UEP with Adaptive Multilevel Embedded Modulation for MIMO-OFDM Systems

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***Abstract*— In this paper, a scalable multimedia system with unequal error probabilities (UEP) is considered. A combination of adaptive modulation and hierarchical (embedded) constellations is implemented to realize three different levels of data priorities. The highest priority data bits spread over all available subcarriers with a relatively low modulation order. The next data levels are embedded on top of each other, according to the given channel gain, forming varying multilevel hierarchical constellations. An arbitrary margin separation is considered between the classes, however the highest priority data should be granted a certain maximum symbol-error rate (SER). The given data rates are allocated adaptively using an optimum bit-loading algorithm. In case of channel feedback errors, our algorithm performs very close to the ideal case, keeping the high priority data even safer. Moreover, the low-complex bit-loading algorithm makes the UEP adaptation concepts suitable for future wireless systems with limited feedback.**

***Index Terms*— Hierarchical constellations, adaptive modulation, limited feedback, and UEP**

I. INTRODUCTION

ADAPTIVE communication is, theoretically, beneficial for spectral efficiency. However, an adaptive chain requires a closed loop system, which means that the channel state information (CSI) has to be readily available at the transmitter. Full channel knowledge at the transmitter is not a practical assumption. Nevertheless, a partial CSI suffices for keeping the performance acceptable, especially when we keep transmitting on the stronger channels only. For this reason, a MIMO-OFDM combination with a restricted beamforming has been considered, where beamforming can be used as a tool to suppress the CSI error effects as in [1]. In addition to this, MIMO-OFDM is also seen as an optimum adaptive scheme, since both frequency and spatial domain can be adapted jointly. Furthermore, data is prioritized to keep the source data significance levels matched to the different channel conditions.

There are several ways to realize UEP in a data transmission link. Coding can be a suitable choice, however, going deeper into the physical layer can bring in simpler methods. UEP bit loading has already been discussed in literature [11]. A more practical approach has been described in [1] using a modified rate-adaptive Chow bit-loading. However, this method is considered to be a sub-optimal solution, since it is based on the non-integer water-filling approach. The other drawback of this method is its inefficient energy utilization, where energy can be wasted due to allocating power on weak subcarriers.

The classical Hughes-Hartogs algorithm is known as the optimum bit-loading algorithm, even though it is a computationally expensive method. It is based on the "greedy optimization algorithm" that aims at finding a global minimum by adding a single bit to the subcarrier that requires the minimum incremental power at each iteration [3]. Therefore, it is optimal for minimizing the energy and maximizing the system margin. However, due to complex search operations in Hughes-Hartogs, less complex methods are preferred. Campello bit-loading [4], which is a linear representation of the Levin bit-loading algorithm [10], is a simple alternative to the Hughes-Hartogs. It achieves *almost the same* optimum power allocation requiring only a fraction of the complexity. The simplification here lies in the quantization of the channel-gain-to-noise ratio g_k. Similar levels of g_k can be gathered into L groups, where L is much smaller than the number of subcarriers. In addition to the simplicity, Campello's bit-loading can be thought of as a practical solution for limited (quantized) channel feedback systems. In this paper, we reduce the complexity even further to make it suitable for wireless applications.

The simple subcarrier partitioning used in [1] to separate different priorities cannot directly be implemented in any of the previous power-minimization bit-loading algorithms, since they allocate the subcarriers with the least incremental power first (based on their current channel gains, the number of bits already allocated, and the required symbol-error ratio). Therefore, hierarchical modulation – combined with the power-minimization bit-loading – is used, where the highest priority class first consumes the good-SNR subcarriers with the minimum incremental power. Thereafter, the next classes are allowed to be superimposed on used or free subcarriers. This leads to a multilevel hierarchical modulation. An arbitrary margin separation is kept constant by using a fixed ratio between the inter-constellation distances d_j for each level j.

This paper is organized as follows. Section 2 discusses the MIMO-OFDM parameters and the used channel model. Section 3 introduces our UEP adaptive hierarchical modulation using Hughes-Hartogs and our simplifications to the UEP-Campello algorithm. Section 4 discusses the results. Finally, we conclude our findings in the last section.

II. MIMO-OFDM Adaptive Model

A MIMO-OFDM combination is selected due to its capability of adapting both frequency and spatial domains. We consider a MIMO $N_R \times N_T$ channel matrix $\mathbf{H}_k$, where N_R is the number of the receive antennas, N_T is the number of the transmit antennas, k is the subcarrier index, and the total number of subcarriers is N

A. The MIMO Channel Model

We modeled the channel impulse responce in time domain ($h^{n_r,n_t}(\tau;t)$) as a Rayleigh fading model with an exponential power delay profile [9]. The channel is composed of L different paths (echos); each path has its own amplitude β_l, delay τ_l, and Doppler frequency f_{D_l}. The time-variant channel impulse response for each channel matrix element can be defined as

$$h^{n_r,n_t}(\tau;t) = \sum_{l=0}^{L-1} \beta_l(t) p_l(t) e^{2\pi f_{D_l} t} \delta(t-\tau_l) \ , \quad (1)$$

where t is the instant of the response $h(\tau;t)$ with a dispersion τ. The variable β_l is considered to be a zero mean independent, identically distributed (i.i.d.) random complex Gaussian variable and p_l is an exponentially fading variable. Therefore, one can assume that the elements of $\mathbf{H}_k$ (in frequency domain) are also i.i.d. and zero mean circularly symmetric complex Gaussian (ZMCSCG) (assuming no antenna correlation), i.e, $\mathbf{H}_k \in \mathcal{C}\ \mathcal{N}(0, \sigma^2_{\mathbf{H}}\mathbf{I})$ [7].

B. Adaptive MIMO-OFDM with a Limited CSI Regime

We assume a partial CSI, which is fed back to the transmitter through a limited feedback channel, i.e., the quantized/outdated channel $\overline{\mathbf{H}}_k$. Therefore, the error matrix is defined as $\mathbf{\Xi}_k = \mathbf{H}_k - \overline{\mathbf{H}}_k$, where $\mathbf{\Xi} \in \mathcal{C}\ \mathcal{N}(0, \sigma^2_{\mathbf{\Xi}}\mathbf{I})$ and $\mathbf{H}_k$ are the instantaneous channel values [7].

Bits and powers are allocated according to the eigenvalues of the erroneous Hermitian matrix $\overline{\mathbf{H}}^{\mathrm{H}}\overline{\mathbf{H}}$ [1]. Furthermore, the total power is directed to the space according to the eigenvectors (beamforming preprocessing matrix) of the same Hermitian matrix. The eigen-beamforming and the eigenvectors at every subcarrier are simply computed using the eigenvalue decomposition (EVD), such that

$$\overline{\mathbf{H}}_k^{\mathrm{H}}\overline{\mathbf{H}}_k = \overline{\mathbf{V}}_k^{\mathrm{H}}\overline{\mathbf{D}}_k^2\overline{\mathbf{V}}_k \quad (2)$$

where $\overline{\mathbf{V}}_k$ is the unitary preprocessing matrix and $\overline{\mathbf{D}}_k^2$ is a diagonal matrix with the eigenvalues λ_i along its diagonal.

Due to CSI errors, the total channel, after using the post-processing matrix, is not diagonalized. Consequently, a spatial equalizer is becoming mandatory to mitigate possible interferences. Therefore, it is obvious that there is no need for the post-processing matrix. One can directly replace it by a linear minimum mean-square error (MMSE) or a zero-forcing (ZF) equalizer [6]. The received signal after equalization is given by

$$\mathbf{Y}_k = (\hat{\mathbf{H}}_k^{\mathrm{H}}\hat{\mathbf{H}}_k)^{-1}\hat{\mathbf{H}}_k^{\mathrm{H}} \overbrace{\mathbf{H}_k\overline{\mathbf{V}}_k\mathbf{P}_k^{1/2}}^{\hat{\mathbf{H}}} \mathbf{X}_k + \tilde{\mathbf{N}}_k \ , \quad (3)$$

where $(\hat{\mathbf{H}}_k^{\mathrm{H}}\hat{\mathbf{H}}_k)^{-1}\hat{\mathbf{H}}_k^{\mathrm{H}}$ is the pseudoinverse of the total channel $(\mathbf{H}_k\overline{\mathbf{V}}_k\mathbf{P}_k^{1/2})$, $\mathbf{P}_k$ is the power loading matrix, and $\tilde{\mathbf{N}}_k$ is the colored noise at the equalizer output. One can still prove that $\hat{\mathbf{H}}_k^{\mathrm{H}}\hat{\mathbf{H}}_k$ is a diagonal matrix if (and only if) $\overline{\mathbf{V}}_k = \mathbf{V}_k$, i.e., if the CSI is perfectly known at the transmitter.

III. The UEP Adaptive Hierarchical Modulation Algorithm

The algorithm is based on the Hughes-Hartogs bit-loading in [2], where the incremental power is calculated based on the maximum allowed symbol-error rate (SER) and the channel eigenbeams. The most important data are allowed to consume the good subcarriers (the ones with the minimum incremental power) first requiring a given SER P_{e_0}. The next classes are allocated to either already used subcarriers, thereby introducing hierarchical modulation, or the free ones based on their eigenvalues and the required margin. P_{e_j} of the less important data are calculated using the given margin separations γ_j and the given P_{e_0}, as in [12]. In Fig. 1, the distances between classes d_j are selected based on γ_j, which may, e.g., be set to 3 dB. The hierarchical construction follows the one in [8].

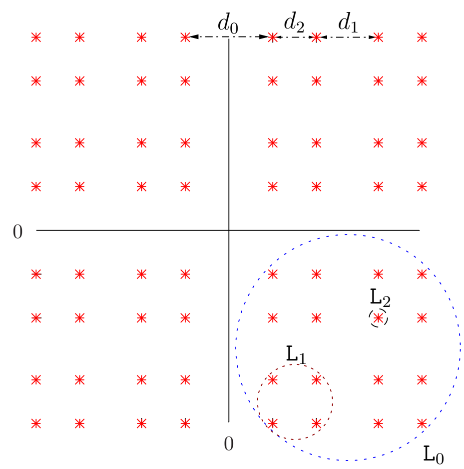

Fig. 1. Hierarchical modulation: a 4-QAM (L_0) is embedded in a 16-QAM (L_1) and the 16-QAM (L_1) is, itself, embedded in a 64-QAM (L_2) with the performance priority ratios $d0/d1 = d1/d2 = \sqrt{2}$.

A. Hughes-Hartogs as a UEP Bit-loading

In the following, we will describe our UEP bit-loading based on the Hughes-Hartogs algorithm. The algorithm is following the margin-adaptive loading criterion, which is defined as:

$$\min_{\mathcal{E}_k} \quad \mathcal{E}_\sigma = \sum_{k=1}^{N} \mathcal{E}_k \quad (4)$$

$$\text{subject to} : \ \overline{\mathcal{B}} = \sum_{k=1}^{N} \log_2\left(1 + \frac{\mathcal{E}_k \mathcal{G}_k}{\Gamma}\right) \ , \quad (5)$$

where $\mathcal{E}_k$ is the power allocated to the k^{th} subcarrier, $\mathcal{E}_{\mathrm{tot}}$ is the given target power, $\mathcal{E}_\sigma$ is the accumulated power,

$\mathcal{G}_k$ is the channel gain ($\lambda_{i,k}$) to noise (σ_n^2) ratio, and the "gap" approximation is given by $\Gamma = \frac{2}{3}\left[\text{erfc}^{-1}\left(\frac{\text{P}_{\text{e}_\text{j}}}{2}\right)\right]^2$, [3]. If the total target rate is tight to a certain value $\mathcal{B}_T$ and $\mathcal{E}_{\text{tot}} > \mathcal{E}_\sigma$, 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

$$\gamma_{\max} = \frac{\mathcal{E}_{\text{tot}}}{\mathcal{E}_\sigma} \ . \tag{6}$$

The complete algorithm is as follows:

1) Generate an array B (with maximum length T) containing the available number of bits, where $B(0) = 0$ and $B(T-1) = b_{\max}$.
2) Set $j = 0$ and generate an array c of length N starts from all zeros and does not exceed the length of B ($0 \leq c_k \leq T$).
3) Compute the incremental power steps $\Delta\mathcal{E}_k$, for every subcarrier, using the following approximate equation (as in [3]):

$$\Delta\mathcal{E}_k = \frac{\frac{2}{3}\left[\text{erfc}^{-1}\left(\frac{\text{P}_{\text{e}_\text{j}}}{2}\right)\right]^2}{\mathcal{G}_k}\left(2^{B(c_k+1)} - 2^{B(c_k)}\right).$$

4) Find the minimum $\Delta\mathcal{E}_k$ among all subcarriers.
5) Increment c_k such that it does not exceed $T-1$, then set $b_{k,j}$ to the value in $B(c_k)$.
6) Increment the power of the k^{th} subcarrier by $\Delta\mathcal{E}_i$.
7) If the target bit-rate of the j^{th} class is not fulfilled,
   - if the sum of the energy is less than the target energy $\mathcal{E}_{tot}$, go to 3),
   - else, stop.
8) Else if the target bit-rate of the j^{th} class is fulfilled and j is less than the number of the given classes $j_{\max}$,
   - if the sum of the energy is less than the target energy $\mathcal{E}_{tot}$, increment j and go to 3),
   - else, stop.
9) Scale-up the allocated energy $\mathcal{E}_k$ using Eq. (6), then

$$\mathcal{E}_i = \mathcal{E}_i \cdot \gamma_{\max} \ .$$

In the following, we extend our work to modify the less complex Campello algorithm in order to realize UEP.

B. Campello Algorithm for UEP Bit-loading

Based on Levin's bit-loading [10], Campello has proposed a very simple linear representation for Hughes-Hartogs using only $\mathcal{O}(N)$ complexity. In addition to this, it achieves the same optimum allocation [5]. However, the consecutive logical iterations, used for computing the approximated bit-loading, make it less efficient. Therefore, we present an easier linear solver to this problem.

This algorithm follows the same margin maximization problem (in Eq. (4-6)), assuming a constant target power $\mathcal{E}_{tot}$ and a target bit-rate $\mathcal{B}_T$. The number of bits on the k^{th} subcarrier is given by

$$b_k = \log_2(1 + \mathcal{E}_k g_k) \ , \tag{7}$$

where g_k is the gain to noise ratio with the gap Γ impeded into it. Therefore, the allocated energy and the incremental energy on each subcarrier are

$$\mathcal{E}_k = \frac{(2^{b_k} - 1)}{g_k} \ , \tag{8}$$

$$\begin{aligned} \Delta\mathcal{E}_k(b_k) &= \mathcal{E}_k(b_k) - \mathcal{E}_k(b_k - 1) \\ &= \frac{2^{b_k}}{2g_k} = 2^{b_k - 1 - \log_2 g_k} \ . \end{aligned} \tag{9}$$

The discrete bit allocation $b \in \mathbb{Z}_0^{b_{\max}}$ is said to be "efficient" and the system margin is maximizied if the following is satisfied [5]:

$$\max_k \Delta\mathcal{E}_k(b_k) < \min_m \Delta\mathcal{E}_m(b_m + 1) \ \ \forall\, n, m = 1, \cdots, N \ . \tag{10}$$

This means that the maximum incremental power required to achieve the current allocation, as in Eq. (10), is less than the minimum power required to add an extra bit on any subcarrier [4]. Accordingly, an approximated discrete bit-allocation b_k that satisfies Eq. (10) is given by [5]

$$b_k = \left[\lfloor \log_2 g_k \rfloor + i_{B_{\text{opt}}}\right]_0^{b_{\max}} \ , \tag{11}$$

where $i_{B_{\text{opt}}} \in \mathbb{Z}$ is computed such that $\sum_k b_k \leq \mathcal{B}_T$. The proof of (11), as in [5], is easly found by substituting (11) into (10). The floor operator ensures a lower bound for bit-allocation. This means that different subcarriers, with different g_k, are allocated to the same number of bits $\bar{b}_i$. Let L be the number of the subcarrier groups allocated to the same $\bar{b}_i \ \forall\, i \in [0, L-1]$. The subcarriers of the i^{th} group are stored in u_i and the number of subcarriers, in this group, is stored in M_i such that

$$M_i = \text{length}\{u_i \in \{1, \cdots, N\} : \lfloor \log_2 g_{u_i} \rfloor = \bar{b}_i\} \ . \tag{12}$$

To keep the value inside the floor operator positive, one has to normalize g_k by dividing it by its minimum value $g_{\min}$, i.e., $K_i = \lfloor \log_2 g_k - \log_2 g_{\min} \rfloor$. The positive quantization error Δ_k on each subcarrier is calculated as

$$\Delta_k = \{\log_2 g_k - \log_2 g_{\min}\} - K_i \ . \tag{13}$$

Therefore, what follows is a good approximation to the required bit-allocation $\mathcal{B}_T$ [4],

$$\bar{\mathcal{B}} = \sum_{k=0}^{N-1} b_k = \sum_{i=0}^{L-1} \left[K_i M_i + i_{B_{\text{opt}}} M_i\right]_0^{b_{\max}} \ . \tag{14}$$

From (14), it is clear that the absolute value $|\mathcal{B}_T - \bar{\mathcal{B}}|$ is convex (but not strictly convex) in $\bar{\mathcal{B}}$, which is a function of $i_{B_{\text{opt}}}$, since $\mathcal{B}_T$ is constant. Therefore, we prefer to relax $i_{B_{\text{opt}}}$ and $\left[K_i + i_{B_{\text{opt}}}\right]$ to the real domain in order to find the optimum bit-loading. This solution is more efficient than the nested logical operations in [5]. The equivalent relaxed problem is defined as follows:

$$\begin{aligned} \text{minimize} \quad & \left|\mathcal{B}_T - \bar{\mathcal{B}}\right| \\ \text{subject to} \quad & i_B \text{ and } (K_i + i_B) \in \mathbb{R} \end{aligned} \tag{15}$$

This makes the computation of the equivalent real value i_B relatively easy. However, this solution is only valid when $\mathcal{B}_T < N b_{\max}$. Now, Eq. (14) can be written as

$$\bar{\mathcal{B}} = \sum_{k=0}^{N-1} b_k \approx i_B \cdot N + \sum_{i=0}^{L-1} K_i \cdot M_i \,. \tag{16}$$

To find the minimum of (15), the absolute value must be set to zero, i.e., $\mathcal{B}_T = \bar{\mathcal{B}}$. Hence, one can directly plug this into (16) and get the real i_B as follows:

$$i_B = \frac{\mathcal{B}_T - \sum_{i=0}^{L-1} K_i \cdot M_i}{N} \in \mathbb{R} \,. \tag{17}$$

To find the integer $i_{B_{\text{opt}}}$, the previous result has to be rounded, however, it may deviate from the global minimum of (15). Therefore, one should not compute (15) using $i_{B_{\text{opt}}}$ found from (17) only, but also using $i_{B_{\text{opt}}} \pm 1$ and find the minimum, i.e., find $i_{B_{\text{opt}}}$ such that

$$\arg \min_{i_{B_{\text{opt}}}+z} \left| \mathcal{B}_T - \sum_{i=0}^{L-1} \left(K_i + i_{B_{\text{opt}}} + z\right) M_i \right| \forall\, z \in \{-1, 0, 1\} \tag{18}$$

The UEP Campello algorithm steps are as follows:

1) Set $j = 0$ and $b_{k,j} = 0, \forall k \in [0, N-1]$
2) Set the target rate to $R_{\text{temp}} = \sum_{l=0}^{j} \mathcal{B}_l$, where $\mathcal{B}_j$ is the individual bit-rate of each class j and R_{temp} is the accumulated rate of the current j classes.
3) Compute $K_i \;\forall\; i \in [0, L-1]$, then compute $i_{B_{\text{opt}}}+z$ using (17) and (18). Thereafter, compute the accumulated bit-loading (of the j classes) b_k using (11) and the accumulated $\mathcal{E}_k$ using (8).
4) If $\mathcal{E}_{\text{tot}} > \sum_k \mathcal{E}_k$
   - If $R_{\text{temp}} < (>) \sum_k b_k$, decrement (increment) b_k corresponds to the subcarriers with the smallest (highest) Δ_k, and change Δ_k to 1 (0).
   - If $j \geq 1$, subtract the bit-loading of the $j-1$ class(es) from the current accumulated b_k, thereby, the bit-loading for each class is $b_{k,j} = b_k - \sum_j b_{k,j-1}$. Else, if $j = 0$, $b_{k,0} = b_k$.
   - If $j < j_{\max}$, increment j and go to 2).
5) Else, the accumulated and the individual bit-loading are found. Accordingly, the energy is recalculated using Eq. (8) and scaled as the last step in the first algorithm.

IV. Results and Discussion

To evaluate the performance of our two proposed UEP bit-loading algorithms, we consider a scalable multimedia system with 3 different layers of priorities. A fixed margin separation between these layers (=3 dB) is considered as a strict constraint. This is easily guaranteed by setting the hierarchical modulation priority ratio to $\sqrt{2}$. The total number of bits for each layer $\mathcal{B}_j$ is chosen to be 2048 bits, resulting in a target bit rate of 6144 bits.

Our simulation model considers a 4×4 MIMO-OFDM system with 512 subcarriers. The bit-rate, the power values, and the eigen-beamforming for each spatial-subcarrier are adapted according to the CSI, which is fed back to the transmitter through a feedback link. According to the feedback link quality, the CSI (combined with the beamforming) is set to one of the following cases:

1- a full dimension eigen-beamforming (4×4) assuming a perfect CSI knowledge at the transmitter,
2- a full dimension eigen-beamforming with a partial CSI, assuming CSI errors with a variance of 0.25 and zero mean [1],
3- and the same amount of CSI errors with a reduced eigen-beamforming with two dimensions, i.e., suppress the weaker eigenbeams such that the MIMO channel is resized to 2×4.

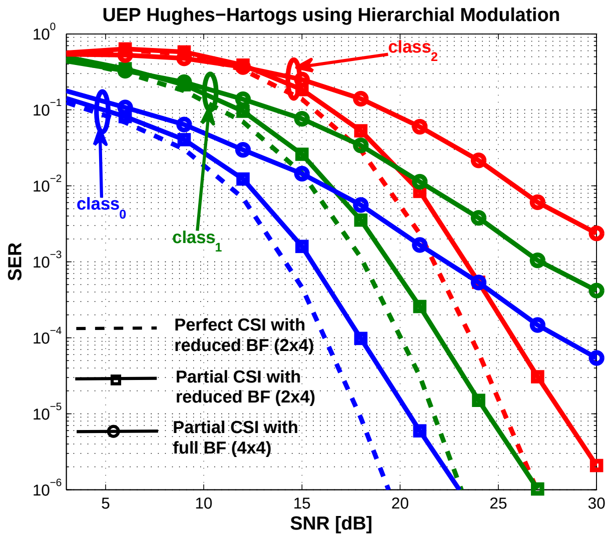

Fig. 2. Adaptive hierarchical modulation using 3 layers of UEP with 3 dB margin separation γ_j. The CSI errors, in case of partial feedback, is assumed to have a variance of 0.25.

A. *SER Performance of UEP Hughes-Hartogs*

Figure 2 depicts the performance of the previous three combinations, considering only the UEP bit-loading based on the Hughes-Hartogs algorithm. As shown in this figure, the two-dimensional beamforming (with CSI errors) outperforms the full eigen-beamforming assuming the same amount of CSI errors. The performance gain (here) is due to suppressing the weaker eigenbeams, which are more susceptible to the inter-eigen interference caused by the CSI errors. This implies that our MIMO channel is, practically, reduced to a 2×4 channel. One can also notice that the margin separation is strictly fulfilled in the perfect CSI conditions and in the partial CSI with reduced beamforming. However, this separation is getting wider in case of full beamforming with the CSI errors.

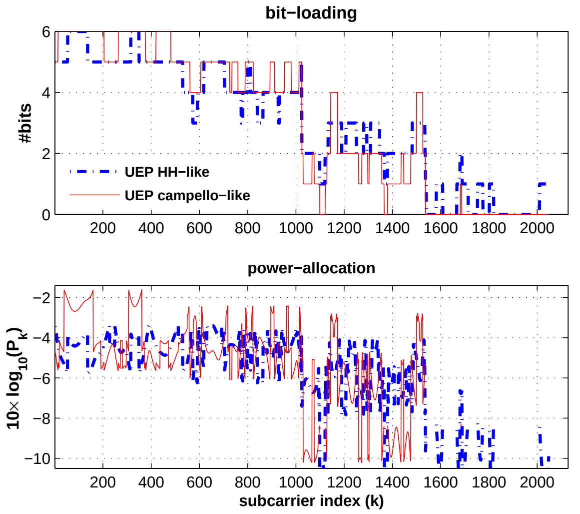

Fig. 3. Accumulated bit-loading and power allocation for 3 hierarchical layers using UEP Campello and UEP Hughes-Hartogs (HH).

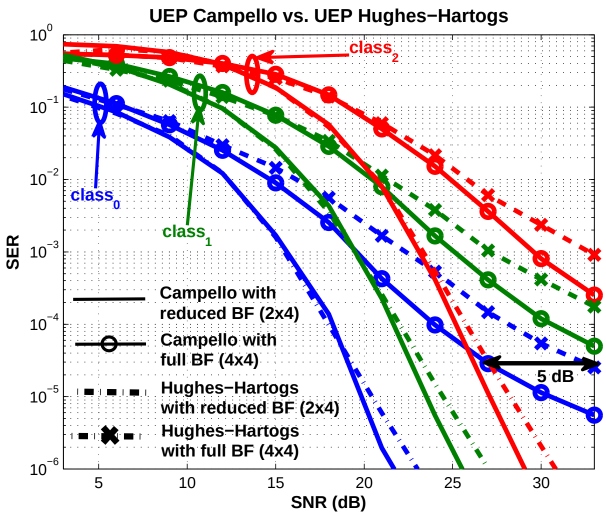

Fig. 4. A Comparison between UEP Hughes-Hartogs and UEP Campello for different beamforming in case of CSI errors.

B. UEP Campello vs. UEP Hughes-Hartogs

Figure 3 shows the power allocation of the two proposed UEP bit-loading algorithms. In this figure, the UEP based on Campello bit-loading algorithm, with its floor operator, allocates more power to the stronger subcarriers (on the stronger eigenbeams). This ensures better performance in case of full eigen-beamforming and imperfect CSI. For this reason, the full beamforming using UEP Campello outperforms the UEP Hughes-Hartogs by 5 dB (at 3×10^{-5}), as shown in Fig. 4.

Apparently, both algorithms perform very close to each other in case of two-dimensional beamforming with CSI errors. However, UEP Campello is still capable of directing more power towards the stronger subcarriers in this case. This explains the limited performance gain (<1 dB) at high SNR. Therefore, we consider this algorithm to be a robust bit-loading in case of channel feedback uncertainties.

V. Conclusions

We described two UEP bit-loading algorithms using hierarchical modulation to allow for arbitrary margin separations. First, we modified the Hughes-Hartogs algorithm to realize UEP. In case of perfect CSI with full beamforming, the margin separations are strictly fulfilled. After introducing CSI errors, the performance is deteriorated and the margin gap is getting wider. However, by reducing the beamforming dimension the performance is getting closer to the perfect CSI conditions and the margin separation is better preserved. This is due to eliminating the weaker eigenbeams, which are more susceptible to inter-eigen interference. Finally, we proposed the UEP Campello bit-loading algorithm due to its capability of achieving the optimum solution using significantly less complexity. Moreover, it invests the given power budget in the stronger subcarriers. This makes it more robust to CSI errors, hence, more efficient for wireless devices.

VI. Acknowledgment

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