for single-user systems over flat-fading channels**, and is suitable for an ideal OFDM case without frequency offsets, in which the overall channel gain matrix is a block-diagonal matrix. This block-diagonal matrix does not hold with frequency offsets. And the fact that different users may have different frequency offsets makes precoding more difficult. The original limited-feedback design in [4]-[7], [14] cannot hence be applied to a more practical OFDM system with frequency offsets.

In this paper, we propose both linear precoding and non-linear limited-feedback THP (LFB-THP) for closed-loop multiuser MIMO OFDM systems with frequency offsets. SM MIMO OFDM with linear receivers and OSTBC MIMO OFDM with maximum likelihood (ML) receivers are also considered. We show that frequency offsets do not impact users' precoding matrices individually, and hence precoding on per-subcarrier basis is possible. Exploiting this property, we generalize the codebook design criterion previously used for single-user flat-fading MIMO systems [5]–[7], [14] to multiuser OFDM systems with frequency offsets. We propose the use of a pre-designed codebook of precoding matrices, available at both the transmitter and each user's receiver. Each user selects optimal matrices at the subcarrier level according to a certain criterion and sends only their indices to the transmitter. Three precoding matrix selection criteria are analyzed: minimum mean squared error (MMSE), maximum singular value (MSV) and maximum mutual information (MMI). To further reduce the number of feedback bits, grouping with interpolation is also introduced. The explicit values of CSI are hence not needed at the transmitter. In our precoders, the feedback load is reduced to only very limited number of bits, which reduces feedback bit rate, and the non-linear processing reduces power efficiency loss inherent in linear precoding, which makes non-linear processing outperform linear precoding. Our precoders also display significant BER improvement for OFDM with frequency offsets over spatially correlated MIMO channels.

A. Organization of the paper

This paper is organized as follows. Section II describes the multiuser MIMO OFDM system model with frequency offsets. Section III proposes linear and non-linear precoding for both SM and OSTBC multiuser MIMO OFDM in the presence of frequency offsets along with the matrix selection criteria. We study the impact of the ICI matrix on the optimal precoding matrix and develop the codebook design scheme. In Section IV, we consider spatially correlated MIMO channels, and analyze the effect of fading correlations on the feedback matrix design. Simulation results for SM MIMO OFDM, OSTBC MIMO OFDM and OFDM in spatially correlated channels are given in Section V. Section VI concludes this paper.

B. Notation

The superscripts ^{T}, ^{H}, * and † stand for transposition, conjugate transposition, element-wise conjugate and Moore-Penrose pseudo-inverse, respectively. Bold symbols denote matrices or vectors. $\jmath = \sqrt{-1}$. The symbols $\otimes$, $\delta(\cdot)$, E and $\mathbf{I}_N$ represent the Kronecker product, Dirac function, expectation operator and the $N \times N$ identity matrix, respectively. The set of integers is indicated as $\mathbb{I} = \{0, \pm 1, \pm 2, \ldots\}$. The $\|\mathbf{A}\|_F$, $\det(\mathbf{A})$, $\gamma(\mathbf{A})$, tr$(\mathbf{A})$ are Frobenius norm, determinant, singular values, trace of the matrix $\mathbf{A}$, respectively. The smallest integer greater than x is $\lceil x \rceil$. The set of $m \times n$ matrices with n orthonormal columns is denoted as $\Theta(m, n)$, and $\Xi(m, n)$ as complex Grassmannian space. An M-ary QAM square signal constellation is defined as $\mathcal{A} = \left\{ a_I + \jmath a_Q \mid a_I,\ a_Q \in \pm 1, \pm 3, \ldots, \pm(\sqrt{M} - 1) \right\}$.

II. System Model

This section will introduce the system model of an N-subcarrier multiuser OFDM downlink system with M_T transmit antennas and U simultaneously active users in the presence of frequency offsets. The u-th user has M_u receive antennas, $M_u < M_T$, and the total number of receive antennas is $M_R = \sum_{u=1}^{U} M_u$.

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

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

where $X_v[k]$ is an M-ary quadrature amplitude modulated (QAM) symbol on the k-th subcarrier sent by the v-th transmit antenna. The input data vector can then be written as $\mathbf{X}_v = \left[X_v[0] \ \ldots \ X_v[N-1] \right]^T$. Assuming that the multipath fading channel contains of L resolvable paths, the time-domain received signal of the u-th user's u_m-th receive antenna can be written as

$$\begin{aligned} y_{u_m}(n) = &\sqrt{\frac{\eta}{M_T}} \sum_{v=1}^{M_T} \sum_{l=0}^{L-1} h_{u_m,v}(l) x_v(n-l) e^{\jmath \frac{2\pi}{N} n \varepsilon_{u_m,v}} \\ &+ w_{u_m,v}(n), \end{aligned} \tag{2}$$

where η is the signal-to-noise power ratio (SNR) at each receive antenna, independent of M_T. The complex path gain $h_{u_m,v}(l)$ represents the l-th path with variance σ_l^2, and $\sum_{l=0}^{L-1} \sigma_l^2 = 1$. $\varepsilon_{u_m,v} = \Delta f_{u_m,v} T_s$ is the normalized frequency offset of the u-th user; $\Delta f_{u_m,v}$ is the frequency offset between the u_m-th receive antenna and v-th transmit antenna, and T_s is the OFDM symbol period. The usual assumption for precoding is that the multipath channel is sufficiently slowly fading, i.e., $h_{u_m,v}(l)$ and $\varepsilon_{u_m,v}$ do not change over several OFDM symbol intervals. $w_{u_m,v}(n)$ is a discrete-time additive white Gaussian noise (AWGN) sample. Theoretically, each transmit-receive antenna pair may have a different frequency offset. In practice, the difference among

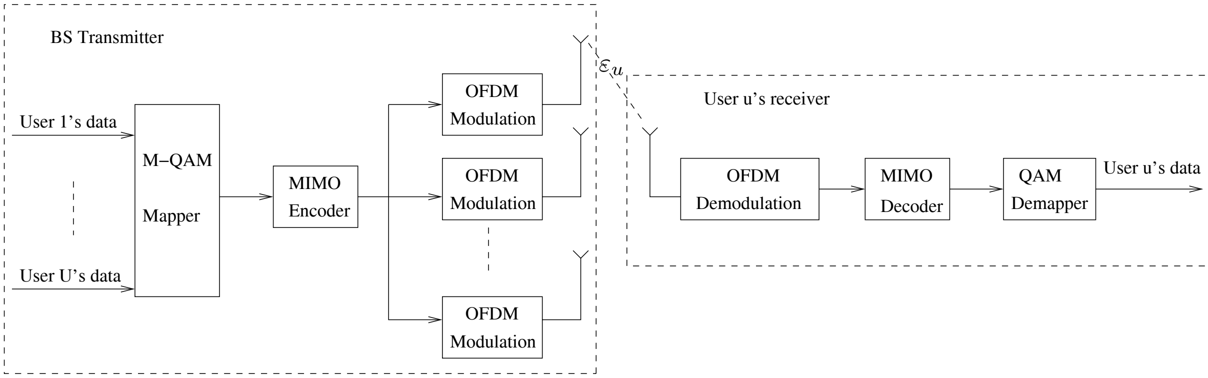

Fig. 1. Block diagram of a multiuser OFDM downlink.

these frequency offsets from different transmit antennas is not significant because even if collocated antennas do not share one RF oscillator, achieving frequency synchronism among collocated oscillators is relatively easy. The frequency offsets for different receive antennas may be different because the receive antennas may belong to different users. In this paper, we consider the case of U different frequency offsets, i.e., for the u-th user, $\varepsilon_{u_m,1} = \varepsilon_{u_m,2} = \cdots = \varepsilon_{u_m,M_T} = \varepsilon_u$, and $\varepsilon_u \neq \varepsilon_{u'}$, $\forall\ u \neq u'$ [13]. A cyclic prefix, which is longer than the expected maximum excess delay, is inserted at the beginning of each time-domain OFDM symbol to prevent intersymbol interference (ISI).

After discarding the cyclic prefix, the demodulated signal is obtained by performing the FFT of $\mathbf{y}_{u_m}$ as

$$Y_{u_m}[k] = \underbrace{\sqrt{\frac{\eta}{M_T}} \sum_{v=1}^{M_T} S_u[0] H_{u_m,v}[k] X_v[k]}_{\text{desired signal}} + \underbrace{\sqrt{\frac{\eta}{M_T}} \sum_{v=1}^{M_T} \sum_{p=0,p\neq k}^{N-1} S_u[p-k] H_{u_m,v}[p] X_u[p]}_{\text{intercarrier interference}} + W_{u_m,v}[k] \tag{3}$$

for $k = 0, 1, \cdots, N-1$, where $W_{u_m,v}[k]$, $\forall\, k$ are independent and identically distributed (i.i.d.) AWGN samples with zero mean and variance σ_W^2; $S_u[p-k]$ is an ICI coefficient given by

$$S_u[p] = \frac{\sin \pi(\varepsilon_u + p)}{N \sin \frac{\pi}{N}(\varepsilon_u + p)} e^{\jmath\pi(1-\frac{1}{N})(\varepsilon_u+p)}, \tag{4}$$

for $p = 1-N, \ldots, 0, \ldots, N-1$; $H_{u_m,v}[k] = \sum_{l=0}^{L-1} h_{u_m,v}(l) e^{-\jmath\frac{2\pi}{N}lk}$. $H_{u_m,v}[k]$ are i.i.d. complex Gaussian random variables with zero mean and variance normalized to unity. For the u-th user, the $NM_u \times NM_T$ channel matrix is

$$\mathbf{G}_u = \mathbf{S}_u \mathbf{H}_u, \tag{5}$$

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

$$\mathbf{H}_u = \begin{bmatrix} \mathbf{H}_{1,1} & \cdots & \mathbf{H}_{1,M_T} \\ \vdots & \ddots & \vdots \\ \mathbf{H}_{M_u,1} & \cdots & \mathbf{H}_{M_u,M_T} \end{bmatrix}, \tag{6}$$

with elements being the $\{u_m, v\}$th channel gain matrix $\mathbf{H}_{u_m,v}$ for the N orthogonal subchannels,

$$\mathbf{H}_{u_m,v} = \text{diag}\left[H_{u_m,v}[0] \quad H_{u_m,v}[1] \quad \ldots \quad H_{u_m,v}[N-1]\right]. \tag{7}$$

And the $NM_u \times NM_u$ ICI matrix is

$$\mathbf{S}_u = \text{diag}\left[\mathbf{S}_1 \quad \cdots \quad \mathbf{S}_{M_u}\right] \tag{8}$$

with

$$\mathbf{S}_{u_m} = \begin{bmatrix} S_{u_m}[0] & S_{u_m}[1] & \ldots & S_{u_m}[N-1] \\ S_{u_m}[-1] & S_{u_m}[0] & \ldots & S_{u_m}[N-2] \\ \vdots & & \ddots & \vdots \\ S_{u_m}[1-N] & S_{u_m}[2-N] & \ldots & S_{u_m}[0] \end{bmatrix}. \tag{9}$$

Each sub-matrix $\mathbf{S}_{u_m}$ is an $N \times N$ unitary matrix [12], [13]. $\mathbf{S}_u$ is therefore also a unitary matrix.

Stacking all the users, the $NM_R \times NM_T$ overall channel matrix $\mathbf{G}$ therefore is

$$\mathbf{G} = \mathbf{S}\mathbf{H}, \tag{10}$$

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

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_1 \\ \vdots \\ \mathbf{H}_U \end{bmatrix}, \tag{11}$$

and the $NM_R \times NM_R$ ICI matrix is

$$\mathbf{S} = \text{diag}\left[\mathbf{S}_1 \quad \cdots \quad \mathbf{S}_U\right]. \tag{12}$$

Similar to $\mathbf{S}_u$, $\mathbf{S}$ is a unitary matrix. For a MIMO OFDM system, the signal on the u_m-th receive antenna is $\mathbf{Y}_{u_m} = \left[Y_{u_m}[0] \ldots Y_{u_m}[N-1]\right]^T$. The received signal vector $\mathbf{Y} = \left[\mathbf{Y}_1^T \ldots \mathbf{Y}_{M_R}^T\right]^T$ can be represented as

$$\mathbf{Y} = \sqrt{\frac{\eta}{M_T}} \mathbf{G}\mathbf{X} + \mathbf{W} = \sqrt{\frac{\eta}{M_T}} \mathbf{S}\mathbf{H}\mathbf{X} + \mathbf{W}; \tag{13}$$

the noise matrix $\mathbf{W} = \left[\mathbf{W}_1^T \ldots \mathbf{W}_{M_R}^T\right]^T$ and $\mathbf{W}_{u_m}$ is the noise vector on the u_m-th receive antenna, $u_m = 1, \ldots, M_R$. The transmitted vector $\mathbf{X} = \left[\mathbf{X}_1^T \ldots \mathbf{X}_{M_T}^T\right]^T$.

A. Orthogonal Space-Time Block Coded OFDM

OSTBC [16], [17] is an effective space-time coding scheme to realize transmit diversity. Specifically, the Alamouti code [16] for two transmit antennas has been adopted in several 3G standards and 3GPP specifications, including WCDMA and cdma2000 [2], [3].

A $T \times M_T$ code matrix $\mathbf{C}$ for OSTBC satisfies

$$\mathbf{C}^H\mathbf{C} = \left(\sum_{t=1}^{P} |c_t|^2\right)\mathbf{I}_{M_T}. \quad (14)$$

The code rate in this case is $R = P/T$, where P represents the number of symbols transmitted over the T time slots. The full-rate codes transmit an average of one symbol per symbol period, i.e., $R = 1$. OSTBC can be directly applied to OFDM at a subcarrier level to offer full spatial diversity gain, if there is no correlation between transmit antennas. For example, the full-rate transmission matrix of Alamouti-coded OFDM is to transmit $\mathbf{C} = \begin{pmatrix} X_1[k] & -X_2^*[k] \\ X_2[k] & X_1^*[k] \end{pmatrix}$ onto the subcarrier k, i.e., $X_1[k]$ and $X_2[k]$ are transmitted over the 1-st and 2-nd antenna at the first time slot, respectively; the $-X_2^*[k]$ and $X_1^*[k]$ are transmitted in the following slots.

III. Limited-Feedback Precoding for MIMO OFDM with Frequency Offsets

Previous work in limited-feedback precoding has focused on the ideal case of single-user MIMO OFDM without frequency offsets [14], [18]. In this case, the $NM_R \times NM_T$ overall channel matrix $\mathbf{G}$ is only determined by channel gains and becomes

$$\mathbf{G}' = \text{diag}\left[\mathbf{H}[0] \quad \dots \quad \mathbf{H}[N-1]\right]. \quad (15)$$

The $M_R \times M_T$ sub-matrix $\mathbf{H}[k]$ is the channel matrix on the subcarrier k with i.i.d. $\mathcal{CN}(0,1)$ entries. Precoding can thus be designed on a subcarrier basis using the limited-feedback approach for flat-fading MIMO systems [6], [7]. However, with non-zero frequency offsets, the overall channel matrix is dependent on both frequency offset and channel gains. Each user only knows the frequency offset and channel gains affecting its receiver. The limited-feedback precoding design in [6], [7], [14], [18] cannot be directly applied.

Before developing our limited-feedback precoder for both SM and OSTBC MIMO OFDM systems with frequency offset, we first investigate the relationship between $\mathbf{G}$ in (10) and $\mathbf{G}'$ in (15). The channel gain matrix $\mathbf{H}$ can be permuted into $\mathbf{G}'$:

$$\mathbf{H} = \mathbf{Q}_1\mathbf{G}'\mathbf{Q}_2, \quad (16)$$

where $\mathbf{Q}_1$ is an $NM_R \times NM_R$ unitary permutation matrix and $\mathbf{Q}_2$ is an $NM_T \times NM_T$ unitary permutation matrix. The singular value decomposition (SVD) of $\mathbf{G}'$ is given by

$$\mathbf{G}' = \mathbf{U}'\mathbf{\Gamma}'\mathbf{V}'^H, \quad (17)$$

where $\mathbf{U}'$ is an $NM_R \times NM_T$ unitary matrix and $\mathbf{U}'^H\mathbf{U} = \mathbf{I}_{NM_T}$; $\mathbf{V}'$ is an $NM_T \times NM_T$ unitary matrix. Since $\mathbf{G}'$ is a block-diagonal matrix, $\mathbf{V}'$ can be constructed by $\mathbf{V}'[k]$, which is generated from the SVD of $\mathbf{H}[k]$ in (15). Therefore, $\mathbf{V}'$ is also a block-diagonal matrix. The singular value matrix $\mathbf{\Gamma}'$ is an $NM_T \times NM_T$ diagonal matrix with real, positive entries γ_n, $n = 1, \dots, NM_T$, in descending order $\gamma_1' \geq \gamma_2' \geq \dots \geq \gamma_{NM_T}' > 0$. The SVD of the overall channel matrix becomes

$$\mathbf{G} = \mathbf{S}\mathbf{H} = \mathbf{S}\mathbf{Q}_1\mathbf{G}'\mathbf{Q}_2 = \mathbf{S}\mathbf{Q}_1\mathbf{U}'\mathbf{\Gamma}'\mathbf{V}'^H\mathbf{Q}_2 = \mathbf{U}\mathbf{\Gamma}\mathbf{V}^H. \quad (18)$$

Since $\mathbf{Q}_1$ and $\mathbf{Q}_2$ are unitary matrices, we have $\mathbf{U} = \mathbf{S}\mathbf{Q}_1\mathbf{U}'$, $\mathbf{\Gamma} = \mathbf{\Gamma}'$ and $\mathbf{V} = \mathbf{Q}_2^H\mathbf{V}'$. Similarly, for the u-th user, the SVD of the channel matrix is

$$\mathbf{G}_u = \mathbf{S}_u\mathbf{Q}_1\mathbf{G}_u'\mathbf{Q}_2 = \mathbf{U}_u\mathbf{\Gamma}_u\mathbf{V}_u^H, \quad (19)$$

where $\mathbf{U}_u = \mathbf{S}_u\mathbf{Q}_1\mathbf{U}_u'$, $\mathbf{\Gamma}_u = \mathbf{\Gamma}_u'$ and $\mathbf{V}_u = \mathbf{Q}_2^H\mathbf{V}_u'$.

For each user, the precoding matrix is an $NM_T \times NM_C$ matrix, i.e., on each subcarrier the incoming data streams are multiplexed into M_C streams and sent over M_T transmit antennas, $M_T \geq M_C$. For convenience of detection, we need $M_u \geq M_C$. In the single-user case, we assume $M_R \geq M_C$. We construct a finite codebook of matrices and choose the precoding 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 back to the transmitter. We analyze the matrix selection criteria, and propose the codebook design algorithms.

A. Precoding Matrix Selection Criteria for Linear Receivers

Here, we analyze the matrix selection criteria for multiuser SM MIMO OFDM with a zero-forcing (ZF) receiver. We assume that the precoding matrix of the user u $\breve{\mathbf{B}}_u = \mathcal{P}(\mathbf{G}_u)$, where $\mathcal{P}(\mathbf{G}_u)$ is a mapping from the channel matrix $\mathbf{G}_u$ to the codebook $\mathcal{B} = \{\breve{\mathbf{B}}_1, \dots, \breve{\mathbf{B}}_K\}$, optimizing the precoding matrix based on some performance criterion; K is the size of the codebook. All users share the same codebook. Once the optimal $\breve{\mathbf{B}}_{\text{u,opt}}$ is chosen, the ZF receiver applies a $NM_C \times NM_u$ matrix $\mathbf{D}_u = \left[\mathbf{G}_u\breve{\mathbf{B}}_{\text{u,opt}}\right]^\dagger$. For simplicity, in the following analysis, the user index u is omitted.

1) Minimum Mean Squared Error (MMSE) Criterion: The MSE for precoding matrix $\breve{\mathbf{B}}$ can be expressed as [6], [19]

$$\begin{aligned}\overline{\text{MSE}}(\breve{\mathbf{B}}) &= E_s\left(\mathbf{I}_{NM_C} + E_s\breve{\mathbf{B}}^H\mathbf{G}^H\mathbf{R}_{WW}^{-1}\mathbf{G}\breve{\mathbf{B}}\right)^{-1} \\ &= E_s\left(\mathbf{I}_{NM_C} + \frac{E_s}{\sigma_W^2}\breve{\mathbf{B}}^H\mathbf{Q}_2^H\mathbf{G}'^H\mathbf{G}'\mathbf{Q}_2\breve{\mathbf{B}}\right)^{-1},\end{aligned} \quad (20)$$

where $\mathbf{R}_{WW} = \text{E}\left[\mathbf{W}\mathbf{W}^H\right]$ is the noise covariance matrix; E_s is the average power of the transmitted symbols. We use (20) to select $\breve{\mathbf{B}}$ from $\mathcal{B}$ according to minimize the MSE

$$\mathcal{P}(\mathbf{G}) = \arg\min_{\breve{\mathbf{B}}_i \in \mathcal{B}} \text{tr}\left[\overline{\text{MSE}}(\breve{\mathbf{B}}_i)\right]. \quad (21)$$

From the relationship between $\mathbf{G}$ and $\mathbf{G}'$ in (18), we can expect the precoding matrix for $\mathbf{G}$ is related to that for $\mathbf{G}'$. Let $\breve{\mathbf{B}}' = \mathbf{Q}_2\breve{\mathbf{B}}$ be the precoding matrix for $\mathbf{G}'$. In [18], [19], for $\mathbf{H}[k]$ in $\mathbf{G}'$ (15), the optimal precoding matrix on the subcarrier k is $\breve{\mathbf{B}}_{\text{opt}}'[k] = \mathbf{V}_C'[k]$, where $\mathbf{V}_C'[k]$ is formed from the first M_C columns of the right matrix $\mathbf{V}'[k]$ yielded by the SVD of the $\mathbf{H}[k]$. Since $\mathbf{G}'$ is a block-diagonal matrix, $\breve{\mathbf{B}}_{\text{opt}}' = \mathbf{V}_C' = \text{diag}\left[\mathbf{V}_C'[0]\dots\mathbf{V}_C'[N-1]\right]$ is also an $NM_T \times NM_C$ block-diagonal matrix, and the desired overall optimal precoding matrix is $\breve{\mathbf{B}} = \mathbf{Q}_2^H\breve{\mathbf{B}}'$.

2) Maximum Singular Value (MSV) Criterion: Using a selection criterion based on the minimum SNR is difficult to implement since it requires the computation of the SNR for every OFDM symbol interval. With pre-equalization by $\breve{\mathbf{B}}$, the system experiences an $NM_R \times NM_C$ effective channel matrix $\mathbf{G}\breve{\mathbf{B}}$. Since the minimum SNR for a ZF receiver is lower bounded by the minimum singular value of the effective channel [6], [19], we can select $\breve{\mathbf{B}}$ such that the minimum singular value is as large as possible. The optimization problem therefore is

$$\mathcal{P}(\mathbf{G}) = \arg\max_{\breve{\mathbf{B}}_i \in \mathcal{B}} \gamma_{\min}\left(E_s \breve{\mathbf{B}}^H \mathbf{G}^H \mathbf{R}_{WW}^{-1} \mathbf{G}\breve{\mathbf{B}}\right). \quad (22)$$

Similarly, the solution is given by $\breve{\mathbf{B}}_{\text{opt}} = \mathbf{Q}_2^H \mathbf{V}_C'$.

3) Maximum Mutual Information (MMI) Criterion: The mutual information under an average transmit power constraint is given by [20]

$$C(\breve{\mathbf{B}}) = \frac{1}{N}\log_2\left[\det(\mathbf{I}_{NM_C} + E_s \breve{\mathbf{B}}^H \mathbf{G}^H \mathbf{R}_{WW}^{-1} \mathbf{G}\breve{\mathbf{B}})\right]. \quad (23)$$

To maximize the mutual information (23), we select the feedback matrix $\breve{\mathbf{B}}$ such that

$$\mathcal{P}(\mathbf{G}) = \arg\max_{\breve{\mathbf{B}}_i \in \mathcal{B}} C(\breve{\mathbf{B}}_i). \quad (24)$$

This optimization problem is equivalent to

$$\mathcal{P}(\mathbf{G}) = \arg\min_{\breve{\mathbf{B}}_i \in \mathcal{B}} \det(\overline{\text{MSE}}(\breve{\mathbf{B}}_i)). \quad (25)$$

Similar to the MMSE criterion (21), the solution is $\breve{\mathbf{B}}_{\text{opt}} = \mathbf{Q}_2^H \mathbf{V}_C'$.

B. Matrix Selection Criterion for Maximum-Likelihood Receiver

Now we consider the precoding matrix selection criterion for a multiuser OSTBC MIMO OFDM. Each user has an ML receiver. The user's index u is omitted here for simplicity.

To minimize the system BER given the channel matrix $\mathbf{G}$, we need to maximize the Frobenius norm of the effective channel $\mathbf{G}\breve{\mathbf{B}}$ [7]. We choose the precoding matrix from the codebook $\mathcal{B}$ according to

$$\mathcal{P}(\mathbf{G}) = \arg\max_{\breve{\mathbf{B}}_i \in \mathcal{B}} \|\mathbf{G}\breve{\mathbf{B}}_i\|_F = \|\mathbf{S}\mathbf{Q}_1\mathbf{G}'\mathbf{Q}_2\breve{\mathbf{B}}_i\|_F. \quad (26)$$

This selection criterion can be easily implemented by performing a matrix multiplication and computing a Frobenius norm for each of the K codebook matrices. With the SVD of $\mathbf{G}'$ in (17), the optimal $\breve{\mathbf{B}}_{\text{opt}}$ is given by $\breve{\mathbf{B}}_{\text{opt}} = \mathbf{Q}_2^H \mathbf{V}_C'$. Each user performs ML decoding on the effective channel $\mathbf{G}\breve{\mathbf{B}}_{\text{opt}}$.

Similarly as proven in [7], limited-feedback precoding in OSTBC systems achieves full-diversity order. Before transmission, an $M_C \times T$ antenna OSTBC matrix is passed through the pre-processing filter and transmitted over M_T antennas ($M_T \geq M_C$). Our precoding can thus be used in OSTBC MIMO OFDM with an arbitrary number of transmit antennas.

C. Limited-Feedback Precoding Design

To find the optimal precoding matrix $\breve{\mathbf{B}}_{\text{u,opt}} = \mathbf{Q}_2^H \mathbf{V}_{u,C}'$, we need to look for $\mathbf{V}_{u,C}'$. Since $\mathbf{V}_{u,C}'$ is a block diagonal matrix, we can build up a codebook $\mathcal{B} = \{\breve{\mathbf{B}}_1, \ldots, \breve{\mathbf{B}}_K\}$ to find $\mathbf{B}_{\text{u,opt}}'[k]$ at a subcarrier level. Therefore, in our codebook $\mathcal{B}_u$, each complex matrix $\breve{\mathbf{B}}_i$ has a size of $M_T \times M_C$, rather than $NM_T \times NM_C$. The optimal $\breve{\mathbf{B}}_{\text{u,opt}}'[k]$ is chosen from the codebook $\mathcal{B}$ at the u-th user's receiver according to the current channel conditions. After the optimal $\mathbf{V}_{u,C}'[k]$ for all subcarriers are obtained, the optimal precoding matrix $\breve{\mathbf{B}}_{\text{u,opt}} = \mathbf{Q}_2^H \text{diag}\left[\mathbf{B}_{\text{u,opt}}'[0] \ldots \mathbf{B}_{\text{u,opt}}'[N-1]\right]$ can be constructed. The overall precoding matrix is built up as follows:

$$\breve{\mathbf{B}}_{\text{opt}} = \left[\breve{\mathbf{B}}_{\text{1,opt}}, \ldots, \breve{\mathbf{B}}_{\text{U,opt}}\right]. \quad (27)$$

The effective overall channel therefore becomes

$$\mathbf{G}\breve{\mathbf{B}}_{\text{opt}} = \begin{bmatrix} \mathbf{G}_1\breve{\mathbf{B}}_{\text{1,opt}} & \cdots & \mathbf{G}_1\breve{\mathbf{B}}_{\text{U,opt}} \\ \vdots & \ddots & \vdots \\ \mathbf{G}_U\breve{\mathbf{B}}_{\text{1,opt}} & \cdots & \mathbf{G}_U\breve{\mathbf{B}}_{\text{U,opt}} \end{bmatrix}. \quad (28)$$

Since the codebook is known at both the transmitter and each user, only the index of the selected precoding matrix needs to be fed back to the BS transmiter. The BS transmitter broadcasts every user's precoding index such that each user will also knows other users' precoding matrix. At the u-th user's receiver, the effective channel is $\mathbf{G}_u \sum_{u=1}^{U} \breve{\mathbf{B}}_{\text{u,opt}}$. Since the index of $\breve{\mathbf{B}}_{\text{u,opt}}$, $\forall\, u$, is available at each user's receiver, the transmitted signal can be detected. For the single-user case, the overall $NM_T \times NM_C$ precoding matrix is $\breve{\mathbf{B}}_{\text{opt}} = \mathbf{Q}_2^H \text{diag}\left[\mathbf{B}_{\text{opt}}'[0] \ldots \mathbf{B}_{\text{opt}}'[N-1]\right]$. Zero-forcing detection can be used at the receiver with the assumption of $M_R \geq M_C$.

For each user, the total $N\lceil \log_2 K \rceil$ feedback bits are required for a codebook with K precoding matrices. When a better performance is required, a larger codebook can be constructed, i.e., more bits can be sent to the transmitter. To reduce the total amount of the feedback information, we can exploit the correlation of precoding matrices on adjacent subcarriers. The significant correlations between adjacent subcarriers lead to substantial correlation between the precoders corresponding to neighboring subcarriers. The neighboring subcarriers can therefore be combined into a group and use the precoding matrix corresponding to the center subcarrier for all the subcarriers in the group. If the N subcarriers are divided into N_b groups, each group includes N/N_b subcarriers. Since the subcarriers near the group boundary may experience performance degradation, an interpolation scheme can be used to improve the performance due to grouping. Interpolation introduces a unitary matrix $\mathbf{Q}_I$ which takes the unitary-invariance of the optimal precoding matrix into consideration. The optimal $\mathbf{Q}_I$ can be determined by the same criteria used for precoding selection, and a codebook $\mathcal{Q}$ can be built up for optimal interpolation matrix selection. Naturally, a larger size of the codebook $\mathcal{Q}$ will lead to a better BER performance. The details of interpolation schemes can be found in [14], [18].

In the following subsections, we develop the matrix selection criteria for both SM and OSTBC MIMO OFDM systems, and propose the codebook design schemes.

D. Codebook Design

We now consider the construction of the precoding codebook $\mathcal{B} = \{\breve{\mathbf{B}}_1, \ldots, \breve{\mathbf{B}}_K\}$. We first show that frequency offsets have no impact on the codebook design, which makes precoding on per-subcarrier basis possible. Next, we generalize the codebook design criterion which is valid only for flat-fading MIMO channels in [5]–[7], [15], to OFDM systems with frequency offset. The user index is omitted here.

1) Impact of the ICI Matrix: Since $\mathbf{S}$ is a unitary matrix, the singular value matrix and the right matrix in (18) are actually generated from the eigenvalue decomposition (EVD) of the channel gain matrix $\mathbf{H}^H\mathbf{H}$:

$$\begin{aligned} \mathbf{G}^H\mathbf{G} = \mathbf{H}^H\mathbf{H} &= \mathbf{Q}_2^H\mathbf{G}'^H\mathbf{Q}_1^H\mathbf{Q}_1\mathbf{G}'\mathbf{Q}_2 \\ &= \mathbf{Q}_2^H\mathbf{G}'^H\mathbf{G}'\mathbf{Q}_2 = \mathbf{V}\mathbf{\Gamma}^2\mathbf{V}^H. \end{aligned} \quad (29)$$

We thus find

Lemma 1: In an OFDM system, if the values of frequency offsets only change on different receive antennas, $\breve{\mathbf{B}}'_{\text{opt}}[k] = \mathbf{V}'_C[k]$ is uniformly distributed on the set $\Theta(M_T, M_C)$, i.e., the frequency offset does not have impact on the codebook design compared to an OFDM system without frequency offset.

Proof: If there is no frequency offset, the overall channel matrix is reduced to a block diagonal matrix $\mathbf{G}'$, and $\mathbf{V}' = \mathbf{Q}_2\mathbf{V} = \text{diag}\left[\mathbf{V}'[0]\ldots\mathbf{V}'[N-1]\right]$ is also a block diagonal matrix. Each subblock matrix is uniformly distributed on $\Theta(M_T, M_T)$ [21], where $\Theta(m, n)$ is the set of $m \times n$ matrices with n orthonormal columns. As in [6], the optimal precoding matrix on the subcarrier k $\breve{\mathbf{B}}'_{\text{opt}}[k] = \mathbf{V}'_C[k]$ is also uniformly distributed on the set $\Theta(M_T, M_C)$. $\square$

2) Codebook Design Criteria: Since $\breve{\mathbf{B}}'_{\text{opt}}[k]$ is uniformly distributed over $\Theta(M_T, M_C)$, we should design the codebook matrices in the set of $\Theta(M_T, M_C)$. The set of all possible column spaces of the matrices $\breve{\mathbf{B}}_i$, $\forall\ i$, in $\Theta(M_T, M_C)$ is a complex Grassmannian manifold $\Xi(M_T, M_C)$, in which $\Xi(m, n)$ is the set of n-dimensional subspaces in an m-dimensional vector space. Considering each codebook matrix generates a subspace, there is a set of subspaces yielded by the codebook matrices $\{\breve{\mathbf{B}}_1, \ldots, \breve{\mathbf{B}}_K\}$. The subspaces in the Grassmann manifold can be related by several different distances: The chordal distance between any two subspaces is defined as

$$\text{d}_{\text{chord}}\left(\breve{\mathbf{B}}_p,\, \breve{\mathbf{B}}_q\right) = \frac{1}{\sqrt{2}}\|\breve{\mathbf{B}}_p\breve{\mathbf{B}}_p^H - \breve{\mathbf{B}}_q\breve{\mathbf{B}}_q^H\|_F. \quad (30)$$

The projection two-norm distance between two subspaces is

$$\text{d}_{\text{norm}}\left(\breve{\mathbf{B}}_p,\, \breve{\mathbf{B}}_q\right) = \|\breve{\mathbf{B}}_p\breve{\mathbf{B}}_p^H - \breve{\mathbf{B}}_q\breve{\mathbf{B}}_q^H\|_2, \quad (31)$$

where $\|.\|_2$ is two-norm distance. The Fubini-Study distance between two subspaces is

$$\text{d}_{\text{FS}}\left(\breve{\mathbf{B}}_p,\, \breve{\mathbf{B}}_q\right) = \arccos|\det\left(\breve{\mathbf{B}}_p^H\breve{\mathbf{B}}_q\right)|. \quad (32)$$

The codebook can be designed according to a specific precoding selection criterion. This set of subspaces is a packing of subspaces in $\Xi(M_T, M_C)$. In OSTBC MIMO OFDM, a packing can be described by the its minimum chordal distance on the Grassmann manifold

$$\text{d}_{\text{min}} = \min_{1\le p<q\le K} \text{d}_{\text{chord}}\left(\breve{\mathbf{B}}_p,\ \breve{\mathbf{B}}_q\right). \quad (33)$$

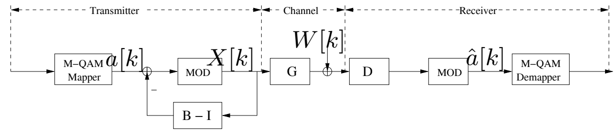

Fig. 2. Tomlinson-Harashima precoding in a MIMO OFDM link.

As in [15], we need to minimize

$$\rho = \max_{1\le p<q\le K} \|\mathbf{B}_p^\dagger\breve{\mathbf{B}}_q\|_F, \quad (34)$$

which is related to a packing in a complex Grassmannian space. Since minimizing ρ in (34) is equivalent to maximizing the minimum distance of (30), the codebook design criterion turns out to be the problem of Grassmannian subspace packing. We therefore choose the set of K subspaces in $\Xi(M_T, M_C)$ such that d_{min} is as large as possible. Each precoding matrix in the set $\breve{\mathcal{B}}$ is given as $\breve{\mathbf{B}}_i = \mathbf{\Phi}^{i-1}\breve{\mathbf{B}}_1$, where $\breve{\mathbf{B}}_1$ is formed by the first M_C columns of a unitary matrix in the set of $\Theta(M_T, M_C)$, and $\mathbf{\Phi}$ is an $M_T \times M_T$ diagonal unitary matrix and $\mathbf{\Phi}^K = \mathbf{I}$, i.e., $\mathbf{\Phi}$ is a K-th root of unity. We select the parameter set $\mathbf{\Psi} = \{\varphi_1, \ldots, \varphi_{M_T}\}$ to achieve

$$\min_{0\le\varphi_1,\ldots,\varphi_{M_T}\le K-1} \rho = \min_{0\le\varphi_1,\ldots,\varphi_{M_T}\le K-1}\ \max_{i=2,\ldots,K} \left\|\breve{\mathbf{B}}_1^\dagger\breve{\mathbf{B}}_i\right\|_F, \quad (35)$$

i.e., the parameter set $\mathbf{\Psi}$ of the diagonal entries of $\mathbf{\Phi}$ is

$$\mathbf{\Psi} = \arg\max \min_{1\le i\le K-1} \text{d}\left(\breve{\mathbf{B}}_1,\ \mathbf{\Phi}^{i-1}\breve{\mathbf{B}}_1\right). \quad (36)$$

Geometrically, this construction rotates an initial M_T-dimensional subspace using a K-th root of unity to form K different M_C-dimensional subspaces.

In SM MIMO OFDM, for MSV and MMSE selection criteria, the codebook can be designed by maximizing the minimum projection two-norm distance between any pair of codeword matrix column space. For capacity selection criterion, the codebook can be designed by maximizing the minimum Fubini-Study distance between any pair of codeword matrix column space [6]. Once the codebook is designed, the matrix on the k-th subcarrier $\breve{\mathbf{B}}[k] \in \mathcal{B}$ using the performance criteria in (21), (22) or (25) is chosen at the receiver. Its index is then delivered to the transmitter using only $\lceil\log_2 K\rceil$ bits.

E. Non-Linear Tomlinson-Harashima Precoding

Non-linear THP, originally proposed for temporal pre-equalization of dispersive channels [10], [11], has recently been extended to flat-fading (multiuser) MIMO systems to mitigate the inter-layer interference [22], [23], reduce ICI in MIMO OFDM systems [12], [13], and mitigate the impact of spatial correlations in OSTBC MIMO OFDM [24]. The non-linear property of THP reduces power efficiency loss in linear precoding, and the precoding structure avoids error propagation typical for the decision feedback equalization (DFE) [23].

The structure of a typical TH precoder in a MIMO OFDM link is shown in Fig. 2. The whole setup involves a receiver-based feedforward filter $\mathbf{D}$, a transmitter-based feedback matrix $\mathbf{B}$, and modulo devices at both the transmitter and the

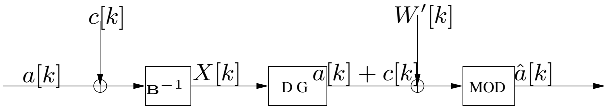

Fig. 3. Equivalent structure of Tomlinson-Harashima precoding.

receiver. The linear pre-distortion $\mathbf{B} = \breve{\mathbf{B}}^{\dagger}$ is equivalent to the feedback structure in Fig. 2. An equivalent structure of conventional THP is shown in Fig. 3. For better clarification, we name $\mathbf{B}$ as the feedback matrix of THP in the rest of the paper, while $\breve{\mathbf{B}}$ as the precoding matrix.

Specific design targets for linear precoding are considered in Section III. A to improve the overall system performance. If the input sequence $a[k]$ is a sequence of i.i.d. symbols with variance E_a, the output of the modulo arithmetic feedback structure is also a sequence of i.i.d. random variables, and the real and imaginary parts are independent, i.e., we can assume $X[k]$ with variance $E_s = \mathrm{E}\left[|X[k]|^2\right], \forall\ k$. E_s is slightly larger than E_a, but the difference between E_s and E_a becomes negligible as the constellation size increases [23]. We can thus apply the linear precoding design criteria in in (21), (22) and (23) to our non-linear precoding selection criteria. Similarly, the feedforward matrix $\mathbf{D} = \left[\mathbf{G}\breve{\mathbf{B}}\right]^{\dagger}$.

Given the data carrying symbols $a[k] \in \mathcal{A}$ (the initial M-ary constellation), the transmitted symbols $X[k]$ are successively calculated via the modulo arithmetic feedback filter. The non-linear modulo $2\sqrt{M}$ reduction, which is applied separately to the real and imaginary parts of the input, can be defined as $\mathrm{MOD}_{2\sqrt{M}}(a) = a + c$, where c is the unique integer multiple of $2\sqrt{M}$ for which $\Re\left\{\mathrm{MOD}_{2\sqrt{M}}(a)\right\}$, $\Im\left\{\mathrm{MOD}_{2\sqrt{M}}(a)\right\} \in \left(-\sqrt{M}, \sqrt{M}\right]$, such that a transmit signal has smaller power. In other words, instead of sending the data symbols $a[k]$ to the linear pre-equalization $\breve{\mathbf{B}}$, the symbols $a[k] + c[k]$ are passed through $\breve{\mathbf{B}}$. At the receiver, a slicer, which applies the same modulo operation as that at the transmitter, is used. The unique estimates of the data symbols can be generated by the modulo operation that takes the periodic extension into account and yields estimates $\hat{c}$. Since the feedback structure is moved to transmitter and the modulo operation is memoryless, no error propagation can occur. Consequently, after passing through the filter $\mathbf{D}$ and discarding the modulo congruence at the receiver, the data symbols $a[k]$ are only corrupted by an additive noise and become $Y[k] = a[k] + W'[k]$. Note that $W'[k]$, the k-th entry of the filtered noise vector $\mathbf{W}' = \mathbf{D}\mathbf{W}$, has individual variance $\sigma^2_{W'_k}$.

IV. Limited-Feedback Precoding over Spatially Correlated Channels

In the section, we focus on OFDM in spatially correlated fading environments. Insufficient scattering around antennas of the BS transmitter and/or the receiver makes the channels spatially correlated. The precoding matrix is restrained to lie in the codebook $\mathcal{B} = \{\breve{\mathbf{B}}_1, \breve{\mathbf{B}}_2, \ldots, \breve{\mathbf{B}}_K\}$. At the receiver, we choose the matrix as a function of the channel conditions to minimize the error rate or maximize the capacity.

A. Correlated MIMO Channels

The correlated channel model builds on previous work reported in [25] and [26]. We assume a uniform linear array (ULA) at the transmitter and the receiver with identical antenna elements, and each tap has the identical fading correlation matrices. As in [25], the entries of the full-rank transmit-antenna correlation $\mathbf{R}_T$ and the receive-antenna correlation matrix $\mathbf{R}_R$ are

$$\begin{aligned}\mathbf{R}_T(m,n) &= \mathcal{J}_0\left(2\pi|m-n|\Delta\frac{d_T}{\lambda}\right)\\ \mathbf{R}_R(m,n) &= \mathcal{J}_0\left(2\pi|m-n|\frac{d_R}{\lambda}\right),\end{aligned} \tag{37}$$

where $\mathcal{J}_0$ is zero-order Bessel function of the first kind. $\lambda = c/f_c$ is the wavelength with center frequency f_c. Δ is the angle of arrival spread, and d_T and d_R are the inter-element distance of the transmit and receive antenna arrays, respectively. For MIMO OFDM systems, the frequency-domain channel gain matrix at the k-th subcarrier is given by

$$\mathbf{H}[k] = \sum_{l=0}^{L-1} \mathbf{r}_R \mathbf{H}_{w,l} \mathbf{r}_T e^{-\jmath\frac{2\pi}{N}lk} = \mathbf{r}_R \mathbf{H}_w[k] \mathbf{r}_T, \tag{38}$$

where the entries in $\mathbf{H}_w[k] = \sum_{l=0}^{L-1} \mathbf{H}_{w,l} e^{-\jmath\frac{2\pi}{N}lk}$ are i.i.d. complex Gaussian random variables with zero mean and unity variance. $\mathbf{r}_R = \mathbf{R}_R^{1/2}$ and $\mathbf{r}_T = \mathbf{R}_T^{1/2}$.

B. Limited-Feedback Precoding in Spatially Correlated MIMO Channels

No correlation is assumed among different users. The receive antennas in this subsection means the antennas of the same user. The user index is omitted.

1) Receive Antenna Correlations Only: We first consider the case that each pair of transmit antennas has sufficient distance, i.e., only the effect of receive correlations needs to be analyzed. The channel is given by

$$\mathbf{G}_R = \mathbf{S}\mathbf{Q}_1\mathbf{G}'_R\mathbf{Q}_2, \tag{39}$$

where $\mathbf{G}'_R = \left(\mathbf{I}_N \otimes \mathbf{r}_R\right)\mathbf{G}'$ is a block diagonal matrix, and $\mathbf{G}'$ is given by (15). The k-th subblock on the main diagonal of $\mathbf{G}'_R$ is $\mathbf{H}_{Rc}[k] = \mathbf{r}_R\mathbf{H}_w[k]$. We thus can still design precoding on a subcarrier basis. The SVD of $\mathbf{H}_{Rc}[k]$ is $\mathbf{H}_{Rc}[k] = \mathbf{U}_{Rc}[k]\mathbf{\Gamma}_{Rc}[k]\mathbf{V}^H_{Rc}[k]$ with singular values of $\gamma_{Rc,1}[k] \geq \ldots \geq \gamma_{Rc,M_T}[k]$. The following lemma describes the distribution of $\mathbf{V}_{Rc}[k]$.

Lemma 2: $\mathbf{V}_{Rc}[k]$ is a uniformly distributed unitary matrix over $\Theta(M_T, M_C)$ and is independent of $\mathbf{\Gamma}_{Rc}[k]$, if the channel gain matrix $\mathbf{H}_w[k]$ (11) has i.i.d. $\mathcal{CN}(0,1)$ entries.

Proof: The entries $H_{u,v}[k]$ of the channel gain matrix $\mathbf{H}_w[k]$ are i.i.d. complex Gaussian random variables with zero mean and unity variance (see details in Section II). The distribution of an uniformly distributed unitary matrix is unchanged when the matrix is multiplied by any deterministic unitary matrix. $\mathbf{H}_{Rc}[k]$ is thus a right-rotationally invariant random matrix, and $\mathbf{V}_{Rc}[k]$ is thus uniformly distributed on $\Theta(M_T, M_T)$ [21], i.e., $\mathbf{\Omega}\mathbf{V}_{Rc}[k]$ is also uniformly distributed if $\mathbf{\Omega} \in \Theta(M_T, M_T)$ [27]. □

Thus, the matrix which is formed by the first M_C columns of $\mathbf{V}_{Rc}[k]$ is uniformly distributed on the set $\Theta(M_T, M_C)$.

The codebook design criterion in Section III is also available for the case that only receive antenna array is correlated. We can choose the feedback matrix from the codebook $\mathcal{B}$ by (21) or (23) for SM MIMO OFDM and (26) for OSTBC MIMO OFDM.

2) Both Receive and Transmit Antenna Correlations: This subsection considers the fully correlated channel model (40). Similarly the overall channel matrix $\mathbf{G}$ can be given by

$$\mathbf{G} = \mathbf{S}\mathbf{Q}_1\mathbf{G}'\mathbf{Q}_2, \tag{40}$$

where the $NM_R \times NM_T$ spatially-correlated channel gain matrix is $\mathbf{G}' = \text{diag}\left[\mathbf{H}[0]\ldots\mathbf{H}[N-1]\right]$ with $\mathbf{H}[k]$ is given in (38). We first need to find the optimal matrix $\mathbf{B}'_{\text{opt}}[k]$ for $\mathbf{H}[k]$, and the desired overfall optimal precoding matrix is $\mathbf{B}_{\text{opt}} = \mathbf{Q}_2^H\mathbf{B}'_{\text{opt}}$.

Intuitively, we want to choose a precoding matrix with the effective channel $\mathbf{H}[k]\breve{\mathbf{B}}'[k]$ to provide a performance approaching to that given by $\mathbf{H}[k]\breve{\mathbf{B}}'_{\text{opt}}[k]$. Since improvement of probability of error or capacity is relative to maximize $\|\mathbf{H}[k]\breve{\mathbf{B}}'[k]\|_F$, we will attempt to minimize

$$\vartheta = \text{E}\left[\|\mathbf{H}[k]\breve{\mathbf{B}}'_{\text{opt}}[k]\|_F^2 - \|\mathbf{H}[k]\breve{\mathbf{B}}'[k]\|_F^2\right]. \tag{41}$$

Eq. (41) can be bounded as

$$\begin{aligned}\vartheta &= \text{E}\left[\|\mathbf{H}[k]\breve{\mathbf{B}}'_{\text{opt}}[k]\|_F^2 - \|\mathbf{G}_R[k]\mathbf{r}_T\breve{\mathbf{B}}'[k]\|_F^2\right]\\ &\le \text{E}\left[\|\mathbf{H}[k]\breve{\mathbf{B}}'_{\text{opt}}[k]\|_F^2 - \gamma_{Rc,1}^2\|\breve{\mathbf{B}}_{Rc,\text{opt}}^H[k]\mathbf{r}_T\breve{\mathbf{B}}'[k]\|_F^2\right],\end{aligned} \tag{42}$$

where $\breve{\mathbf{B}}_{Rc,\text{opt}}[k]$ is the $M_T \times M_C$ optimal precoding matrix of $\mathbf{H}_{Rc}[k]$ in $\mathbf{G}'_R$ (39). Since $\mathbf{V}_{Rc}[k]$ is a uniformly distributed unitary matrix over $\Theta(M_T, M_C)$ and independent of the singular value matrix, the distribution of (42) only has one term which depends on the codebook. As a consequence, it is sufficient to only consider $\mathbf{r}_T^H\breve{\mathbf{B}}_{Rc,\text{opt}}[k]$, which is a subspace correlation as it only depends on the column space of $\breve{\mathbf{B}}_{Rc,\text{opt}}[k]$, $\breve{\mathbf{B}}_{Rc,\text{opt}}[k] \in \Theta(M_T, M_C)$. All column spaces in $\Theta(M_T, M_C)$ generate the complex Grassmannian manifold $\Xi(M_T, M_C)$. As shown in Section III, the Grassmannian subspace packing problem in uncorrelated channels can be described by designing a codebook $\mathbb{B} = \{\mathbf{b}_1, .., \mathbf{b}_K\}$ to maximize the minimum distance

$$\text{d}_{\min} = \min_{1\le m\le n\le K} \text{d}(\mathbf{b}_m, \mathbf{b}_n). \tag{43}$$

We obtain our codebook in correlated channels from multiplying each element of $\mathbb{B}$ by $\mathbf{r}_T$. Therefore, we design the codebook $\mathcal{B}$ by picking $\{\mathbf{b}_1, .., \mathbf{b}_K\}$ that maximize (43) and normalize the codebook to meet the average transmit power constraint

$$\breve{\mathbf{B}}_i = \frac{\mathbf{r}_T^H\mathbf{b}_i}{\|\mathbf{r}_T^H\mathbf{b}_i\|}. \tag{44}$$

This design criterion adapts the precoder to the correlated-channel conditions. The transmit spatial correlation matrix $\mathbf{R}_T$ can be estimated at the transmitter so that no feedback for the correlation matrix is needed.

V. Simulation Results

We run simulations to verify the BER performance of our proposed precoders. The codebook is known a priori at both the transmitter and users' receivers. A MIMO 4-QAM-OFDM system with 64 subcarriers on a 6-tap Rayleigh fading channel is considered. The vehicular B channel specified by ITU-R M. 1225 [28], is used where the channel tap gains are zero-mean complex Gaussian random processes with variances of -4.9 dB, -2.4 dB, -15.2 dB, -12.4 dB, -27.6 dB, and -18.4 dB relative to the total power gain. The excess delays of the channel taps are 0, 300 ns, 8900 ns, 12900 ns, 17100 ns, and 20000 ns, respectively. For many wireless systems, the multipath channels fade slowly in comparison to the symbol or block duration. In this work we have assumed channel gains to be constant over several OFDM symbol intervals. Each user has M_u receive antennas and different frequency offsets. M_R different values of normalized frequency offsets are assumed to be uniformly distributed in two intervals $\mathbb{I} = (0, 0.1]$ and $\mathbb{II} = (0.1, 0.3]$.

A. Spatially Multiplexed OFDM

This subsection considers spatially uncorrelated MIMO channels.

1) Perfect CSI at the Receiver: Here, we assume that the user has perfect knowledge of CSI, including frequency offset and channel gains. Fig. 4 shows the BER of 2-user 4-QAM OFDM for two groups; $M_T = 4$, $M_C = 2$, $M_u = 2$. First, the codebook consists of 64 matrices, i.e., 6 bits are transferred to the transmitter; while 32 matrices are included in the second group. The BER of an OFDM system with zero-frequency offset is shown as a reference. The MMSE criterion is used to select the precoding matrix. Our precoders improve BER notably: even with normalized frequency offsets in the interval $\mathbb{II}$, OFDM with the proposed precoder performs as well as the zero-frequency offset reference, i.e., the BER increase due to ICI has been eliminated completely. Our non-linear LFB-THP outperforms its linear counterpart. The system has lower BER when the number of feedback bits increases. We also consider the case when grouping with the interpolation scheme is used. The N subcarriers are divided into 8 groups, and each group has 8 subcarriers. The size of the codebook $\mathcal{Q}$ for interpolation is 4 as in [14]. Thus, the total number of feedback bits is $8 \times (\log 64 + \log 4) = 64$ bits. If the grouping scheme is not used, $64 \times \log 64 = 384$ bits are needed. Due to grouping, there is 2.5 dB performance loss at the BER of 10^{-4}. However, we save more than 80% feedback bits. The optimal higher-order interpolation design (large-size of $\mathcal{Q}$) is more complicated and still under investigation.

2) Impact of Inaccurate Channel Estimation: Fig. 5 presents BER when each user has imperfect knowledge of channel gains and frequency offset. The MMSE selection criterion is used and the size of the codebook is 64. We assume that the channel does not change much for several consecutive OFDM symbol durations. The frequency offset is then estimated using the frequency offset correction and tracking scheme as in [1]. At an SNR of 20 dB, the average normalized MSE of the frequency offset estimate is 1.44×10^{-3} for 10% normalized frequency offset, and 6.30×10^{-3} for 30%

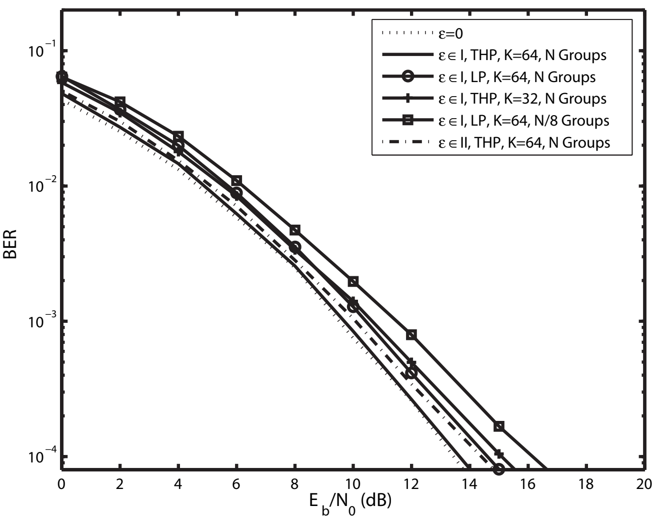

Fig. 4. BER of LFB-THP and LFB linear precoding (LFB-LP) with MMSE codebook selection criterion as a function of the SNR for different values of the normalized frequency offset and 64-subcarrier 2-user 4-QAM-OFDM; perfect CSI at the receiver; $M_T = 4$, $M_C = 2$, $M_u = 2$.

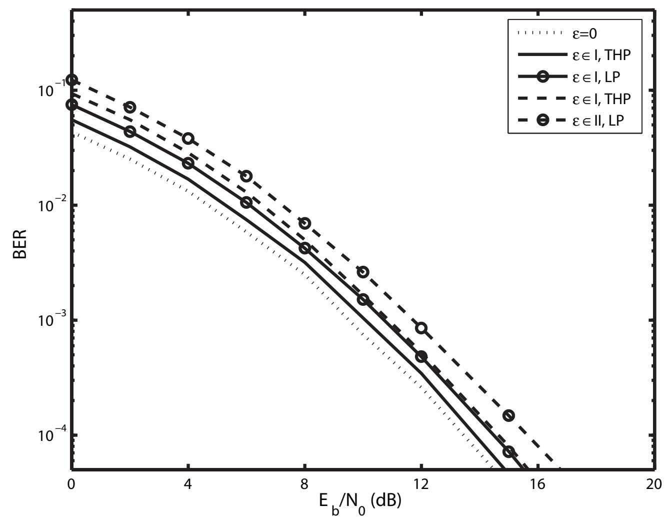

Fig. 5. BER of LFB-THP and LFB linear precoding (LFB-LP) with MMSE codebook selection criterion as a function of the SNR for different values of the normalized frequency offset and 64-subcarrier 2-user 4-QAM-OFDM; $M_T = 4$, $M_C = 2$, $M_u = 2$; estimated CSI at the receiver. $K = 64$.

normalized frequency offset. With the estimated frequency offset assumed constant over at least one OFDM symbol period, the channel gain is estimated using pilot symbols as in [29]. In order to guarantee reasonable performance of the channel estimator, every OFDM symbol is followed by a pilot block of length $2N_{CP}$, where N_{CP} is the length of cyclic prefix. The throughput loss incurred due to the pilot blocks is $2N_{CP}/(N + N_{CP})$. In our case, the number of subcarriers is 64 and the length of cyclic prefixes is 16. At an SNR of 20 dB, the average normalized MSE of the channel gain estimates is around 0.036 with a normalized frequency offset of 10%, and 0.047 with a normalized frequency offset of 30%. The value of MSE decreases as SNR increases.

The BER of OFDM with zero-frequency offset is also shown for comparison. Clearly, limited-feedback linear precoding is more sensitive to estimation errors than the proposed THP. Compared with the ideal case of zero-frequency offset, the BER degradation is small for the imperfect channel and frequency offset estimates. Consequently, even with imperfect CSI, our precoders still improve BER significantly.

3) Different Performance Criteria: The BERs of two-user 4-QAM uncoded and coded OFDM systems with non-linear limited-feedback precoding are shown in Fig. 6. The codebook size K is set to 64. Each user has 2 receive antennas. The BER of OFDM with zero-frequency offset is shown as a reference. In Fig. 6, the MMSE, MSV and MMI selection criteria are used in an SM MIMO OFDM system; $M_T = 3$, $M_C = 2$, $M_u = 2$. As before, our precoder reduces ICI significantly; even for the normalized frequency offsets in the interval $\mathbb{III}$, the BER is significantly improved. Furthermore, the system using the MMSE criterion outperforms the systems using the MSV and the MMI criteria by 0.6 and 1.8 dB, respectively, at a BER of 10^{-4} with normalized frequency offsets in $\mathbb{I}$.

B. OSTCB OFDM

Fig. 7 presents the BER of 2×4 and 3×4 single-user OFDM using the MMSE criterion and shows that our

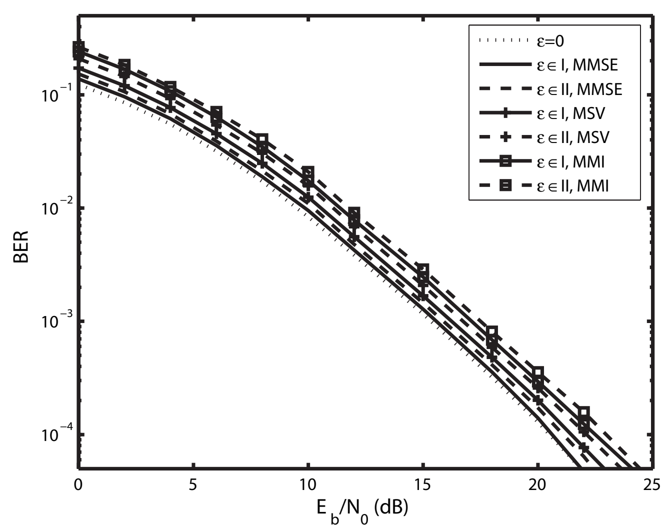

Fig. 6. BER for LFB-THP as a function of the SNR for different values of the normalized frequency offset and 64-subcarrier 2-user 4-QAM SM OFDM; $M_T = 3$, $M_C = 2$, $M_u = 2$. The MMSE, MSV and MMI codebook selection criteria are compared.

precoder can be used in OSTBC systems. The Alamouti code is considered with code rate of 1. The BER increase due to ICI is significantly reduced by our precoder, and the Alamouti-coded OFDM achieves a 3 dB gain over the uncoded SM MIMO OFDM at a BER of 10^{-4} in 2×4 systems. In 3×4 Alamouti-coded OFDM, once again, ICI is almost completely suppressed. As a result, our precoder can be used for OSTBC MIMO OFDM with an arbitrary number of transmit antennas.

C. Spatially Correlated MIMO Channels

In Fig. 8, a correlated Rayleigh fading channel is simulated for a two-user OFDM system; $M_T = 3$, $M_C = 2$, $M_u = 2$. The MMSE selection criterion is used and the codebook size is 64. We set the angle of arrival spread Δ in (37) as 0.1; the transmit and each user's receive antenna spacings are 4λ and 0.45λ, respectively. No correlation is assumed between

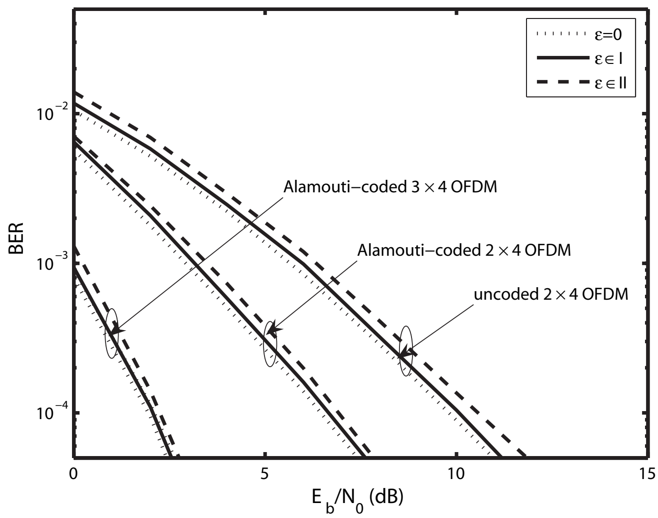

Fig. 7. BER of LFB-THP with the MMSE selection criterion as a function of the SNR for different values of the normalized frequency offset and 64-subcarrier 2×4 4-QAM uncoded (SM) OFDM, 4-QAM Alamouti-coded OFDM, and 3×4 4-QAM Alamouti-coded OFDM.

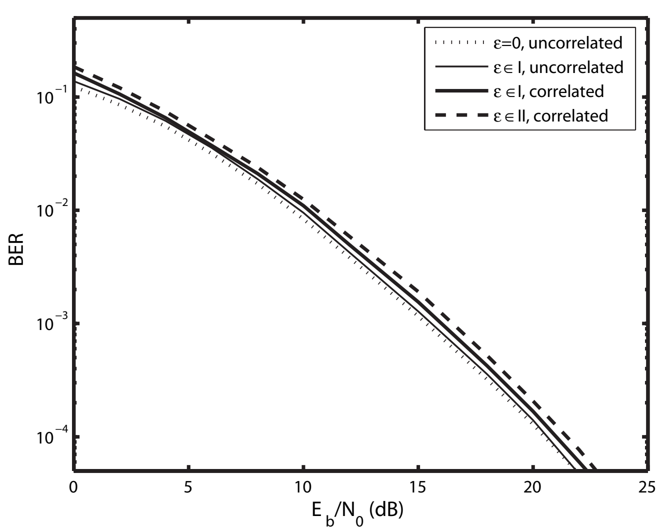

Fig. 8. BER for LFB-THP with the MMSE criterion as a function of the SNR for different values of the normalized frequency offset and 64-subcarrier 4-QAM SM MIMO OFDM in spatially correlated channels; $M_T = 3$, $M_C = 2$, $M_u = 2$.

different users. The transmit antenna correlation matrix $\mathbf{R}_T$ is known at both the transmitter and the receiver. Our LFB-TH precoder reduces ICI due to frequency offset. A marginal BER loss occurs as a result of antenna correlations, i.e., our LFB-THP works well in spatially-correlated MIMO channels.

VI. Conclusion

Linear and non-linear limited-feedback precoding have been developed for both multiuser SM OFDM and OSTBC MIMO OFDM in the presence of frequency offsets. We have shown that the ICI matrix does not influence precoding matrices for individual users, allowing precoding design on per-subcarrier basis. Exploiting this property, we have proposed the limited-feedback codebook design algorithm and derived the precoding matrix selection criteria. The proposed limited-feedback precoders reduce the feedback requirement. Non-linear precoding outperforms linear precoding. The results demonstrate that limited-feedback precoding significantly reduces the BER degradation due to frequency offset. Furthermore, the proposed precoder can also be used in OSTBC MIMO OFDM systems with an arbitrary number of transmit antennas and spatially correlated MIMO channels.

To further reduce the feedback load, grouping with interpolation has been introduced. Higher-order interpolation design will lead to a better performance, but is more complicated. There is a trade-off among the complexity of the interpolation design, performance, and the feedback load. It thus remains a future research topic to develop a practical size of the interpolation codebook.

References

[1] G. L. Stüber, J. R. Barry, S. W. McLaughlin, Y. G. Li, M. A. Ingram, and T. G. Pratt, "Broadband MIMO-OFDM wireless communications," *Proc. IEEE,*, vol. 92, no. 2, pp. 271–294, Feb. 2004.

[2] Telecommunications Industry Association, TIA/EIA Physical Layer Standard for cdma2000 Spread Spectrum Systems, Revision C. TIA/EIA/IS-2000-2.

[3] 3rd Generation Partnership Project, Technical Specification Group Radio Access Network: Physical Channels and Mapping of Transport Channels onto Physical Channels (FDD) (release 1999). v.3.2.0, 2000-2003.

[4] D. J. Love and R. W. Heath Jr., "What is the value of limited feedback for MIMO channels?" *IEEE Commun. Mag.*, vol. 42, no. 10, pp. 54–59, Oct. 2004.

[5] ——, "Grassmannian beamforming for multiple-input multiple-output wireless systems," *IEEE Trans. Inform. Theory*, vol. 49, no. 10, pp. 2735–2747, Oct. 2003.

[6] ——, "Limited feedback precoding for spatial multiplexing systems," *IEEE Trans. Inform. Theory*, vol. 51, no. 8, pp. 2967–2976, Aug. 2005.

[7] ——, "Limited feedback unitary precoding for orthogonal space-time codes," *IEEE Trans. Signal Processing*, vol. 53, no. 1, pp. 64–73, Jan. 2005.

[8] T. Pollet, M. V. Bladel, and M. Moeneclaey, "BER sensitivity of OFDM systems to carrier frequency offset and Wiener phase noise," *IEEE Trans. Commun.*, vol. 43, no. 2/3/4, pp. 191–193, Feb./Mar./Apr. 1995.

[9] S. - Y. Kung, Y. Wu, and X. Zhang, "Bezout space-time precoders and equalizers for MIMO channels," *IEEE Trans. Signal Processing*, vol. 50, no. 10, pp. 2499–2514, Oct. 2002.

[10] M. Tomlinson, "New automatic equalizer employing modulo arithmetic," *Electron. Lett.*, vol. 7, pp. 138–139, Mar. 1971.

[11] H. Harashima and H. Miyakawa, "Matched-transmission technique for channels with intersymbol interference," *IEEE Trans. Commun.*, vol. COM-20, pp. 774–780, Aug. 1972.

[12] Y. Fu, W. A. Krzymień, and C. Tellambura, "Transmitter precoding for ICI reduction in OFDM systems," in *Proc. IEEE VTC'05-Spring*, vol. 3, Stockholm, Sweden, May 2005, pp. 1788–1792.

[13] Y. Fu, C. Tellambura, and W. A. Krzymień, "Transmitter precoding for ICI reduction in closed-loop MIMO OFDM systems," *IEEE Trans. Veh. Technol.*, vol. 56, no. 1, pp. 115–125, Jan. 2007.

[14] J. Choi and R. W. Heath Jr., "Interpolation based unitary precoding for spatial multiplexing MIMO-OFDM with limited feedback," in *Proc. IEEE Globecom'04*, vol. 1, Dallas, TX, Nov. 2004, pp. 214–218.

[15] B. M. Hochwald, T. L. Marzetta, T. J. Richardson, W. Sweldens, and R. Urbanke, "Systematic design of unitary space-time constellation," *IEEE Trans. Inform. Theory*, vol. 46, no. 6, pp. 1962–1973, Sept. 2000.

[16] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, no. 8, pp. 1451–1458, Oct. 1998.

[17] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, "Space-time block codes from orthogonal designs," *IEEE Trans. Inform. Theory*, vol. 45, no. 5, pp. 1456–1467, July 1999.

[18] D. J. Love and R. W. Heath Jr., "Interpolation based transmitter beamforming for MIMO-OFDM with limited feedback," *IEEE Trans. Signal Processing*, vol. 53, no. 11, pp. 4125–4135, Nov. 2005.

[19] A. Scaglione, P. Stoica, S. Barbarossa, G. B. Giannakis, and H. Sampath, "Optimal designs for space-time linear precoders and decoders," *IEEE Trans. Signal Processing*, vol. 50, no. 5, pp. 1051–1064, May 2002.

[20] H. Bölcakei, D. Gesbert, and A. J. Paulraj, "On the capacity of OFDM-based spatial multiplexing systems," *IEEE Trans. Commun.*, vol. 50, no. 2, pp. 225–234, Feb. 2002.

[21] A. T. James, "Distribution of matrix variates and latent roots derived from normal samples," *Ann. Math. Statist.*, vol. 35, pp. 457–501, 1964.

[22] C. Windpassinger, R. F. H. Fischer, T. Vencel, and J. B. Huber, "Precoding in multiantenna and multiuser communications," *IEEE Trans. Wireless Commun.*, vol. 3, no. 4, pp. 1305–1315, July 2004.

[23] R. F. H. Fischer, *Precoding and Signal Shaping for Digital Transmission*. New York: Wiley, 2002.

[24] Y. Fu, W. A. Krzymień, and C. Tellambura, "Transmitter precoding for orthogonal space-time block-coded OFDM in transmit-antenna and path-correlated channels," in *Proc. IEEE VTC'06-Fall*, Montreal, QC, Canada, 2006.

[25] D. Shiu, G. J. Foschini, M. J. Gans and J. M. Kahn, "Fading correlation and its effect on the capacity of multi-element antenna systems," *IEEE Trans. Commun.*, vol. 48, no. 3, pp. 502–512, Mar. 2000.

[26] H. Bölcskei, M. Borgmann, and A. J. Paulraj, "Impact of the propagation environment on the performance of space-frequency coded MIMO-OFDM," *IEEE J. Select. Areas Commun.*, vol. 21, no. 3, pp. 427–439, Mar. 2003.

[27] T. L. Martetta and B. M. Hockwald, "Capacity of a mobile multiple-antenna communication link in Rayleigh flat fading," *IEEE Trans. Inform. Theory*, vol. 45, no. 1, pp. 139–157, Jan. 1999.

[28] International Telecommunication Union, Recommendation ITU-R M. 1225, Guidelines for Evaluation of Radio Transmission Technologies for IMT-2000, Feb. 2002.

[29] Y. - S. Choi, P. J. Voltz, and F. A. Cassara, "On channel estimation and detection for multicarrier signals in fast and selective Rayleigh fading channels," *IEEE Trans. Commun.*, vol. 49, no. 8, pp. 1375–1387, Aug. 2001.

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