Prioritized Multi-layer Transmission for MIMO-OFDM Multiuser Broadcast Channel

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Abstract— **In this paper, three different users, with three different qualities of service (QoSs) are allowed to share the same frequency using different space and modulation layers. A multiple-input multiple-output (MIMO) antenna transceiver combined with orthogonal frequency division multiplexing (OFDM) is implemented to allow space-frequency multiple access. Moreover, the users' symbols are allowed to share the same frequency via embedding their constellations in a hierarchical fashion (with non-homogeneous constellation sizes). This ensures arbitrary QoS granularities and guarantees the same information transmission for each user. The latter characteristic can be exploited further to limit the multiuser interference (MUI) in conjunction with an optimized pre- and post-filter design. Simulation results show a fixed separation at different channel state information (CSI) accuracies.**

I. Introduction

Space division multiple access (SDMA) exploits the available spatial channel dimensions in case of multiple antenna transmission. Thereby, in a multiuser environment, potentially high spectral efficiencies can be achieved. However, due to sharing the frequency and time resources, MUI becomes a crucial limiting factor that determines the system performance. Therefore, the pre- and post-filter have to be jointly optimized in order to limit this interference. Several research works already consider this problem, [2]-[**?**]. The block-diagonalization scheme avoids the MUI interference completely by allowing each user to transmit multiple data streams on the other users' null-spaces [2] and [3]. This changes the problem to an equivalent one with parallel (non-interfering) users. In this particular case, the total number of transmit antennas at the base station (BS) must be greater than or (at least) equal to the sum of all receive antennas at the users' mobile sets (MS); providing sufficiently many null-space vectors [2]. The minimization of the MUI mean square error (MSE), [5]-[**?**], has been proposed to keep the interference as low as possible under a fixed power constraint. This scheme relaxes the constraint regarding the number of antennas mentioned before. However, increasing the number of users or their receive antennas (strictly speaking, their data streams) requires an additional coding scheme [4], which is not considered in our work.

Certainly, optimizing the pre-processing matrix requires the current channel state information (CSI) to be readily available at the BS. In a time division duplexing (TDD) regime, the reciprocity between the uplink and the downlink can be exploited to guarantee the presence of CSI values based on the existing reverse channel estimation without any extra feedback information. However, long reception durations may result in channel delays. This leads to erroneous CSI which can be a serious threat for keeping the users orthogonal (block-diagonalized). Nevertheless, it could be sufficient to keep the performance acceptable for the most important users by exploiting their stronger subchannels or automatically switch to another multiplexing scheme, i.e., switch to frequency division multiple access (FDMA).

Our QoS criterion is fulfilled by devoting arbitrary data-rates and arbitrary SER to each user. This is realized by modifying the sub-optimal hierarchical bit-loading algorithm in [1] to devote different constellation sizes to each user to fulfill the required SER. Without loss of generality, the SER steps between users are chosen to fulfill the arbitrary margin separation proposed in [1], which can be further modified to suite many practical cases. According to this scheme, the highest priority user is allowed first to consume the stronger subcarriers with a relatively low modulation order and a suitable symbol energy that fulfills his SER requirements. The next users are allowed to embed their data on the already used subcarriers (in a hierarchical fashion as in [1]) or consume the unused ones based on the required SER and the given channel knowledge. This means that all users share the same subcarriers are receiving, physically, the same information symbols (multicast) with different hierarchical code book. Certainly, this criterion minimizes the MUI in case of CSI uncertainties. Furthermore, the strongest eigenvalue of the orthogonal block diagonalized matrix can be utilized to ensure robust transmission for the given data streams. This physically reduces the interuser-intereigen interference that could influence the weaker eigenchannels. However, a robust adaptation can be designed by considering the channel uncertainties during the bit-loading process. For the sake of comparison, the multicast scheme proposed in [13] has been modified to accommodate hierarchical multiuser multiplexing satisfying a minimum mean square error (MMSE) criterion.

The rest of this paper is organized as follows. Section 2 discusses the block diagonalized multiuser MIMO-OFDM system and the MMSE multicast scheme. Section 3 discusses our adaptive hierarchical modulation. Section 4 discusses some results. Finally, we conclude our findings in the last section.

II. Multiuser MIMO-OFDM Adaptive Model

A. Channel Model

We consider an $(MN_R) \times N_T$ MIMO channel matrix $\mathbf{H}_{k,u}$, where N_R is the number of the receive antennas of each user, M is the number of users, N_T is the number of the transmit antennas, k is the subcarrier index, and u is the user index. The total number of subcarriers is N. We assume a partial CSI with the delayed channel version $\overline{\mathbf{H}}_{k,u}$. The error matrix can be defined as $\boldsymbol{\Xi}_{k,u} = \mathbf{H}_{k,u} - \overline{\mathbf{H}}_{k,u}$, where $\boldsymbol{\Xi} \in \mathcal{C}\ \mathcal{N}(0, \sigma^2_{\boldsymbol{\Xi}}\mathbf{I})$ [1]. The channel matrix entries are uncorrelated zero mean circularly symmetric complex Gaussian (ZMCSCG) values and modeled as independent Rayleigh fading blocks with exponential decaying [7].

B. Formulation and Null-Space Constraint

Assuming OFDM modulation, the received vector at the i^{th} receiver and the k^{th} subcarrier is given by

$$\mathbf{Y}_{k,i} = \mathbf{H}_{k,i}\mathbf{F}_{k,i}\mathbf{X}_{k,i} + \overbrace{\mathbf{H}_{k,i}\sum_{u=1,u\neq i}^{M}\mathbf{F}_{k,u}\mathbf{X}_{k,u}}^{\text{MUI}} + \mathbf{N}_{k,i} \tag{1}$$

where $\mathbf{F}_{k,i}$ is the pre-processing matrix at the transmitter side, $\mathbf{X}_{k,i}$ is the transmitted vector, and $\mathbf{N}_{k,i}$ is a zero mean additive white Gaussian noise (AWGN) with a variance σ^2_n. To simplify our notations, we henceforth omit the subcarrier indices k.

To approach the sum capacity of the broadcast channel (BC) with a multiuser MIMO system, the MUI has to be completely mitigated [6]. Precoding with block diagonalization is one of the schemes that completely suppresses the MUI by projecting each user into the null-space of the others [2], i.e., $\mathbf{H}_{u\neq i}\mathbf{F}_i = \mathbf{0}$. This means that $\mathbf{F}_i$ spans the L null spaces (zero eigenvalues) of the matrix

$$\mathbf{H}_i^{\text{null}} = \left[\mathbf{H}^T_{u=1} \cdots \mathbf{H}^T_{u\neq i} \cdots \mathbf{H}^T_{u=M}\right]^T \ \forall\ u \neq i\ , \tag{2}$$

where $L \leq N_R$. The downside of this null-space approach is the strict requirement that allows only the total number of receive antennas (MN_R) to be less than or equal to the number of transmit antennas. In addition, $\mathbf{F}_i^{(N_T\times L)}$ contains the L eigenvectors (across its columns) spanning the L zero eigenvalues of $\mathbf{H}_i^{\text{null}}$.

Accordingly, the optimal power- and bit-loading are computed based on the eigenvalue decomposition of the new resultant channel

$$\mathbf{H}^\star_{k,i} = \mathbf{H}_i\mathbf{F}_i\boldsymbol{\Psi}_i\ . \tag{3}$$

Henceforth, $\mathbf{H}^\star_{k,i}$ can be treated as a non-interfering single-user MIMO channel and $\boldsymbol{\Psi}_i$ is the spectral shaping (diagonal) matrix spans the spatial substreams $j\ \forall\ j \in [1\cdots L]$. Since the system is reduced to the single user case, the authors in [2] suggested to adapt the spectral shaping $\rho_{i,j}$ according to MMSE criterion. This minimizes the intereigen interference for each user, where $\rho_{i,j}$ is found to be

$$\rho_{i,j} = \left[\mu\frac{\sigma_n}{\Sigma_{i,j}} - \frac{\sigma_n^2}{\Sigma_{i,j}^2}\right]^+ \ ,\ \forall\ j = 1\cdots L \tag{4}$$

where $[x]^+$ denotes that $x \geq 0$, μ is a Lagrangian multiplier that can be found iteratively such that $\sum_j^L \rho^2_{i,j} = 1$, and $\Sigma^2_{i,j}$ is the eigenvalues of $\mathbf{H}_i\mathbf{F}_i$.

Hence, the eigenvalue decomposition of $\mathbf{H}_i^{\text{H}\star}\mathbf{H}_i^\star$ results in the beamforming vectors $\mathbf{V}_i^\star$ and the channel eigenvalues $\boldsymbol{\Sigma}_i^\star$. In case of CSI delays, $\overline{\mathbf{H}}_i^\star$ is introduced as the erroneous channel matrix that is deviated from the instant channel values by $\boldsymbol{\Xi}_i$, i.e., $\overline{\mathbf{H}}_i^\star = \mathbf{H}_i^\star + \boldsymbol{\Xi}_i$. Accordingly, the pre-processing and the null-space matrices are the erroneous ones $\overline{\mathbf{V}}_i^\star$ and $\overline{\mathbf{F}}_i$, respectively. Finally, the received symbol is given by

$$\mathbf{Y}_i = \overbrace{\mathbf{H}_i\overline{\mathbf{F}}_i\overline{\boldsymbol{\Psi}}_i}^{\overline{\mathbf{H}}_i^\star}\overline{\mathbf{V}}_i^\star\mathbf{P}_i^{1/2}\mathbf{X}_i + \overbrace{\boldsymbol{\Xi}_i\sum_{u=1,l\neq i}^{M}\overline{\mathbf{F}}_u\overline{\boldsymbol{\Psi}}_u\overline{\mathbf{V}}_u^\star\mathbf{P}_l^{1/2}\mathbf{X}_u}^{\text{residual MUI}} + \mathbf{N}_i\ , \tag{5}$$

where $\mathbf{P}_i^{1/2}$ is a diagonal matrix containing the power allocations based on $\boldsymbol{\Sigma}_i^\star$. By assuming hierarchical modulation alphabets for each user, the total transmitted symbol $\mathbf{X}$ and the power allocation matrix $\mathbf{P}$ are the same for every user. Now, Eqn. 5 can be re-written as follows

$$\begin{aligned}\mathbf{Y}_i &= \overbrace{\left(\overline{\mathbf{H}}_i^\star\overline{\mathbf{V}}_i^\star + \sum_{u=1,l\neq i}^{M}\boldsymbol{\Xi}_i\overline{\mathbf{F}}_u\overline{\boldsymbol{\Psi}}_u\overline{\mathbf{V}}_u^\star\right)}^{\text{equivalent channel}}\mathbf{P}_i^{1/2}\mathbf{X}_i + \mathbf{N}_i \\ &= \mathbf{H}_{\text{eq}}\mathbf{P}^{1/2}\mathbf{X}_i + \mathbf{N}_i\ . \end{aligned} \tag{6}$$

In here, Eqn. (6) reduces our multiuser system to a simple single user MIMO system with a full (diagonal dominant) channel matrix that can be decoded using a ZF or a MMSE receiver. However, for sever channel uncertainties, the matrix $\mathbf{H}_{\text{eq}}$ is not a diagonal dominant any more. Accordingly, ZF or MMSE will not be the optimal receivers. Therefore, successive interference cancellation or space-time block codes can be an alternative, which is beyond the scope of this paper.

C. Broadcasting the Same Information with MMSE

We modified the vector broadcast (VB) design, in [13], to compare it to our multilevel block diagonalized multiuser MIMO system. When the channel matrix is diagonalizable, as in the case of OFDM that employs a cyclic prefix, the optimization reduces to a simple MMSE water-filling-like solution (similar Eqn. (4)). For the case of a circulant channel matrix $\mathbf{H}^{N\times N}$ that is simultaneously diagonalized at the receiver and the transmitter using discrete Fourier transform (DFT) and inverse discrete Fourier transform (IDFT) matrices, respectively, the problem is described as follows

$$\min_{\rho_k} \quad \sum_{k=1}^{N} \frac{\sigma_n^2}{\rho_k \sum_i^M \frac{\lambda_i^2}{M} + \sigma_n^2} \tag{7}$$
$$\text{subject to} \quad \sum_{k}^{N} \rho_k \le P_T \; ; \quad \rho_k \ge 0 \; ,$$

where λ_i is the channel coefficient, and ρ_k is the power spectral shaping across the given subcarriers. Solving (8) using the Karush-Kuhn-Tucker (KKT) conditions, we get

$$\rho_k = \left[\mu \frac{\sigma_n}{\sqrt{\sum_{i=1}^{M} \lambda_k^2 / M}} - \frac{\sigma_n^2}{\sum_{i=1}^{M} \lambda_k^2 / M} \right]^+ , \forall j = 1 \cdots L \tag{8}$$

where $[\ .\ ]^+$ denotes a positive or zero value, μ is the Lagrangian multiplier that can be found iteratively such that $\sum_{k=1}^{K} \rho_k = \mathcal{E}_{\text{tot}}$.

In [13], authors consider a single transmit antenna and multiple users, each with a single receive antenna. However, in our deployment, we accommodate N_T antennas at the transmitter sending the same information sympol with a total transmit power $\mathcal{E}_{\text{tot}}$. Furthermore, each information symbol is encoded in a hierarchical fashion, allowing each user to receive different information from the same symbol.

III. Adaptive Prioritized Multiuser Multilevel Modulation

The algorithm is based on the multilevel adaptive modulation proposed in [1], which is a modified version of Levin's and Campello's bit-loading algorithms in [12] and [11], respectively. In the original algorithm in [1], the bits of the most important class are allocated according to a required symbol-error rate (SER). The less important classes' bits are allocated based on the calculated SER that allows for a given $\Delta\gamma$ margin separation, which can be seen in the SER-vs-SNR curves. Thereby, we adopted this scheme to realize fixed margin separations between different users. This criterion succeeds in keeping the relative QoS separation between users fixed under different channel conditions and different user target data rates. Similar to [1], the user with the highest QoS is allowed to consume the good-SNR subcarriers first. Thereafter, the bits of less important users are allowed to be allocated to either already used subcarriers in a hierarchical fashion if the quantized channel-to-noise ratios and the remaining power of these subcarriers are sufficient to accommodate more bits. However, if not, free subcarriers can instead be used based on the margin separation $\Delta\gamma$. This algorithm can be considered as a margin-adaptive bit-loading following this definition:

$$\min_{\mathcal{E}_k} \quad \mathcal{E}_\sigma = \sum_{k=1}^{N} \sum_{i=1}^{M} \sum_{j=1}^{L} \mathcal{E}_{k,i,j} \tag{9}$$
$$\text{subject to}: \ \overline{\mathcal{B}} = \sum_{k=1}^{N} \sum_{i=1}^{M} \sum_{j=1}^{L} \log_2 \left(1 + \frac{\mathcal{E}_{k,i,j} \mathcal{G}_{k,i,j}}{\Gamma} \right) \; , \tag{10}$$

where $\mathcal{E}_{k,i,j}$ is the power allocated to the k^{th} subcarrier, the j^{th} stream and the i^{th} user, $\mathcal{E}_{\text{tot}}$ is the given target power, $\mathcal{E}_\sigma$ is the accumulated power, $\mathcal{G}_{k,i,j}$ is the channel gain ($\Sigma_{k,i,j}^2$) to noise (σ_n^2) ratio, and the "gap" approximation is given by $\Gamma = \frac{2}{3} \left[\text{erfc}^{-1} \left(\frac{\text{P}_{\text{e}_\text{j}}}{2} \right) \right]^2$, [9]. If the total target rate is tight to a certain value $\mathcal{B}_T$ and $\mathcal{E}_{\text{tot}} > \sum_k \mathcal{E}_k$, then the performance can be further enhanced by scaling up the effective power allocation $\mathcal{E}_k$ by the ratio $\mathcal{E}_{\text{tot}} / \mathcal{E}_\sigma$. This is called "margin maximization", where the maximum system margin is defined as

$$\gamma_{\max} = \frac{\mathcal{E}_{\text{tot}}}{\sum_k \mathcal{E}_k} \; . \tag{11}$$

Therefore, a certian target SER is given for the most important user, i.e., P_{e_0}. Thereafter, and according to the given $\Delta\gamma_i$, P_{e_i} of the less important users are calculated using the "gap" value $\left(\Gamma = \frac{2}{3} \left[\text{erfc}^{-1} \left(\frac{\text{P}_{\text{e}_\text{j}}}{2} \right) \right]^2 \right)$ and Eqn. (10) to optain the following

$$P_{e_i} = 2 \ \text{erfc} \left(\sqrt{ \left(10^{\frac{\Delta\gamma_i}{10}} \right) \left[\text{erfc}^{-1} \left(\frac{P_{e_i}}{2} \right) \right] } \right) \tag{12}$$

A. Hierarchical Multilevel Modulation

In hierarchical modulation, different symbols with unequal priorities can be embedded on each other creating different Euclidean distances d_i between different users with different QoS. The relative margin separations between users are adjusted using the ratios of their constellation distances ($\frac{d_{u_i}}{d_{u_j}}$, where u_i and u_j are two different users). In Fig. 1, the distances between hierarchical users d_{u_i} are selected based on γ_i, which may, e.g., be set to 3 dB, i.e., $\frac{d_{u_i}}{d_{u_j}} = \sqrt{2}$. The proposed hierarchical construction follows the one in [10].

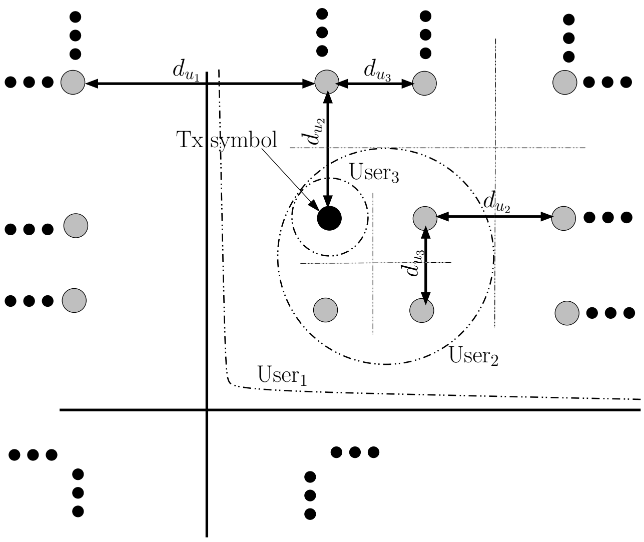

Fig. 1. Hierarchical modulation 4/16/64-HQAM: the decision boundaries of user$_0$ is a 4-QAM is embedded in the 16-QAM of user$_1$; user$_1$ decision boundaries is a 4-QAM embedded in the 64-QAM of user$_2$, where user$_2$ himself also detects the smallest 4-QAM hierarchical level.

Our multilevel modulation scheme automatically allows for different hierarchical levels based on the SNR of each user, where the maximum number of hierarchical levels is limited by the number of users M. Even more, homogeneous constellations (non-hierarchical modulation) are also allowed on some subcarriers if their SNRs (for a certain user) are far better than the equivalent subcarriers of other users. This is automatically decided based on the following prioritized adaptation algorithm.

B. Adaptive Algorithm for Prioritized QoS Bit-loading

Based on [1], we propose a multiuser computationally efficient bit-loading algorithm that quantizes the channel-to-noise ratio and adjusts the bit-loading to a certain target bit rate $\mathcal{B}_T$ in order to minimize the total power. The number of bits on any given subcarrier k is

$$b_k = \log_2(1 + \mathcal{E}_k g_k) \ , \tag{13}$$

where g_k is the channel-to-noise ratio with the gap Γ included into it. Therefore, the allocated energies and the incremental energies on each subcarrier are

$$\mathcal{E}_k = \frac{\left(2^{b_k} - 1\right)}{g_k} \ , \tag{14}$$

$$\begin{aligned} \Delta\mathcal{E}_k(b_k) &= \mathcal{E}_k(b_k) - \mathcal{E}_k(b_k - 1) \\ &= \frac{2^{b_k}}{2g_k} = 2^{b_k - 1 - \log_2 g_k} \ . \end{aligned} \tag{15}$$

According to [11], the discrete bit-allocation b_k is

$$b_k = \left[\lfloor \log_2 g_k \rfloor + i_{B_{\rm opt}}\right]_0^{b_{\max}} \ , \tag{16}$$

where $i_{B_{\rm opt}} \in \mathbb{Z}$ is an arbitrary constraint adjusted such that $\sum_k b_k \leq \mathcal{B}_T$. The floor operator allows different subcarriers, with different g_k, to take the same bit value b_μ, i.e., there will be groups of similar subcarriers due to the quantization process. Hence, let $\mathcal{L}$ be the maximum number of the subcarrier groups allocated to the same $\bar{b}_\mu$ $\forall\ \mu \in [0, L-1]$. The subcarriers forming the μ^{th} group (with the same bit value b_μ after quantization) are stored in v_μ and their indices are stored in S_μ such that

$$S_\mu = \text{length}\{v_\mu \in \{1, \cdots, N\} : \lfloor \log_2 g_{v_\mu} \rfloor = \bar{b}_\mu\} \ . \tag{17}$$

Therefore, a positive quantization error Δ_k on each subcarrier is calculated, using $K_\mu = \lfloor \log_2 g_k - \log_2 g_{\min} \rfloor$, as

$$\Delta_k = \{\log_2 g_k - \log_2 g_{\min}\} - K_\mu \ . \tag{18}$$

Based on the linear programming in [1] and the relaxation to real values, an estimated value for i_B is calculated as

$$i_B = \frac{\mathcal{B}_T - \sum_{\mu=0}^{\mathcal{L}-1} K_\mu \cdot M_\mu}{N} \in \mathbb{R} \ , \tag{19}$$

where it might be not the optimal value due to the quantization in K_μ. Therefore, it is suggested, in [1], to calculate $i_B + z$, $\forall\ z \in \{-1, 0, 1\}$. Thereafter, the algorithm searches for the optimal z (that results in $i_{B_{\rm opt}}$) that minimizes $|\mathcal{B}_T - \overline{\mathcal{B}}|$, i.e., find $i_{B_{\rm opt}}$ such that

$$\arg \min_{i_{B_{\rm opt}}+z} \left| \mathcal{B}_T - \sum_{\mu=0}^{\mathcal{L}-1} \left(K_\mu + i_{B_{\rm opt}} + z\right) S_\mu \right| \ . \tag{20}$$

The UEP Campello algorithm steps are as follows:

1) Set the user index $i = 1$ and $b_{k,j} = 0$, $\forall k \in [0, N-1]$
2) Set the target rate to $R_{\rm temp} = \sum_{l=0}^{i} \mathcal{B}_l$, where $\mathcal{B}_i$ is the individual bit-rate of each user i and $R_{\rm temp}$ is the accumulated rate of this user.
3) Compute $K_\mu\ \forall\ \mu \in [0, -1]$, then compute $i_{B_{\rm opt}} + z$ using (19) and (20). Thereafter, compute the accumulated bit-loading b_k (of the user$_i$) using (16) and the accumulated $\mathcal{E}_k$ using (14).
4) If $\mathcal{E}_{\rm tot} > \sum_k \mathcal{E}_k$
   - If $R_{\rm temp} < (>) \sum_k b_k$, decrement (increment) b_k corresponds to the subcarriers with the smallest (highest) Δ_k, and change Δ_k to $+\infty$ $(-\infty)$, i.e., so that these locations will not be incremented (decremented) further.
   - If $i \geq 1$, subtract the bit-loading of the $i-1$ user(s) from the current accumulated b_k, thereby, the bit-loading for each user is $b_{k,i} = b_k - \sum_i b_{k,i-1}$. Else, if $i = 0$, $b_{k,0} = b_k$.
   - If $i < i_{\max}$, increment i and go to 2).
5) Else, the correct bit-loading values are found. Accordingly, the energy has to be recalculated using Eqn. (14) and scaled to the fixed total energy $\mathcal{E}_{\rm tot}$.

IV. Results and Discussion

To evaluate our algorithm, we consider three users with 3 different priorities. A fixed margin separation between these users (=3 dB) is considered as a strict constraint. This is easily guaranteed by setting the hierarchical modulation priority ratio to $\sqrt{2}$ at those subcarriers where hierarchical modulation is actually used. However, when hierarchical modulation is not implemented, our bit-loading algorithm takes care of these 3 dB separations. In our simulations, we assumed that the total number of subcarriers N is 512. Additionally, the homogeneously modulated subcarriers are allowed to allocate quadrature amplitude modulation (QAM) with a constellation size of a maximum of 8 bits. The hierarchically modulated ones are allowed to have a maximum of 6 bits on each layer.

As seen in Fig 2, assuming 1024 bits/user, the margin separation in case of perfect CSI is preserved for the single eigen-beamforming as well as for the full eigen-beamforming. However, at higher CSI errors, the separation is getting wider and the performance deteriorates. Nevertheless, the same user ordering is preserved. From the same figure, the vector broadcast (VB) design (the dashed black curves) seems to be an upper bound to our scheme with a residual error floor. Figure 3 depicts the full eigen-beamforming performance under CSI error (σ^2_{Ξ}) of 0.25 with (and without) spectral shaping $\mathbf{\Psi}$. The one with the spectral shaping outperforms the other by almost 7 dBs, since the spectral shaping proposed in Eqn. (4) suppresses the weaker channels. In this case, they are not allocated during the bit-loading process. As could be also seen in this figure, the single eigen-beamforming already outperforms the full eigen-beamforming, even with the spectral shaping, in case of CSI errors. This shows the

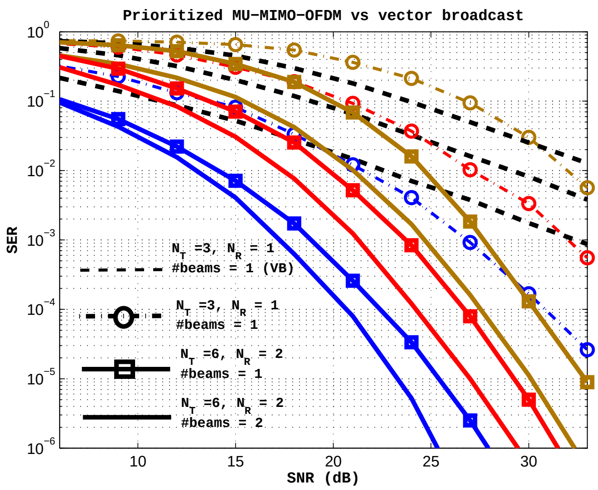

Fig. 2. Adaptive multiuser MIMO-OFDM with margin separations of 3 dB, different transmit-receive antenna setup, and 1024/user.

drawback (the suboptimality) of Eqn. (4). This means that some residual intereigen interferences still exist.

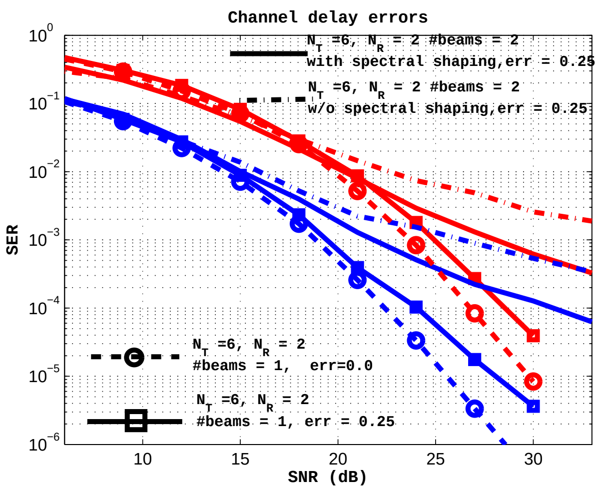

Fig. 3. Different CSI errors, assuming the same margin separation (3 dB), with/without spectral shaping and different beamforming modes.

Finally, Fig. 4 depicts the performance of the reduced target bit-rate, only 512 bits/user. One can notice that the performance of the full/single beamforming dramatically improved. However, the single beamforming still performs the best amongest the other setups.

V. Conclusions

We implemented a hierarchical multiuser system with an orthogonal SDMA to suppress the multiuser interference. We succeeded to preserve our given QoS constraint under perfect channel conditions. However, the margin separations became wider in case CSI errors, where the overall performance deteriorates due to the limitations of the ZF approach. The performance is dramatically enhanced when the target bit-rate is considerably reduced. However, our adaptive modulation scheme outperforms the vector broadcast design using the same data-rates and assuming the same channel conditions.

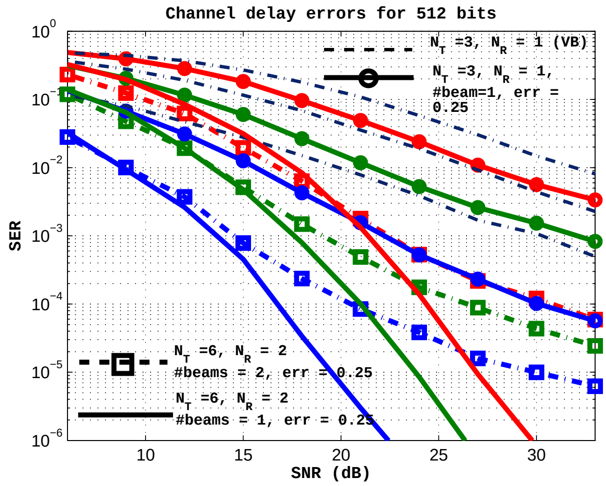

Fig. 4. The performance under reduced bit-rate (512 bits/user).

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