# Calculation of the Spatial Preprocessing and Link Adaption Feedback for 3GPP UMTS/LTE

Stefan Schwarz, Christian Mehlführer and Markus Rupp
Institute of Communications and Radio-Frequency Engineering, Vienna University of Technology
Gusshausstrasse 25/389, A-1040 Vienna, Austria
Email: {sschwarz, chmehl, mrupp} @nt.tuwien.ac.at

***Abstract*—This paper presents an efficient method for calculating the Precoding Matrix Indicator (PMI), Rank Indicator (RI) and Channel Quality Indicator (CQI) at a Long Term Evolution (LTE) User Equipment (UE). The indicators are required for spatial preprocessing and link adaption in the downlink of a 3GPP UMTS/LTE system. To reduce the computational burden for the UE, our method decomposes the problem into two separate steps, one of jointly evaluating the PMI and RI based on a mutual information metric and one of choosing the CQI value to achieve a given target Block Error Ratio (BLER) constraint. The performance of the method is evaluated utilizing an LTE downlink physical layer simulator. The influence of estimated channel knowledge on the feedback choice is investigated for Least Squares (LS) and Linear Minimum Mean Squared Error (LMMSE) channel estimators.**

***Index Terms*—LTE, MIMO, Precoding, Link-Adaption**

## I. INTRODUCTION

In 3GPP's future mobile communication system UMTS Long Term Evolution (LTE) [1] the feedback for channel adaption comprises three values (Channel Quality Indicator (CQI), Rank Indicator (RI), Precoding Matrix Indicator (PMI)) in the so-called closed-loop spatial multiplexing transmission mode [2]. By the CQI the transmitter selects one of 15 modulation alphabet and code rate combinations for transmission. The RI informs the transmitter about the number of useful spatial transmission layers for the current MIMO channel (which is not more than four in the current standard), and the PMI signals the codebook index of the prefered precoding matrix [3]. Finding a jointly optimal solution for these three values will in many cases not be feasible due to feedback delay constraints and limited signal processing hardware. It is therefore necessary to reduce computational complexity, which we achieve by separating the overall optimization process into several steps of finding local independent optima for the three values, thereby sacrificing overall optimality.

LTE is an Orthogonal Frequency Division Multiple Access (OFDMA) system with a system bandwidth of up to 20 MHz (1200 subcarriers). The time schedule for User Equipment (UE) feedback is given by the duration of one subframe (1 ms). The time-frequency grid spanned by the subcarriers and temporal samples within such a subframe is divided into several Resource Blocks (RBs) consisting of 12 subcarriers and 6 temporal samples each. Depending on the mode of operation the feedback granularity ranges from one PMI and CQI value per Resource Block (RB) up to just one value for the full system bandwidth [2]. The optimal feedback values will depend on the subcarrier and time instant, which necessitates some kind of majority decision.

In [4] we already introduced a PMI feedback scheme that is based on maximizing the sum mutual information over subcarriers. This method will be specialized and optimized here for linear receivers and will be used to jointly evaluate the optimal RI and PMI value in Section III.

The CQI value is chosen to achieve a given Block Error Ratio (BLER) target (BLER $\leq 0.1$, a typical operating point for mobile communication systems). This choice is based on a mapping between post equalization Singal to Interference and Noise Ratio (SINR) and CQI for a Single Input Single Output (SISO) AWGN channel, which is evaluated by simulations. It is therefore necessary to map the SINR experienced on every fading subcarrier to an equivalent AWGN channel Signal to Noise Ratio (SNR), a method that is already well known from link level abstraction e.g. [5], [6], [7]. For this purpose we apply Mutual Information Effective SINR Mapping (MIESM) as well as Exponential Effective SINR Mapping (EESM) in Section IV. The system model is introduced in Section II and simulation results for the full feedback scheme and different antenna configurations are presented in Section V. In Section VI we analyze the algorithm and show where it gains in complexity compared to jointly optimizing RI, PMI and CQI. We suggest some additional modifications to further reduce the computational effort.

## II. SYSTEM MODEL

LTE converts a frequency selective channel into a number of narrowband frequency flat channels, by adopting OFDM. The input-output relation on subcarrier $k$, assuming $M_R$ receive and $N_T$ transmit antennas, at sampling time instant $n$ is given by

$$\mathbf{y}_{k,n} = \mathbf{H}_{k,n}\mathbf{W}_i\mathbf{x}_{k,n} + \mathbf{n}_{k,n}, k \in 1, ...K, n \in 1, ...N. \quad (1)$$

$\mathbf{y}_{k,n} \in \mathbb{C}^{M_\mathrm{R}\times 1}$ is the received symbol vector, $\mathbf{H}_{k,n} \in \mathbb{C}^{M_\mathrm{R}\times N_\mathrm{T}}$ is the channel matrix experienced on subcarrier $k$ at time instant $n$, $\mathbf{x}_{k,n} \in \mathcal{A}^{L\times 1}$ is the transmit symbol vector with $\mathcal{A}$ being the utilized symbol alphabet and $\mathbf{n}_{k,n} \sim \mathcal{CN}(0, \sigma_n^2 \cdot \mathbf{I})$ is white, complex-valued Gaussian noise with variance $\sigma_n^2$. The channel matrix and noise variance are assumed to be known by the receiver. The dimension of the transmit symbol vector

978-1-4244-7071-6/10/$26.00 ©2010 IEEE Crown

depends on the number of useful spatial transmission layers $L$.

Spatial preprocessing is carried out with the precoding matrix $\mathbf{W}_i \in \mathcal{W}$. Here $i$ denotes the index within the codebook of precoding matrices $\mathcal{W}$, defined in [3]. Depending on the feedback granularity, the precoder $\mathbf{W}_i$ will be either constant over only one RB or over the total system bandwidth and subframe duration.

The received symbol vector $\mathbf{y}_{k,n}$ is filtered by a linear equalizer given by a matrix $\mathbf{F}_{k,n} \in \mathbb{C}^{L \times M_\mathrm{R}}$. The output of this filter is the post-equalization symbol vector $\mathbf{r}_{k,n}$

$$\mathbf{r}_{k,n} = \mathbf{F}_{k,n}\mathbf{y}_{k,n} = \underbrace{\mathbf{F}_{k,n}\mathbf{H}_{k,n}\mathbf{W}_i}_{\mathbf{K}_{k,n} \in \mathbb{C}^{L \times L}} \mathbf{x}_{k,n} + \mathbf{F}_{k,n}\mathbf{n}_{k,n}. \tag{2}$$

The linear receiver is typically chosen according to a zero forcing or minimum mean square error design criterion [8]. The input signal vector is normalized to unit power.

## III. PMI AND RI FEEDBACK

We have already presented the calculation of the PMI in [4], but we will specialize results here for linear receivers and also extend the idea to allow for the evaluation of the RI as well.

The basic idea is to choose the precoder that maximizes the mutual information for a specific subcarrier- $(1 \ldots K)$ and temporal-range $(1 \ldots N)$ of interest (which is at least a single RB and can be up to the full system bandwidth and subframe duration). Denoting $I_{k,n}$ the mutual information of the resource element $(k, n)$ we obtain

$$\mathbf{W}_j = \arg \max_{\mathbf{W}_i \in \mathcal{W}} \sum_{k=1}^{K} \sum_{n=1}^{N} I_{k,n}(\mathbf{W}_i). \tag{3}$$

In [4] we have considered the pre-equalization mutual information for this choice, which achieves optimal results for maximum likelihood receivers, but not for linear ones in combination with MIMO systems (for interference free MISO systems linear receivers are optimal as well). Therefore we will now use the post-equalization mutual information which is given in terms of the post-equalization $\mathrm{SINR}_{k,n,l}$ as

$$I_{k,n} = \sum_{l=1}^{L} \log_2(1 + \mathrm{SINR}_{k,n,l}) \tag{4}$$

in bits per channel use, with $L$ denoting the number of spatial transmission layers.

The post-equalization SINR on layer $l$ equals

$$\mathrm{SINR}_{k,n,l} = \frac{|\mathbf{K}_{k,n}(l,l)|^2}{\sum_{i \neq l} |\mathbf{K}_{k,n}(l,i)|^2 + \sigma_n^2 \sum_i \mathbf{F}_{k,n}(l,i)}, \tag{5}$$

where $\mathbf{K}_{k,n}(l, i)$ refers to the element in the $l$th row and $i$th column of matrix $\mathbf{K}_{k,n}$ (see Equation (2)). The first term in the denominator corresponds to inter stream interference and the second term to noise enhancement. This expression assumes perfect channel knowledge and no inter cell interference. Of course such impairments can also be considered in the expression. For example, including a channel estimator in the system will just increase the effective noise variance by the mean square error of the channel estimator $\mathrm{MSE}_{k,n}$

$$\mathrm{SINR}_{k,n,l} = \frac{|\mathbf{K}_{k,n}(l,l)|^2}{\sum_{i \neq l} |\mathbf{K}_{k,n}(l,i)|^2 + \tilde{\sigma}_n^2 \sum_i \mathbf{F}_{k,n}(l,i)} \tag{6}$$

$$\tilde{\sigma}_n^2 = \sigma_n^2 + \mathrm{MSE}_{k,n}. \tag{7}$$

Simulation results in [4] have shown that it suffices to calculate the mutual information (4) just for a single channel matrix value per RB to come up with the optimal PMI. This approach is also adopted here. The single channel value can be obtained by averaging the channel over the corresponding Resource Elements (REs). This considerably reduces the computational complexity of the feedback calculation, but as Section V shows, it also entails a rate loss.

The feedback strategy for the PMI and RI values involves two steps (assuming different PMIs on every RB):

1) Calculate the post-equalization mutual information (4) for all possible precoders from $\mathcal{W}_L$ and spatial layer numbers $L = 1 \ldots \min(M_R, N_T)$ for all resource blocks.
2) Find the combination of layer number and precoders that maximizes the sum mutual information over all resource blocks. The RI is given by this layer number and the PMIs by the codebook indices of the precoders. Of course a layer number $L$ can only be combined with the corresponding precoders from $\mathcal{W}_L$.

## IV. CQI FEEDBACK

LTE uses the same modulation order and code rate (corresponding to a CQI value) for all resources allocated to a UE. Nevertheless, RB dependent CQI feedback is supported to give the scheduler the opportunity to schedule users on favourable resources. There are transmission modes defined in [3] that allow for independent codewords per spatial layer, but also for a single codeword for several layers. All these possibilities must be captured by a reasonable feedback strategy.

Our feedback strategy is based on averaging the post-equalization SINR over all resources of interest. This can include SINRs corresponding to single or multiple RBs per layer but also to RBs of different layers. Effective SINR Mapping (ESM) methods map several SINR values to an equivalent SNR value of a SISO AWGN channel (see [5], [6], [7] for details). This equivalent AWGN channel has similar BLER performance as the original OFDM system. In our work we have considered the EESM and MIESM methods. Mathematically the mapping is given by

$$\mathrm{SNR}_{\mathrm{eff}} = \beta f^{-1}\left(\frac{1}{R}\sum_{r=1}^{R} f\left(\frac{\mathrm{SINR}_r}{\beta}\right)\right), \tag{8}$$

where $R$ corresponds to the number of resources of interest. For EESM the function $f$ corresponds to an exponential, for MIESM it is given by the Bit Interleaved Coded Modulation (BICM) capacity [9]. Both methods require the calibration of the CQI dependent $\beta$ value that adjusts the mapping to the different code rates and modulation alphabets. The goal of the calibration is to obtain a close match between the BLE of the equivalent AWGN channel and the BLER of the re fading channel. This calibration was carried out according t a relatively low complex procedure explained in [10]. The values obtained for EESM and MIESM can be found in o LTE physical layer simulator that can be downloaded at [11

The mapping from $\mathrm{SNR}_{\mathrm{eff}}$ to a corresponding CQI valu is carried out such that a BLER lower than 0.1 is achieve For this purpose SISO AWGN simulations have been carrie out for each CQI value that delivered this mapping, whic turned out to be a linear function (see [12] for details on th procedure).

The CQI feedback value is the highest possible value (whic delivers the highest throughput) with BLER $\leq$ 0.1 for th equivalent SISO AWGN channel.

The described method was also compared to another poss bility that jointly chooses RI and CQI to maximize the number of transmitted bits. Both methods have shown equivalent performance in all test cases investigated, but the described method has lower complexity as the ESM SINR averaging only has to be performed once.

## V. Simulation Results

This section presents simulation results obtained with a standard compliant LTE physical layer simulator [13]. Simulations are carried out for a $2 \times 1$, $2 \times 2$ and $4 \times 2$ antenna system. A block fading channel model is assumed; that is, the channel is constant during one subframe duration and is fading independently between subframes. The feedback is sent to the transmitter with a delay of 0, meaning that the feedback values are calculated before the actual transmission. Antennas are assumed to be spatially uncorrelated. Simulations are carried out with a single UE occupying the full system bandwidth. As mentioned in Section III the channel is averaged over one RB before calculating the SINRs according to (5). The main simulation parameters are summarized in Table I.

TABLE I
Simulation Parameters

| Parameter | Value |
|---:|:---|
| System bandwidth | 1.4 MHz |
| Number of subcarriers K | 72 |
| Feedback delay | 0 TTI |
| Channel Model | ITU-T VehA [14] |
| Antenna configurations | 2 transmit, 1 receive ($2 \times 1$)<br>2 transmit, 2 receive ($2 \times 2$)<br>4 transmit, 2 receive ($4 \times 2$) |
| Receiver | Zero Forcing ZF |
| Feedback granularity | full bandwidth |
| Channel estimator | perfect channel knowledge |

### A. Antenna Configuration: $2 \times 1$

We first consider the $2 \times 1$ antenna configuration. As there is no source of interference in a single cell $2 \times 1$ system, the zero forcing receiver is equivalent to the Maximum Likelihood (ML) receiver. The feedback values, calculated according to Sections III and IV, are directly applied at the transmitter. Just a single PMI and CQI value is used for the full system bandwidth. The RI is equal to one.

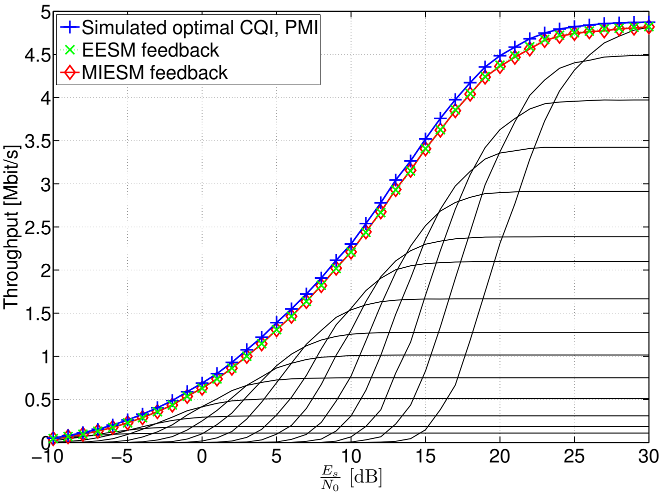

Fig. 1. Throughput over symbol energy to noise power spectral density for a $2 \times 1$ VehA channel.

Figure 1 shows simulated throughput versus transmit energy to noise power spectral density (SNR) obtained for this setup. For every SNR value 5000 channel realizations were simulated. The blue line with plus markers corresponds to an ideal choice of PMI and CQI that maximizes throughput. This choice is obtained by simulating every channel and noise realization with all possible combinations of PMI and CQI values and storing the result of the best combination. The red diamond marked and green cross marked lines correspond to the proposed feedback scheme, when applying MIESM or EESM for SINR averaging. There is virtually no difference between the two methods if they are well calibrated. The black lines show the throughput for fixed CQI values using just PMI feedback. The line with the smallest throughput at 30 dB corresponds to CQI 1 (4 QAM, code rate $\sim$ 0.076), and the one with largest throughput to CQI 15 (64 QAM, code rate $\sim$ 0.925). Adapting the CQI value to the current channel conditions brings a large gain of about $4-5$ dB. Our proposed feedback method looses about $0.5-1$ dB compared to the optimal choice.

Figure 2 shows the BLER obtained during the same simulation. As can be seen, the BLER target is achieved if the SNR is larger than approximately $-5$ dB. Below that value even a CQI value of 1 delivers a higher BLER. From 0 dB SNR upwards, the BLER fluctuates around 0.01. This is because in every subframe the CQI is adjusted to achieve BLER $\leq$ 0.1. Because the SNR range that is mapped to a certain CQI value has a width of approximately 2 dB (see [12]), in most cases the BLER is well below 0.1. The figure also shows the 95% confidence intervals for the MIESM simulation (similar for EESM). The slight tedency to increasing BLER with increasing SNR is caused by the calibration of MIESM or by the choice of the SNR-CQI mapping intervals.

In the next step a channel estimator is included in the

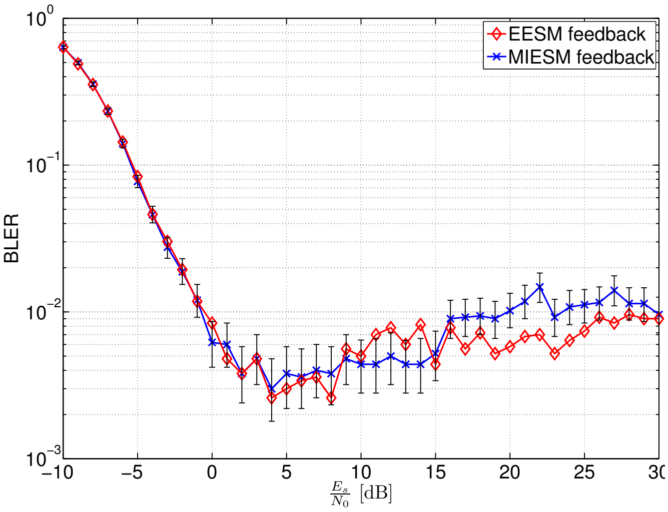

Fig. 2. Block error ratio over symbol energy to noise power spectral density for a 2 × 1 VehA channel.

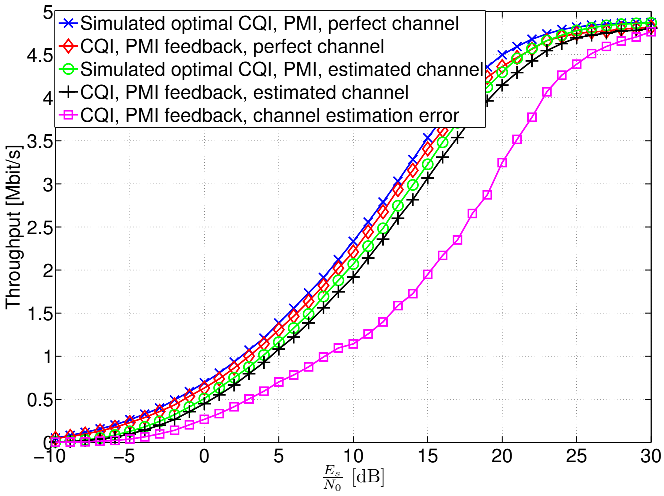

Fig. 3. Throughput over symbol energy to noise power spectral density for a 2 × 1 VehA channel with perfect and estimated channel knowledge.

system. For this purpose Least Squares (LS) [15] and Linear Minimum Mean Squared Error (LMMSE) [16] channel estimators are employed. Figure 3 compares the throughput curves obtained in this case with the ones with perfect channel knowledge for the LS channel estimator. The channel estimation error is incorporated into the post-equalization SINR expression as in (6). Due to the noise enhancement caused by the channel estimator there is a performance loss of about $1-1.5\,$dB for the optimal choice of the feedback values as well as our proposed feedback strategy (compare the lines with cross and diamond markers and the ones with circle and plus markers). The magenta circle marked line shows the behaviour if the channel estimation MSE is not considered in the feedback calculation. The performance drops because the effective channel and therefore the CQI value is overestimated. Also the BLER can not be kept below 0.1 in this case.

When employing the LMMSE channel estimator, the throughput degradation is much less severe as Figure 4 shows. There is almost no difference between the optimal choice lines employing estimated or perfect channel knowledge. The feedback method looses about $0.5-1\,$dB with knowlegdge of the channel estimation error and $0.5-2\,$dB without.

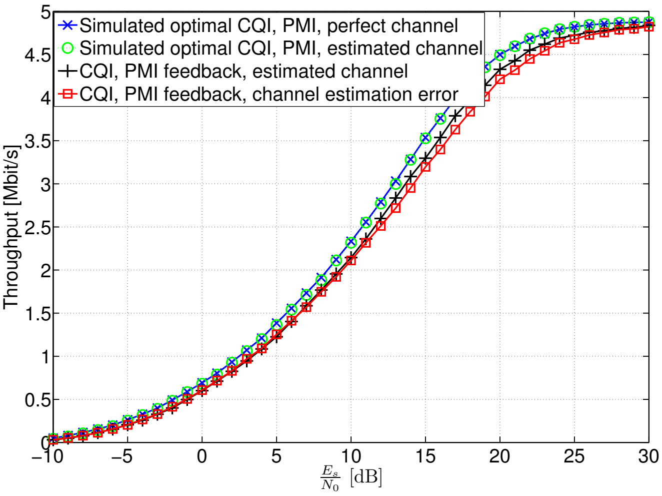

Fig. 4. Throughput over symbol energy to noise power spectral density for a 2 × 1 VehA channel with perfect and estimated channel knowledge.

### *B. Antenna Configuration:* 2 × 2

Next a $2\times2$ system is investigated. In this case also RI feedback is supported and the number of spatial layers is adapted according to the feedack. Figure 5 shows the simulation results obtained for this case. The green line with plus markers uses all feedback capabilities (PMI, RI and CQI). For this result the channel is averaged over an RB before calculating SINR values. The red line with circle markers shows the performance of the proposed full feedback scheme if the channel is not averaged over an RB, but individual SINRs are calculated for every Resource Element (RE). At high SNR values $\sim 30\,$dB this method delivers a performance gain of almost $2\,$dB and should therefore be considered whenever complexity is not an issue (in the $2\times1$ case no gain was observed). The optimal performance, obtained by exhausitive search, is shown in blue with cross markers. Our feedback method looses about $0.5-1\,$dB in SNR if channel averaging is not applied. The black line with diamond markers employs PMI and CQI feedback but fixes the number of spatial streams to RI $=2$ while the magenta square marked line fixed it to RI $=1$. A spatial stream number of two results in poor performance at low SNR. At $1\,$Mbit/s throughput the dual stream mode looses about $4\,$dB compared to the stream adaptive mode and also the single stream mode. In this regime the performance is therefore dominated by single stream operation. This shows that it is beneficial to exploit diversity and array gain instead of multiplexing gain at low SNR.

In Figure 6 the BLERs corresponding to the throughput curves in Figure 5 are depicted. If all feedback values are adapted or if RI is fixed to one the BLER meets the target

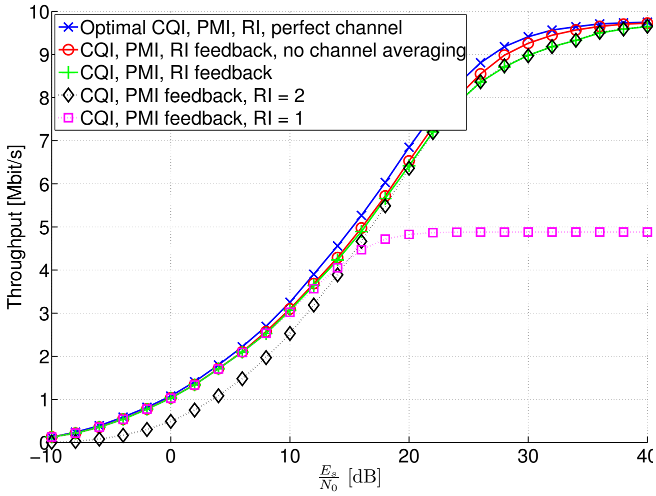

Fig. 5. Throughput over symbol energy to noise power spectral density for a $2 \times 2$ VehA channel.

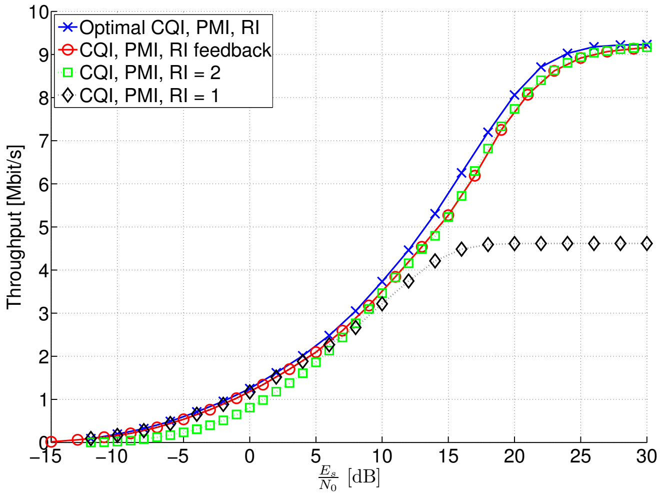

Fig. 7. Throughput over symbol energy to noise power spectral density for a $4 \times 2$ VehA channel.

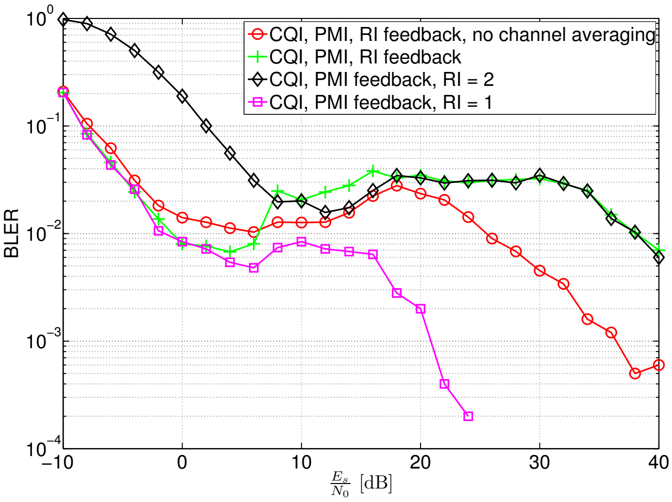

Fig. 6. BLER over symbol energy to noise power spectral density for a $2 \times 2$ VehA channel.

(BLER $\leq 0.1$) already for SNR $\geq -8$ dB, while it requires more than 2 dB of SNR if the stream number is fixed to two. As soon as dual stream transmission becomes more dominant (at approximately 7 dB SNR) the BLER increases, but still meets the target (green plus marked line). The figure shows that the BLER when employing no channel averaging (red circle marked line) starts decreasing at approximately 25 dB while this happens only at 35 dB (green plus marked line) when the channel is being averaged. It is also this regime in which the throughput degradation of channel averaging occurs.

### *C. Antenna Configuration:* $4 \times 2$

In the previous configurations with two transmit antennas, the number of possible precoders according to the standard [3] is small. For two transmit antennas there are just four precoders for single layer transmission and two precoders for dual layer transmission. With four transmit antennas, the amount of precoders grows to sixteen for every layer number, which makes the choice more complex. Nonetheless, our feedback method works well as Figure 7 shows. Channel averaging is applied in the feedback calculation procedure. The loss in SNR compared to the optimal choice is similar to the $2 \times 2$ case and is approximately $0 - 1.5$ dB depending on the throughput. Again, at low SNR the transmission is dominated by single stream operation and at $\sim 7$ dB, dual stream operation outperforms single stream operation. The BLER performance is similar as in the $2 \times 2$ configuration. Comparing the throughput performance of the $2 \times 2$ and $4 \times 2$ configurations shows that the $4 \times 2$ system gains approximately 3 dB SNR at 3 Mbit/s throughput. In saturation at 30 dB SNR the throughput of the $4 \times 2$ system is less because there are more reference symbols due to the larger amount of transmit antennas.

## VI. Computational Complexity Gains

The complete feedback algorithm consists of the following steps:

1) Computation of the post-equalization SINRs and mutual informations for all rank and precoding matrix combinations and all RBs using (4), (5) (up to 32 combinations for $4 \times 2$).
2) Choice of the rank and precoding matrix combination that maximizes the sum mutual information (3).
3) Calculation of the effective SNR using ESM (8) and mapping to a corresponding CQI value.

The first step has considerable computational complexity, as it requires computing the receive filter for all precoders and ranks. Complexity is reduced here, by not considering every subcarrier and temporal sample on its own, but just a single value per RB. This reduces the amount of computations by a factor of 72. For channels with low delay spread, this amount can be even further reduced without degrading the performance (cf. [4]). The complexity of the second step is negligible. The third step requires SNR averaging for all

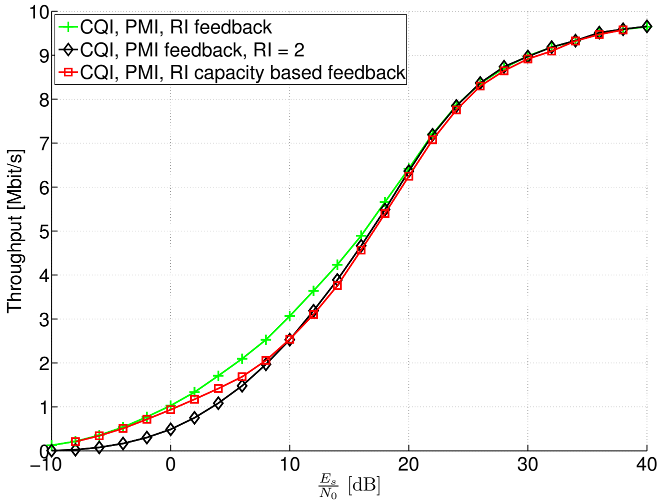

Fig. 8. Throughput over symbol energy to noise power spectral density for a 2 × 2 VehA channel.

possible CQI values (15 in our case), which is also complex. A feedback method that jointly optimizes RI, PMI and CQI would need to repeat this task for all rank and precoder combinations. Therefore, by choosing RI and PMI separately from CQI, we gain here up to a factor of 32 in complexity, depending on the antenna configuration.

A further reduction of computational complexity is possible by choosing PMI and RI from the theoretical capacity given by

$$I_{k,n} = \log_2 \det \left( \mathbf{I}_L + \frac{1}{\sigma_n^2} \mathbf{W}_i^H \mathbf{H}_{k,n}^H \mathbf{H}_{k,n} \mathbf{W}_i \right). \tag{9}$$

In [4] we show that this entails an SNR loss of $0 - 1.5\,\text{dB}$ for the PMI choice depending on the antenna configuration. The receive filter then only needs to be calculated for the chosen combination of PMI and RI to find the appropriate CQI. Figure 8 compares the performance of this method with the previous one for a $2 \times 2$ channel. At low SNR ($\sim -5\,\text{dB}$) both methods transmit in single stream mode and choose the same precoder. But at around $5\,\text{dB}$ the theoretical capacity is a too optimistic estimate for the performance of the zero forcing receiver. The feedback method switches to rank 2, which would deliver better performance with an ML receiver, but not with the linear receiver. Nevertheless the choice of the precoder is almost perfect ($\sim 0.1\,\text{dB}$ loss due to wrong precoder choice). In a $2 \times 1$ system both methods perform similar, because the linear receiver obtains ML performance.

## VII. CONCLUSION

In this paper we present a suboptimal, reduced complexity PMI, RI and CQI feedback method for 3GPP UMTS/LTE. These feedback values are used for spatial preprocessing and link adaption at the transmitter (eNodeB). We show that our method performs close to optimal (in terms of throughput) for different antenna configurations and that the imposed BLER target is met. We also investigate the influence of channel estimation errors on our method and see that the performance is similar if the estimation error is included in the feedback calculation. Neglecting the channel estimation error deterioriates the performance of the method considerably.

## ACKNOWLEDGEMENT

This work has been funded by A1 Telekom Austria AG and the Institute of Communications and Radio-Frequency Engineering, Vienna University of Technology.

## REFERENCES

[1] 3GPP, “Technical Specification Group Radio Access Network; (E-UTRA) and (E-UTRAN); Overall description; Stage 2,” September 2008. [Online]. Available: http://www.3gpp.org/ftp/Specs/html-info/36300.htm.

[2] 3GPP, “Technical Specification Group Radio Access Network; Evolved Universal Terrestrial Radio Access (E-UTRA); Physical layer procedures (Release 8),” March 2009. [Online]. Available: http://www.3gpp.org/ftp/Specs/html-info/36213.htm.

[3] 3GPP, “Technical Specification Group Radio Access Network; Evolved Universal Terrestrial Radio Access (E-UTRA); Physical Channels and Modulation (Release 8),” September 2009. [Online]. Available: http://www.3gpp.org/ftp/Specs/html-info/36211.htm.

[4] S. Schwarz, M. Wrulich, and M. Rupp, “Mutual Information based Calculation of the Precoding Matrix Indicator for 3GPP UMTS/LTE,” in Proc. IEEE Workshop on Smart Antennas 2010, (Bremen, Germany), February 2010.

[5] L. Wan, S. Tsai, and M. Almgren, “A Fading-Insensitve Performance Metric for a Unified Link Quality Model,” in Proc. IEEE Wireless Communications & Networking Conference WCNC, 2006.

[6] X. He, K. Niu, Z.He, and J. Lin, “Link Layer Abstraction in MIMO-OFDM System,” in Proc. International Workshop on Cross Layer Design, 2007.

[7] R. Sandanalakshmi, T. Palanivelu, and K. Manivannan, “Effective SNR Mapping for Link Error Prediction in OFDM based Systems,” in Proc. IET-UK International Conference on Information and Communication Technology in Electrical Sciences ICTES, 2007.

[8] D. Tse and P. Viswanath, Fundamentals of Wireless Communications. Cambridge University Press, 2008.

[9] G. Caire, G. Taricco, and E. Biglieri, “Capacity of bit-interleaved channels,” Electron. Lett., vol. 32, issue 12, pp. 1060–1061, June 1996.

[10] A. M. Cipriano, R. Visoz, and T. Sälzer, “Calibration Issues of PHY Layer Abstractions for Wireless Broadband Systems,” in Proc. IEEE Vehicular Technology Conference Fall VTC2008-Fall, (Calgary, Canada), September 2008.

[11] [Online]. Available: http://www.nt.tuwien.ac.at/ltesimulator/.

[12] J. Ikuno, M. Wrulich, and M. Rupp, “System level simulation of LTE networks,” in Proc. 71st Vehicular Technology Conference VTC2010-Spring, 2010.

[13] C. Mehlführer, M. Wrulich, J. C. Ikuno, D. Bosanska, and M. Rupp, “Simulating the Long Term Evolution Physical Layer,” in Proc. 17th European Signal Processing Conference EUSIPCO 2009, (Glasgow, Scotland), August 2009. [Online]. Available: http://publik.tuwien.ac.at/files/PubDat_175708.pdf.

[14] ITU, “Recommendation ITU-R M.1225: Guidelines for Evaluation of Radio Transmission Technologies for IMT-2000,” tech. rep., ITU, 1997.

[15] J. J. van de Beek, O. Edfors, M. Sandell, S. K. Wilson, and P. O. Borjesson, “On channel estimation in OFDM systems,” in Proc. IEEE 45th Vehicular Technology Conference VTC1995, vol. 2, pp. 815–819, 1995.

[16] S. Omar, A. Ancora, and D. T. M. Slock, “Performance analysis of general pilot-aided linear channel estimation in LTE OFDMA systems with application to simplified MMSE schemes,” in Proc. IEEE 19th International Symposium on Personal, Indoor and Mobile Radio Communications, pp. 1–6, September 2008.