Interpolation-Based Multi-Mode Precoding for MIMO-OFDM Systems with Limited Feedback

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***Abstract*— Spatial multiplexing with multi-mode precoding provides a means to achieve both high capacity and high reliability in multiple-input multiple-output orthogonal frequency-division multiplexing (MIMO-OFDM) systems. Multi-mode precoding uses linear transmit precoding that adapts the number of spatial multiplexing data streams or modes, according to the transmit channel state information (CSI). As such, it typically requires complete knowledge of the multi-mode precoding matrices for each subcarrier at the transmitter. In practical scenarios where the CSI is acquired at the receiver and fed back to the transmitter through a low-rate feedback link, this requirement may entail a prohibitive feedback overhead. In this paper, we propose to reduce the feedback requirement by combining codebook-based precoder quantization, to efficiently quantize and represent the optimal precoder on each subcarrier, and multi-mode precoder frequency down-sampling and interpolation, to efficiently reconstruct the precoding matrices on all subcarriers based on the feedback of the indexes of the quantized precoders only on a fraction of the subcarriers. To enable this efficient interpolation-based quantized multi-mode precoding solution, we introduce (1) a novel precoder codebook design that lends itself to precoder interpolation, across subcarriers, followed by mode selection, (2) a new precoder interpolator and, finally, (3) a clustered mode selection approach that significantly reduces the feedback overhead related to the mode information on each subcarrier. Monte-Carlo bit-error rate (BER) performance simulations demonstrate the effectiveness of the proposed quantized multi-mode precoding solution, at reasonable feedback overhead.**

***Index Terms*— Multiple-input multiple-output (MIMO), OFDM, limited feedback, vector quantization, mode selection, linear precoding, matrix interpolation under a unitary constraint.**

I. INTRODUCTION

MULTIPLE-input multiple-output orthogonal frequency-division multiplexing (MIMO-OFDM) is widely recognized today as a key technology to achieve high spectral efficiency and high link reliability over space- and frequency-selective wireless channels, at a reasonable hardware cost. MIMO-OFDM capitalizes on the reduced-complexity per-subcarrier processing of OFDM to enable an easier fulfillment of the high capacity and/or diversity promises of MIMO systems. The actual fulfillment of the benefits of MIMO is achieved through a variety of MIMO processing techniques [1], [2], [3], [4]. One of these techniques is spatial multiplexing, which transmits multiple independent parallel data streams on the MIMO channel, in an attempt to approach the MIMO capacity [1], [2]. Its performance, however, is sensitive to the rank behavior of the MIMO channel. Hence, linear precoding was proposed, which improves the robustness of spatial multiplexing to rank deficiencies of the MIMO channel [5], [6].

Spatial multiplexing with linear precoding simply multiplies the spatial data streams, prior to transmission, by a precoding matrix that is designed according to some form of transmit channel state information (CSI). This precoding matrix adapts the transmission to the preferred directions of the MIMO channel, which inherently adds resilience against channel ill-conditioning. The state-of-the-art optimization criteria for designing the precoding matrix include minimizing the mean squared error [5], [6], maximizing the minimum distance between two received data vectors [7], maximizing the minimum signal-to-noise ratio (SNR) [8], [9], [10], [11], and maximizing the mutual information [12], [13], [6]. The resulting precoding matrix invariably consists of a unitary matrix, which beamforms into the eigenmodes of the available transmit CSI, and a power loading matrix, which distributes the available transmit power across these eigenmodes according to the considered optimization criterion. Nevertheless, except for the mutual information-based design, all linear precoding solutions assumed a fixed number of streams as well as fixed coding and modulation across these spatial streams. Such conventional fixed-mode linear precoding was shown to be suboptimal [14], [9], [15]. Alternatively, multi-mode precoding was introduced [14], [9], [15], which optimally adapts the number of spatial multiplexing streams according to the channel conditions, leading to significant performance improvements over its fixed-mode counterparts. Unfortunately, to realize its full gains, spatial multiplexing with multi-mode precoding requires full CSI at the transmitter; an assumption, which is often severely challenged in practical systems.

We herein consider a practical MIMO-OFDM system which pursues a closed-loop approach to transmit CSI acquisition. This closed-loop approach, which is applicable to both time-division duplexing (TDD) and frequency-division duplexing

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(FDD) systems, estimates the CSI at the receiver side and conveys it to the transmitter side through a feedback channel [16], [17], [18], [19]. However, one of the challenges for this approach is to come up with a practical multi-mode precoding solution that incurs minimal feedback overhead, which is the topic of this paper.

Our approach is to quantize the space of precoding matrices into a finite multi-mode precoder codebook [20], [17], [21], [15], [11], [22]. This approach allows for the most effective compression of the feedback overhead, through exploiting the special structure of the precoding matrices, and their limited number of underlying degrees of freedom [18], [23], [24], [19], [22]. However, all state-of-the-art quantized multi-mode precoding solutions have considered flat-fading channels. Their straightforward extension to frequency-selective channels, using OFDM, would entail an unacceptable amount of feedback bits, whose number scales linearly with the number of subcarriers. This feedback requirement can be significantly reduced through precoder frequency down-sampling and interpolation [25], [26]. Unfortunately, interpolation cannot be used in combination with state-of-the-art multi-mode precoder codebook designs [15], [11], [20], [17], [21], [22], due to the fact that these codebooks exhibit a unitary ambiguity. This unitary ambiguity, inherited from the various codebook design criteria, compromises the ordering of the precoder's singular vectors, which is crucial for successful multi-mode precoding on the interpolated precoders. Consequently, an efficient interpolation-based extension of the existing multi-mode precoder codebooks is not feasible. To enable such a solution, we propose to construct a new precoder codebook that preserves the ordering of the singular vectors, corresponding to the quantization of the unitary right singular matrices of the MIMO channels on the different subcarriers. Based on this precoder codebook, the receiver only needs to feed-back the optimal indexes of the codewords on a fraction of the subcarriers. The precoders on the remaining subcarriers can then be recovered using our proposed conditional interpolator. Finally, on each reconstructed unitary precoder related to a certain subcarrier, we enforce the optimal number of streams, following a *clustered* mode selection approach, where the receiver defines a few subcarrier-clusters and feeds back only a single mode information per subcarrier-cluster.

The remainder of this paper is organized as follows: Section II introduces the system model as well as the fundamentals of multi-mode precoding. Section III examines the limited-feedback constraint and the resulting quantized multi-mode precoding framework. It also highlights the shortcomings of the state-of-the-art quantized multi-mode precoding solutions, in the context of OFDM-based systems. To overcome these shortcomings, we propose a novel interpolation-based quantized multi-mode precoding system concept, in Section IV. Its performance is assessed in Section V. Finally, Section VI draws the concluding remarks.

Notation: In all the following, normal letters designate scalar quantities, boldface lower-case letters indicate column vectors, and boldface capitals represent matrices. Normal brackets indicate time-domain signals, whereas square brackets denote frequency-domain quantities. $\mathbf{I}_p$ and $\mathbf{0}_p$ are the $p \times p$ identity and all-zero matrices, respectively. Moreover, $trace(\mathbf{M})$, $[\mathbf{M}]_{i,j}$, $[\mathbf{M}]_{.,j}$, $[\mathbf{M}]_{.,1:j}$, respectively, stand for the trace, the $(i,j)^{th}$ entry, the j^{th} column, the j first columns of matrix $\mathbf{M}$. $(.)^H$, $(.)^\dagger$, $(.)^{-1}$ denote the conjugate transpose, the pseudo-inverse, the inverse of a matrix, respectively. Finally, $diag(\mathbf{m})$ is the operator that creates a diagonal matrix with $\mathbf{m}$ on its main diagonal.

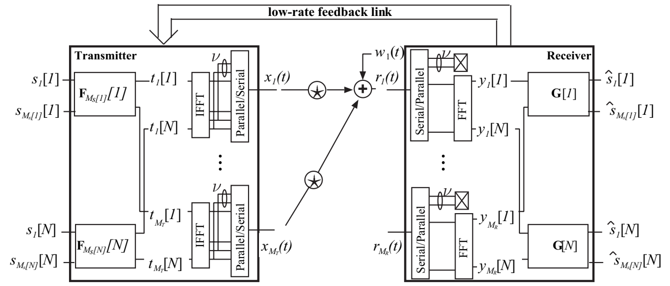

Fig. 1. The considered MIMO-OFDM linear precoding system with limited feedback.

II. System Overview and Background

A. System model

The spatial multiplexing OFDM-based WLAN communication system, under consideration, is depicted in Figure 1. It consists of an M_T-antenna transmitter, an M_R-antenna receiver, and comprises N subcarriers. It deploys a multi-mode linear precoding transmission scheme based on quantized transmit channel state information (CSI), which is conveyed back from the receiver through a low-rate feedback link.

We assume that R bits are conveyed per channel use, under an average transmit power constraint $\mathcal{P}_T$. We further assume that both the rate, R, and the average transmit power, $\mathcal{P}_T$, are evenly distributed among the N subcarriers. Accordingly, for each channel use, on the n^{th} subcarrier, the transmitter demultiplexes R/N bits into an $M_s[n]$-dimensional vector, whose entries are then modulated using a constellation $\mathcal{S}[n]$ to form the $M_s[n]$-dimensional spatial data vector $\mathbf{s}[n] = [s_1[n] \cdots s_{M_s[n]}[n]]^T$, where $M_s[n] \leq \min(M_T, M_R)$, so that $E\{\mathbf{s}[n]\mathbf{s}[n]^H\} = \frac{\mathcal{P}_T}{NM_s[n]}\mathbf{I}_{M_s[n]}$. The transmitter then maps the symbol data vector, $\mathbf{s}[n]$, onto the M_T transmit antennas, using the linear precoder, $\mathbf{F}_{M_s[n]}[n]$. This linear precoder, $\mathbf{F}_{M_s[n]}[n]$, is designed based on the quantized transmit CSI acquired through a low-rate feedback link. On each transmit antenna m_T, the transmit symbols, $t_{m_T}[n]$, are subsequently grouped into blocks of N symbols, converted to the time-domain via an N-tap IFFT, prefixed with a cyclic prefix of length ν, and finally serialized. The resulting stream $x_{m_T}(t)$ is launched into the convolutional channels, $\{[\mathbf{H}]_{m_R,m_T}(t)\}_{1\leq m_R \leq M_R}$, which denote the equivalent baseband representations of the multipath propagation channels to the M_R receive antennas. At the output of each receive antenna m_R, the receiver acquires the convolutional mixture $r_{m_R}(t) = \sum_{m_T=1}^{M_T}[\mathbf{H}]_{m_R,m_T}(t) \star x_{m_T}(t) + w_{m_R}(t)$, where $w_{m_R}(t)$ is an AWGN term. Subsequently, it removes the cyclic prefix and takes the N-tap FFT. We end up with M_R received sequences $\{y_{m_R}[n]\}_{1\leq m_R \leq M_R}$ on each subcarrier n. The corresponding frequency-domain M_R-dimensional received signal vector $\mathbf{y}[n]$ is then postprocessed to estimate [1] the frequency-domain counterparts of the $M_T M_R$ convolutional channels $\{[\mathbf{H}]_{m_R,m_T}(t)\}_{m_R,m_T}$, on the n^{th} subcarrier, in order to subsequently provide estimates $\hat{\mathbf{s}}[n]$ for the transmitted data symbol vector $\mathbf{s}[n]$, as well as generate the quantized CSI to be fed-back to the transmitter.

If ν is larger than the length of the channels $\{[\mathbf{H}]_{m_R,m_T}(t)\}_{m_R,m_T}$, the linear convolutions are observed as cyclic by the receiver. Thus, in the frequency domain, they become equivalent to multiplication with the discrete Fourier transform (DFT) of these channels, given by $\{[\mathbf{H}]_{m_R,m_T}[n]\}_{m_R,m_T}$ on each subcarrier n. Consequently, the data model on subcarrier n simply reads

$$\mathbf{y}[n] = \boldsymbol{\mathcal{H}}[n].\mathbf{F}_{M_s[n]}[n].\mathbf{s}[n] + \mathbf{w}[n] \quad (1)$$

where $\mathbf{y}[n] = [y_1[n] \cdots y_{M_R}[n]]^T$, $\mathbf{s}[n] = [s_1[n] \cdots s_{M_s[n]}[n]]^T$ and $\mathbf{w}[n] = [w_1[n] \cdots w_{M_R}[n]]^T$. $\mathbf{w}[n]$ is the zero-mean spatially-white complex Gaussian receiver noise with covariance matrix $N_0 \mathbf{I}_{M_R}$. $\boldsymbol{\mathcal{H}}[n]$ is the $M_R \times M_T$ matrix whose $(m_R, m_T)^{th}$ entry is simply $[\mathbf{H}]_{m_R,m_T}[n]$. Clearly, OFDM modulation decouples the convolutional composite channel into a set of N orthogonal flat-fading composite channels, on the N subcarriers. This property is exploited to carry out data precoding and detection on each subcarrier independently. Accordingly, on subcarrier n, $\hat{\mathbf{s}}[n]$ is recovered based on the perfectly-estimated frequency-domain composite channel, $\boldsymbol{\mathcal{H}}^c[n] = \boldsymbol{\mathcal{H}}[n]\mathbf{F}_{M_s[n]}[n]$, on that subcarrier, using the linear minimum mean squared error (MMSE) receiver

$$\mathbf{G}[n] = \left(\frac{N M_s[n] N_0}{\mathcal{P}_\mathcal{T}} \mathbf{I}_{M_s[n]} + \boldsymbol{\mathcal{H}}^c[n]^H \boldsymbol{\mathcal{H}}^c[n] \right)^{-1} \boldsymbol{\mathcal{H}}^c[n]^H.$$

B. Optimal per-subcarrier multi-mode precoder design

Assuming linear MMSE reception, a sensible criterion to design the linear precoder, $\mathbf{F}_{M_s[n]}[n]$, is the minimization of the sum mean squared error across the received spatial data symbol vector, $\hat{\mathbf{s}}[n]$, or equivalently the minimization of the trace of the mean squared error matrix

$$\mathbf{F}_{M_s[n]}[n] = \min_{\tilde{\mathbf{F}}:||\tilde{\mathbf{F}}||_F^2 = M_s[n]} trace \left[\frac{\mathcal{P}_\mathcal{T}}{N M_s[n]} \left(\mathbf{I}_{M_s[n]} + \frac{\mathcal{P}_\mathcal{T}}{N M_s[n] N_0} \tilde{\mathbf{F}}^H \boldsymbol{\mathcal{H}}[n]^H \boldsymbol{\mathcal{H}}[n] \tilde{\mathbf{F}} \right)^{-1} \right]. \quad (2)$$

The constraint on the Frobenius norm of the precoder, $\mathbf{F}_{M_s[n]}[n]$, translates the average transmit power constraint on the n^{th} subcarrier ($\mathcal{P}_\mathcal{T}/N$), given the considered covariance matrix of the $M_s[n]$-dimensional spatial data symbol vector, $\mathbf{s}[n]$ ($E\{\mathbf{s}[n]\mathbf{s}[n]^H\} = \frac{\mathcal{P}_\mathcal{T}}{N M_s[n]} \mathbf{I}_{M_s[n]}$). Let $\boldsymbol{\mathcal{H}}[n] = \mathbf{U}[n]\boldsymbol{\Sigma}[n]\mathbf{V}[n]^H$ be the singular value decomposition (SVD) of $\boldsymbol{\mathcal{H}}[n]$. The optimal linear precoder, solution to (2), was shown to be $\mathbf{F}_{M_s[n]}[n] = \mathbf{V}[n]_{:,1:M_s[n]} \boldsymbol{\Sigma}_{M_s[n]}[n] \mathbf{Q}[n]$ [5], [6]. It consists of the combination of $\mathbf{V}[n]_{:,1:M_s[n]}$, the $M_T \times M_s[n]$ transmit eigen-beamformer into the $M_s[n]$ strongest modes of $\boldsymbol{\mathcal{H}}[n]$, $\boldsymbol{\Sigma}_{M_s[n]}[n]$, the $M_s[n] \times M_s[n]$ optimal spatial power loading matrix across these modes, and $\mathbf{Q}[n]$, an optional $M_s[n] \times M_s[n]$ unitary matrix. The optional unitary matrix $\mathbf{Q}[n]$ does not alter the system's sum MMSE on the n^{th} subcarrier, denoted $MMSE[n]$, but it can be used to enforce additional constraints on the mean squared errors (MSE) of the individual spatial streams on that subcarrier [27], [28], [29]. For instance, it was shown that choosing $\mathbf{Q}[n]$ such that its entries have a constant modulus equal to $1/\sqrt{M_s[n]}$ (for instance, an $M_s[n] \times M_s[n]$ normalized IFFT or Hadamard matrix), guarantees equal minimum uncoded BER on all spatial streams for a given number of streams $M_s[n]$ and a fixed constellation across these streams $\mathcal{S}[n]$ [27], [28]. In fact, such constant-modulus $\mathbf{Q}[n]$ simply enforces an even MSE across the spatial streams, which is then equal to $MMSE[n]/M_s[n]$. The resulting design is known as *min-BER* linear precoding. In this contribution, we assume a uniform spatial power allocation. This is motivated by the fact that spatial power allocation only leads to small performance and capacity gains [30] while significantly increasing the feedback requirements. Consequently, for a given number of spatial streams, $M_s[n]$, our optimal linear precoder has orthonormal columns ($\mathbf{F}_{M_s[n]}[n]^H \mathbf{F}_{M_s[n]}[n] = \mathbf{I}_{M_s[n]}$). We further assume that $\mathbf{Q}[n] = \mathbf{I}_{M_s[n]}$, and thus our optimal linear precoder is simply given by $\mathbf{F}_{M_s[n]}[n] = \mathbf{V}[n]_{:,1:M_s[n]}$. Although the corresponding *min-MMSE* linear precoding exhibits a worse uncoded BER figure than the min-BER design, it was found to actually outperform the latter design in the presence of channel coding [10]. This is why we have opted for the min-MMSE design. Nevertheless, our interpolation-based multi-mode linear precoding can be applied to the min-BER design, as will be discussed in Section V.

A crucial element in the design of the optimal precoder, $\mathbf{F}_{M_s[n]}[n]$, is the choice of the number of transmit data streams, $M_s[n]$, or, equivalently, the number of used channel eigen-modes, depending on the channel state, $\boldsymbol{\mathcal{H}}[n]$. Since the number of bits per channel use and per subcarrier is fixed, the adaptation of the number of data streams, $M_s[n]$, requires also the adaptation of the used constellation size, $card(\mathcal{S}[n])$. We have previously shown that the selection/adaptation of the pair $\{M_s[n], card(\mathcal{S}[n])\}$, based on the knowledge of $\boldsymbol{\mathcal{H}}[n]$, enables significant performance improvements over its non-optimized counterparts [14], [9], [15], [11]. In this contribution, we call *multi-mode precoder*, a precoder that carries out this selection/adaptation of the optimal pair $\{M_s[n], card(\mathcal{S}[n])\}$, depending on the channel state $\boldsymbol{\mathcal{H}}[n]$. The optimality criterion pertains to the minimization of an upper-bound on the symbol-vector error rate [9], [10], [11], which can be shown to be achieved through the minimization of the symbol-error rate of the weakest spatial stream [8]. Let $SINR_{min}(\mathbf{F}_{M_s[n]}[n])$ denote the Signal-to-Interference and Noise Ratio (SINR) of the weakest spatial stream on the n^{th} subcarrier [11]

$$SINR_{min}(\mathbf{F}_{M_s[n]}[n]) = \min_{1 \le k \le M_s[n]} \frac{\mathcal{P}_\mathcal{T}/(N M_s[n])}{\left[\left(\boldsymbol{\mathcal{H}}^c[n]^H \boldsymbol{\mathcal{H}}^c[n] + \frac{N M_s[n] N_0}{\mathcal{P}_\mathcal{T}} \mathbf{I}_{M_s[n]} \right)^{-1} \right]_{k,k}} - 1. \quad (3)$$

For the used Gray-encoded square QAM constellation of size $card(\mathcal{S}[n])$ and average transmit symbol energy

[1] This estimation is assumed to be error-free. Indeed, the aim of this contribution is to investigate the individual impact of feedback quantization.

$$BER_{min}(\mathbf{F}_{M_s[n]}[n]) = \frac{4}{log_2(card(\mathcal{S}[n]))}\left(1-\frac{1}{\sqrt{card(\mathcal{S}[n])}}\right)Q\left(\sqrt{\frac{3SINR_{min}(\mathbf{F}_{M_s[n]}[n])}{card(\mathcal{S}[n])-1}}\right).$$

$\mathcal{P}_T/N/M_s[n]$, the average bit-error rate of the weakest spatial stream $BER_{min}(\mathbf{F}_{M_s[n]}[n])$ is well approximated by the equation at the top of the next page [31]. Consequently, the mode-selection criterion, for the n^{th} subcarrier, reads

$$\begin{cases} M_s[n] = \max_{M_s} \frac{1}{card(\mathcal{S}[n])-1} SINR_{min}(\mathbf{F}_{M_s}[n]) \\ M_s[n] \cdot log_2(card(\mathcal{S}[n])) = \frac{R}{N} \end{cases}. \quad (4)$$

Since $\{\boldsymbol{\mathcal{H}}[n]\}_{1\le n\le N}$ are perfectly known only at the receiver, the multi-mode precoder selection is performed at the receiver, on all subcarriers. The result of this selection is then communicated, via feedback, to the transmitter, which then adapts its transmission format accordingly. The goal of this paper is to propose a quantized multi-mode precoding system solution, which significantly reduces the feedback requirements, or, equivalently, enables a more efficient use of the available feedback bandwidth, compared to state-of-the-art solutions.

III. Multi-Mode Precoding for OFDM-Based Systems with Limited Feedback

A. The limited feedback constraint

As aforementioned in Subsection II-B, given the perfect channel knowledge, $\boldsymbol{\mathcal{H}}[n]$, on each subcarrier n, the receiver identifies the optimal multi-mode precoder, $\mathbf{F}_{M_s[n]}[n]$, to be $\mathbf{V}[n]_{:,1:M_s[n]}$, the eigen-beamformer into the $M_s[n]$ strongest eigen-modes of the MIMO channel $\boldsymbol{\mathcal{H}}[n]$. $M_s[n]$ is optimally selected according to (4). However, the optimal multi-mode precoder solutions $\{\mathbf{F}_{M_s[n]}[n]\}_{1\le n\le N}$ must be fed-back to the transmitter through a low-rate feedback link. This feedback link is assumed to have a capacity of B bits, constraining the maximum number of feedback bits per transmit-CSI update period. Consequently, the optimal multi-mode precoder solutions $\{\mathbf{F}_{M_s[n]}[n]\}_{1\le n\le N}$ must inevitably be quantized such that their feedback can be accommodated by the available B bits of feedback. To characterize the maximum achievable performance of the proposed multi-mode precoding design for OFDM-based systems with limited feedback, we first assume that the feedback link is error free and has zero delay. This assumption is later on relaxed in Section V, where we evaluate the performance of the proposed design in the presence of errors and delay in the feedback link.

B. State-of-the-art quantized multi-mode precoding

Quantized precoding for spatial multiplexing MIMO systems has been explored recently in [18], [23], [24], [19], [22] for precoding matrices with orthonormal columns. These works essentially capitalize on the special structure of the precoders, and their limited number of underlying degrees of freedom to reduce the feedback requirements, compared to direct MIMO channel quantization [32], [33], [34], [35]. More specifically, they propose to directly quantize the limited number of underlying independent parameters of the precoders [19], or, more effectively, to quantize the space of precoding matrices into a finite precoder codebook [18], [23], [24], [22]. More recently, quantized precoding with multi-mode capability has been investigated, which further significantly improves either the BER performance [15], [11], or the capacity [20], [17], [21], [15], [11], [22] of spatial multiplexing MIMO systems. All these contributions invariably advocate a precoder codebook approach, where the precoding matrix $\mathbf{F}$ can be chosen only from an offline-computed finite codebook $\mathcal{F} = \{\mathbf{F}_1, \cdots, \mathbf{F}_{2^B}\}$, given the available B bits of feedback. That is, based on the knowledge of the channel state $\mathbf{H}$, the receiver chooses the optimal $\mathbf{F}_j$, according to a certain selection criterion, and relays the corresponding index back to the transmitter. Nevertheless, these contributions can still be substantially differentiated based on their respective approaches to deploy the multi-mode capability. Indeed, [20], [17], [21], [22] propose to provide multiple precoder codebooks corresponding to multiple modes (i.e number of spatial multiplexing streams) and to switch between them depending on the observed SNR. The additional required complexity and selection of this multiple-codebook approach is avoided by the SNR-independent multi-mode precoder codebook of [15], [11] which contains entries corresponding to different modes.

So far, all state-of-the-art quantized multi-mode precoding solutions have considered flat-fading channels. Their straightforward extension to frequency-selective channels using OFDM would require a multiplicative scaling of the number of feedback bits that is proportional to the number of subcarriers N. This unacceptable amount of feedback bits can be significantly reduced through precoder interpolation [25], [26]. This approach feeds back the indexes of the precoders on a fraction of the subcarriers, and interpolates at the transmitter to recover the precoders on the remaining subcarriers. This frequency down-sampling and interpolation of the precoders is justified by the observed coherence of the precoders on adjacent subcarriers [25], [26]. Unfortunately, interpolation cannot be used in combination with state-of-the-art multi-mode designs [15], [11], [20], [17], [21], [22], due to the fact that these codebooks exhibit a unitary invariance/ambiguity. This unitary invariance, which is inherited from the various codebook design criteria [24], [15], compromises the ordering of the precoder's singular vectors. This ordering is crucial for successful multi-mode precoding on the interpolated precoders. An efficient interpolation-based extension of the existing multi-mode precoder codebooks is consequently not feasible. Hence, we subsequently consider a precoder codebook design that preserves the ordering of the singular vectors, and, consequently, enables an efficient interpolation-based multi-mode precoding solution for OFDM-based systems.

IV. Our Quantized Multi-Mode Precoding System Concept

We essentially propose to construct a precoder codebook corresponding to the vector quantization of the complete $M_T \times M_T$ unitary right singular matrix $\mathbf{V}[n]$ on any subcarrier n, based on the observation that the flat-fading MIMO channels on all subcarriers exhibit the same statistics [36].

This codebook fulfills two fundamental requirements to enable interpolation-based quantized multi-mode precoding. First, it preserves the ordering of the quantized right singular vectors. Second, through quantizing the complete singular matrix, it provides sufficient information to the transmitter for successful interpolation and mode selection. It is computed off-line and known to both the transmitter and the receiver.

Based on the proposed precoder codebook, our interpolation-based multi-mode precoding system is described as follows. Subsequently to estimating the frequency-domain MIMO channels $\{\boldsymbol{\mathcal{H}}[n]\}_n$ and the corresponding optimal unquantized precoders $\{\mathbf{V}[n]\}_n$, the receiver selects and feeds back the indexes of the codewords, which better quantize the optimal precoders *only on a subset of U subcarriers* $\{\mathbf{V}_Q[n_i]\}_{1\leq i\leq U}$. It also identifies and feeds back the optimal mode for every subcarrier based on (4). Upon reception of the feedback information, the transmitter retrieves the relevant codebook entries $\{\mathbf{V}_Q[n_i]\}_{1\leq i\leq U}$, interpolates them to reconstruct the precoders on the remaining subcarriers and enforces the optimal mode on every subcarrier. To also reduce the amount of feedback related to the mode information on all subcarriers, we introduce a simple clustered mode selection approach, which induces only a marginal performance degradation. The design of the precoder codebook is detailed in Subsection IV-A. Precoder interpolation and clustered mode selection are discussed in Subsection IV-B and Subsection IV-C, respectively.

A. Per-subcarrier Precoder codebook design

As aforementioned, the optimal unquantized precoder $\mathbf{V}[n]$, on any subcarrier n, must be quantized into $\mathbf{V}_Q[n]$ due to the limited feedback constraint. Given B_p [2] bits of feedback, $\mathbf{V}_Q[n]$ can only be picked from the 2^{B_p} different but fixed entries in the precoder codebook $\mathcal{V} = \{\mathbf{V}_1, \cdots, \mathbf{V}_{2^{B_p}}\}$ [3]. Determining the optimal $\mathbf{V}_Q[n]$ requires the joint design of the codebook $\mathcal{V} = \{\mathbf{V}_1, \cdots, \mathbf{V}_{card(\mathcal{V})}\}$, and the mapping of each value of $\mathbf{V}[n]$ to one of the codebook entries. To design our per-subcarrier precoder codebook, we assume that the MIMO channel on the n^{th} OFDM subcarrier $\boldsymbol{\mathcal{H}}[n]$ is Rayleigh fading. Hence, the right singular matrix $\mathbf{V}[n]$ is isotropically-distributed on $\mathcal{U}(M_T, M_T)$ [37]. Consequently, the considered precoder codebook design essentially amounts to defining $card(\mathcal{V})$ regions $\{\mathcal{R}_1, \cdots, \mathcal{R}_{card(\mathcal{V})}\}$ in the space of unitary matrices $\mathcal{U}(M_T, M_T)$, spanned by $\mathbf{V}[n]$ [36], and identifying the optimal codebook entry to represent each of these regions

$$\mathbf{V}_Q[n] = \mathbf{V}_j \quad \text{if} \quad \mathbf{V}[n] \in \mathcal{R}_j, \quad j = 1, \cdots, card(\mathcal{V}). \tag{5}$$

A natural approach to determine the quantization regions $\{\mathcal{R}_j\}_{1\leq j\leq card(\mathcal{V})}$ and codebook $\mathcal{V}$ is to minimize the average quantization distortion, measured by the mean squared error between the optimal unquantized precoder $\mathbf{V}[n]$ and its quantized version, over $\mathcal{U}(M_T, M_T)$

$$\{\mathcal{R}_j, \mathbf{V}_j\}_{1\leq j\leq card(\mathcal{V})} = \arg\min_{\mathbf{V}_Q[n]} \quad E\left\{||\mathbf{V}[n] - \mathbf{V}_Q[n]||_F^2\right\}. \tag{6}$$

[2] B_p is the number of feedback bits allocated to the feedback of the square quantized precoder on each of the U pilot subcarriers. Obviously, $U \cdot B_p < B$.

[3] In all the following, 2^{B_p} will be replaced by $card(\mathcal{V})$.

The previous MMSE optimization problem can be further developed into

$$\{\mathcal{R}_j, \mathbf{V}_j\}_{1\leq j\leq card(\mathcal{V})} = \arg\min_{\{\mathcal{R}_k, \mathbf{V}_k\}_k} \sum_{k=1}^{card(\mathcal{V})} E\{||\mathbf{V}[n] - \mathbf{V}_k||_F^2 | \mathbf{V}[n] \in \mathcal{R}_k\} p(\mathbf{V}[n] \in \mathcal{R}_k) \tag{7}$$

where $E\{.|\mathbf{V}[n] \in \mathcal{R}_k\}$ denotes the expectation conditioned on the event that $\mathbf{V}[n]$ is in the quantization region $\mathcal{R}_k$, and $p(\mathbf{V}[n] \in \mathcal{R}_k)$ stands for the probability of this event. The MMSE optimization problem of (7) cannot be solved in closed-form due to the geometrical complexity of the partition regions $\{\mathcal{R}_j\}_{1\leq j\leq card(\mathcal{V})}$. Thus, we resort to the iterative Lloyd's algorithm as in [16], [20], which sequentially optimizes the quantization regions $\{\mathcal{R}_j\}_{1\leq j\leq card(\mathcal{V})}$ and $\{\mathbf{V}_j\}_{1\leq j\leq card(\mathcal{V})}$. It is interesting to note that, for each iteration of the quantization regions $\{\mathcal{R}_j\}_{1\leq j\leq card(\mathcal{V})}$ in $\mathcal{U}(M_T, M_T)$, the corresponding codebook entries are easily identified as [16]

$$\mathbf{V}_j = \arg\min_{\mathbf{V}\in\mathcal{U}(M_T,M_T)} E\{||\mathbf{V}[n] - \mathbf{V}||_F^2 | \mathbf{V}[n] \in \mathcal{R}_j\} \tag{8}$$

with $j = 1, \cdots, card(\mathcal{V})$. In other words, each $\mathbf{V}_j$ is simply the *mean unitary matrix in the Euclidean sense* [4] *over the quantization region $\mathcal{R}_j$*. This mean unitary matrix conveniently turns out to be the orthogonal projection of the arithmetic mean over $\mathcal{R}_j$ in the linear space $\mathcal{C}^{M_T\times M_T}$, onto the space of unitary matrices $\mathcal{U}(M_T, M_T)$ [38]. This crucial result is exploited to allow for simple analytical solutions inside the iterations of the Lloyd algorithm, as subsequently illustrated. The codebook design steps are as follows:

1) We generate a training set with N_{tr} independent and isotropic $M_T \times M_T$ unitary matrices $\{\mathbf{T}^{(k)}\}_{1\leq k\leq N_{tr}}$.
2) Starting with an initial codebook $\mathcal{V} = \{\mathbf{V}_1, \cdots, \mathbf{V}_{card(\mathcal{V})}\}$ (randomly chosen, or, if available, the best codebook of the previous iteration), the two following sub-steps are iteratively performed.
   - We assign each $\mathbf{T}^{(k)}$ to one of the regions using the rule
   $$\mathbf{T}^{(k)} \in \mathcal{R}_i \quad \text{if} \quad \|\mathbf{T}^{(k)} - \mathbf{V}_i\|_F^2 < \|\mathbf{T}^{(k)} - \mathbf{V}_j\|_F^2, \forall j \neq i. \tag{9}$$
   The training unitary matrices $\mathbf{T}^{(k)}$, for which equality occurs, can be assigned either way as this does not change our performance measure. This first sub-step is referred to as *the nearest neighbor rule* [39], and defines the regions $\{\mathcal{R}_j\}_{1\leq j\leq card(\mathcal{V})}$.
   - For each region $\mathcal{R}_j$, we find the optimal codebook according to (8). As aforementioned, $\mathbf{V}_j$ conveniently turns out to be the orthogonal projection of the arithmetic mean over $\mathcal{R}_j$ in $\mathcal{C}^{M_T\times M_T}$, onto the space of unitary matrices $\mathcal{U}(M_T, M_T)$ [38]. Let the SVD of the arithmetic mean over $\mathcal{R}_j$ be
   $$[\mathbf{A}_j, \mathbf{S}_j, \mathbf{B}_j] = \text{SVD}\left(\frac{1}{N_{tr,j}} \sum_{\mathbf{T}^{(k)}\in\mathcal{R}_j} \mathbf{T}^{(k)}\right) \tag{10}$$

[4] the mean is associated with the Euclidean distance $\text{d}(\mathbf{A}, \mathbf{B}) = ||A - B||_F$.

where $N_{tr,j}$ denotes the fraction of the training set's elements which has been assigned to $\mathcal{R}_j$ following the nearest neighbor rule of (9). The projection of the arithmetic mean over $\mathcal{R}_j$, onto the space of unitary matrices $\mathcal{U}(M_T, M_T)$ is simply [40], [38]

$$\mathbf{V}_j = \mathbf{A}_j \mathbf{B}_j^H. \tag{11}$$

This second sub-step is called *the centroid condition* [39], and identifies the codebook entries $\{\mathbf{V}_j\}_{1 \le j \le card(\mathcal{V})}$.

3) In general, there is no guarantee that Lloyds' algorithm will converge to the global optimum. In our simulations, we repeat each codebook iteration 100 times, each time starting with a random set of initial centroids. At the end of the algorithm, we pick the one that gives us the minimum average quantization distortion.

Based on the resulting precoder codebook, the optimal mapping rule of the unquantized precoder $\mathbf{V}[n]$ is then simply

$$\mathbf{V}_Q[n] = \arg \min_{\{\mathbf{V}_j\}_{1 \le j \le card(\mathcal{V})}} ||\mathbf{V}[n] - \mathbf{V}_j||_F^2. \tag{12}$$

The Lloyd algorithm is carried out off-line and ahead of time to determine the precoder codebook. The resulting precoder codebook is then stored at both the transmitter and the receiver. Note that the underlying assumption that the linear precoder $\mathbf{V}[n]$, on subcarrier n, is isotropically-distributed in $\mathcal{U}(M_T, M_T)$ was shown to hold for delay-spread MIMO channels whose time-domain taps are spatially-uncorrelated circularly symmetric zero-mean complex Gaussian variables [36][5]. Nevertheless, this assumption still makes sense in the presence of other types of delay-spread channels, as the isotropical distribution represents a worst-case distribution for the linear precoder $\mathbf{V}[n]$.

Finally, a similar MMSE quantization criterion was previously used in [16] to quantize the waterpouring transmit covariance matrix. However, the resulting quantization regions and codebook were only used as "good" initial values that are further on redefined based on a capacity criterion.

B. Precoder Interpolation

As previously announced, the precoder codebook $\mathcal{V}$, is now used to efficiently feedback the $M_T \times M_T$ unitary precoders only on a limited set of U subcarriers, subsequently referred to as *pilot subcarriers*. Based on these pilot precoders, we seek to recover the unitary precoders on the remaining subcarriers through interpolation. This is reasonable because the frequency correlation exhibited by the MIMO channels across subcarriers was shown [25], [26] to hold for the unitary precoders on these subcarriers. More specifically, we propose a novel *conditional* interpolation strategy that exploits this frequency correlation or smoothness to interpolate the available unitary pilot precoders $\{\mathbf{V}_Q[n_i]\}_{1 \le i \le U}$, under a unitary constraint. It is inspired from pilot-aided linear MIMO channel estimation for OFDM.

Our conditional interpolation approach first considers channel interpolation, based on the MIMO channel matrices acquired on the U pilot subcarriers. Based on the results of this channel interpolation and the knowledge of the structure of the optimal precoders on the remaining subcarriers, this approach tries to identify an "inherited" precoder interpolation. Based on the MIMO channels on the U pilots, it is easy to reconstruct the MIMO channel on the k^{th} subcarrier as [41]

$$\boldsymbol{\mathcal{H}}[k] = \sum_{i=1}^{U} (\mathbf{T}\mathbf{T}_U^\dagger)_{k,n_i} \boldsymbol{\mathcal{H}}[n_i] \tag{13}$$

where $\mathbf{T}$ represents the $N \times N$ DFT matrix and $\mathbf{T}_U$ is the $U \times N$ partial DFT matrix which corresponds to the U pilot positions. For notational brevity, we subsequently write $\alpha_{k,n_i} = (\mathbf{T}\mathbf{T}_U^\dagger)_{k,n_i}$. Since U as well as the position of the pilots are known both at the transmitter and the receiver, these parameters $(\alpha_{k,n_i})_{1 \le k \le N; 1 \le i \le U}$ are also known at the transmitter. Based on the previous interpolation expression, we try to extract the optimal precoder on the k^{th} subcarrier based on the knowledge of the precoders on the U pilots. Since the optimal precoder on the k^{th} subcarrier is given by $\mathbf{V}[k]$, where $\mathbf{V}[k]$ contains the eigenvectors of $\boldsymbol{\mathcal{H}}^H[k]\boldsymbol{\mathcal{H}}[k]$, we now detail the expression of $\boldsymbol{\mathcal{H}}^H[k]\boldsymbol{\mathcal{H}}[k]$ based on (13)

$$\begin{aligned}\boldsymbol{\mathcal{H}}^H[k]\boldsymbol{\mathcal{H}}[k] =& \sum_{i=1}^{U} |\alpha_{k,n_i}|^2 \mathbf{V}[n_i]\boldsymbol{\Sigma}^2[n_i]\mathbf{V}^H[n_i] \\ &+ \sum_{i \ne j} \alpha_{k,n_i}^* \alpha_{k,n_j} \boldsymbol{\mathcal{H}}^H[n_i]\boldsymbol{\mathcal{H}}[n_j].\end{aligned} \tag{14}$$

Clearly, the calculation of the optimal precoder on the k^{th} subcarrier would require not only the knowledge of the precoders on the pilots $(\mathbf{V}[n_i])_{1 \le i \le U}$ but also the knowledge of the eigenvalues $(\boldsymbol{\Sigma}^2[n_i])_{1 \le i \le U}$ and that of the complete SVD of $(\boldsymbol{\mathcal{H}}^H[n_i]\boldsymbol{\mathcal{H}}[n_j])_{i<j}$. Since the former information is the only one available at the transmitter, we propose to consider the optimal precoder conditioned on the knowledge of the precoders on the U pilots. Rather than considering (14), the right expression to be evaluated is $E_{cond}\{\boldsymbol{\mathcal{H}}^H[k]\boldsymbol{\mathcal{H}}[k]\}$, where $E_{cond}\{.\}$ denotes $E_{\boldsymbol{\mathcal{H}}|(\mathbf{V}[n_i])_{1 \le i \le U}}\{.\}$. This expression can be shown to reduce to (cf. Appendix)

$$E_{cond}\{\boldsymbol{\mathcal{H}}^H[k]\boldsymbol{\mathcal{H}}[k]\} = \sum_{i=1}^{U} |\alpha_{k,n_i}|^2 \mathbf{V}[n_i] E_{cond}\{\boldsymbol{\Sigma}^2[n_i]\}\mathbf{V}^H[n_i].$$

Since the calculation of $E_{cond}\{\boldsymbol{\Sigma}^2[n_i]\}$ only requires the knowledge of the channel statistics, it can easily be acquired or made available beforehand at the transmitter. Finally, the optimal precoder, given only the knowledge of the precoders on the pilots, is given by

$$\mathbf{T}_{opt} = \text{eigenvectors of} \left(\sum_{i=1}^{U} |\alpha_{k,n_i}|^2 \mathbf{V}[n_i] E_{cond}\{\boldsymbol{\Sigma}^2[n_i]\}\mathbf{V}^H[n_i] \right). \tag{15}$$

As a final step, we calculate $E_{cond}\{\boldsymbol{\Sigma}^2[n_i]\}$ or equivalently $E_{\boldsymbol{\mathcal{H}}[n_i]}\{\boldsymbol{\Sigma}^2[n_i]\}$ through using the joint probability density function of the ordered eigenvalues of $\boldsymbol{\mathcal{H}}^H[n_i]\boldsymbol{\mathcal{H}}[n_i]$ [42]. It is worthwhile mentioning that this conditional interpolation is invariant with respect to the orientation of the singular vectors. As such, it avoids the additional optimization of these orientations needed by other state-of-the-art interpolators [26], [43].

[5]Such as the IEEE 802.11 TGn non line-of-sight indoor MIMO channels considered in Section V.

C. Clustered Mode Selection

So far, based on the precoder codebook of Subsection IV-A, the transmitter has retrieved the $M_T \times M_T$ quantized precoding matrices on the U pilot subcarriers, $\{\mathbf{V}_Q[n_i]\}_{1\leq i\leq U}$, whose indexes had been fed-back from the receiver using $U \cdot log_2(card(\mathcal{V}))$ bits. Furthermore, it has interpolated these U unitary pilot precoding matrices, under a unitary constraint, to recover the $M_T \times M_T$ precoding matrices on all subcarriers, $\{\mathbf{V}_Q[n]\}_{1\leq n\leq N}$. To be able to deploy multi-mode precoding, the transmitter additionally requires the feedback of the optimal number of streams on each subcarrier. Ideally, the receiver would identify the optimal number of streams, $M_s[n]$, or equivalently the optimal number of first columns of the interpolated precoding matrix $\mathbf{V}_Q[n]$, to be used on each subcarrier n, according to (4). The additional feedback required by this per-subcarrier mode-selection strategy amounts to $N \cdot \lceil log_2(\min(M_T, M_R))\rceil$, which turns out to be too high, for practical values of N.

Consequently, we alternatively propose a *clustered mode-selection* approach to significantly reduce the feedback overhead related to mode selection. This approach capitalizes on the frequency correlation of the MIMO channels as well as the precoders across subcarriers, to cluster adjacent subcarriers and use the optimal number of streams related to the middle subcarrier for the whole cluster. Without loss of generality, we assume that the U pilot subcarriers are equidistant. Moreover, we consider the U resulting subcarrier-clusters for the application of our clustered mode selection. The related feedback overhead is then reduced to $U \cdot \lceil log_2(\min(M_T, M_R))\rceil$. Finally, the overall feedback overhead required by multi-mode quantized precoding is given by $B = U \cdot log_2(card(\mathcal{V})) + U \cdot \lceil log_2(\min(M_T, M_R))\rceil$, where $card(\mathcal{V})$ is the cardinality of the precoder codebook, $\mathcal{V}$. The choice of the parameters B, U and $card(\mathcal{V})$ will obviously impact the performance of our interpolation-based multi-mode precoding system. This impact is subsequently investigated for various MIMO system dimensions.

V. Performance Results

In this section, we investigate the BER performance of our multi-mode precoding solution for spatial multiplexing MIMO-OFDM systems with limited feedback. More specifically, we first illustrate the limited BER performance degradation introduced by our quantized multi-mode precoding solution, when the feedback bandwidth is large enough to accommodate the quantized information on all subcarriers. Then, we proceed to confirm that multi-mode precoding still outperforms the conventional fixed-mode precoding solutions, when the feedback is limited such that precoder interpolation and clustered multi-mode selection are required. In all the following, we consider a MIMO-OFDM system with $N = 64$ subcarriers and a guard interval of $\nu = 16$ samples. Furthermore, we use the MIMO channel model provided by the IEEE 802.11 TGn [44] assuming the following parameters: channel model B for the downlink and non line-of-sight propagation, antenna spacing at both the transmitter and the receiver is λ, where λ is the carrier wavelength at 5.2 GHz, and a sampling rate of 20 MHz. At this sampling rate, channel model B (rms delay spread 15 ns) exhibits $L = 10$ samples. Finally, the BER plots were obtained by averaging over more than 1000 channel realizations.

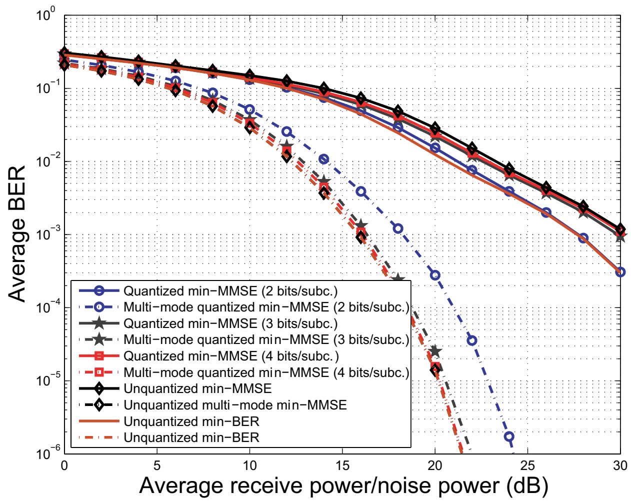

Fig. 2. Average uncoded BER comparison for a 2×2 MIMO-OFDM set-up at $R = 64$ Mbps with channel B and $U = 64$ pilot precoders and $card(\mathcal{V}) = \{4, 8, 16\}$.

Impact of precoder quantization: To evaluate the individual impact of our codebook-based precoder quantization on the performance of multi-mode precoding, we consider a scenario where the feedback bandwidth is large enough to accommodate the indexes of the quantized precoders, in the off-line computed precoder codebook, as well as the mode information on all $N = 64$ subcarriers. Figure 2 illustrates the BER performance degradation due to the quantization of the unitary precoders using $\{2, 3, 4\}$ bits per subcarrier (bits/subc.), for a 2×2 MIMO set-up. It confirms that the larger the quantization word-length, and, equivalently, the cardinality of the precoder codebook, $card(\mathcal{V})$, the smaller the induced BER degradation. This is clearly observed in the performance of multi-mode precoding, where the dominant mode is single-stream transmission. In the fixed-mode precoding, the uncoded BER performance appears to be degraded when the quantization accuracy is increased. This can be explained by the fact that a worse accuracy, in fact, means an imperfect diagonalization of the channel, which basically amounts to a kind of linear precoding of the two transmit streams across the two eigenmodes of the 2×2 MIMO channel. Such "unintentional" linear precoding was shown to exhibit a better uncoded BER performance, yet, a worse coded BER performance [10].

Figure 3 shows similar results, for a 3×2 MIMO set-up using $\{4, 5, 6\}$ bits per subcarrier. More importantly, Figures 2 and 3 show that as few as 2 and 4 bits per subcarrier are required to quantize 2×2 and 3×3-dimensional unitary matrices, respectively, for reasonable multi-mode precoding performance. It is worthwhile highlighting that the superior performance of multi-mode precoding over conventional fixed-mode precoding is maintained whatever the resolution of the precoder quantization is. Consequently, these exhibited results assess the effectiveness of codebook-based precoder quantization as a means of efficiently representing the unitary precoders, in view of their feedback to the transmitter. The analytical impact analysis of the limited-feedback constraint

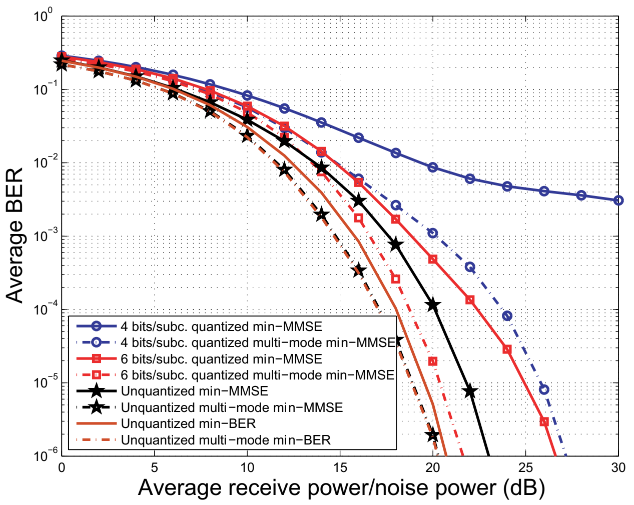

Fig. 3. Average uncoded BER performance for a 3×2 MIMO-OFDM set-up at $R = 64$ Mbps with channel B and $U = 64$ pilot precoders and $card(\mathcal{V}) = \{16, 32, 64\}$.

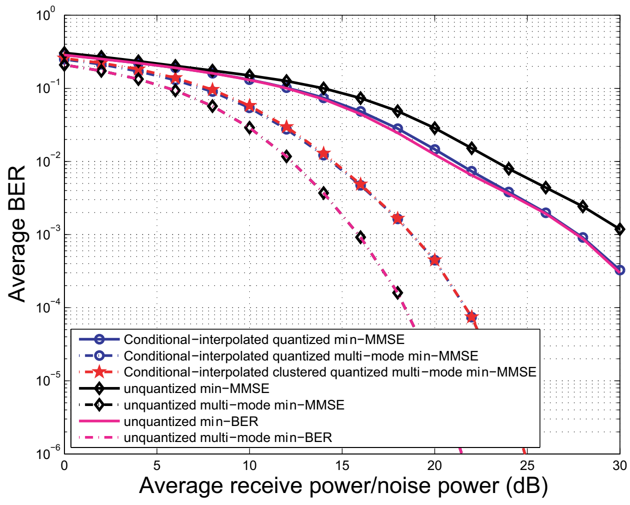

Fig. 4. Average uncoded BER comparison for a $(2, 2)$ MIMO-OFDM set-up at rate 64 Mbps with channel B and 8 pilot precoders and $card(\mathcal{V}) = 4$.

on the system BER/MMSE performance is the topic of our current investigations.

Although we have focused on designing quantized min-MMSE precoding, minimum BER performance can be achieved provided that we feedback to the transmitter on each subcarrier n, in addition to the precoder entry and the mode information, whether a constant-modulus unitary matrix $\mathbf{Q}[n]$ is required or not to enforce the min-BER property on that subcarrier [29]. Note that we would need to feedback this information on all N subcarriers, which would substantially increase the feedback overhead.

Impact of quantized precoder interpolation: In combination with codebook-based precoder quantization, we now introduce precoder frequency down-sampling and interpolation, as an additional feedback-reduction measure. We illustrate our proposed conditional interpolator. Obviously, the performance of our proposed interpolator depends on the number of used pilot precoders, U, and how this number compares to the length of the delay-spread channel, L. For $U \geq L$, the interpolator will succeed in reconstructing the precoders, based on the pilots. Whereas, when $U < L$, the interpolator will make errors on the precoders and will consequently lead to a degradation of the BER performance. The latter case is clearly the most relevant, as it allows a higher reduction in the feedback requirements. Consequently, the illustrated performance results are dedicated to scenarios, where $U < L$. Figure 4 depicts the average BER performance of multi-mode selection when our proposed conditional interpolation solution is used to recover the unitary precoders, based on only $U = 8$ pilot precoders, drawn from a precoder codebook of cardinality $card(\mathcal{V}) = 4$, for IEEE 802.11 TGn channel B.

As it turns out, the BER performance of our conditional interpolator falls within 0.6 dB SNR loss at BER=10^{-3}, with respect to the non-interpolated quantized multi-mode performance of Figure 2. The latter performance, itself, exhibits a 3 dB SNR degradation, compared to the unquantized multi-mode solution. The marginal BER performance loss, related to this 8-fold precoder down-sampling, illustrates the feedback-reduction potential of capitalizing on the precoders correlation across frequency, to deploy precoder frequency down-sampling combined with our proposed interpolator. Figure 4 also confirms that our interpolation-based quantized multi-mode precoding solution remains attractive, with respect to fixed-mode solutions. It is worthwhile mentioning that we have previously shown that our conditional interpolator is particularly attractive in highly frequency-selective channels [43]. Indeed, we have shown that the conditional interpolator significantly outperforms other state-of-the-art precoder interpolators, namely the geodesic interpolator of [43] and the projection-based interpolator of [26]. This is due to the fact that the conditional interpolator exploits all U pilots to reconstruct the intermediate precoders, while the other two interpolators only use 2 adjacent pilots. This difference may not be relevant for rather frequency-flat channels, it is however fundamental in highly frequency-selective channels.

Impact of clustered mode selection: In all previous simulations, we have assumed that the feedback bandwidth is large enough to accommodate the mode information on all subcarriers,i.e, $N \cdot \lceil log_2(\min(M_T, M_R)) \rceil$ bits. We now include our clustered mode selection proposal to further decrease this amount of feedback to $U \cdot \lceil log_2(\min(M_T, M_R)) \rceil$. Figure 4 shows that the introduced clustered mode selection hardly degrades the BER performance, compared to feeding back the mode information on all subcarriers, for a 2×2 MIMO system using channel model B, with $card(\mathcal{V}) = 4$ and $U = 8$. This is due to the existence of a dominant mode, which will be selected on most subcarriers. The existence of this dominant mode was evidenced in [45] for any given $M_T \times M_R$ i.i.d. flat-fading MIMO channel and any given rate R. Since the MIMO channels across the OFDM subcarriers are well characterized as i.i.d. flat-fading [36], they consequently exhibit a common dominant mode. Thus, our clustered mode selection will only lead to a marginal BER performance degradation. The total feedback required, in this case, amounts to $B = 8 \cdot 2 + 8 \cdot 1 = 24$ bits, compared to $64 \cdot 2 + 64 \cdot 1 = 192$ bits without interpolation and clustered mode selection.

Impact of channel estimation errors and feedback delay: In all previous simulations, we have considered an error-free zero-delay feedback link, to focus on the impact of precoder

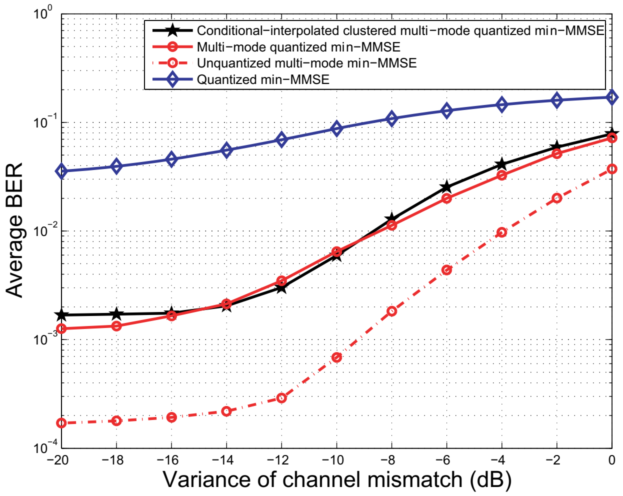

Fig. 5. Average uncoded BER comparison for a (2, 2) MIMO-OFDM set-up at rate 64 Mbps with channel B and 8 pilot precoders and $card(\mathcal{V}) = 4$, at SNR=15 dB in the presence of channel mismatches.

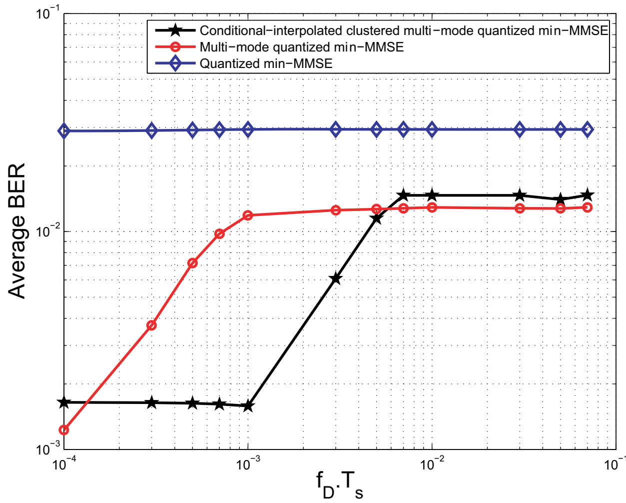

Fig. 6. Average uncoded BER performance for a (2, 2) MIMO-OFDM set-up at rate 64 Mbps with channel B and 8 pilot precoders and $card(\mathcal{V}) = 4$, at SNR=15 dB in the presence of feedback delay.

quantization and interpolation errors on the performance of our proposed interpolation-based multi-mode quantized precoding scheme. In real systems, however, there are also errors due to channel estimation errors at the receiver and/or delays in the feedback channel. To study this effect, we subsequently use the conditional channel mean feedback model introduced in [32], which characterizes the unknown actual CSI $\boldsymbol{\mathcal{H}}[n]$, at subcarrier n, in terms of the available imperfect CSI $\hat{\boldsymbol{\mathcal{H}}}[n]$. Under the assumption of dense scattering in the vicinity of both transmitter and receiver, the conditional channel mean feedback model is described as

$$\boldsymbol{\mathcal{H}}[n] \sim \mathcal{N}^c\left(\rho\hat{\boldsymbol{\mathcal{H}}}[n], \sigma_h^2(1-|\rho|^2)\mathbf{I}_{M_R M_T}\right) \quad (16)$$

where ρ is the common correlation between the coefficients of the true and the available imperfect CSI, and σ_h^2 is the common variance of the channel coefficients. The correlation coefficient ρ depends on the nature of the available imperfect CSI (i.e. estimated, outdated or predicted). Based on (16), the variance of the mismatch between $\boldsymbol{\mathcal{H}}[n]$ and $\hat{\boldsymbol{\mathcal{H}}}[n]$ is given by

$$E\left\{\left(\boldsymbol{\mathcal{H}}[n]-\hat{\boldsymbol{\mathcal{H}}}[n]\right)(\boldsymbol{\mathcal{H}}[n]-\hat{\boldsymbol{\mathcal{H}}}[n])^H\right\} = \left(2-2|\rho|^2\right)\mathbf{I}_{M_R}. \quad (17)$$

Figure 5 reports the BER degradation caused by the channel mismatches due to estimation errors or feedback delay at SNR=15 dB. Clearly, the BER performance of all linear precoding schemes degrades in the presence of channel/CSI mismatches. Moreover, the BER advantage provided by our multi-mode precoding, over fixed-mode linear precoding, reduces with increasing channel mismatch variance. Nonetheless, Figure 5 confirms that multi-mode precoding remains beneficial even at large channel mismatches. It also corroborates that our interpolation-based clustered mode selection quantized precoding approach consistently leads to very little degradation with respect to multi-mode quantized precoding on all subcarriers ($U = 64$).

Figure 6 illustrates and compares the uncoded BER performance of all considered precoding schemes in time-varying channels. Note that $f_D T_s$ denotes the Doppler frequency multiplied by the OFDM symbol duration T_s. It is assumed that all precoding methods use the same feedback ratio. Since these schemes have different feedback requirements, a fair comparison implies the use of different frame lengths for each scheme. Quantized multi-mode precoding on all subcarriers, for example, uses a very long frame length since it has more bits to send while our interpolation-based clustered multi-mode precoding can use a very short frame length and thus update the precoders more often. Taking into account the feedback requirements discussed earlier in this Section, the frame length was set to 40, 212 and 320 symbols for our proposed approach (24 bits of feedback), fixed-mode quantized precoding (128 bits) and multi-mode quantized precoding without interpolation and clustering (192 bits), respectively. Our proposed approach is shown to be the most advantageous of the three techniques over a large range of feedback delays, which actually corresponds to the relevant range for using linear precoding techniques.

VI. Conclusions

We have proposed a novel multi-mode precoding system concept for MIMO-OFDM systems with limited feedback. It combines 3 key features to minimize the feedback requirements with limited BER performance penalty. First, a precoder codebook design that efficiently quantizes the optimal unitary precoder on each subcarrier and, more importantly, lends itself to precoder interpolation across subcarriers for further feedback reduction. Second, a novel conditional interpolator that actually performs this precoder interpolation to reconstruct the precoding matrices on all the subcarriers, based on the feedback of the quantized precoders only on a fraction of the subcarriers. Finally, a clustered mode selection approach, which enables a significant reduction of the feedback overhead required by the mode information on each subcarrier. Monte-Carlo simulations have been provided that show that the proposed multi-mode precoding solution exhibits good performance at reasonable feedback requirements. The analytical impact analysis of the limited feedback constraint on the system BER/MMSE performance is the topic of our current investigations.

APPENDIX

Based on (14), $E_{cond}\{\boldsymbol{\mathcal{H}}^H[k]\boldsymbol{\mathcal{H}}[k]\}$ develops into

$$E_{cond}\{\boldsymbol{\mathcal{H}}^H[k]\boldsymbol{\mathcal{H}}[k]\} = \sum_{i=1}^{U} |\alpha_{k,n_i}|^2 \mathbf{V}[n_i] E_{cond}\{\mathbf{\Sigma}^2[n_i]\}\mathbf{V}^H[n_i] + \sum_{i\neq j} \alpha^*_{k,n_i}\alpha_{k,n_j} E_{cond}\{\boldsymbol{\mathcal{H}}^H[n_i]\boldsymbol{\mathcal{H}}[n_j]\} \quad (18)$$

where $E_{cond}\{.\}$ denotes $E_{\boldsymbol{\mathcal{H}}|(\mathbf{V}[n_i])_{1\leq i\leq U}}\{.\}$. We now explicitly calculate, the conditional part of the second term in the sum in (18). For $i \neq j$, we can write

$$E_{cond}\{\boldsymbol{\mathcal{H}}^H[n_i]\boldsymbol{\mathcal{H}}[n_j]\} = \mathbf{V}[n_i] E_{cond}\{\mathbf{\Sigma}[n_i]\mathbf{U}^H[n_i]\mathbf{U}[n_j]\mathbf{\Sigma}[n_j]\}\mathbf{V}^H[n_j]. \quad (19)$$

Exploiting the statistical independence of the singular values and the singular vectors of i.i.d Rayleigh-fading MIMO channels [37], [36] and assuming that the i^{th} and j^{th} pilot subcarriers are far apart such that their fadings are independent, we can further develop (19) into

$$E_{cond}\{\boldsymbol{\mathcal{H}}^H[n_i]\boldsymbol{\mathcal{H}}[n_j]\} = \mathbf{V}[n_i] E_{cond}\{\mathbf{\Sigma}[n_i]\} E_{cond}\{\mathbf{U}^H[n_i]\} \cdot E_{cond}\{\mathbf{U}[n_j]\} E_{cond}\{\mathbf{\Sigma}[n_j]\}\mathbf{V}^H[n_j].$$

Recalling that, for the i.i.d Rayleigh-fading MIMO channel $\boldsymbol{\mathcal{H}} = \mathbf{U}\mathbf{\Sigma}\mathbf{V}^H$, $\mathbf{U}$ and $\mathbf{V}$ are isotropically-distributed [37], [36] in the group of unitary matrices $\mathcal{U}(M_R, M_R)$ and $\mathcal{U}(M_T, M_T)$, respectively, we can state that $E_{cond}\{\mathbf{U}^H[n_i]\} = \mathbf{0}_{M_R}$. Consequently, (19) finally reduces to $E_{cond}\{\boldsymbol{\mathcal{H}}^H[n_i]\boldsymbol{\mathcal{H}}[n_j]\} = \mathbf{0}_{M_T}$. Hence, (18) becomes

$$E_{cond}\{\boldsymbol{\mathcal{H}}^H[k]\boldsymbol{\mathcal{H}}[k]\} = \sum_{i=1}^{U} |\alpha_{k,n_i}|^2 \mathbf{V}[n_i] E_{cond}\{\mathbf{\Sigma}^2[n_i]\}\mathbf{V}^H[n_i].$$

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