Precoding for Orthogonal Space–Time Block-Coded OFDM Downlink: Mean or Covariance Feedback?

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***Abstract*—This paper presents linear and nonlinear precoding design for error-rate improvement in orthogonal space–time block-coded (OSTBC) multiple-input–multiple-output (MIMO) orthogonal frequency-division-multiplexed (OFDM) downlink, where both the conditional mean of the channel gain matrix and the channel gain covariance matrix may be available at the transmitter. The conditional means of the channel matrix are derived for a general transmit-antenna-correlated frequency-selective fading MIMO channel with estimation errors and feedback delay. Mean-feedback linear precoding and nonlinear Tomlinson–Harashima precoding (THP) are developed to maximize the signal-to-noise power ratio (SNR). The intuition that when the mean feedback becomes accurate the mean-feedback precoding outperforms covariance precoding is confirmed. Dual-mode precoding is also proposed, in which the novel mean-feedback precoding or covariance precoding is adaptively chosen at the receiver. The precoding-mode switching metric is the maximized SNR, which is an indicator of the error rate. The receiver calculates its metric, selects the mode that achieves a higher SNR, and decides whether mean feedback is necessary. Our proposed precoders (both mean feedback and adaptive) significantly reduce the system error rate. Nonlinear precoding is shown to outperform linear precoding. Adaptive precoding outperforms both mean-feedback precoding and covariance precoding if individually applied in OSTBC OFDM.**

***Index Terms*—Correlated antennas, covariance precoding, mean-feedback precoding, multiple-input–multiple-output (MIMO) orthogonal frequency-division-multiplexed (OFDM) downlink, orthogonal space–time block coding (OSTBC).**

I. Introduction

THIS paper considers precoding for multiple-input–multiple-output (MIMO) orthogonal space–time block-coded (OSTBC)[1] orthogonal frequency-division-multiplexed (OFDM) downlink, where both the conditional mean of the channel gain matrix (the first-order channel statistics) and the channel gain covariance matrix (the long-term/second-order channel statistics) are available at the transmitter (these two cases are referred to as *mean-feedback precoding* and *covariance precoding*, respectively). The application of the original OSTBC is constrained by insufficient antenna spacing at the transmitter, which leads to transmit-antenna correlations. Precoding, which can optimize MIMO transmission, is, thus, needed to offer the original OSTBC the flexibility of adapting to correlated MIMO channels [1]–[4].

A typical precoding design needs either the channel covariance matrix [3]–[5] or full channel state information at the transmitter (CSIT) [6]–[11]. CSIT, thus, can be an outdated (due to channel time variations) and imperfect (due to estimation and feedback errors) estimate. Given CSIT and channel statistics, the conditional mean of the actual channel gain matrix can be calculated and used for mean-feedback precoding. On the other hand, since the channel covariance matrix is primarily determined by the antenna correlation, which can readily be evaluated at the transmitter, the feedback requirement for covariance precoding is much smaller than for mean-feedback precoding. Naturally, the quality of mean-feedback precoding will be degraded due to estimation errors, and, in general, it is more sensitive to channel time variations and feedback delay than covariance precoding. In contrast, covariance precoding may become less effective when the mean feedback is accurate.

Mean-feedback precoders have been developed for OSTBC systems over an uncorrelated flat-fading MIMO channel in [6] and [7] with the assumption of perfect channel state information at the receiver (CSIR) or imperfect CSIT with additive estimation noise but no feedback delay. A more general model has not yet been considered. If the mean feedback and the channel covariance matrix are available, then [8] and [9] study the precoding schemes designed to approach capacity in spatially uncorrelated flat-fading MIMO channels and multiuser multiple-input–single-output (MISO) OFDM downlink (each user has one receive antenna). However, from the available results, it is still not clear whether the mean feedback can be helpful to reduce error rates. This paper aims to examine whether and when the mean feedback is necessary to achieve a lower error rate when considering precoding strategies for a general OSTBC OFDM system with channel estimation errors and feedback delay over a spatially correlated frequency-selective fading MIMO channel. We assume that the long-term channel statistics, including the variance of the estimation error and feedback delay, are available at the transmitter. In the

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

mean-feedback model, the complete channel matrix is sent back so that the transmitter can calculate the conditional mean of the actual channel matrix given the channel statistics. In the covariance precoding model, the channel covariance matrix is available at the transmitter; a complete CSIT is not necessary.

A general system model is assumed, in which the receiver imperfectly estimates the channel state information (CSI) and sends the inaccurate estimates back to the transmitter via a feedback channel, which introduces delay. We propose mean-feedback precoding to maximize the signal-to-noise power ratio (SNR) in a general OSTBC OFDM system; SNR determines the error rate in OSTBC systems. Both linear precoding (LP) and nonlinear Tomlinson–Harashima precoding (THP) [12], [13] are considered. The proposed precoders take into account the estimation errors and channel time variations over feedback delay and offer optimal power allocation. We confirm the intuition of the fundamental tradeoff between mean-feedback precoding and covariance precoding. As the mean feedback becomes accurate, mean-feedback precoding can achieve a lower bit error rate (BER) than the covariance precoding considered in [3], i.e., mean feedback is helpful. Nonlinear precoding is shown to outperform LP. Adaptive dual-mode precoding, which switches between the proposed mean-feedback precoding and covariance precoding [3], is also developed. The maximum achievable SNR is derived and used as the precoding-mode selection metric. The receiver calculates the metric and decides whether mean feedback is necessary. The decision is sent back to the transmitter using only 1 bit per subcarrier. Our adaptive precoding significantly reduces the required capacity of the feedback link since only the channel covariance matrix available at the transmitter is used when the channel conditions become severe, and it outperforms both mean-feedback precoding and covariance precoding when they are applied individually.

II. System Model

This section will introduce the system model of an N-subcarrier OFDM downlink system with M_T transmit antennas and M_R receive antennas in the presence of transmit antenna correlations. The receive antennas can be considered uncorrelated if the antenna spacing is greater than half a wavelength [14]. This condition is easily satisfied by practical systems that use carrier frequencies of about 2 GHz or higher.

A. MIMO OFDM With Transmit-Antenna Correlations

The channel between the vth transmit antenna and the uth receive antenna is a wideband frequency-selective fading channel with L resolvable paths. The path gains are independent identically distributed (i.i.d.) zero-mean complex Gaussian random variables (Rayleigh fading) with variances σ_l^2. At time i, the set of the lth path spatial gain can be represented by an $M_R \times M_T$ matrix $\mathbf{H}_l[i]$ with entries $h_{u,v,l}[i]$. The sum of channel tap power gains is normalized to unity. The channel gains remain constant over several OFDM symbol intervals. The $M_R \times M_T$ channel on the kth subcarrier at time i can be represented as [15]

$$\mathbf{H}[k,i] = \sum_{l=0}^{L-1} \mathbf{H}_l[i] e^{-j\frac{2\pi}{N}kl} \mathbf{R}_T^{1/2}. \tag{1}$$

$\mathbf{R}_T$ is the transmit antenna correlation matrix with entries that are expressed by [15]

$$R_T(p,q) = \mathcal{J}_0\left(2\pi|p-q|\zeta_T\right) \tag{2}$$

where $\mathcal{J}_0$ is the zero-order Bessel function of the first kind, p and q are antenna indexes, $\zeta_T = \Delta(d_T/\lambda_c)$, $\lambda_c = c/f_c$ is the wavelength at the center frequency f_c, Δ is the angle of arrival spread, and the transmit antennas are spaced by d_T. $\mathbf{R}_T = \mathbf{R}_T^H$.

At the receiver, the kth received signal vector can, thus, be given by

$$\mathbf{Y}[k,i] = \mathbf{H}[k,i]\mathbf{X}[k,i] + \mathbf{W}[k,i] \tag{3}$$

where $\mathbf{Y}[k,i]$ is an M_R-dimensional vector, and $\mathbf{W}[k,i]$ is the noise vector with the entries $W_u[k,i]$ being i.i.d. additive white Gaussian noise (AWGN) samples with zero mean and variance σ_W^2. The transmitted data vector is $\mathbf{X}[k,i] = [X_1[k,i] \quad \cdots \quad X_{M_T}[k,i]]^T$; $X_v[k,i]$ denotes an M-ary quadrature amplitude modulation (QAM) symbol on the kth subcarrier sent by the vth transmit antenna at time i. Stacking all the N received signal vectors, the $NM_R \times NM_T$ overall channel matrix is then

$$\mathbf{H}[i] = \operatorname{diag}\left[\mathbf{H}[1,i], \ldots, \mathbf{H}[N,i]\right]. \tag{4}$$

This paper considers the structure described by (3), for which precoding can individually be designed for each subcarrier.

B. Models for Channel Covariance and Mean Feedback

In this section, we introduce general MIMO channel models accounting for imperfect channel estimates, multiplicative time-varying effects, and transmit-antenna correlations. The conditional mean and covariance of the channel matrix are derived by exploiting the channel statistics.

1) Channel Covariance Model: For simplicity, the sum of the channel tap power gains is normalized to unity. The channel gain matrix on subcarrier k in (1) is assumed to be a zero-mean complex Gaussian matrix with autocovariance

$$\mathbf{C_{HH}}[k] = E\left[\mathbf{H}^H[k,i]\mathbf{H}[k,i]\right] = \mathbf{R}_T. \tag{5}$$

Hence, the autocovariance matrix is independent of the subcarrier indexes. $\mathbf{R}_T$ is mainly dependent on the angle of arrival spread and antenna spacing, which are easily available at the transmitter. $\mathbf{C_{HH}}[k]$ can also be readily determined at the receiver from the channel estimates. Evaluation of $\mathbf{C_{HH}}[k]$ needs to be done very infrequently because of the very low rate of change of $\mathbf{R}_T$.

2) Mean-Feedback Model: In this model, the complete complex channel gain matrix is sent back to the transmitter, and the transmitter calculates the conditional mean of the actual channel matrix given the channel statistics. It is called mean feedback in the rest of this paper. The receiver has inaccurate estimates $\mathbf{H}_R[k,i]$ of the current actual but unknown channel gain matrix $\mathbf{H}[k,i]$. The imperfect estimates are sent to the transmitter via a feedback channel that introduces a delay τ. The transmitter, thus, has the inaccurate estimate $\mathbf{H}_T[k,i]$ of the actual (unknown) but outdated channel matrix $\mathbf{H}[k,i-\tau]$, which is τ seconds older than the current channel matrix $\mathbf{H}[k,i]$. The variances of the estimation error and feedback delay are available at both the transmitter and the mobile receiver.

In the time-variant frequency-selective channel modeled, the entries in a tap gain vector for the $\{u,v\}$th antenna pair $\mathbf{h}_{u,v}[i]=[h_{u,v,0}[i],\ldots,h_{u,v,L-1}[i]]^T$ are time varying according to Clarke's 2-D isotropic scattering model with maximum Doppler shift f_D [14]. Since $\mathbf{h}_{u,v}[i-\tau]$ is a delayed version of $\mathbf{h}_{u,v}[i]$, they are jointly Gaussian with an autocovariance matrix

$$E\left[\mathbf{h}_{u,v}[i]\mathbf{H}_{u,v}^H[i-\tau]\right]=J\mathbf{R}_P \tag{6}$$

where $\mathbf{R}_P=\mathrm{diag}[\sigma_0^2,\ldots,\sigma_{L-1}^2]$, $J=\mathcal{J}_0(2\pi\epsilon)$ is the normalized autocovariance, and $\epsilon=f_D\tau$ is the normalized maximum Doppler shift with respect to the feedback delay τ. From (5) and by analogy to (6), the following covariance matrix can be obtained:

$$\mathbf{C}_T=E\left[\mathbf{H}^H[k,i-\tau]\mathbf{H}[k,i]\right]=J\mathbf{R}_T. \tag{7}$$

The channel estimates at the receiver can be expressed as $\mathbf{H}_R[k,i]=\mathbf{H}[k,i]+\mathbf{E}_{\mathrm{err}}[k]$, where $\mathbf{E}_{\mathrm{err}}[k]$ is the estimation error matrix with i.i.d. entries $E_{u,v}[k]\sim\mathcal{CN}(0,\Omega_e)\ \forall i,k$; $\mathbf{E}_{\mathrm{err}}[k]$ is independent of all the other stochastic processes. Therefore, the cross-covariance matrix $\mathbf{C}_{\mathbf{H}\mathbf{H}_R}$ of $\mathbf{H}[k,i]$ and $\mathbf{H}_R[k,i]$ is $\mathbf{R}_T$. The channel matrix estimate $\mathbf{H}_T[k,i]$ available at the transmitter can be modeled by $\mathbf{H}_T[k,i]=\mathbf{H}_R[k,i-\tau]=\mathbf{H}[k,i-\tau]+\mathbf{E}_{\mathrm{err}}[k]$. Obviously, $\mathbf{H}[k,i]\neq\mathbf{H}_T[k,i-\tau]\neq\mathbf{H}_R[k,i]$. $\mathbf{C}_{\mathbf{H}_T\mathbf{H}_R}=J\mathbf{R}_T+\Omega_e\mathbf{I}_{M_T}$ can be obtained similar to (5)–(7).

Given $\mathbf{H}_T[k,i]$, the transmitter can obtain the conditional mean and covariance matrices of the user following the approach in [16] as

$$\begin{aligned}\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k,i]&=J\mathbf{H}_T[k,i](\mathbf{R}_T+\Omega_e\mathbf{I}_{M_T})^{-1}\mathbf{R}_T\\ \mathbf{C}_{\mathbf{H}|\mathbf{H}_T}&=\mathbf{R}_T-J^2\mathbf{R}_T(\mathbf{R}_T+\Omega_e\mathbf{I}_{M_T})^{-1}\mathbf{R}_T.\end{aligned} \tag{8}$$

A more detailed derivation of (8) can be found in [11] and [17]. Since the user's receiver has the information $\mathbf{H}_R[k,i-\tau]=\mathbf{H}_T[k,i]$, the conditional mean of the channel matrix at the transmitter $\mathbf{H}_T[k,i]$ can be calculated at both the transmitter and the receiver, and the variance is dependent on the correlation matrix $\mathbf{R}_T$. Similarly, given $\mathbf{H}_R[k,i]$, the conditional channel and covariance matrices can be obtained as

$$\begin{aligned}\mathbf{H}_{\mathbf{H}|\mathbf{H}_R}[k,i]&=\mathbf{H}_R[k,i](\mathbf{R}_T+\Omega_e\mathbf{I}_{M_T})^{-1}\mathbf{R}_T\\ \mathbf{C}_{\mathbf{H}|\mathbf{H}_R}&=\Omega_e\mathbf{R}_T(\mathbf{R}_T+\Omega_e\mathbf{I}_{M_T})^{-1}\mathbf{R}_T.\end{aligned} \tag{9}$$

The conditional means $\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k,i]$ and $\mathbf{H}_{\mathbf{H}|\mathbf{H}_R}[k,i]$ can be described as equivalent channels that exploit the channel statistics and uncertainty structure to mitigate the impact of imperfect CSI at the transmitter and the receivers [18], [19]. The conditional covariance matrices $\mathbf{C}_{\mathbf{H}|\mathbf{H}_T}$ and $\mathbf{C}_{\mathbf{H}|\mathbf{H}_R}$ indicate the CSIT and CSIR channel uncertainty given by the equivalent channels, respectively. Clearly, the CSIR uncertainty is determined by the transmit antenna correlation matrix and the estimation error. At the transmitter, beside $\mathbf{R}_T$ and estimation errors, the CSIT uncertainty also depends on the autocovariance J, which is the function of the normalized maximum Doppler shift. As the maximum Doppler shift increases, which may be caused by the increasing velocity of the user, the mean-feedback uncertainty becomes significant.

III. Mean-Feedback and Covariance Precoding

In this section, mean-feedback precoding is proposed to maximize SNR in the general OSTBC OFDM system. By exploiting the conditional channel mean and covariance, our precoders take into account the channel uncertainty due to estimation errors and channel time variations. Both LP and nonlinear THP are presented.

A. Mean-Feedback Precoding

In this section, a general case with multiplicative time-varying effects and imperfect CSIR is considered. With the precoding matrix $\mathbf{E}_{\mathrm{MF}}[k]$, the effective transmitted signal is $\mathbf{E}_{\mathrm{MF}}[k]\mathbf{X}[k]$. For simplicity, the time index is omitted.

If the channel's mean and covariance matrices are available, then given $\mathbf{H}_T[k,i]$, the transmitter can calculate $\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k,i]$ of the actual channel matrix [see (8)] and perform singular value decomposition (SVD), which yields $\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k]=\widetilde{\mathbf{U}}[k]\widetilde{\Gamma}[k]\widetilde{\Gamma}^H[k]$. $\widetilde{\mathbf{U}}[k]$ and $\widetilde{\Gamma}[k]$ are unitary matrices, and the diagonal singular value matrix $\widetilde{\Gamma}[k]$ has real nonnegative entries. As in [7] and [20], a general form of the precoding matrix $\mathbf{E}_{\mathrm{MF}}[k]$ can be given by

$$\mathbf{E}_{\mathrm{MF}}[k]=\widetilde{\mathbf{V}}[k]\mathbf{\Lambda}_{\mathrm{MF}}[k]\widetilde{\mathbf{V}}^H[k] \tag{10}$$

where $\mathbf{\Lambda}_{\mathrm{MF}}[k]$ is an $M_T\times M_T$ diagonal matrix representing the power distribution with the main diagonal entries. In OSTBC systems, the minimum SNR (which should be maximized) determines the BER performance. With precoding [see (10)], the SNR on the kth subcarrier in OSTBC OFDM can be given by [21]

$$\begin{aligned}&\mathrm{SNR}_{\mathrm{MF}}[k]\\&=\frac{E_s}{\sigma_W^2}E\left[\mathrm{tr}\left(\mathbf{\Lambda}_{\mathrm{MF}}^H[k]\widetilde{\mathbf{V}}^H[k]\mathbf{H}^H[k]\mathbf{H}[k]\widetilde{\mathbf{V}}[k]\mathbf{\Lambda}_{\mathrm{MF}}[k]\right)\right]\end{aligned} \tag{11}$$

where E_s is the average energy of the transmitted symbols; $\mathrm{tr}(\cdot)$ denotes the trace of a matrix. Since the actual channel has the conditional mean $\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k]$ and covariance $\mathbf{C}_{\mathbf{H}|\mathbf{H}_T}$, as shown in (8), the error-rate minimization problem becomes

$$\begin{aligned}\mathbf{\Lambda}_{\mathrm{MF}}[k]_{\mathrm{opt}}&\\ =\arg\max_{\mathbf{\Lambda}_{\mathrm{MF}}[k]}&\ \mathrm{tr}\Big[\mathbf{\Lambda}_{\mathrm{MF}}^H[k]\widetilde{\mathbf{V}}^H[k]\\ &\times\left(\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}^H[k]\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k]+\mathbf{C}_{\mathbf{H}|\mathbf{H}_T}\right)\\ &\times\widetilde{\mathbf{V}}[k]\mathbf{\Lambda}_{\mathrm{MF}}[k]\Big]\end{aligned} \tag{12}$$

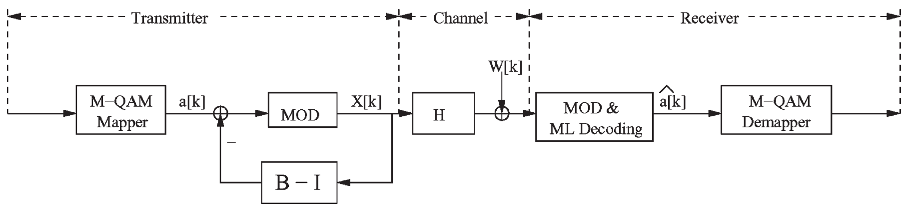

Fig. 1. THP in OSTBC OFDM downlink.

subject to $\mathrm{tr}(\mathbf{E}_{\mathrm{MF}}^H[k]\mathbf{E}_{\mathrm{MF}}[k]) = \mathrm{tr}(\mathbf{\Lambda}_{\mathrm{MF}}^2[k]) = M_T$, which is a normalization condition guaranteeing that precoding will not increase the transmitted power. The function in (12) is linear, and, therefore, it is concave in $\mathbf{\Lambda}_{\mathrm{MF}}[k]$. The objective of (12) is to find the power allocation $\mathbf{\Lambda}_{\mathrm{MF}}[k]$ that maximizes the SNR. This maximum SNR will depend on the quality of the conditional mean $\mathbf{H}_T[k]$. As $\mathbf{H}_T[k]$ approaches the actual channel matrix $\mathbf{H}[k]$, $\widetilde{\Gamma}[k]$ approaches $\Gamma[k]$ (the right unitary matrix from the SVD of the actual channel matrix), and $\mathbf{\Lambda}_{\mathrm{MF}}[k]$ is primarily determined by the singular values of the channel matrix. The optimization problem of (12) can numerically or analytically be solved. Furthermore, it has been shown that for some specific uncertainties, the problem can be simplified, and closed-form solutions exist [6], [7]. Several software packages are available including *optimization toolbox* in MATLAB. For example, the function *fmincon* can be used, which offers efficient computation within a polynomial time.

B. Nonlinear THP

The structure of the proposed precoder is illustrated in Fig. 1. Zero-forcing (ZF) THP is considered. The receiver consists of a maximum likelihood (ML) decoder and a modulo arithmetic device. The transmitter includes a modulo arithmetic feedback structure consisting of the matrix $\mathbf{B}[k]$ by which the transmitted symbols $X[k]$ are successively calculated for the data symbols $a[k]$ drawn from the initial M-ary QAM signal constellation. Ignoring the modulo device, the feedback structure is equivalent to $\mathbf{B}^{-1}[k]$, which can optimally be designed using (12) as $\mathbf{B}[k]_{\mathrm{opt}} = \mathbf{E}_{\mathrm{MF}}^{-1}[k]_{\mathrm{opt}}$. The effective channel is then $\mathbf{H}[k]\mathbf{E}_{\mathrm{MF}}[k]$, and ML decoding is used at the receiver. The overall precoding matrix can be written as $\mathbf{B} = \mathrm{diag}[\mathbf{B}[1], \ldots, \mathbf{B}[N]]$.

THP employs modulo operation at both the transmitter and the receiver. The modulo $2\sqrt{M}$ reduction at the transmitter, which is separately applied to the real and imaginary parts of the input, is to restrict the transmitted signals to the range of $\Re\{X[k]\} \in (-\sqrt{M}, \sqrt{M}]$ and $\Im\{X[k]\} \in (-\sqrt{M}, \sqrt{M}]$. If the input sequence $a[k]$ is a sequence of i.i.d. symbols, then the output of the modulo device is also a sequence of i.i.d. random variables, and the real and imaginary parts are independent, i.e., $E[\mathbf{X}[k]\mathbf{X}^H[k]] = E_s\mathbf{I}_{M_T}\ \forall k$ [13]. At the receiver, a slicer, which applies the same modulo operation as that at the transmitter, is used. After ML decoding and discarding the modulo congruence, the unique estimates of the data symbols $\hat{a}[k]$ can be generated. Further details of the THP operation are described in [13].

C. Covariance Precoding

When mean feedback is not available, the transmitter only knows the channel covariance matrix [see (5)]. The pairwise error probability (PEP) optimal covariance precoding introduced in [3] has the following form:

$$\mathbf{E}_{\mathrm{CP_{opt}}} = \mathbf{V}_T\mathbf{\Lambda}_{\mathrm{CP}}\mathbf{V}_T^H \tag{13}$$

where Γ_T is the right $M_T \times M_T$ unitary matrix from the SVD of the transmit-antenna correlation matrix $\mathbf{R}_T$. The optimal $\mathbf{\Lambda}_{\mathrm{CF}}$ can be found as discussed in [3].

The precoding design, which involves either mean-feedback precoding [see (12)] or covariance precoding [see (13)], aims to minimize the system error rate, which is determined by SNR in OSTBC systems. Mean-feedback precoding requires a high-capacity feedback link and is more sensitive to channel time variations and feedback delay. On the other hand, covariance precoding is a one-size-fits-all solution that does not reflect the instantaneous and varying channel conditions. Covariance precoding, thus, becomes less helpful when mean feedback can be accurate. Naturally, the quality of the mean and covariance matrices used in precoding determines the achievable error probability.

IV. Mean-Feedback or Covariance Precoding?

This section develops adaptive precoding, which switches between mean-feedback precoding [see (12)] and covariance precoding [see (13)]. The maximum achievable SNR in both cases of (12) and (13) is derived. The SNRs are calculated at the receiver and used as the selection metric, according to which the precoding mode is chosen.

A. Maximum Achievable SNR

In a precoded OSTBC system, the maximized SNR on the kth subcarrier can be given by

$$\begin{aligned}\mathrm{SNR}[k] =& \frac{E_s}{\sigma_W^2} E\left[\mathrm{tr}\left(\mathbf{H}^H[k]\mathbf{H}[k]\right)\right]\\ =& \frac{E_s}{\sigma_W^2}\mathrm{tr}\Big[\mathbf{E}_{\mathrm{opt}}^H\Big(\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}^H[k]\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k]\\ &+ \mathbf{C}_{\mathbf{H}|\mathbf{H}_T}\Big)\mathbf{E}_{\mathrm{opt}}\Big].\end{aligned} \tag{14}$$

If only the channel covariance matrix is available, then $\mathbf{C}_{\mathbf{H}|\mathbf{H}_T} = \mathbf{C}_{\mathbf{HH}}$, and the conditional mean is an all-zero matrix. The SNR is then

$$\mathrm{SNR}_{\mathrm{CP}} = \frac{E_s}{\sigma_W^2} \mathrm{tr}\left(\mathbf{E}^H_{\mathrm{CP}_{\mathrm{opt}}} \mathbf{C}_{\mathbf{HH}} \mathbf{E}_{\mathrm{CP}_{\mathrm{opt}}}\right). \tag{15}$$

Clearly, the SNR of the covariance precoded OSTBC OFDM is independent of the index of subcarriers and the channel time variations.

When mean feedback is available, the transmitter can calculate the equivalent channel matrix $\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k]$, and the SNR becomes

$$\begin{aligned}\mathrm{SNR}_{\mathrm{MF}}[k] &\\ = \frac{E_s}{\sigma_W^2} \mathrm{tr}\Big(&\mathbf{E}^H_{\mathrm{MF}}[k]_{\mathrm{opt}}\Big(\mathbf{H}^H_{\mathbf{H}|\mathbf{H}_T}[k]\mathbf{H}_{\mathbf{H}|\mathbf{H}_T}[k] \\ &+ \mathbf{C}_{\mathbf{H}|\mathbf{H}_T}\Big)\mathbf{E}_{\mathrm{MF}}[k]_{\mathrm{opt}}\Big). \end{aligned} \tag{16}$$

The SNR for a mean-feedback precoded OSTBC system is a function of J. The precoding matrix [see (12)] is determined by $\mathbf{H}_T[k]$ and $\mathbf{R}_T$ in (8), whose accuracy is dominated by J and Ω_e. The SNR in (16) is sensitive to channel-estimation errors and channel variations.

B. When to Use Mean Feedback

The SNRs are used as an indicator of the accuracy needed for mean feedback to give a lower BER than covariance precoding. The accuracy of the mean feedback can be gauged by its correlation with the actual channel matrix, which primarily depends on the normalized autocovariance J and estimation errors Ω_e. There may exist certain values of J and Ω_e such that for some channel realizations, the SNR of the mean-feedback precoded systems given by (16) is greater than the SNR of the covariance precoded systems given by (15), i.e., mean feedback is helpful for achieving a lower BER.

The mean-feedback precoding on subcarrier k will outperform covariance precoding if

$$\mathrm{SNR}_{\mathrm{MF}}[k] > \mathrm{SNR}_{\mathrm{CP}}. \tag{17}$$

The mode-selection inequality [see (17)] is the selection criterion used by our proposed adaptive precoding scheme. The inequality depends on J and Ω_e, which indicate when mean feedback has better performance. The receiver calculates (15) and (16) and adaptively selects the precoding mode that achieves a higher SNR. The same selection criterion is used for nonlinear precoding. The decision is sent back to the transmitter; one bit per subcarrier is required. The transmitter allows different subcarriers to use different precoding modes. When the channel time variation is fast or the channel conditions are poor, covariance precoding may outperform mean-feedback precoding. The channel information then does not need to be sent back to the transmitter, which significantly reduces the feedback requirements. If the mean feedback can be accurate, then mean-feedback precoding may offer a higher SNR. The receiver can, thus, use the mean feedback to achieve a lower error rate.

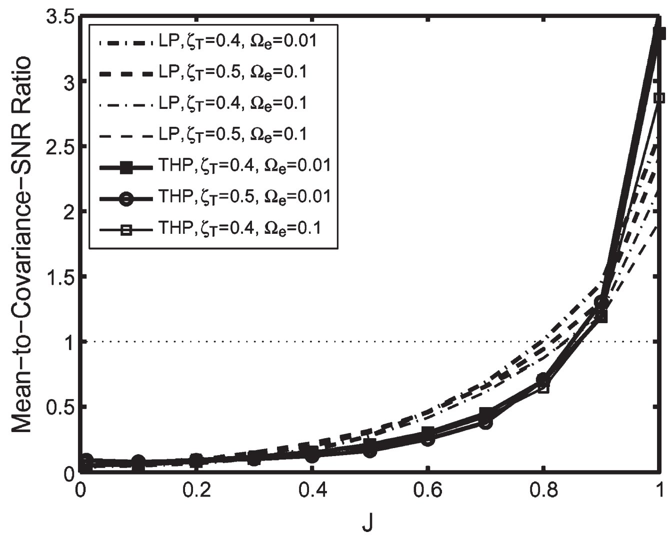

Fig. 2. MCSR with LP and THP as a function of the normalized autocovariance for 4-QAM Alamouti-coded 64-subcarrier 2×2 OFDM systems over transmit-antenna-correlated channels with different values of $\zeta_T = \Delta d_T/\lambda_c$, $\Omega_e = 0.01$, and $\Omega_e = 0.1$.

To further reduce the feedback load, subcarrier grouping with interpolation can be introduced. Higher order interpolation design will lead to a better performance, but it is more complex. There is a tradeoff among complexity, performance, and feedback load. Development of a simpler mode switching metric remains as a future research topic.

V. Simulation Results

This section presents simulation results to show how our proposed precoders improve the performance of a 64-subcarrier OSTBC OFDM system with transmit-antenna correlations. ZF precoding is considered. The vehicular B-channel specified by ITU-R M. 1225 [22] is used, where the channel taps represent zero-mean complex Gaussian random processes with variances of −4.9, −2.4, −15.2, −12.4, −27.6, and −18.4 dB relative to the total power gain of the frequency-selective channel. The excess delays of the channel taps are 0, 300, 8900, 12 900, 17 100, and 20 000 ns, respectively. ML decoding is used at the receiver. The transmitter and the receiver know the correlation matrix $\mathbf{R}_T$ with the correlation parameter $\zeta_T = \Delta(d_T/\lambda)$; the angle of arrival spread is assumed to be 12°, i.e., $\Delta \approx 0.2$.

The normalized autocovariance J values are assumed to be uniformly distributed in the interval $\mathbb{I} = [0.9, 1]$. In the interval $\mathbb{I} = [0.9, 1]$, the Doppler shifts $\epsilon = f_D\tau$ normalized with respect to the feedback delays τ are in the range [0, 0.1]. This range of normalized Doppler shifts corresponds to the range of mobile velocities from 0 to 216 km/h (at 5-GHz carrier frequency and 100-μs feedback delay).

Fig. 2 compares the maximum achievable SNR of mean-feedback precoding and covariance precoding in the Alamouti-coded 2×2 OFDM systems. We consider $J = \mathcal{J}_0(2\pi\epsilon)$. The variances of the estimation error are 0.1 and 0.01. When J is 0.998, the user's speed is around 30 km/h (assuming 5-GHz carrier frequency and $\tau = 100$ μs, as before). The value of J decreases as the speed increases. The mean-feedback precoding

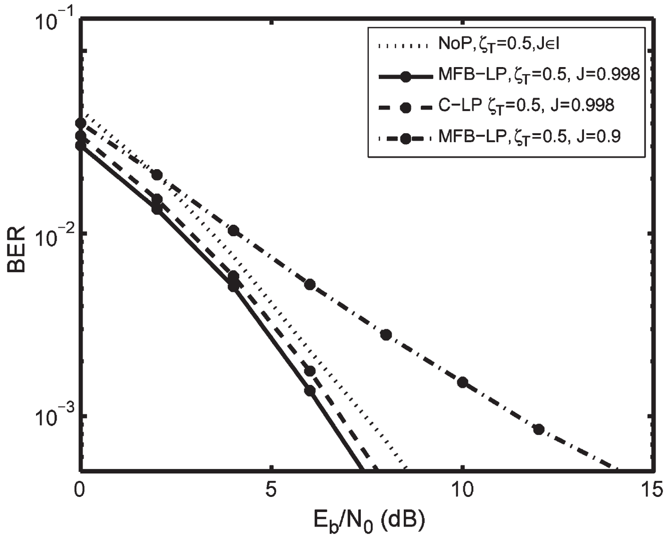

Fig. 3. BER as a function of SNR for MFB-LP, C-LP, and NoP in 64-subcarrier, 4-QAM Alamouti-coded 2×2 OFDM systems with perfect feedback. $J \in \mathbb{I}$, $J = 0.9$, and $J = 0.998$.

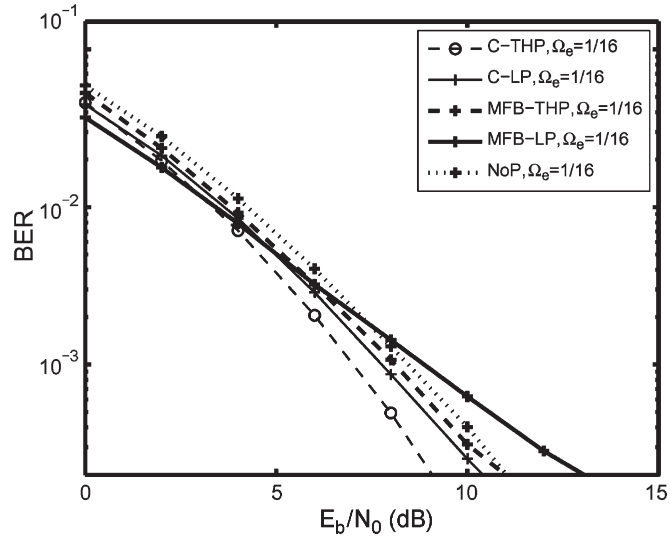

Fig. 4. BER as a function of SNR for MFB-LP, MFB-THP, C-LP, C-THP, and NoP in 64-subcarrier, 4-QAM Alamouti-coded 2×2 OFDM systems with perfect and imperfect channel estimation. $\zeta_T = \Delta d_T/\lambda_c = 0.4$, and $J \in \mathbb{I}$.

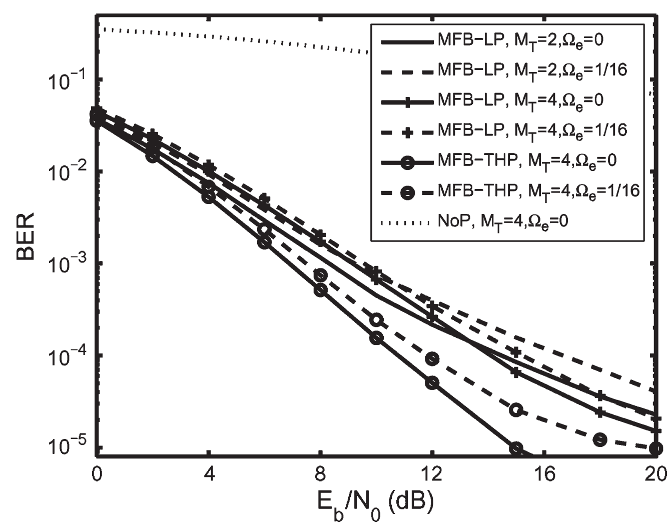

Fig. 5. BER as a function of SNR for MFB-LP, MFB-THP, and NoP in 64-subcarrier, 16-QAM, 1/2-rate-OSTBC OFDM and 4-QAM Alamouti-coded OFDM systems with perfect and imperfect channel estimation. $\zeta_T = \Delta d_T/\lambda_c = 0.3$, $M_R = 2$, and $J \in \mathbb{I}$.

SNR to covariance precoding SNR ratio (MCSR) shows the gain that mean-feedback precoding can obtain over covariance precoding. As the variance becomes large, the gain of mean-feedback precoding over covariance precoding decreases. On the other hand, as the correlation parameter ζ_T increases, i.e., the transmit-antenna correlation decreases, MCSR also increases. When ζ_T is greater than 0.2, and J is greater than 0.8, the value of MCSR is larger than 1, i.e., the proposed mean-feedback precoding achieves a higher SNR than covariance precoding. In this case, mean feedback is sufficiently accurate to achieve a lower error rate. Furthermore, THP performs better than LP.

Fig. 3 shows the BERs of the proposed mean-feedback LP (MFB-LP) in Alamouti-coded 2×2 OFDM with perfect channel estimation. The normalized autocovariances J of 0.9 and 0.998 are considered. The BER of the covariance LP (C-LP) in [3] is shown for reference. When the user is moving at the speed of 30 km/h ($J = 0.998$), the proposed mean-feedback precoder offers a 0.4-dB gain over covariance precoding at the BER of 10^{-3}. Clearly, as CSIT becomes more accurate, mean-feedback precoding outperforms covariance precoding. The BER increases with decreasing J. At J of 0.9, the BER of mean-feedback precoding drastically increases, i.e., covariance precoding is more suitable in this case.

Fig. 4 shows the performance of our proposed MFB-LP and mean-feedback THP (MFB-THP) and the impact of imperfect channel estimation. 4-QAM Alamouti-coded 2×2 OFDM is considered. The variance of the channel estimation error Ω_e is 1/16. Our MFB-THP outperforms the no-precoding (NoP) case even if the channel's time-variation range is as high as [0.9, 1]. Clearly, nonlinear precoding outperforms LP. At a BER of 10^{-3}, covariance THP (C-THP) has a 0.6-dB gain over C-LP, and MFB-THP has about a 1.2-dB gain over MFB-LP.

Fig. 5 shows the impact of the number of transmit antennas on the proposed mean-feedback precoding. We consider 16-QAM, 1/2-rate-OSTBC 4×2 OFDM and 4-QAM, full-rate OSTBC 2×2 OFDM systems when the antenna correlation coefficient ζ_T is 0.3. Clearly, Fig. 5 confirms the intuition that a large number of transmit antennas improves the BER in the high SNR region and increases robustness against channel estimation errors. Once again, nonlinear THP outperforms LP in both perfect and imperfect CSIR cases.

Fig. 6 shows the BERs of the proposed adaptive linear precoder for 16-QAM 1/2-rate OSTBC 4×2 OFDM when the correlation parameter ζ_T is 0.4. Perfect CSIR is considered. Adaptive precoding switches between the mean-feedback and covariance precoding modes, which offers an additional selection diversity gain. As a result, it achieves almost 1.5 dB improvement over pure covariance precoding. Also, adaptive precoding eliminates error floors at a BER of 10^{-4} present with mean-feedback precoding.

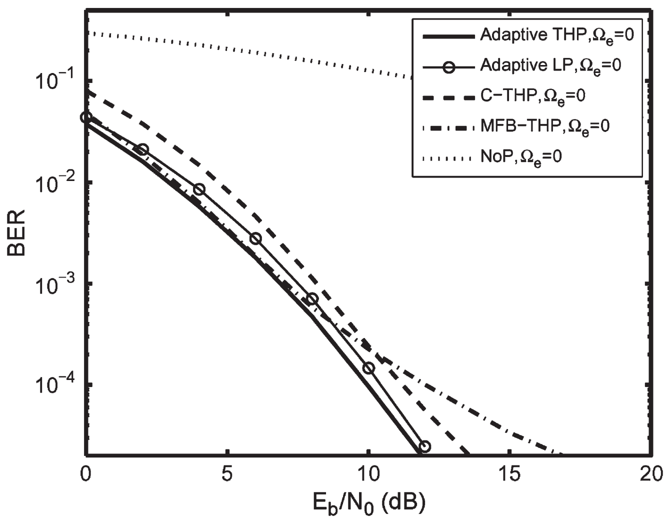

Fig. 6. BER as a function of SNR for mean-feedback THP (MFB-THP), covariance THP (C-THP), adaptive LP and THP, and no precoding (NoP) in a 64-subcarrier 16-QAM 1/2-rate OSTBC 4×2 OFDM system. $\zeta_T = \Delta d_T/\lambda_c = 0.4$, $J \in \mathbb{I}$.

VI. CONCLUSION

This paper has considered an OSTBC OFDM downlink system with estimation errors and feedback delay over a general transmit-antenna-correlated frequency-selective fading MIMO channel. SNR-maximizing mean-feedback precoding and adaptive precoding have been proposed. Depending on the channel conditions, the proposed adaptive dual-mode precoding switches between either new mean-feedback precoding or covariance precoding. The receiver calculates the precoding-mode switching metric and decides whether mean feedback is necessary. If mean feedback is sufficiently accurate, then it improves the system performance. Our proposed precoders (both the mean-feedback precoder and the adaptive dual-mode precoder) reduce the error rate. Adaptive precoding outperforms both mean-feedback precoding and covariance precoding, which were individually applied.

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