Mutual Information based Calculation of the Precoding Matrix Indicator for 3GPP UMTS/LTE

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Abstract—**This paper presents an efficient method for calculating the Precoding Matrix Indicator (PMI) at the receiver. The PMI is required for MIMO precoding in the downlink of a 3GPP UMTS/LTE system. Our method is based on maximizing the mutual information between the transmitted and received symbols with respect to the precoding matrix applied at the transmitter. The advantage of this method is that it is independent of the symbol alphabet (4/16/64 QAM) and code rate applied, which are signaled by the channel quality indicator (CQI). Although this paper only focuses on the selection of the optimal PMI, such a procedure eventually allows to decouple both problems (CQI and PMI calculation), thereby reducing complexity. The proposed method also provides means to obtain the number of useful MIMO transmission layers, signaled in form of the Rank indicator (RI), by maximizing mutual information also with respect to this value. The performance of the method is evaluated utilizing an LTE downlink physical-layer simulator.**

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

I. Introduction

3GPP UMTS Long Term Evolution (LTE) [1] requires the calculation of three different feedback values (CQI, RI, PMI) at the receiver to perform channel adaption 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. The RI informs the transmitter about the number of useful 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 precoding matrix (depending on the number of transmit antennas) that shall be applied at the transmitter [3]. Optimizing over all possible combinations of 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 complexity, which can be achieved by separating the overall optimization process into several steps of finding local independent optimal values for the three feedback parameters, thereby sacrificing overall optimality.

A lot of different criteria for identifying an—in the respective sense—optimal precoder from the given codebook have already been proposed in [4]. Most of these criteria are receiver specific with the exception of the capacity selection criterion that chooses the precoder that maximizes capacity. Furthermore, in [4] the channel was assumed to be frequency flat, which in most situations will not be the case for the high data rate LTE system. However, because LTE is an orthogonal frequency division multiplexing (OFDM) system, every subcarrier will experience a frequency flat channel and the criterion can be applied at every subcarrier. Depending on the system bandwidth and the number of users simultaneously served, the number of subcarriers can vary from 12 for a single resource block to 1200 for the full 20 MHz system bandwidth. Not all of these carriers will experience the same precoding to be optimal, which requires some kind of majority decision.

The paper is organized as follows: Section II reviews the mutual information based method of [4] to choose the prefered precoder and its natural extension to an OFDM system. Then the lower complexity *mean mutual information based* method for evaluating the precoding matrix is introduced. Section III shows performance results in terms of simulated uncoded bit error ratios (BER) and coded block error ratios (BLER) over different types of channels and for different antenna configurations. Next in Section IV the mean mutual information based method is investigated in more detail. The influence of the number of subcarriers, over which the mean channel is calculated, is investigated, leading to the conclusion that averaging can be achieved without any performance degradation up to a certain number of subcarriers. The same is then also repeated in the time domain in Section V, showing that averaging over the whole LTE subframe (1 ms duration) is possible without performance degradation until very high user equipment (UE) speeds respectively very low coherence time. Lastly Section VI investigates the performance degradation with a feedback delay in multiples of LTE transmission timing intervals (TTIs) present and Section VII concludes the paper.

II. Mutual Information based Precoding

LTE is an OFDM system that converts a broadband frequency selective channel into K narrowband frequency flat channels with the help of a discrete fourier transform (DFT) and application of a cyclic prefix. Assuming M_R receive and N_T transmit antennas, the input-output relation between the transmit symbol vector $\mathbf{x}_k \in \mathcal{A}^{L\times 1}$ and the received symbol vector $\mathbf{y}_k \in \mathbb{C}^{M_\mathrm{R}\times 1}$ on subcarrier k, at an arbitrary sampling time instant, is given by

$$\mathbf{y}_k = \mathbf{H}_k \mathbf{W}_i \mathbf{x}_k + \mathbf{n}_k, \tag{1}$$

with $\mathcal{A}$ being the utilized symbol alphabet and L the number of spatial transmission layers. In Equation (1), $\mathbf{H}_k \in \mathbb{C}^{M_\mathrm{R}\times N_\mathrm{T}}$ is the channel matrix experienced on subcarrier k and $\mathbf{n}_k \sim$

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$CN(0, \sigma_n^2 \cdot \mathbf{I})$ is white, complex-valued Gaussian noise with variance σ_n^2 added at the M_R receive antennas. Both values, σ_n^2 and $\mathbf{H}_k$, are assumed to be known by the receiver. The transmit symbols $\mathbf{x}_k$ are precoded with the same precoding matrix $\mathbf{W}_i \in \mathcal{C}$ in the whole frequency interval $\{1, 2, ..., K\}$ of interest (for LTE this is typically several resource blocks in width). Here i denotes the index within the codebook of precoding matrices $\mathcal{C}$, defined in [3].

With Gaussian signalling on each antenna (which in practice of course is not fullfilled) the capacity on subcarrier k in bits per channel use is

$$I_k(\mathbf{W}_i) = \log_2 \det \left(\mathbf{I}_L + \frac{1}{\sigma_n^2} \mathbf{W}_i^H \mathbf{H}_k^H \mathbf{H}_k \mathbf{W}_i \right) \tag{2}$$

according to [7], where ^{H} denotes conjugate complex transposition. Furthermore, the transmit power is assumed to be one. For this definition to make sense the channel must vary sufficiently slow, such that a large codeblock experiences the same channel conditions. Although in many cases, especially in high mobility environments, this requirement will be violated, the above definition can nevertheless be viewed as a measure for instantaneous channel quality albeit the calculated rate might not be reached.

In order to come up with a single precoder for the whole bandwidth of K subcarriers under consideration the natural extension to the above described method is to simply sum up the mutual information on every carrier k and maximize this sum rate with respect to the precoding matrix.

$$\mathbf{W}_j = \arg \max_{\mathbf{W}_i \in \mathcal{C}} \sum_{k=1}^{K} I_k(\mathbf{W}_i) \tag{3}$$

Only the index j of the optimal precoder $\mathbf{W}_j = \mathbf{W}_i$ is then fed back from the receiver to the transmitter.

The burden of this method is, that the computational effort can be prohibitivly large if the number of subcarriers grows large. To reduce complexity the *mean mutual information based* method splits up the total set of subcarriers $\{1, 2, ..., K\}$ into $D \leq K$ disjoint subsets $\mathcal{D}_d$, with $\bigcup_{d=1}^{D} \mathcal{D}_d = \{1, 2, ..., K\}$. The idea is to choose the width of these subsets such that the capacity formula (2) within each subset can be replaced by its linear Taylor approximation around the arithmetic mean of the channel matrix $\overline{\mathbf{H}}_d$ over the corresponding set of subcarriers

$$\overline{\mathbf{H}}_d = \frac{1}{|\mathcal{D}_d|} \sum_{k \in \mathcal{D}_d} \mathbf{H}_k. \tag{4}$$

Instead of using this channel averaging it is simpler to just use channel subsampling, meaning that the channel inside one interval $\mathcal{D}_d$ is replaced by a sample value $\mathbf{H}_j$ taken from the interval

$$\hat{\mathbf{H}}_d = \mathbf{H}_j, \; j \in \mathcal{D}_d. \tag{5}$$

For both cases let I_d denote the capacity corresponding to this substitute channel. The sum rate R_d of the interval $\mathcal{D}_d$

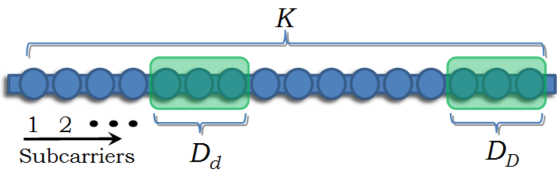

Fig. 1. Illustration of the splitting of the total subcarrier interval of size K into subsets $\mathcal{D}_d, \; d = 1 \ldots D$.

can then simply be calculated as

$$R_d = I_d \cdot |\mathcal{D}_d|, \tag{6}$$

where $|\cdot|$ denotes cardinality of the set. This idea is illustrated in Fig. 1. The *mean mutual information based* method involves just a summation over these D sum rates R_d to find the precoding matrix

$$\mathbf{W}_j = \arg \max_{\mathbf{W}_i \in \mathcal{C}} \sum_{d=1}^{D} R_d. \tag{7}$$

The design of the method suggests that the frequency intervals $\mathcal{D}_d$ must be small (depending on the strength of channel variations over frequency) to produce accurate results, such that the linear Taylor approximation for the capacity is valid. But, as Section IV shows, the codebook constraint for the precoding matrices allows to relax this limitation and increase the width of the frequency intervals without influencing the choice of the precoder.

Note that the described method does not work for the 2×2 antenna case when transmitting two spatial streams simultaneously, as in this case the unitary precoding matrices do not influence the channel capacity.

If not mentioned otherwise, the simulation results presented in the following sections are obtained with channel averaging according to Equation (4).

III. Performance Comparison

This section presents simulation results obtained with a standard compliant LTE physical-link simulator [6]. The main simulation parameters are summarized in Table I. The code rate and modulation alphabet correspond to a channel quality

TABLE I
SIMULATION PARAMETERS

System bandwidth:	1.4 MHz
Number of subcarriers K:	72
Feedback delay:	0 TTI
Channel Model:	ITU-T VehA ITU-T PedA
Antenna configuration:	4 transmit, 2 receive (4×2), L = 2 2 transmit, 1 receive (2×1), L = 1
Receiver:	Soft Output Sphere Decoder SSD Zero Forcing ZF
Modulation alphabet:	QPSK
Code rate:	0.3
UE speed:	0 km/h

indicator value CQI $= 4$ according to [2]. The Soft Output Sphere Decoder (SSD) implementation is based on [8] with performance of an a posteriori probability detector. A feedback delay of 0 TTI (transmission time intervals) means that the precoding matrix indicator is actually calculated before transmission. As the user equipment (UE) does not move, the channel does not change during a subframe but channel realizations of different subframes are drawn independently from each other to average performance results (block fading). Different entries of the channel matrix are uncorrelated, meaning that there is no correlation between different antenna elements. For the eNodeB this assumption can be well fullfilled in practice while for the user equipment it might be too stringent, but can be justified by assuming cross polarized antennas.

Fig. 2 shows uncoded bit error ratios obtained with different precoding schemes when simulating a 2×1 PedA [9] scenario. In this case the ZF receiver already achieves maximum likelihood performance and therefore performs as good as the SSD. The line labeled *ZF optimal choice* is obtained by simulating every subframe with all possible precoding matrices—while keeping the channel and noise realizations constant—and taking the best result, in terms of minimal uncoded BER. *ZF no feedback* corresponds to the result obtained, when keeping the precoding matrix constant at $\mathbf{W}_1 = \frac{1}{\sqrt{2}}\begin{pmatrix}1\\1\end{pmatrix}$, but the other precoders defined in [3] deliver the same result. The result obtained with the *mutual information based* choice of the precoder is labeled *ZF mutual information* and the one labeled *ZF mean mutual information* corresponds to the *mean mutual information based* method, when averaging the channel over all 72 subcarriers. The figure clearly shows

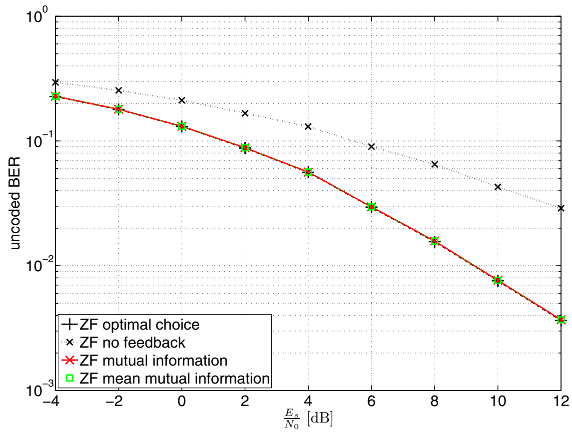

Fig. 2. Uuncoded BER over symbol energy to noise power spectral density for various precoder selection schemes for a 2×1 PedA channel.

that both feedback methods are practically optimal for this case and an improvement of ~ 4 dB is obtained over fixed precoding at a BER of 10^{-1}. Also in terms of coded block error ratio the performance is optimal as Figure 3 shows. The hybrid ARQ mechanism, included in the LTE specification, was deactivated to obtain these results. The typical operating

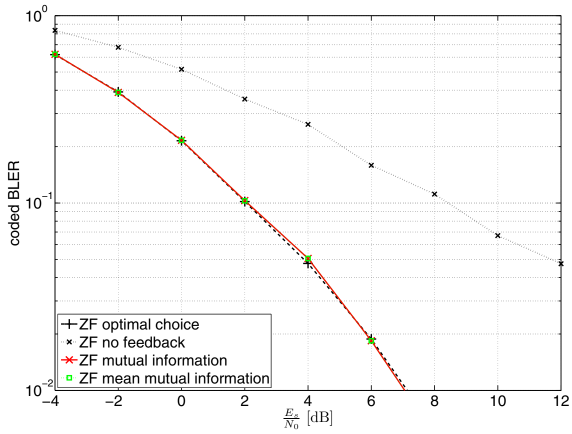

Fig. 3. Coded BLER over symbol energy to noise power spectral density for various precoder selection schemes for a 2×1 PedA channel.

point of a mobile communications system is at a BLER of 0.1 for the first transmission. For this value the gain is even larger than 6 dB. It is also remarkable that there is no loss in performance when averaging the channel over the whole system bandwidth. The r.m.s. delay spread of the PedA channel equals $\tau_{\mathrm{rms}} = 45\,\mathrm{ns}$ which amounts to a coherence bandwidth of $B_{\mathrm{c}} = \frac{1}{2\tau_{\mathrm{rms}}} = 11\,\mathrm{MHz}$ [10].

Fig. 4 shows similar results for a 4×2 VehA channel ($\tau_{\mathrm{rms}} = 370\,\mathrm{ns}$, $B_{\mathrm{c}} = 1.35\,\mathrm{MHz}$) when transmitting two spatial streams simultaneously. The BER is equal for both streams. The improvements of the *mutual information based* method are similar substantial, but the *mean mutual information based* method looses significantly, especially with the ZF receiver. Simulation results in terms of uncoded BLERs are given in

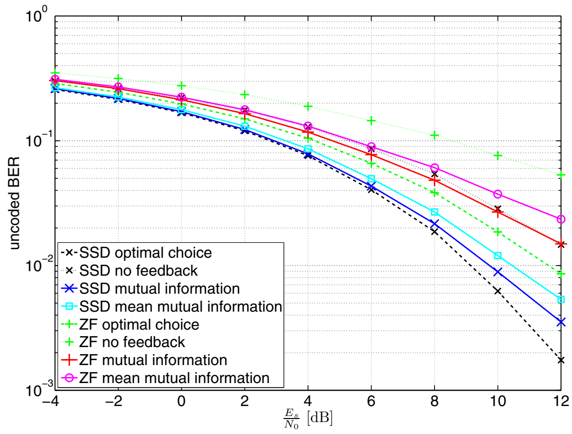

Fig. 4. Uncoded BER over symbol energy to noise power spectral density for various precoder selection schemes for a 4×2 VehA channel.

Fig. 5 for the same setup. For a BLER of 10^{-1} the *mutual information based* feedback scheme performs optimally with the SSD and looses about 0.5 dB with the ZF receiver compared to the optimal values. The *mean mutual information* *based* feedback scheme looses about 0.5 dB with the SSD and almost 1.5 dB with the ZF compared to the optimal values. Nevertheless, combining a ZF receiver with mean mutual information based feedback still performs as good as the SSD without feedback. The differences between VehA and PedA channels can be explained by considering the lower coherence bandwidth of the VehA channel, which causes the channel mean not to reproduce the actual channel well for all subcarriers. Figures 2-5 can be reproduced by running

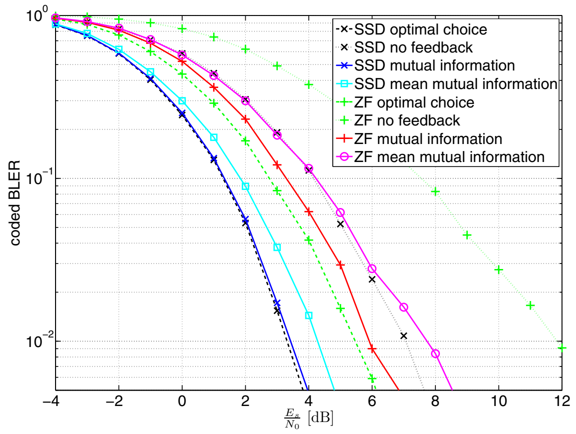

Fig. 5. Coded BLER over symbol energy to noise power spectral density for various precoder selection schemes for a 4×2 VehA channel.

the Matlab script file **WSA_schwarz_batch** available in [6] version v.1.3.

In Figure 6 a comparison of channel averaging and channel subsampling (Equations (4) and (5)) is given over system bandwidth. The results are given in terms of uncoded BER and are obtained for a 2×1 VehA channel using a ZF receiver, independent block fading, the *mean mutual information based* feedback method and a SNR of 10 dB. A single interval $(D = 1)$ spans the whole bandwidth. For channel sampling the sampling point is chosen in the middle of the interval. The figure shows that for small interval bandwidths channel subsampling performs as good as channel averaging, while for bandwidths $\geq$ 1.4 MHz channel averaging outperforms subsampling. Of course these values also depend on the channel under consideration, especially on the strength of channel variation over bandwidth. Practical systems will typically not just use a single channel coefficient to compute the feedback for the total system bandwidth. If the channel is divided into sufficiently small intervals, channel subsampling can be used instead of channel averaging without performance loss.

IV. Influence of Averaging Bandwidth

The previous section has shown that the *mean mutual information based* method can perform as good as the *mutual information based* method (c.f. Figure 3) if the channel is divided in sufficiently small intervals. To further elaborate on this fact this section presents simulation results with varying averaging bandwidth. This means that the total number of

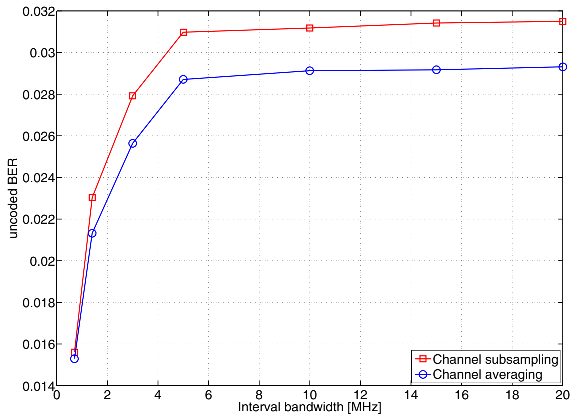

Fig. 6. Uncoded BER over interval bandwidth when using channel averaging or channel sampling.

subcarriers K is divided into D intervals, according to Section II, over which the channel is averaged whereby the width of all intervals is the same. Simulations are carried out with different types of channels at constant SNR. Table II presents the main simulation parameters.

TABLE II
Simulation Parameters

System bandwidth:	10 MHz
Number of subcarriers K:	600
Feedback delay:	0 TTI
Channel Model:	ITU-T VehA ITU-T PedB 3GPP TU
Antenna configuration:	2 transmit, 1 receive (2×1), L = 1
Receiver:	Soft Output Sphere Decoder SSD
CQI:	4
E_s/N_0:	6 dB
UE speed:	0 km/h

Figure 7 presents the simulated uncoded BER over averaging bandwidth respectively number of intervals D for different channel types and a 2×1 antenna configuration.

The figure clearly shows a threshold behaviour of the bit error ratio. It is possible to increase the averaging bandwidth until a certain, channel dependent threshold without increasing the BER. The threshold value for a BER increase of 1% lies at about 1 MHz for the VehA channel, at 0.7 MHz for the PedB channel and at 0.5 MHz for the TU channel. This means that the number of intervals, for which the capacity according to Equation (2) must be calculated, decreases from 600 (mutual information based method) to 20 for the TU channel and to even just 10 for the VehA channel. This behaviour does not change when simulating other system bandwidths and antenna configurations. Although the BER seems to saturate at high averaging bandwidths, it is still much less than that when using no feedback ($\sim$ 0.09 for all three channel types).

The open question is, how to find this channel dependent

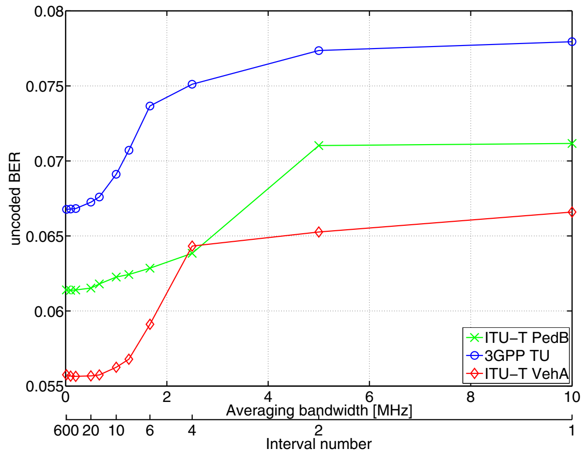

Fig. 7. Uncoded BER over averaging bandwidth for various channel types.

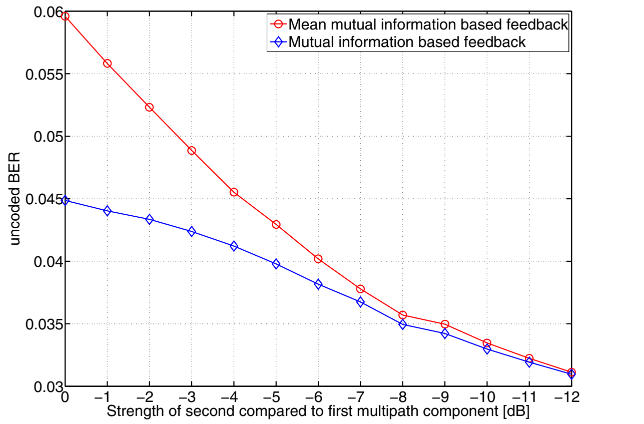

Fig. 8. Uncoded BER over strength of second channel tap for a two tap channel model.

threshold value for a given channel. Inspecting the coherence bandwidths of the three simulated channels (VehA: $B_c = 1.35$ MHz, TU: $B_c = 1$ MHz, PedB: $B_c = 0.8$ MHz) shows that there is no direct relation between the threshold bandwidth and the coherence bandwidth, as the threshold bandwidth is smallest for the TU channel while the coherence bandwidth is smallest for PedB. The reason for this is that multipath components that are weaker than about -8 dB do not influence the choice of the precoding matrix anymore. This has been found by simulating a channel with only two multipath components. The channel parameters are summarized in Table III.

TABLE III
SIMULATION PARAMETERS

System bandwidth:	1.4 MHz
Number of subcarriers K:	72
Feedback delay:	0 TTI
Channel model: Delays: Expected tap strength:	two tap channel [0,800] ns variable
Antenna configuration:	2×1
Receiver:	Zero Forcing
CQI:	4
E_s/N_0:	6 dB
UE speed:	0 km/h

The channel gains of the two taps are Rayleigh distributed and the expected strength of the second component is varied, while keeping the total channel gain constant. The delays have been chosen such that a total period of the channel variation over frequency is experienced inside the system bandwidth. Fig. 8 shows how the uncoded BER behaves with decreasing strength of the second multipath component for the mutual and mean mutual information based feedback methods, with averaging over all 72 subcarriers. The BER of both feedback methods as well as the BER gap between the two decreases with decreasing strength of the second channel tap.

A simple method to come up with an estimate of the threshold bandwidth value for an arbitrary channel realization is to assume that for a given variation of the absolute value of the channel gain over frequency one needs a constant number of averaging intervals to resolve this variation. So if the channel varies by ΔH_k over Δk subcarriers one needs

$$i_k = c \cdot \Delta H_k \tag{8}$$

intervals in the frequency range Δk. This amounts to a total number of

$$I_k = i_k \cdot \frac{K}{\Delta k} = c \cdot \frac{\Delta H_k}{\Delta k} K \tag{9}$$

intervals for the total system bandwidth consisting of K subcarriers.

The algorithm for estimating the necessary number of intervals consists of three steps

1) Find minima and maxima of the absolute value of the channel matrix over frequency for every channel matrix element. ΔH_k is given by the difference between neighbouring extrema.
2) Calculate the number of necessary intervals according to (9) for all ΔH_k.
3) Take the mean of these interval numbers for every channel matrix element and find the maximum over the matrix elements. This value is used as the number D of averaging intervals.

The constant c has been evaluated by simulating different channel types and antenna configurations. A value of $c = 1.2$ gives good performance. With this value of c for the *mean mutual information based* method the BER does not increase compared to the *mutual information based* method. Still the algorithm is a bit to conservative because it takes some more intervals than would actually be necessary. Improvements could be achieved by adjusting the widths of the individual averaging intervals according to the channel variation. The point is to trade-off computational complexity for finding the ideal value of averaging intervals against the one incurred by having to calculate channel capacity more often.

V. Influence of Averaging Duration

Until now the channel was assumed to be constant during a subframe (block fading) so temporal changes of the capacity according to (2) did not need to be considered. Dual to splitting the total number of subcarriers into $D \leq K$ intervals and calculating the rate on every such interval it is also possible to split a subframe, which consist of 14 OFDM symbols, into several intervals. This section shows that such a strategy can lead to a better choice of the precoder in terms of BER, but only if the UE speed and therefore the Doppler shift, is already high.

To simulate temporal channel variations, a correlated fast fading channel model is applied, where channel realizations are correlated according to the improved version of Rosa Zheng's model [11]. For the following simulation results a similar approach for calculation of the precoding matrix indicator as described in Section II is used, but the time and frequency domains are swapt. One subframe ($\hat{K} = 14$) is divided into $\hat{D}$ intervals of width $\hat{\mathcal{D}}_d$ in OFDM symbols and the channel average is calculated over these intervals according to (4). The channel is also averaged with respect to the subcarriers. The simulation parameters are summarized in Table IV.

TABLE IV
SIMULATION PARAMETERS

Carrier frequency f_c:	2.1 GHz
System bandwidth:	1.4 MHz
Number of subcarriers K:	72
Subframe duration:	1 ms
Number of OFDM symbols per subframe:	14
Feedback delay:	0 TTI
Channel model:	ITU-T VehA
Antenna configuration:	2×1
Receiver:	Zero Forcing
CQI:	4
E_s/N_0:	6 dB
UE speed v:	variable: 0…260 km/h
# intervals $\hat{D}$ per subframe:	1, 2, 14

Figure 9 presents uncoded BERs obtained for the described setup. Until a speed of about 50 km/h the BER is not influenced by temporal channel variations. Then the BER starts to degrade, as the chosen precoder does not match the channel for the whole subframe duration anymore. At 100 km/h the performance of the feedback method using just one temporal averaging interval starts to degrade in comparison to the methods using two or more intervals. The performance difference between two or fourteen intervals is even at 260 km/h negligible. The coherence time given in the figure is calculated according to

$$T_c = \frac{c_0}{2 f_c \cdot v}, \tag{10}$$

with c_0 being the speed of light.

VI. Influence of Feedback Delay

The simulation results of this section are produced with the *mean mutual information based* method using the algorithm described in Section IV to dynamically adjust the number of averaging intervals. The frequency domain granularity of PMI feedback is one value for the whole system bandwidth. A subframe is divided into two time slots, according to the LTE specification, and for every time slot a distinct PMI value is calculated. The section is intended to investigate the influence of a feedback delay in multiples of TTIs. As the maximum number of HARQ processes of LTE is 8, this is also the highest possible feedback delay value in TTIs. Simulations are carried out for two different channel types and antenna configurations (see Table V).

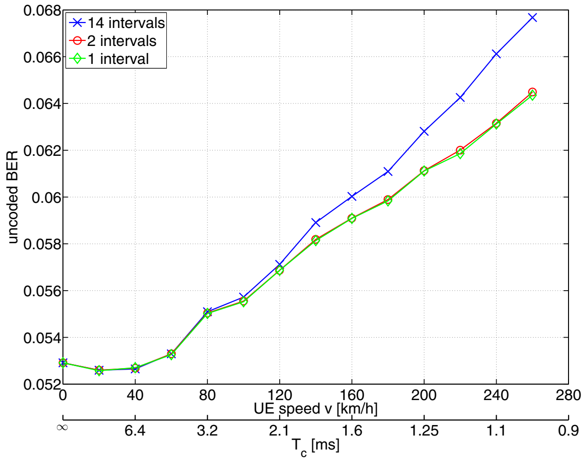

Fig. 9. Uncoded BER over UE speed respectively coherence time for different number of temporal averaging intervals.

TABLE V
SIMULATION PARAMETERS

Carrier frequency f_c:	2.1 GHz
System bandwidth:	1.4 MHz
Number of subcarriers K:	72
Subframe duration:	1 ms
# OFDM symbols per subframe:	14
Feedback delay:	variable
Channel model:	VehA @ 50 km/h, $T_c = 5.1$ ms PedA @ 8 km/h, $T_c = 32.1$ ms
Antenna configuration:	2×1 4×2
Receiver:	SSD
CQI:	4
E_s/N_0:	6 dB

Figure 10 presents simulation results obtained for variable feedback delay in terms of uncoded BER. Clearly, the BER increases as expected with feedback delay. The degradation is larger for higher UE speed respectively shorter channel coherence time. Also it can be seen that there is a saturation effect in the BER degradation. Comparing the results with the ones obtained with feedback delay zero (c.f. figures 2, 4) shows that even at a feedback delay of 8 TTI the BER has not degraded to the value obtained with no feedback. The amount of BER degradation seems to be correlated with the UE speed. This is further investigated in Figure 11. Here

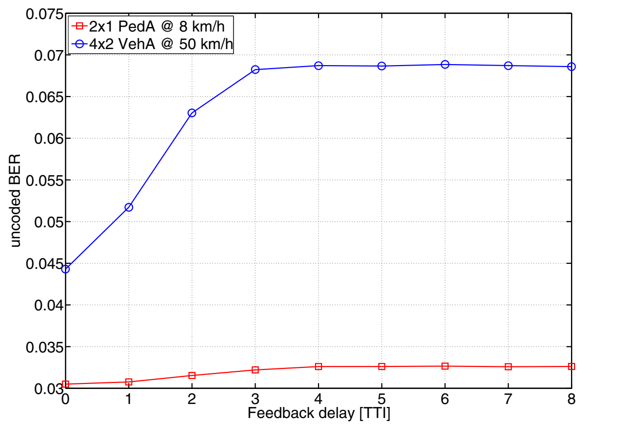

Fig. 10. Uncoded BER over feedback delay for different channel types and antenna configurations.

the 2×1 PedA channel is simulated with varying UE speed and a constant feedback delay of 8 TTI. The amount of BER

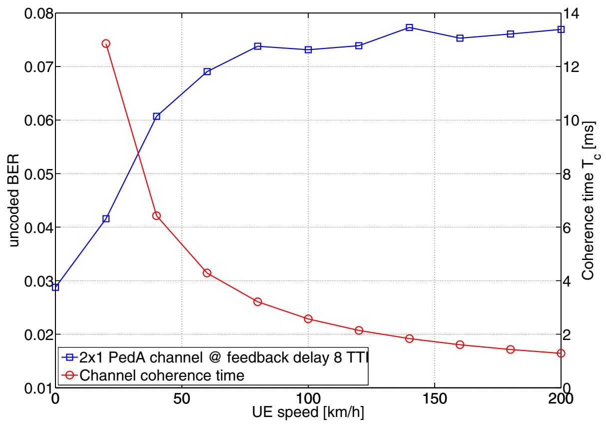

Fig. 11. Uncoded BER over UE speed at constant feedback delay of 8 TTI and channel coherence time, according to equation (10) over UE speed.

degradation clearly depends on the UE speed. For speed zero the coherence time is infinite and the channel does not change. Therefore the result is the same as obtained with feedback delay 0 (the result is averaged over many independent channel realizations). The saturation of BER degradation with speed is related to a corresponding saturation in the coherence time, as also shown in the figure.

VII. Conclusions

3GPP UMTS/LTE, in the closed loop spatial multiplexing mode, requires the UE to feedback a precoding matrix indicator in order to help the eNodeB in making decision about which precoder to apply. In this paper we present a mutual information based method to evaluate this value. Our method chooses the optimal precoder by maximizing the sum rate over the set of subcarriers of interest. Simulation results confirm that this method achieves optimal results if the channel variation over the target bandwidth is not too strong. Next, a lower complexity method is derived, which divides the total number of subcarriers into several intervals, computes a channel reproducer coefficient (single sample or arithmetic mean) and corresponding mutual information for each interval and calculates the optimal precoder from the sum of these values. It is shown that this method can drastically reduce the computational complexity without influencing the BER performance by reducing the number of costly capacity computations. This succeeds until the number of averaging intervals reaches a certain lower bound. A simple algorithm is presented to estimate this threshold value. Averaging the channel not over the entire subframe but dividing the subframe into several slots is further shown to achieve a BER gain for high UE speeds $v \geq 100$ km/h. There is no gain in using more than two time slots for channel averaging until very high UE speeds $v \leq 260$ km/h, but this does not mean that there is no gain if independent PMIs are fed back for these time slots (as done in LTE). Lastly, the influence of a feedback delay is investigated, showing a BER degradation with increasing delay and UE speed.

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References

[1] 3GPP, "Technical Specification Group Radio Access Network; (E-UTRA) and (E-UTRAN); overall description; stage 2," Sep. 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)," Sep. 2009. [Online]. Available: http://www.3gpp.org/ftp/Specs/html-info/36211.htm
[4] D. J. Love, R. W. Heath, "Limited Feedback Precoding for Spatial Multiplexing Systems," IEEE Global Telecommunications Conference, 2003.
[5] C. Mehlführer, M. Wrulich, J. Colom Ikuno, D. Bosanska, and M. Rupp, "Simulating the Long Term Evolution Physical Layer," in Proc. of the 17th European Signal Processing Conference (EUSIPCO 2009), Glasgow, Scotland, Aug. 2009, [Online].
Available: http://publik.tuwien.ac.at/files/PubDat_175708.pdf
[6] [Online]. Available: http://www.nt.tuwien.ac.at/ltesimulator/
[7] I. E. Telatar, "Capacity of multi-antenna Gaussian channels," Europ. Trans. Telecommun., Vol. 10, No. 6, 1999.
[8] C. Studer, M. Wenk, A. Burg, H. Bölcskei, "Soft-Output Sphere Decoding: Performance and Implementation Aspects," Fortieth Asilomar Conference on Signals, Systems and Computers, 2006. ACSSC '06.
[9] ITU, "Recommendation ITU-R M.1225: Guidelines for Evaluation of Radio Transmission Technologies for IMT-2000," Technical Report, 1997.
[10] D. Tse, P. Viswanath, "Fundamentals of Wireless Communications," Cambridge University Press, 2008.
[11] Y. R. Zheng, C. Xiao, "Simulation Models with Correct Statistical Properties for Rayleigh Fading Channels," IEEE Transactions on Communications, vol. 51, no. 6, pp. 920-928, June 2003.