Performance Limitation of HSDPA MIMO by Pre-Coding Induced Phase Distortion

Wolfgang Aichmann
Nokia Siemens Networks
D-89081 Ulm, Germany
Email: wolfgang.aichmann@nsn.com

***Abstract*—In multiple-input multiple-output (MIMO) antenna systems both strength of the data layers as well as separation of them is achieved by pre-coding the transmitted signal with different factors for different layers and different antennas. In W-CDMA systems, the selection of these factors is decided anew every transmission time interval, which causes a phase distortion with every change. Due to raised cosine filtering this phase distortion is broadened even if the physical channel is flat in frequency, which distorts the affected symbols and thus limits the maximum bit rate to be transmitted in such cases if no counter measures are taken. If the pre-coding weights of adjacent time intervals are known by the receiver, the distortion can be removed by the equalizer; otherwise high code rates should be avoided by the scheduler.**

***Keywords*-W-CDMA; HSDPA; MIMO; pre-coding; raised cosine filter;**

I. Introduction

Multiple-input multiple-output is a well known method to increase the capacity of radio systems on the air interface [1]–[5] and was introduced in 3G high speed downlink packet access (HSDPA) with Release 7 of 3GPP TS 25.214 [6] in 2007 and in high speed uplink packet access (HSUPA) with Release 11 [7] in 2012. Two modes of operation are defined, namely single and dual layer transmission. The distribution of data to the transmitting antennas is controlled by so called pre-coding vectors. The effect of this pre-coding is twofold: In case of single data layer transmission, the pre-coding vector is adjusted such as to optimize the power at the receiving antennas. This is possible as shifting the phase difference between the signals fed to the transmitting antennas changes the interference behaviour at the receiving antennas, and constructive interference results in higher power.

If two layers are transmitted, orthogonal pre-coding vectors are applied, which allow to separate the layers at the receiving side, typically by a linear minimum squared equalizer in space and time as it is, e.g., described in [8]. Non-linear receiver algorithms such as interference cancellation improve the de-coding performance on receiver side but are quite sensitive to very good channel knowledge [9], [10].

As the interference conditions are changing with time (fast fading), especially for moving user equipments (UE), one tries to always find and use the best pre-coding vector out of the set of allowed ones, and to change it if necessary. Switching between single and dual stream as well as between MIMO and non-MIMO can be regarded as a generalization of switching the pre-coding vector: In non-MIMO mode, the weights for the second transmitting antenna are 0 and single layer transmission can formally be regarded as dual layer transmission with the same data and the same pre-coding for both layers.

Any change in the pre-coding vector however causes a superposition of signals coded with the old and ones coded with the new pre-coding vector in the range of channel delay due to multi-path propagation including raised cosine filtering. This has some impact on the detection of the affected bits and might even prohibit at all the correct block detection for high coding rates. In real networks switching of MIMO mode as well as switching the pre-coding vector within a MIMO mode is a quite frequent operation, which leads to a significant restriction of the achievable maximum data rate even in very good radio conditions. This could be one of the reasons why HSDPA MIMO isn't present in commercial networks so far besides other reasons such as the chicken or egg (here better base station or UE) dilemma and the need for a second pilot signal reducing thus the power available for data transmission.

The remainder of the paper is organized as follows: In Section II, we investigate the impact of pre-coding induced phase distortions on the received symbols. We show then in Section III that a smooth transition of equalizer weights at the boundaries of transmission time intervals (TTI) pre-coded with different vectors is able to remove this distortion. Unfortunately, the pre-coding vector of adjacent TTIs is known on receiving side only if they bear data for the same UE. Therefore, it is useful to take counter measures on transmitter side if TTIs are dedicated to different UEs. Some of them are introduced in Section IV followed by some concluding remarks in Section V.

II. Basic Considerations

In order to understand the principles of MIMO pre-coding, the basic concept of MIMO transmission in HSDPA is displayed in Fig. 1.

The signal s_λ of each data layer $\lambda \in \{1,2\}$ is fed to the $m \in \{1,2\}$ transmitting antennas with weights $w_1 = 1/\sqrt(2)$ and $w_2 \in W_2 \equiv \{(1+i)/2, (1-i)/2, (-1+i)/2, (-1-i)/2\}$ [6]; w_1 and w_2 build the pre-coding vector

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and the pre-coding index (PCI) defines which component w_2 out of the code book W_2 is taken.

If only one data layer is transmitted, w_2 typically is chosen such as to achieve the maximum channel power at receiving side as this is related to the highest possible throughput if we ignore here that also the post-equalizer interference power may depend on the applied pre-coding vector. In case of two layers being transmitted the two pre-coding vectors have to be orthogonal in order to enable the receiver to separate the layers and hence $w_2|_{\lambda=2} = -\, w_2|_{\lambda=1}$. As exchanging the PCI values within two layers transmitted in parallel has no impact on the aggregated throughput this means that only two choices remain for w_2, namely either PCI 1 and 4 or PCI 2 and 3, and the aggregated possible throughput is used to decide which of these two possibilities is taken.

The signal r_j received at antenna j and time k counted in chip length ($0 \le k < n$) consists of contributions of all transmitted data layers and can be expressed as

$$\begin{aligned} r_j(k) &= \underbrace{\sum_{\delta=0}^{n-1} \left[w_1 h_{1j}(\delta) + w_2 h_{2j}(\delta)\right] s_1(k-\delta)}_{\text{layer } \lambda=1} \\ &+ \underbrace{\sum_{\delta=0}^{n-1} \left[w_1 h_{1j}(\delta) - w_2 h_{2j}(\delta)\right] s_2(k-\delta)}_{\text{layer } \lambda=2} \qquad (1) \end{aligned}$$

To achieve high data rates it is necessary the coherence time of the channel impulse response to be large compared to the delay spread in the presence of multi-path propagation, and large compared to the averaging length required for appropriate channel estimation. Practical experience shows that code rates of 80% or higher require a coherence time of at least one slot, which is achieved for UEs moving not faster than about $3km/h$.

However, this constraint is violated if the pre-coding vector is changed: The effective channel impulse response becomes unsteady at the TTI boundary and hence the coherence time approaches zero, at least in the time range of the delay spread. It has to be emphasized that the raised cosine (RC) filter has to be included explicitly in this considerations as on transmitting side, the pre-coding usually is executed before the root raised cosine (RRC) filtering and on receiving side the equalizer algorithm containing the inverse pre-coding is fed with data, which have already passed the filter.

To demonstrate the impact of RC filtering we have investigated the positions of symbols in the phase space diagram, modulated with 64QAM and passing a one-tap channel with additional RC filter, i.e. a RRC filter was applied on both transmitting and receiving side, but without equalizing. Four different scenarios have been taken into account, namely keeping the pre-coding vector constant at PCI=1, as well as switching it in the middle of a data block with 2560 chips to another pre-coding vector out of W_2. The distance of the

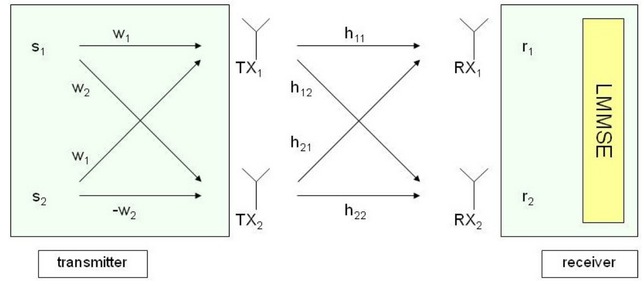

Figure 1. MIMO dual layer transmission, schematic overview.

positions of the symbols in the phase space diagram from the position they would have if the complete data block had been sent with the new pre-coding vector is then measured. The result is shown in Fig. 2 displaying the symbol shifts for an RRC filter with oversampling factor 4 and tail length[1] of 20 chips.

Fig. 2a shows the constellation of the symbols in the phase space. The circles mark the symbol positions for constant pre-coding whereas crosses are placed at the effective symbol positions. Symbols beyond the RRC coherence length are drawn in blue and those within the coherence length around the switching time are red colored. Fig. 2b displays the distance of shift of the symbols with respect to the position without shift, normalized to the average symbol distance from the origin. Please note the reduced scale on ordinate for the case of no switching. It is obvious that symbols within the tail length are affected heavily by switching the pre-coding vector, whereas symbols beyond that range relax quickly to the equilibrium position.

Additional equalizing turns and stretches the symbols such as to have them at the same positions as on the transmitting side, but it does *not* remove the distortion induced by switching the pre-coding vector, i.e., the relative positions of the red crosses to the blue circles.

Two major effects are worth to be noted, namely that the symbol shift increases with increasing length of the RRC filter and that especially for larger tail lengths a phase shift of π is worse than a shift of $\pi/2$. Whether the latter effect also results in higher bit error rates is not clear so far as a longer distance from the ideal position has no more influence on the bit error rate when the bit is detected wrongly anyway.

Furthermore it has to be noted that on transmitting side the pre-coding filter usually is applied before the RRC filter, which causes the disturbance by switching the filter to be twice as long as the length of the RRC filter. If the order of filtering and pre-coding is changed, the length of disturbance is halved.

Link level simulations with a Pedestrian A channel at $3km/h$ for MIMO single stream (see Fig. 3a) show how

[1] Tail length denotes the length of RRC filter, i.e., the length from middle position of the filter to the edge.

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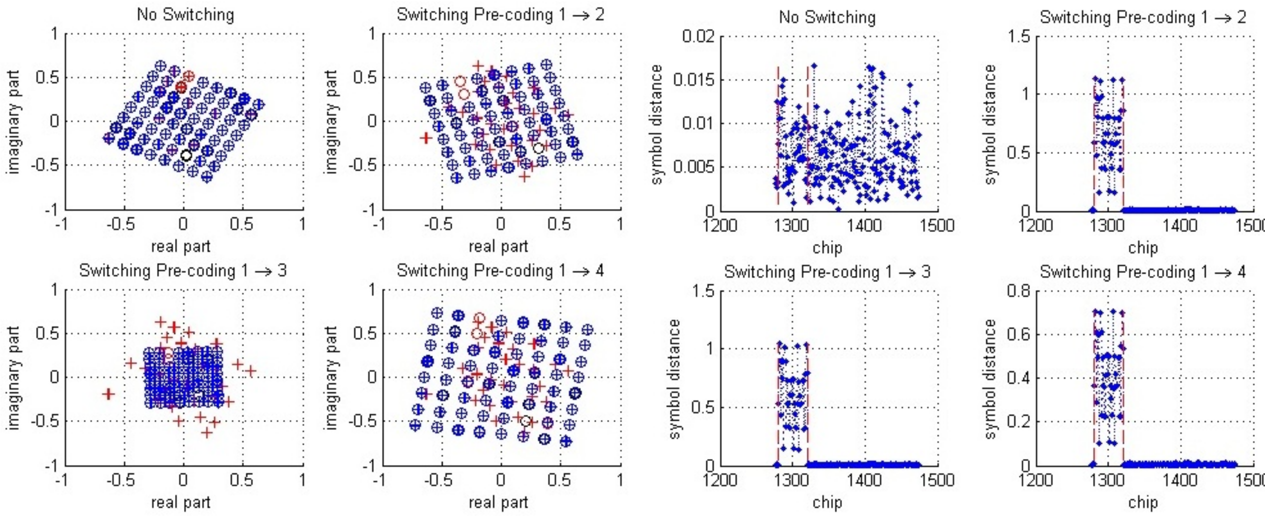
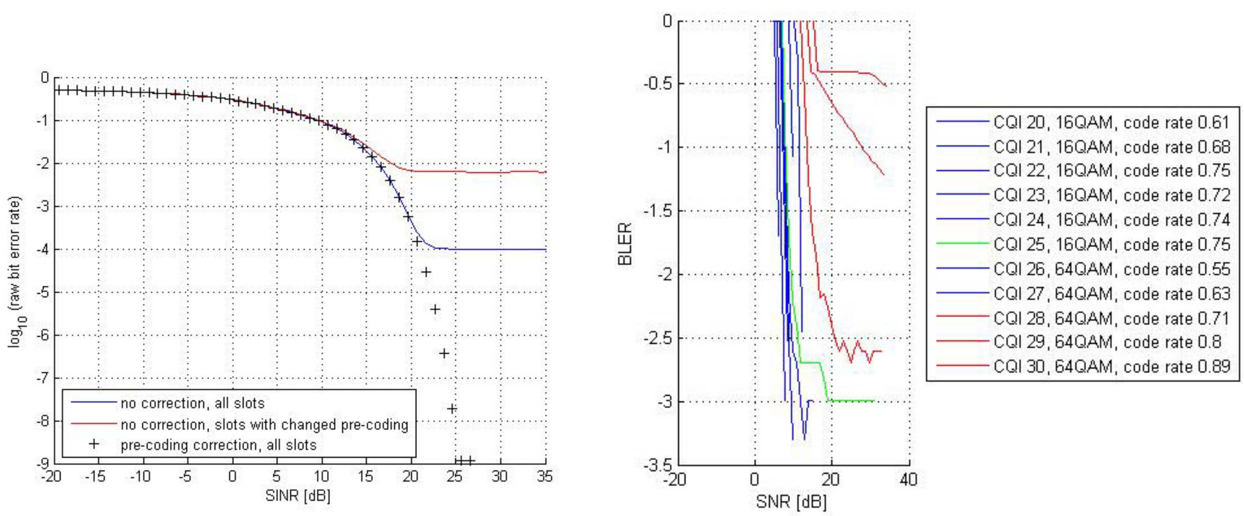


(a) (b)

Figure 2. Shifting of 64QAM symbol position due to switching pre-coding vector for a one tap channel and RRC filter with tail length 20. In (a), the expected (o) and measured (+) symbol positions in the complex phase space are shown. In a range around the pre-coding change (red), determined by the maximum channel delay including RC filter, the symbols are distorted. The absolute value of the dislocations as a function of chip index is shown in (b).



(a) (b)

Figure 3. Link level simulations for a Pedestrian A channel at 3km/h, UE category20 and CQI values according table G in [7] (a) Raw bit error rate for CQI 30. Slots with changed pre-coding vector show a significantly higher error floor compared to all slots. Correct handling of these pre-coding changes (see III-B) removes the error floor completely. (b) High code rates show an error floor in the block error rate (BLER) which can even be above the working point of 10% BLER.

switching the pre-coding vector causes an error floor of the raw bit error rate of about 0.6% in affected slots. These bit errors directly cause erroneous decoding of the blocks they belong to in case of coding rates higher than about 0.75 as is demonstrated in Fig. 3b.

III. Equalizer with Smooth Weight Transition

A. Calculation of Symbol Distortion

To see how the distortion induced by switching the pre-coding vector in the presence of channel impulse response h including the RC filtering can be properly taken into account on receiver side, let us start with (1) but restricted to one data layer only, namely:

$$r_j(k) = \sum_{\delta=0}^{n-1} \left[w_1(k-\delta) h_{1j}(\delta) + w_2(k-\delta) h_{2j}(\delta) \right] s_1(k-\delta) \tag{2}$$

We have thus generalized the pre-coding vector to depend on time via $k-\delta$. Let us now introduce a single switching of the pre-coding vector at time 0, i.e., $s_1(k<0)$ is pre-coded with $\tilde{w}$ and $s_1(k \geq 0)$ with w:

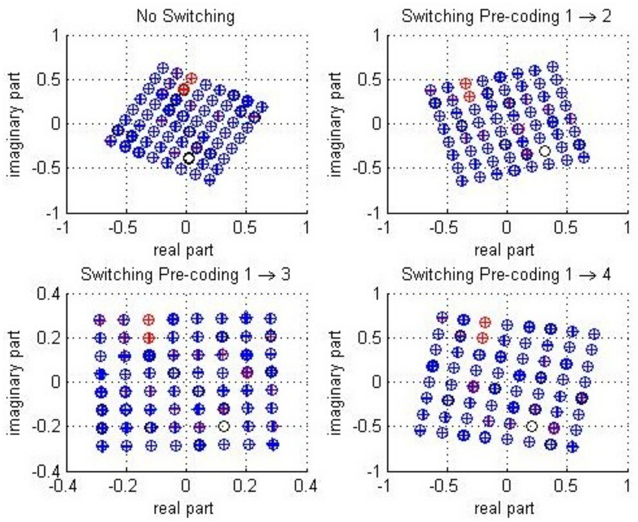


Figure 4. Corrected positions of 64QAM symbols, which are distorted by switching the pre-coding vector in presence of a one tap channel and RRC filter with filter length 20 and OSF 4.

$$\begin{aligned} r_j\left(k \geq 0\right) &= \sum_{\delta=0}^{k}\left[w_1 h_{1j}\left(\delta\right) + w_2 h_{2j}\left(\delta\right)\right] s_1\left(k-\delta\right) \\ &+ \sum_{\delta=k+1}^{n-1}\left[\tilde{w}_1 h_{1j}\left(\delta\right) + \tilde{w}_2 h_{2j}\left(\delta\right)\right] s_1\left(k-\delta\right) \\ &= \bar{r}_j + \Delta r_j \qquad (3) \\ \bar{r}_j(k) &\equiv \sum_{\delta=0}^{n-1}\left[w_1 h_{1j}\left(\delta\right) + w_2 h_{2j}\left(\delta\right)\right] s_1\left(k-\delta\right) \quad (4) \\ \Delta r_j(k) &\equiv \sum_{\delta=k+1}^{n-1}\left[\tilde{w}_2 - w_2\right] h_{2j}\left(\delta\right) s_1\left(k-\delta\right) \qquad (5) \end{aligned}$$

The last step in (3) holds because the same phase factor is applied to the first transmitting antenna for all pre-coding vectors, i.e., $w_1 \equiv \tilde{w}_1$. One can see that received symbols up to index $k = n-2$ are affected from the switch in the pre-coding vector. Hence we can limit the further investigations to $0 \geq k \geq n-2$. It remains $\Delta r_j(k)$ as error correction to the 'static' calculation $\bar{r}_j(k)$ of $r_j(k)$.

One can hence use (4) to calculate $\bar{r}_j$ assuming ideal knowledge of the transmitted symbols $s_1(k < 0)$ also on receiver side, i.e., symbols of the data block preceding the switch of the pre-coding vector, as is shown in Fig. 4.

The assumption of ideal symbol knowledge demonstrates the impact of switching the pre-coding index, but it doesn't help to remove the distortion in real scenarios as of course the symbols aren't known a priori except of pilot measurements, which is beyond the current scope. We must therefore find another solution of removing the symbol distortion.

B. Smooth Weight Transition

In a linear minimum mean square equalizer (LMMSE) the weights are typically re-calculated every slot or every TTI but always with the pre-coding weights as they are relevant for the current data block. If the pre-coding vector is changed on transmitting side also the equalizer weight vector obtained with the old pre-coding weights is replaced as a whole by the new one now calculated also with the new pre-coding weights. This implies that there is a transition interval, in which either data originating from the 'old' data block are equalized with weights obtained for 'new' pre-coding or data already originating from the 'new' data block are still equalized with weights obtained for 'old' pre-coding or a mixture of both depending on when exactly the replacement of the weight vector takes place.

Heuristically, this error can be avoided if the weight vector isn't replaced as a whole but element by element. This means that as soon as the new data block enters the equalization procedure the corresponding element of the 'old' equalizer vector (either the first or the last one, depending on the ordering convention) is replaced by the element of the 'new' weight vector. Then, the elements are exchanged one by one with every chip until the weight vector consists completely of elements calculated with the 'new' pre-coding.

In order to motivate the validity of this approach we remember the calculation of space time LMMSE equalizer weights as, e.g., elaborated in [8]. The weights are proportional to a generalized matrix of channel impulse response, Γ_E, but with Γ_E appearing in a quadratic form in the denominator[2]. If all data layers are transmitted with the same power as is the case in HSDPA MIMO, this quadratic form becomes independent of the applied pre-coding. The latter persists only in the numerator and makes the elements of the weight vector proportional to the respective pre-coding weights, leading directly to the heuristic approach described above.

Applying this procedure to full link level simulations a significant gain can be obtained at the high data rate edge as can be seen in Fig. 5. Although the working point is only slightly shifted to lower SNR, the error floor present without any correction is avoided completely. Compared to the simulation with filtering the pre-coding switching with the goal to reduce the number of transitions, see Fig. 6, the averaged working point is nearly 1dB better in the scenario investigated here.

The formalism above can easily be extended to switching between MIMO single and dual stream: Single stream transmission is formally the same as dual stream transmission with identical pre-coding vectors (and of course identical data). To obtain single layer transition weights from dual layer ones the latter have have to be summed up whereas

[2]We use here the term denominator for an inverted matrix in order to simplify the verbal description.

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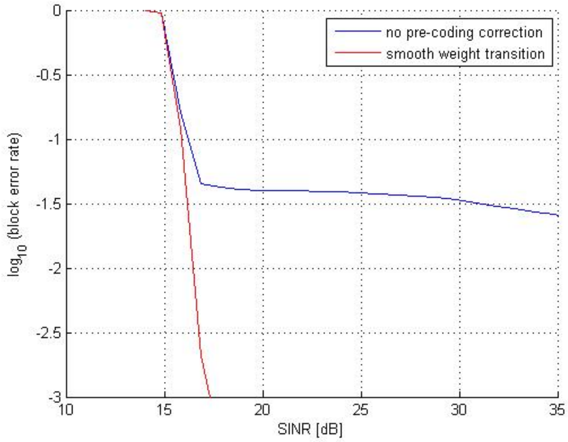

Figure 5. Performance improvement of MIMO Single Stream by smooth transition of equalizer weights. The simulation was executed for a Pedestrian A channel at 3km/h, UE category 20 and CQI 30 according table G in [7]

dual layer transition weights can be obtained from single layer ones be dividing them into two equal parts, of course always with proper scaling such as to conserve the overall transmit power.

As also non-MIMO data transfer is affected if switching between this mode and any of the MIMO modes occurs it has also to be included by defining it to be a single stream transmission with constant pre-coding vectors and vanishing pre-coding weight for the second transmitting antenna.

IV. Other Means of Avoiding Systematic Block Errors

There are many methods, which can help to avoid the occurrence of systematic block errors induced by changing the pre-coding weights in case the receiver is not able to handle the corresponding symbol distortion properly. Some of them, e.g., the introduction of a cyclic prefix as in long term evolution (LTE), would require to change the specification which would make it impossible for legacy UEs to be operated. They are beyond the scope of the current paper. Others can be implemented either on base station or UE side within the valid specification framework. Such proposals are discussed now.

A. Attenuation of Switching Pre-Coding Vector

One quite simple strategy to decrease the negative impact of switching the pre-coding vector is to attenuate the switching itself. We have investigated a simple filter algorithm on receiver side for single layer transmission, which is working in two steps. This approach is easily extensible to the dual layer case. It is based on an ordinary calculation of channel power for all possible pre-coding vectors, i.e.,

$$\begin{aligned} P_s^{(k)} &= \sum_{n=1}^{N_{Rx}} \left[H_{kn} * H_{kn}^{\dagger} \right] \\ H_{kn} &= \sum_{m=1}^{N_{Tx}} \left[w_m^{(k)} \sum_{\delta=1}^{L} h_{mn}(\delta) \right] \end{aligned}$$

where $w^{(k)}$ denotes the k^{th} element out of the code book W.

Without filtering, the pre-coding vector is chosen as

$$k_{0,s} = \max_{\{k\}} \left(P_s^k \right). \tag{6}$$

In the first step of the proposed algorithm, we compare the current optimum channel power with the previous one and introduce a threshold $\zeta > 0$ for signaling a changed pre-coding vector:

$$\tilde{k}_{0,s}(s) = \begin{cases} k_{0,s} & P_s^{k_{0,s}} > (1+\zeta) P_s^{k_{0,s-1}} \\ k_{0,s-1} & \text{otherwise} \end{cases} \tag{7}$$

In the second step, we then introduce an attenuation filter, which remembers the last F pre-coding indices, $k_{0,s}, s \in \{s_0, s_0-1, \ldots, s_0-F+1\}$ and resets the pre-coding index to the previously used one as long as not at least $F_\zeta, 1 < \zeta < F$ times the new index occurs.

Fig. 6 shows simulation results for different modulation and coding schemes as defined in the single layer channel quality indicator (CQI) table 7G [6] for an additive white Gaussian noise (AWGN) channel with filtering where we have used $\zeta = 0.1$, $F = 10$ and $F_\zeta = 7$. It can be seen clearly that the error floor occurring in absence of the threshold filter, see Fig. 3, vanishes with the proposed filter design.

For single stream without filtering the changing of the pre-coding vector, a reasonable block error rate can be achieved for coding rates not too high. In case of 64QAM, an error floor starts to occur for coding rates between 0.7 and 0.8 and for 16QAM in the range of 0.7. This implies to apply the suggested filter only in case of used modulation and coding scheme providing coding rates higher than the mentioned thresholds.

B. Limiting 'CQI' to Specific Code Rate if Pre-Coding Is Changed

As demonstrated in section IV-A, the proper data block detection is destroyed by changing the pre-coding vector only for high coding rates. It seems therefore reasonable to reduce the coding rate to a value below the corresponding threshold via corresponding CQI selection if the pre-coding vector is changed. Although this mechanism can be implemented completely on the transmitting side the receiver should know about such an algorithm in order to avoid

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too many changes anyway. A corresponding signaling might therefore to be useful.

This CQI limitation should be applied only in case of the receiver is not able to correct the switching errors itself (see section III-B). A powerful application of this proposal therefore requires the definition of a new UE class or a parameter indicating the ability of the UE for this correction.

V. Conclusion

We have shown that switching the pre-coding vector applied to data transferred from multiple TX antennas leads to a broad distortion of the channel impulse response including the raised cosine filter. In this context, also switching between non-MIMO, MIMO single layer and MIMO dual layer transmission can formally be regarded as switching the pre-coding vector. For high data rates this effect leads to a high probability that data blocks adjacent to the switching border will be lost.

There are several possibilities to avoid or at least to reduce the negative impact, but most of them cause other constraints, e.g. avoiding changes of the pre-coding vector and / or reducing the coding rate. Other measures, e.g., dynamically shortening the disturbance length by reducing the length of root raised cosine filter if the pre-coding vector is changed might be inapplicable at all.

Therefore, the best strategy seems to remove the disturbance induced by the discontinuity in phase by the equalizer. However, if the receiver isn't able to do this, e.g. because of missing pre-coding information, which is signaled encrypted, a reduction of the coding rate to a reasonable value of affected data blocks is an acceptable strategy to avoid systematic block errors.

The calculations were done for HSDPA MIMO, but they are applicable to a wider range and affect many scenarios with MIMO and / or transmit diversity. Although the smooth transition of equalizer weights assumes phase-only precoding an extension of this approach might be used if precoding with modifications in the amplitude are applied.

Finally we want to mention that the impact of the distortion to LTE systems is low or absent at all because of the cyclic prefix, which is introduced there for other reasons.

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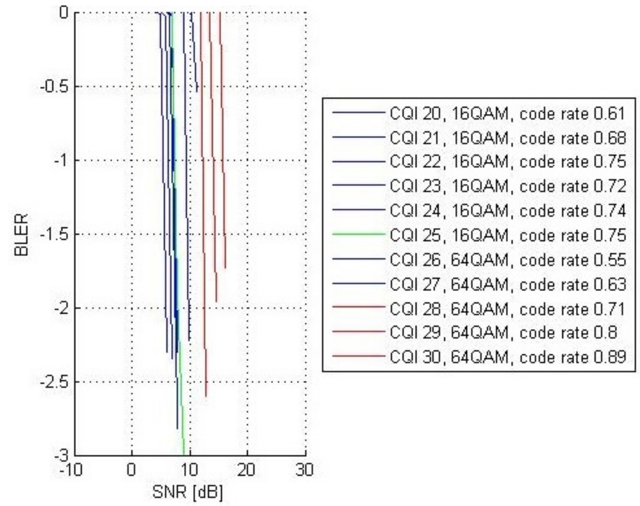

Figure 6. Performance improvement of MIMO Single Stream in AWGN channels by attenuating the changing of the pre-coding vector. The error floor as shown in Fig. 3 can be avoided.

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