# CHANNEL ADAPTIVE OFDM SYSTEMS WITH PACKET ERROR RATIO ADAPTATION

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## ABSTRACT

MIMO-OFDM systems with pre-filtering as well as bit and power loading have gained a lot of interest recently. These systems are based on theoretically optimal solutions maximizing the uncoded throughput and constraining the maximum uncoded BER. In this paper, we extend the existing frameworks to coded systems with bit and packet error ratio adaptation. We evaluate the performance of this approach in a standard compliant WiMAX link level simulator.

## 1. INTRODUCTION

The transmitter design adapted to the propagation channel is capable of improving both performance and throughput of a network. Therefore, an increased interest has been given to channel adaptive transmissions in MIMO-OFDM, e.g. in [1] there is a channel adaptive transmitter with Alamouti coding, in [2] they applied a beamformer with an appropriate power allocation policy but without bit loading. The application goes even further to the cross-layer optimized algorithms for OFDM systems [3, 4] where the user power and bit allocation policy works similarly as in [1]. To the best of our knowledge, all investigations have not yet assumed channel coding. Our algorithm is based on the problem formulation and solution from [1], but we extend this to coded systems and packet error ratio adaptation.

The rest of the paper is organized as follows. In Section 2, we present the MIMO-OFDM system model, followed by the formulation of the constrained optimization problem for the coded user throughput maximization in Section 3. In Section 4, the power cost and threshold metric required by the adaptive algorithm are derived for the $[133\ 171]_8$ half-rate convolutional code. This derivation is based on the bit error probability approximation. The optimal Greedy Power Allocation Algorithm is described in Section 5. The simulation results are obtained using a standard compliant WiMAX link level simulator and presented in Section 6. Finally, Section 7 concludes this paper.

## 2. MIMO-OFDM SYSTEM MODEL

We consider MIMO-OFDM systems equipped with $K = 192$ data subcarriers (FFT size is 256), $N_\mathrm{T}$ transmit, and $N_\mathrm{R}$ receive antennas. Figure 1 shows the discrete time baseband model that is obtained under the following considerations. Let $\mathbf{h}_{n_\mathrm{r}n_\mathrm{t}}$ be the baseband equivalent FIR channel between the $n_\mathrm{t}$-th transmit and the $n_\mathrm{r}$-th receive antenna during the given block with channel order $N$. Let $k \in \{1, \ldots, K\}$ denote the subcarrier index. The frequency flat channel $H_{n_\mathrm{r}n_\mathrm{t}}[k]$ between the $n_\mathrm{t}$-th transmit and the $n_\mathrm{r}$-th receive antenna on the $k$th subcarrier is given by the Discrete Fourier Transform (DFT). The full MIMO frequency flat channel on the $k$th subcarrier is then modeled by the $N_\mathrm{R} \times N_\mathrm{T}$ channel matrix $\mathbf{H}[k]$ with rank $R$

$$\mathbf{H}[k] = \begin{bmatrix} H_{11}[k] & \cdots & H_{1N_\mathrm{T}}[k] \\ \vdots & \ddots & \vdots \\ H_{N_\mathrm{R}1}[k] & \cdots & H_{N_\mathrm{R}N_\mathrm{T}}[k] \end{bmatrix}. \quad (1)$$

We assume that the channel $\mathbf{H}[k]$ changes slowly and is fully known at the transmitter and at the receiver. If the latter condition cannot be satisfied, almost ideal performance can be achieved by applying advanced LMMSE channel estimators with acceptable complexity [5]. By stacking the received symbols of all $N_\mathrm{R}$ receive antennas

$$\mathbf{y}[k] = [y_1[k], y_2[k], \ldots, y_{N_\mathrm{R}}[k]]^\mathrm{T} \quad (2)$$

we can calculate the received symbol vector as

$$\mathbf{y}[k] = \mathbf{H}[k]\mathbf{V}[k]\tilde{\mathbf{a}}[k] + \mathbf{n}[k], \quad (3)$$

where $\mathbf{n}[k]$ is zero mean circularly symmetric complex Additive White Gaussian Noise (AWGN) with covariance $\mathrm{E}\{\mathbf{n}[k]\mathbf{n}[k]^\mathrm{H}\} = N_0\mathbf{I}_{N_\mathrm{R}}$, $\mathbf{V}[k]$ is a pre-filtering matrix, and $\tilde{\mathbf{a}}[k]$ is a power-loaded transmit symbol vector obtained as

$$\tilde{\mathbf{a}}[k] = \mathbf{p}[k] \odot \mathbf{a}[k], \quad (4)$$

where $\odot$ denotes an element-wise multiplication between the power loading vector $\mathbf{p}[k] = \left[\sqrt{P_1[k]}, \ldots, \sqrt{P_R[k]}\right]^\mathrm{T}$ and the vector $\mathbf{a}[k]$. Each information symbol $a_r[k], r = 1, \ldots, R$ is an element of a given symbol alphabet $\mathcal{A}_r[k]$, consisting of $M_r[k] = 2^{2m}, m = 1, 2, \ldots$ symbols for square QAMs.

For exploiting the channel knowledge even further, linear pre-filtering is used in (3). The pre-filtering matrix $\mathbf{V}[k]$ contains analyzing basis vector directions obtained by the Singular Value Decomposition (SVD) of the channel matrix $\mathbf{H}[k] = \mathbf{U}[k]\mathbf{\Sigma}[k]\mathbf{V}^\mathrm{H}[k]$. When (3) is left-multiplied by $\mathbf{U}^\mathrm{H}[k]$ and the SVD of the channel matrix per subcarrier $k$ is applied, (3) reduces to

$$\tilde{\mathbf{y}}[k] = \mathbf{\Sigma}[k]\tilde{\mathbf{a}}[k] + \tilde{\mathbf{n}}[k], \quad (5)$$

where $\tilde{\mathbf{n}}[k] = \mathbf{U}^\mathrm{H}[k]\mathbf{n}[k]$ and $\tilde{\mathbf{y}}[k] = \mathbf{U}^\mathrm{H}[k]\mathbf{y}[k]$.

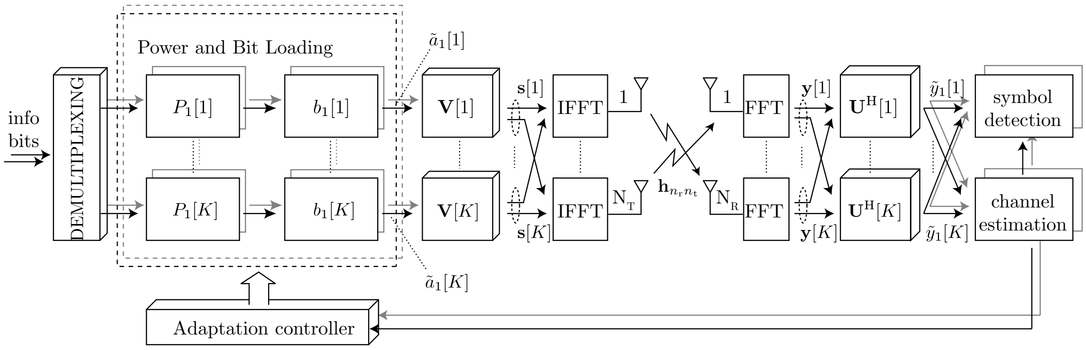


**Fig. 1**. Discrete-time equivalent baseband MIMO OFDM model.

## 3. PROBLEM FORMULATION

The goal is to optimize the coded throughput of the MIMO-OFDM system depicted in Figure 1. The power levels $P_r[k]$ to be assigned to the subcarriers with index $k$ and to the spatial sub-channels with index $r$ are grouped into the transmit power matrix

$$\mathbf{P} = \begin{bmatrix} P_1[1] & \cdots & P_1[K] \\ \vdots & \ddots & \vdots \\ P_R[1] & \cdots & P_R[K] \end{bmatrix}, \tag{6}$$

so are the number of bits $b_r[k]$ grouped into the matrix

$$\mathbf{B} = \begin{bmatrix} b_1[1] & \cdots & b_1[K] \\ \vdots & \ddots & \vdots \\ b_R[1] & \cdots & b_R[K] \end{bmatrix}. \tag{7}$$

The objective is to maximize the coded user throughput $T$ subject to a total transmit power constraint $\bar{P}$ and discrete modulation levels while maintaining a target bit error ratio $\mathrm{BER}_r^{(0)}[k]$ on each subcarrier and in each sub-channel. This objective can thus be formulated as the following constrained optimization problem:

$$\begin{aligned} \text{maximize } T(\mathbf{P},\mathbf{B}) &= \sum_{r=1}^{R}\sum_{k=1}^{K} b_r[k] && (8)\\ \text{subject to coded } \overline{\mathrm{BER}}_r[k] &\leq \mathrm{BER}_r^{(0)}[k] && (9)\\ \sum_{r=1}^{R}\sum_{k=1}^{K} P_r[k] &\leq \bar{P} && (10)\\ P_r[k] &\geq 0 && (11)\\ b_r[k] &\in \{0,1,2,3,4,\ldots\} && (12) \end{aligned}$$

The target BER is assumed to be identical across subcarriers and sub-channels and equal to $\mathrm{BER}^{(0)}$. The relationship between the initial $\mathrm{BER}^{(0)}$ and the resulting coded $\mathrm{PER}^{(1)}$ in the first iteration of the PER adaptation to the target packet error ratio $\mathrm{PER}^{(0)}$ is given by

$$\mathrm{PER}^{(1)} = 1-(1-\mathrm{BER}^{(0)})^{\left(\sum_{r=1}^{R}\sum_{k=1}^{K} b_r^{(1)}[k]\cdot R_{\mathrm{c}}\cdot N_{\mathrm{s}}\right)}, \tag{13}$$

where $N_{\mathrm{s}}$ is the number of OFDM data symbols in a WiMAX frame representing one codeword, and $R_{\mathrm{c}}$ is the coding rate.

## 4. THRESHOLD METRIC AND POWER COST FOR A CONVOLUTIONAL CODE AND $M$-QAM

Note that in (12) both rectangular ($b[k] \in \{1,3,5,\ldots\}$) and square ($b[k] \in \{2,4,6,\ldots\}$) QAMs are allowed ($b[k]=0$ refers to the case of an unused subcarrier). However, we want to reduce the possible constellations just to square ones, what significantly simplifies an adaptive demapper. This excludes also the BPSK modulation, which is used in the first AMC mode of the WiMAX standard. By doing this, one bit at most is lost from the throughput per one OFDM symbol. The proof is given in [1]. Let $d_{\min}[k]$ denote the minimum square Euclidean distance for the given constellation on the $k$th subcarrier. For QAM constellations, it holds that [6]

$$d_{\min}^2[k] = 4g(b[k])E_{\mathrm{s}}[k] = 4g(b[k])P[k]T_{\mathrm{s}}, \tag{14}$$

where $E_{\mathrm{s}}[k]=P[k]T_{\mathrm{s}}$ is the average energy of the constellation chosen for the $k$th subcarrier, $T_{\mathrm{s}}$ is the useful symbol duration (OFDM symbol duration without cyclic prefix) and the constant $g(b[k])$ depends on whether the given constellation is square or rectangular QAM:

$$g(b[k]) := \begin{cases} \frac{6}{5\cdot 2^{b[k]}-4}, & b[k]=1,3,5,\ldots; \\ \frac{6}{4\cdot 2^{b[k]}-4}, & b[k]=2,4,6,\ldots. \end{cases} \tag{15}$$

From the definition of the constant $g(b[k])$ in (15) it is clear that the square QAMs are more power efficient than the rectangular QAMs. Thus, with K subcarriers, it is always possible to avoid the usage of less efficient rectangular QAMs and save the remaining power for other subcarriers to use higher order square QAM [9]. Our exclusion of the rectangular constellations is thus justified.

Equation (14) simply gives us the decreasing minimum Euclidean distance in the constellation with the increasing modulation order and can be easily verified. According to (14) replacing $d_{\min}^2[k]$ by term $d_{\mathrm{avg}}^2[k]=c(b[k])\cdot g(b[k])P[k]$, where $T_{\mathrm{s}}$ is equal to one second, the required power $P[k]$ to transmit $b[k]$ bits/s/Hz is equal to

$$P[k] = \frac{d_{\mathrm{avg}}^2[k]}{c(b[k])\cdot g(b[k])} = \frac{d_0^2[k]}{g(b[k])}, \tag{16}$$

because $d_{\mathrm{avg}}^2[k]/d_0^2[k] = c(b[k])$ and $d_0^2[k]=d_{\min}^2[k]/4$, where $d_0[k]$ is our threshold metric or the unit grid in the

$M$-QAM constellation. The constant $c(b[k])$ that depends on the modulation order is set to be higher than four and it expresses the deviation from the worst-case upper bound on a bit error probability of the convolutional code. This worst-case upper bound refers to the case when $d^2_{\text{avg}}[k] = d^2_{\text{min}}[k]$. In this way, the bit error probability approximation is obtained for the $[133\ 171]_8$ half-rate convolutional code with $M$-QAM modulation for an AWGN channel

$$P_{\text{b}}[k] \approx \frac{1}{n} \sum_{d=d_{\text{free}}}^{d_{\text{free}}+4} \beta(d)[k] \frac{1}{5} \exp\left(\frac{-c(b[k]) \cdot d_0^2[k] \cdot \hat{d}(b[k])}{4N_0}\right), \tag{17}$$

where $n$ is the $n = 1$ input bits of the convolutional encoder and $\hat{d}(b[k]) = d_{\text{free}} + 4$, thus making the exponential term of this approximation independent from $d$. More details about this probability approximation and its verification can be found in [7, 8]. We decided to use for the simulations $c(b[k]) = 8, 10, 20$ for 4-,16-, and 64-QAM, respectively, in order to achieve the bit error ratio in the range of $10^{-6}$ to $10^{-5}$ [8]. The threshold metric or the unit grid $d_0^2[k]$ can be expressed from (17) as follows

$$d_0^2[k] = \left[\ln\left(\sum_{d=d_{\text{free}}}^{d_{\text{free}}+4} \beta(d)[k]\right) - C\right] \frac{4N_0}{c(b[k]) \cdot \hat{d}(b[k]) \cdot \sigma^2[k]}, \tag{18}$$

where $C = \ln(5P_{\text{b}})$ and $\sigma[k]$ is a nonzero eigenvalue of the channel on a subcarrier $k$. This constant $C$ is replaced by $C = \ln 5\text{BER}^{(0)}$ for practical purposes, as the bit error probability can be only achieved in the simulations when the number of transmitted bits goes to infinity. The power cost incurred when loading the $l$th and $(l-2)$th bits (two bits are always loaded in one step) to the $k$th subcarrier and the $r$th sub-channel is from (18) and (16) for coded systems

$$p_r[k,l] = \frac{d^2_{0,r}[k,l]}{g_r[l]} - \frac{d^2_{0,r}[k,l-2]}{g_r[l-2]},\ l = 2,4,6;\ \forall k,r. \tag{19}$$

So this cost is quantified by the additional power needed to maintain the target bit error ratio performance. For $l = 2$, $g_r[l-2]$ is set to $\infty$ and $c_r[0] = \hat{d}_r[0] = 1$.

## 5. GREEDY POWER ALLOCATION ALGORITHM

In this section, the optimal Greedy Power Allocation Algorithm with the packet error ratio adaptation is described. This greedy algorithm converts the frequency-selective channel into a frequency-flat one, what is the main idea behind the water-filling algorithm. Therefore, the performance of the variable-rate $M$-QAM modulation with convolutional encoding in an AWGN channel may be considered per subcarrier, as it was already done in Section 4. Moreover, the target packet error ratio $\text{PER}^{(0)}$ per each codeword can be maintained. Remember that the same target bit error ratio $\text{BER}^{(0)}$ and target packet error ratio $\text{PER}^{(0)}$ across all subcarriers are assumed.

Two iteration loops are given, in the inner $n$th iteration, it is decided whether to load two more bits or not, in the outer $m$th iteration it is checked if the target packet error ratio is achieved. In the first iteration step (m = 1), we start with the target bit error ratio $\text{BER}^{(0)}$ set to $10^{-5}$, if the target packet error ratio equal to $10^{-2}$ is of interest. It is important to have the starting value of BER to obtain the first estimation of the codeword length. In addition, the expression for the coded power cost in (19) depends on the target bit error ratio. It follows from the simulation results, plotted in [8], that for the $\text{PER}^{(0)} = 10^{-2}$, the bit error ratio has to be in the range of $10^{-6}$ to $10^{-5}$ (also the bit error ratio approximation in (17) was previously optimized for this BER region).

In this greedy algorithm, $P_{\text{rem}}$ denotes the remaining power after each iteration, $b_r^{(n)}[k]$ the number of bits loaded at the $k$th subcarrier and the $r$th sub-channel in the $n$th iteration, and $P_r^{(n)}[k]$ to denote the power level in iteration step $n$ on the $k$th subcarrier and in the $r$th sub-channel. The total number of loaded data bits $N_{\text{total}}$ in the $n$th iteration can then be calculated by summing the loaded bits over $R$ sub-channels, $K$ subcarries and $N_{\text{s}}$ OFDM symbols and taking also the coding rate $R_{\text{c}}$ into account

$$N_{\text{total}} = \sum_{r=1}^{R} \sum_{k=1}^{K} b_r^{(n)}[k] \cdot R_{\text{c}} \cdot N_{\text{s}}. \tag{20}$$

The optimal Greedy Power Allocation Algorithm can be performed according to the following steps:

1. In the first (outer) iteration step $m = 1$, $\text{BER}^{(1)}$ is equal to the initial target bit error ratio $\text{BER}^{(0)}$, otherwise

$$\text{BER}^{(m)} = 1 - (1 - \text{PER}^{(0)})^{1/N_{\text{total}}}. \tag{21}$$

2. Initialization step $n = 1$: Set the remaining power equal to the power constraint $P_{\text{rem}} = \bar{P}$. For each subcarrier and sub-channel, set $b_r^{(n)}[k] = P_r^{(n)}[k] = 0$.

3. Compute $p_r(k, b_r^{(n)}[k]+2])$ for all subcarriers and sub-channels, where $b_r^{(n)}[k] \neq 6$. If $b_r^{(n)}[k] = 6$, then set $p_r(k, b_r^{(n)}[k]) = \infty$. Choose the subcarrier and sub-channel that needs the least power to load two additional bits, i.e. select

$$\{r_0, k_0\} = \arg\min_{r,k} p_r(k, b_r^{(n)}[k] + 2). \tag{22}$$

4. If there is not enough power remaining, i.e. if $P_{\text{rem}} < p_{r_0}(k_0, b_{r_0}^{(n)}[k_0] + 2)$ (it always happens if all subcarriers and sub-channels are loaded with 6 bits ($b_r^{(n)}[k] = 6\ \forall r, k$)), then jump to Step 5. Otherwise, load two bits to the $k_0$th subcarrier and the $r_0$th sub-channel, and update iteration variables:

$$\begin{aligned} P_{\text{rem}} &= P_{\text{rem}} - p_{r_0}(k_0, b_{r_0}^{(n)}[k_0] + 2) && (23)\\ P_{r_0}^{(n)}[k_0] &= P_{r_0}^{(n)}[k_0] + p_{r_0}(k_0, b_{r_0}^{(n)}[k_0] + 2) && (24)\\ b_{r_0}^{(n)}[k_0] &= b_{r_0}^{(n)}[k_0] + 2. && (25) \end{aligned}$$

Loop back to step 3 with $n = n + 1$.

5. Calculate

$$\mathrm{PER}^{(\mathrm{m})} = 1 - (1 - \mathrm{BER}^{(m)})^{1/N_{\mathrm{total}}} \tag{26}$$

If $\mathrm{PER}^{(m)} \leq \mathrm{PER}^{(0)}$, then exit with

$$\left\{P_r[k] = P_r^{(n)}[k], b_r[k] = b_r^{(n)}[k]\right\}_{r=1,k=1}^{R,K}, \tag{27}$$

otherwise loop back to step 1 with $m = m + 1$.

Applying this iterative algorithm, the maximum number of iterations for the packet error ratio adaptation is usually equal to two (m = 2). It may happen that the pre-set target bit error ratio is already sufficient to obtain the target packet error ratio after the first outer iteration. Otherwise, the pre-set target bit error ratio is corrected accordingly to meet the PER requirement after the second outer iteration.

The optimal power levels obtained from this algorithm can be only then assigned when the individual SISO channels of the MIMO system are accessible. This is guaranteed by the optimal linear pre-filtering as can be seen in (5)

## 6. SIMULATION RESULTS

The simulation results presented in this section are obtained using a standard compliant IEEE 802.16-2004 WiMAX link level simulator. First, we derive the expression for the ergodic capacity that will be used as a performance bound for the simulated data throughput. For this purpose, the capacities $\bar{C}'[k]$ and $\bar{C}''[k]$ per subcarrier $k$ of a MIMO channel are defined as the ensemble average of the information rate over the $N_{\mathrm{sim}}$ simulated random realizations of the channel matrix $\mathbf{H}[k]$. It is assumed that the channel matrix $\mathbf{H}[k]$ is full-rank, that is, $R = N_{\mathrm{T}} = N_{\mathrm{R}} = N$ and the channel eigenvalues are $\lambda_r = \sigma_r^2 > 0$, for $r = 1, \ldots, N$, what is always fulfilled in the simulations with an uncorrelated Pedestrian B channel.

The ergodic capacity $\bar{C}'$ of the OFDM system can be calculated by summing the individual capacities $\bar{C}'[k]$ of all data subcarriers (192 data subcarriers are used in the WiMAX standard)

$$\bar{C}' = \frac{F}{N_{\mathrm{sim}}} \sum_{s=1}^{N_{\mathrm{sim}}} \sum_{k=1}^{192} \sum_{r=1}^{N} \log_2 \left(1 + \frac{P_r^{(s)}[k]\lambda_r^{(s)}[k]}{N_0}\right), \tag{28}$$

in case of the channel knowledge at the transmitter, and

$$\bar{C}'' = \frac{F}{N_{\mathrm{sim}}} \sum_{s=1}^{N_{\mathrm{sim}}} \sum_{k=1}^{192} \sum_{r=1}^{N} \log_2 \left(1 + \frac{\lambda_r^{(s)}[k]}{N_0}\right), \tag{29}$$

when no channel knowledge is assumed at the transmitter. The ergodic capacity in (28) and the mutual information in (29) was derived from the capacity of the MIMO channel expressed as the sum of the capacities of $N$ SISO channels and can be found in [9]. Since the transmission of an OFDM signal requires also the transmission of a cyclic prefix to avoid inter-symbol interference, and a preamble for the synchronization and channel estimation, the ergodic capacity and mutual information given by Equations (28) and (29) include the

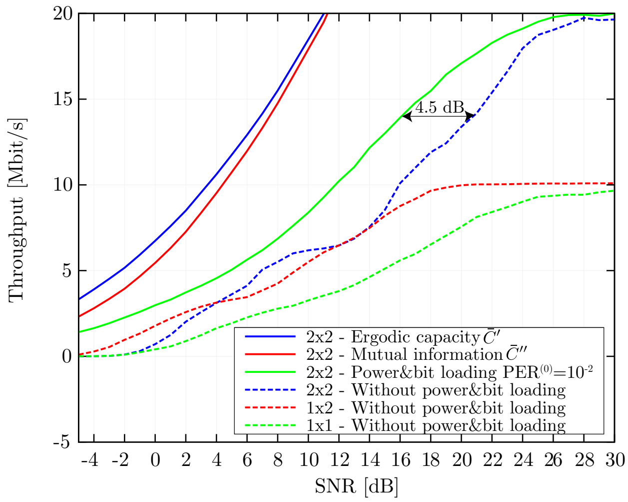


**Fig. 2**. Performance of channel adaptive algorithms in an uncorrelated Pedestrian B channel with $\mathrm{PER}^{(0)} = 10^{-2}$.

correction factor $F$. This correction term accounts for these inherent system losses and is defined according to [10] as follows

$$F = \frac{1}{1+G} \cdot \frac{1/T_{\mathrm{sample}}}{256} \cdot \frac{N_{\mathrm{s}}}{N_{\mathrm{total}}}, \tag{30}$$

where G (1/4 in the simulations) corresponds to the ratio of the cyclic prefix time and useful OFDM symbol time $T_{\mathrm{s}}$, $N_{\mathrm{s}}$ (44 in the simulations) is the number of OFDM data symbols, $N_{\mathrm{total}}$ (47 in the simulations) is the total number of OFDM symbols in one transmission frame (also with the preamble), and $T_{\mathrm{sample}}$ is the sampling rate of the transmit signal. Therefore, the factor $(1/T_{\mathrm{sample}})/256$ is equivalent to the available bandwidth per subcarrier (the FFT size is equal to 256) [10].

In Figure 2, the channel adaptive algorithm for a 2×2 MIMO system with power and bit loading is compared to the ergodic capacity $\bar{C}'$ and mutual information $\bar{C}''$, as well as a 2×2, a 1×2, and a 1×1 system without power and bit loading. The half rate convolutional code of the WiMAX standard is used in combination with 4-,16- and 64-QAM. The 2×2 system using this adaptive algorithm outperforms the standard non adaptive 2×2 system by maximum 4.5 dB. The difference in the low SNR region from the ergodic capacity $\bar{C}'$ is as low as 5 dB. This loss is due to the non-optimal channel coding. With increasing SNR, the loss is growing significantly because of the low rate of the convolutional code and could be reduced by larger symbol alphabets and/or higher coding rates. As expected, the ergodic capacity $\bar{C}'$ with the channel known at the transmitter is higher than the mutual information $\bar{C}''$ in the case of unknown channel. This advantage reduces at higher SNR, as can be observed in Figure 2 and is proved in [9]. Interestingly, although the capacity advantage is only small, the gain in throughput is quite significant.

In Figure 3 the resulting frame error ratio is plotted for the target packet error ratio equal to $10^{-2}$. The adaptation performance of the optimal Greedy Power Allocation Algorithm is satisfactory over a large SNR range of 34 dB.

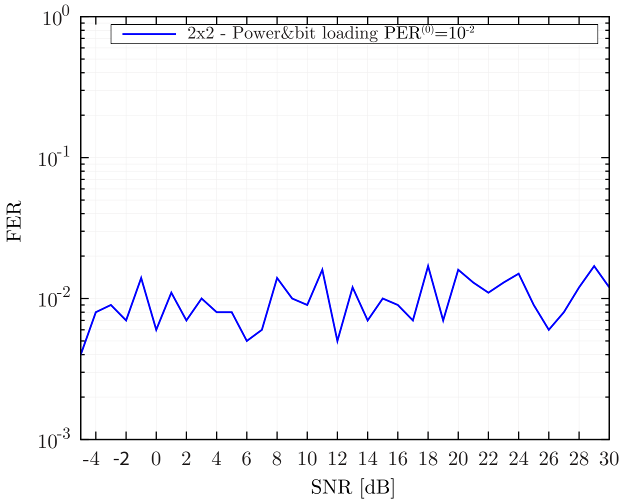


**Fig. 3**. Frame error ratio adaptation for the target packet error ratio $\mathrm{PER}^{(0)} = 10^{-2}$.

## 7. CONCLUSION

In this paper, we create a framework for the bit error ratio and packet error ratio adaptation. In slow fading channels, we find this to be a very promising approach, because the achievable capacity can be reached by applying an appropriate modulation and coding scheme. Moreover, the PER of the system can be controlled. To the best of our knowledge, this is the first attempt to evaluate the performance of the water-pouring based solution in the more realistic scenario of a coded WiMAX system. It remains to further explore the behavior of the proposed algorithm and the possible gains when more efficient channel coding is deployed. We also plan to implement this framework into our Long Term Evolution (LTE) link level simulator [11] that is also based on MIMO-OFDM and is currently under development.

Finally, the complexity of the optimal Greedy Power Allocation Algorithm grows linearly with the number of bits and the number of subcarriers. Obviously, the complexity is considerably large when the number of bits and subcarriers is large. The maximum number of iterations for one particular bit and power level assignment is for example for the $2{\times}2$ MIMO OFDM system with 192 data subcarriers equal to $3 \cdot 2 \cdot 192 = 1152$, where 3 is the maximum number of bit loading steps for the $k$th subcarrier and the $r$th subchannel (4-QAM $\rightarrow$ 16-QAM $\rightarrow$ 64-QAM) . For every assignment of two bits, the power cost has to be calculated across all subcarriers and sub-channels. For the adaptive transmitter design, it is recommended in [1] to use in practice the fast Lagrange bi-sectional search proposed in [12] that also provides an optimal solution with lower complexity.

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