THROUGHPUT ANALYSIS OF PER-ANTENNA RATE CONTROL WITH LINEAR CHIP-LEVEL EQUALIZATION

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ABSTRACT

In this paper the achievable performance gains of Per-Antenna Rate Control (PARC) for Code Division Multiple Access (CDMA) transmission with different linear chip-level equalization algorithms are investigated. High Speed Downlink Packet Access (HSDPA) as a solution of a full packet-based transmission standard utilizes extensive multicode operation especially in low-mobility scenarios. Due to possible large delay spreads resulting from multipath propagation, the orthogonality of the spreading codes is destroyed, which leads to severe performance degradations. To cope with such properties of mobile radio channels, chip-level equalization was suggested to be an advanced receiver technique in future user equipments. Simulations for PARC with chip-level equalization, which include adaptive modulation and coding as well as feedback errors and delays, show that throughput can be significantly increased at high signal-to-interference-and-noise ratios compared to common Rake reception if multiple codes are assigned to a single user.

I. INTRODUCTION

Multiple-Input Multiple-Output (MIMO) systems as a candidate for the High Speed Downlink Packet Access (HSDPA) enhancement concerning the third generation Universal Mobile Telecommunication System (UMTS) downlink with frequency division duplex (FDD) promise a large increase in spectral efficiency, thus leading to a higher throughput [1]. However, applying multiple antennas at transmitter and receiver side and additionally allowing extensive multicode transmission for single users prevents low-complexity receivers, like conventional Rake, from fully exploiting their additional diversity gain due to self- and multiple-access interference (MAI) caused by multipath propagation. To improve the performance chip-level equalization is considered as a practical approach for such CDMA systems [2]. There, the corresponding filtering procedures are processed before despreading of the received data. A detailed analysis concerning the robustness and computational requirements of such different linear filters was published in [3]. As most of these equalizers strongly depend on the degree of prior information about the channel, especially if adaptive filters are employed, the resulting throughput performance of such equalizers provides information about the co-existence with already defined channel estimation algorithms. Therefore, we present some throughput results for Per-Antenna Rate Control (PARC) regarding multicode transmission modes in fast fading frequency-selective environments with linear chip-level equalization techniques.

This paper is organized as follows. In Section II, we describe the PARC system model, which originates from [1]. Section III deals with the investigated chip-level equalization techniques. The simulation results and additional comments on the implementation can be found in Section IV. Finally, the paper is concluded in Section V.

II. SYSTEM DESCRIPTION

We consider a multicode MIMO-CDMA system with N_T transmit and N_R receive antennas. Due to the similarity of PARC to the well-known V-BLAST algorithm [4] the proposition $N_R \geq N_T$ must hold. At first, the high-speed data stream is demultiplexed into N_T groups, each consisting of K substreams, where K is the number of simultaneously applied spreading codes used for transmission. The spreading factor G of each spreading code is 16 according to the HSDPA standard [5] and the maximum number of usable spreading codes is related to $G-1$ due to the occupied branch in the channelization code tree used by control channels and to ensure backward compatibility with the Rel. '99 standard channels. The number of bits belonging to each group depends on their selected modulation and turbo coding scheme (MCS) based on the estimated signal-to-interference-and-noise ratio (SINR) after filtering at the receiver. Thus, the rate of all groups can be adjusted independently. Consequently, the bits of the corresponding $K \cdot N_T$ substreams plus Cyclic Redundancy Check (CRC)

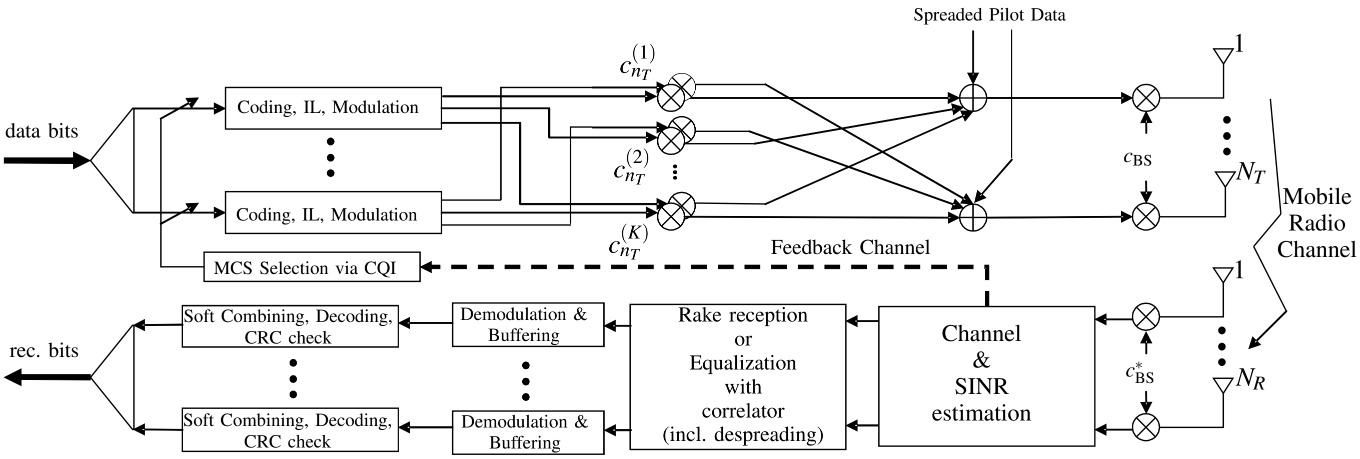

Fig. 1. Block diagram of the PARC system using multicode transmission

bits are independently coded, possibly with different code rates, followed by the interleaving (IL) and modulation processes [6]. The K substreams of each transmit antenna are segmented to the so-called high-speed physical data channels (HS-PDSCH), combined and subsequently scrambled using a common long complex scrambling code. The n_T-th group chip-level signal plus antenna specific pilot symbols $d_{n_T}^{(P)}(g)$ are transmitted by the n_T-th transmit antenna. This signal can be described by

$$s_{n_T}(i)=c_{\mathrm{BS}}(i)\left(\sum_{k=1}^{K} d_{n_T}^{(k)}(g)\,c_{n_T}^{(k)}(i)+d_{n_T}^{(P)}(g)\,c_{n_T}^{(P)}(i)\right), \quad (1)$$

where $c_{n_T}^{(k)}(i)$ and $c_{n_T}^{(P)}(i)$ are the spreading codes for the k-th substream and the n_T-th pilot sequence at the i-th chip, respectively. $c_{\mathrm{BS}}(i)$ corresponds to the cell-specific scrambling sequence and g accordingly denotes the symbol index. $d_{n_T}^{(k)}(g)$ describes the information symbols on the high-speed physical data channels.

The spatial channel is derived from a classical tap delay line model, taking L independent time-varying taps with a classical Jakes Doppler spectrum for each spatial subchannel. Therefore, we assume uncorrelated scattering and wide-sense stationarity (WSSUS model) [7], [8]. Using above notation, the received chip-level signal at the n_R-th receive antenna is given by

$$r_{n_R}(i)=\sum_{n_T=1}^{N_T}\sum_{\ell=0}^{L-1} h_{n_R,n_T}^{(\ell)}(i)\,s_{n_T}(i-\ell)+n(i)\ . \quad (2)$$

$h_{n_R,n_T}^{(\ell)}(i)$ denotes the channel coefficient between transmit antenna n_T and receive antenna n_R at received path ℓ with L being the overall number of received paths. The unit of ℓ is chip length to describe channel-specific delay spreads. $n(i)$ denotes a zero-mean additive white Gaussian noise chip including the received interference from other cells and thermal noise. Fig. 1 shows the block diagram of the PARC system. The complete system description can then be formulated in matrix-vector notation

$$\mathbf{r}=\mathbf{H}\mathbf{s}+\mathbf{n}\,, \quad (3)$$

if we consider the instantaneous channel coefficients in the vector $\mathbf{h}_{n_R,n_T}=\left[h_{n_R,n_T}^{(0)},h_{n_R,n_T}^{(1)},\ldots,h_{n_R,n_T}^{(L-1)}\right]$. The MIMO channel matrix $\mathbf{H}\in\mathbb{C}^{(N_R\cdot N)\times(N_T\cdot(N+L-1))}$, where $L-1$ is the maximum delay spread in chip length, is a block Toeplitz matrix

$$\mathbf{H}=\begin{bmatrix} \mathbf{h}_{1,1} & & \mathbf{0} & & \mathbf{h}_{1,N_T} & & \mathbf{0} \\ & \ddots & & \cdots & & \ddots & \\ \mathbf{0} & & \mathbf{h}_{1,1} & & \mathbf{0} & & \mathbf{h}_{1,N_T} \\ \vdots & & & \ddots & & & \vdots \\ \mathbf{h}_{N_R,1} & & \mathbf{0} & & \mathbf{h}_{N_R,N_T} & & \mathbf{0} \\ & \ddots & & \cdots & & \ddots & \\ \mathbf{0} & & \mathbf{h}_{N_R,1} & & \mathbf{0} & & \mathbf{h}_{N_R,N_T} \end{bmatrix}, \quad (4)$$

which now fully describes the frequency selective channel for received sample sets of length N. $\mathbf{r}$ is the stacked vector of all received signals, while $\mathbf{s}$ being a stacked vector of transmitted chips from all transmit antennas. $\mathbf{n}$ corresponds to the noise vector.

III. Equalization

The main purpose of introducing additional complexity with chip-level equalizers (CE) is to restore the orthogonality of the spreading codes, which is lost due to multipath propagation. These equalizers estimate the transmitted chip samples by a set of N_T FIR filters consisting of L_{eq} coefficients each per receive antenna. In our investigations the number of filter taps L_{eq} equals our chosen sample set length N. It should be noted that the filter length is not fixed in general. In this section we briefly describe the considered algorithms, which are all based on linear approaches. As shown in many proposals, efficient implementation of these algorithms can be realized especially by applying signal processing in the frequency domain [9].

A. Linear MMSE solution

The optimal solution for the filter coefficients is given by the minimum mean squared error (MMSE) criteria and can be formulated as

$$\mathbf{w}^{\text{opt}} = \arg\min_{\mathbf{w}} \mathrm{E}\left\{\left|\mathbf{w}^H\mathbf{r}-\mathbf{s}\right|^2\right\} , \tag{5}$$

where $\mathrm{E}\{\cdot\}$ denotes the expectation. The resulting estimation for the coefficients $\mathbf{w}^{\text{opt}}$ with regard to the output covariance matrix Φ_{rr} and the SINR is [9]

$$\begin{aligned} \mathbf{w}^{\text{opt}} &= \Phi_{rr}^{-1}\mathbf{H}\mathbf{f}_{D_c} && (6)\\ &= \left(\mathbf{H}\mathbf{H}^H + \frac{N_T}{\text{SINR}}\mathbf{I}_{L_{\text{eq}}N_R}\right)^{-1}\mathbf{H}\mathbf{f}_{D_c} && (7) \end{aligned}$$

where $\mathbf{f}_{D_c}$ consists of N_T stacked all zero vectors with a single one at the $D_c \cdot n_T$-th position of the corresponding antenna n_T. These positions were chosen in a compromising way to deal with minimum as well as maximum phase channels. Therefore, D_c is the combined equalizer and channel delay. An estimate of the entries of the channel matrix $\mathbf{H}$ can be obtained via slot-averaged channel estimation. $\mathbf{I}_{L_{\text{eq}}N_R}$ denotes an identity matrix of size $L_{\text{eq}}N_R \times L_{\text{eq}}N_R$, while $(\cdot)^H$ is the Hermitian of a matrix. Eq. (7) is valid because we assume uncorrelated transmit data here.

B. Conjugate gradient algorithm (CGA)

The dominating complexity in (6) is the calculation of the inverse of the correlation matrix Φ_{rr}. One possibility to avoid the direct-matrix inverse is to use a conjugate gradient algorithm [10]. This algorithm is an approach to approximate (6) or (7) iteratively within J iterations. It can be used because the correlation matrix has Hermitian and positive definite properties. Here, Algorithm 1 was applied to all receive antennas N_R, where $\mathbf{v}(j)$ is the residual vector and $\mathbf{p}(j)$ the current search direction with j being the iteration index. According to [10] it is sufficient to update these filter coefficients or the resulting coefficients from Sec. III-A once per WCDMA slot to cope with a wide range of mobile speeds. However, since HSDPA is designed for lower speeds and has a short slot duration of 0.667ms, it should be possible to decrease this update rate.

Algorithm 1 Conjugate Gradient Algorithm [10]

$\mathbf{w}_{\text{CGA}}(0) = \mathbf{H}\mathbf{f}_{D_c}$
for $n_R = 1 : N_R$ **do**
  $\mathbf{v}(0) = \mathbf{w}_{\text{CGA}}^{(n_R)}(0) - \Phi_{rr}\mathbf{w}_{\text{CGA}}^{(n_R)}(0)$
  $\mathbf{p}(0) = \mathbf{v}(0)$
  $\beta(0) = 0$
  for $j = 1 : J$ **do**
    $\mathbf{p}(j) = \mathbf{v}(j-1) + \beta(j-1)\,\mathbf{p}(j-1)$
    $\mathbf{z}(j) = \Phi_{rr}\mathbf{p}(j)$
    $\alpha(j) = \frac{\mathbf{v}^H(j-1)\mathbf{v}(j-1)}{\mathbf{p}^H(j)\mathbf{z}(j)}$
    $\mathbf{w}_{\text{CGA}}^{(n_R)}(j) = \mathbf{w}_{\text{CGA}}^{(n_R)}(j-1) + \alpha(j)\,\mathbf{p}(j)$
    if $j < J$ **then**
      $\mathbf{v}(j) = \mathbf{v}(j-1) - \alpha(j)\,\mathbf{p}(j)$
      $\beta(j) = -\frac{\mathbf{z}^H(j)\mathbf{v}(j)}{\mathbf{p}^H(j)\mathbf{z}(j)}$
    end if
  end for
end for

C. Normalized least mean squares algorithm (NLMS)

One disadvantage of the aforementioned linear equalization is the need for channel state information (CSI) at the receiver. Normally, in HSDPA CSI is estimated using code-multiplexed common pilot symbols. An alternative of exploiting the channel information are adaptive equalization techniques like the least mean squares (LMS) or recursive least squares (RLS) algorithm. The equalizer coefficients of which are trained by the reference pilot chips to adaptively track channel variations [11]. As a result, channel estimation becomes unnecessary. For comparison, we consider a normalized version of the LMS approach in frequency domain by applying $2 \cdot L_{\text{eq}}$-point FFTs and overlap-add method. The received signal is separated into several blocks of length L_{eq} and padded with zeros before frequency transform. Equalization is done in each of the resulting $2 \cdot L_{\text{eq}}$ subbands, where the updating rule for the frequency domain filter coefficients in the j-th iteration step for subband u is

$$\mathbf{W}_u(j) = \mathbf{W}_u(j-1) + \mu \cdot \frac{\mathbf{R}_u(j)\,\mathbf{e}_u^T(j)}{\mathbf{R}_u^H(j)\,\mathbf{R}_u(j)} . \tag{8}$$

Here μ denotes the algorithm step size. $\mathbf{W}_u$ is a matrix of dimensions $N_R \times N_T$ containing the frequency domain filter coefficients. $\mathbf{R}_u$ and $\mathbf{e}_u$ are column vectors, which correspond to the L_{eq} received samples and the error vector in the frequency domain, respectively. $(\cdot)^T$ denotes the complex conjugate transpose of a vector.

TABLE I

SIMULATION PARAMETERS

Parameter	Value
Channel type	ITU Pedestrian A & B
HS-PDSCH power	-1.55 dB (70%)
No. of spreading codes	5 / (10)
P - CPICH power	-10 dB (10%)
Channel estimation	correlation based (CPICH)
Turbo decoding	max-log-MAP - 8 iterations
Target FER	10%
Max. no. of retransmissions	4
OCNS	on
Feedback bit error rate	4 %
Feedback delay	1 TTI
No. of filter taps	32

TABLE II

ADAPTIVE MODULATION AND CODING SCHEMES

MCS level	Modulation	Code rate
1	-	-
2	QPSK	1 / 4
3	QPSK	1 / 2
4	QPSK	3 / 4
5	16-QAM	1 / 2
6	16-QAM	3 / 4

IV. SIMULATION RESULTS

In this section, we investigate the throughput performance of different MIMO-CDMA antenna and receiver configurations. A comparison is done for HSDPA systems using common Rake reception and chip-level equalization in a single-user case.

A. Requirements

The first simulations in this section are based on the parameters depicted in Table I. For simplicity, one single physical layer hybrid-ARQ scheme was implemented using a fixed number of spreading codes. Consequently, the transmission is independent from specific changes in the transport block size as well as any reference power adjustment apart from modulation and coding [12]. Table II summarizes all six possible MCS levels which were used. Different code rates are achieved via parity bit puncturing or repetition of the standard 1/3 turbo code [6]. MCS level 1 for a group actually means ”no transmission”. The chosen level for each transmit antenna depends on the estimated SINR after filtering based on reference thresholds at a specified frame error rate (FER). Such reference thresholds for 10% FER can be found in [13]. The modulation and coding properties of the antennas are updated every transmission time interval (TTI), which is defined as the inter-arrival time between different transport blocks. In HSDPA this TTI is set to 2ms, which is five times shorter than the minimum TTI in the conventional Rel. ’99 UMTS standard. The TTI contains three slots, which are also referred to as ”frames” during this work. If a packet is received with negative acknowledgement, namely if a CRC check fails, the maximum number of retransmissions was set to four, using SINR weighted soft combining of corresponding packets before decoding [14].

For single transmit antenna systems the number of available feedback bits, also known as Channel Quality Indicator (CQI) bits, for signalling of CSI is currently fixed to max. 5 bit/TTI in HSDPA [6]. To maintain this specification, the possible MCS combinations for multiple transmit antennas have to be quantized since we only have a maximum of $2^5 = 32$ signal states available. For example, $6^2 = 36$ states are needed for full signalling in a $2 \times N_R$ antenna case, whereas the 5 bits are adequate to signal all $6^1 = 6$ states in a single transmit antenna system. Hence, we decided to remove some MCS combinations for lower SINR values, when we transmit over multiple antennas. For all receiver types these feedback properties are the same.

The interference of the control channels and other users, affecting the transmission of all transmit antennas, is modelled with the so called Orthogonal Channel Noise Simulator (OCNS) as defined in [15]. The channel is estimated based on correlation using pilot chips of the primary common pilot channel (P-CPICH). For multiple transmit antenna systems the CPICH definition in [16] is used.

B. Performance Analysis

As depicted in Fig. 2 chip-level equalization using the MMSE criterion always performs best for all antenna configurations. The resulting maximum throughput with 2×2 PARC is doubled at high SINR due to the second

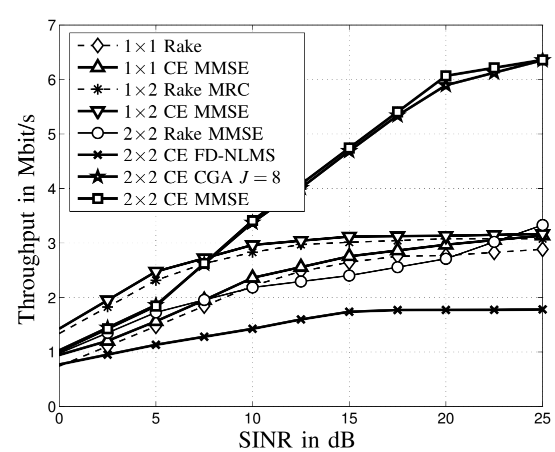

Fig. 2. Throughput results using using single-input and 2×2 PARC transmission for the ITU Pedestrian A channel and 3 kph UE speed

transmit antenna, which can be well separated from the other antenna with a linear chip-level equalizer, as the temporal displacement of the spreading codes is reversed and the MAI is suppressed. But a good performance for this configuration can only be observed with the linear MMSE solution and its approximation using CGA algorithm. Nearly the same performance of the optimum can be achieved with a small number of iterations ($J = 8$). In contrast, the NLMS equalizer in the frequency domain (FD-NLMS) shows even worse performance than a spatial MMSE receiver with Rake reception [17], which only suppresses the MAI caused by the multiple transmit antennas but *not* the inter-code interference caused by multipath propagation. The step size for this FD-NLMS algorithm was $\mu = 0.001$. As this parameter is dependent on the scenario, its application is not reasonable for a fixed parameter setting in general. This again leads to a slow convergence behavior in our scenarios, which results in a moderate equalization at the beginning of each transmission. Thus, we observe a higher frame error rate unlike the preset value. The convergence may be improved if the step size is adjusted adaptively.

It is worth noting, that the performance of a 1×2 system with equalization, as well as Rake reception, is better than the 2×2 PARC system for SINR values smaller 7 dB, only by applying additional receive diversity. For single transmit antenna systems the gain for equalization is marginal regarding the small delay spread of 2 chip lengths ($0.5\mu s$) in the ITU Pedestrian A channel.

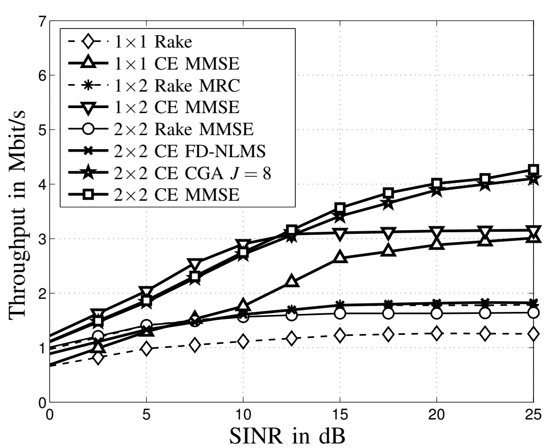

Fig. 3. Throughput results using single-input and 2×2 PARC transmission for the ITU Pedestrian B channel and 3 kph UE speed

Fig. 3 depicts the same investigation for a channel with a maximum delay spread of 14 chip lengths ($3.6\mu s$). Linear MMSE and CGA equalization again show the best performance for high SINR of the user. The achievable rates for 2×2 PARC at 25 dB are less in comparison to the ITU Pedestrian A channel. This is consistent with [18] as less multipath should perform better for high SINR. The higher non-orthogonality of the received physical data channels results in a poor performance for a conventional Rake receiver. In contrast, the MMSE equalizer remains good with a gain of 1.5 Mbit/s at 15 dB SINR in the 1×1 case. Thus, the additional complexity introduced by equalization is really profitable in cases, where delay spread in chip length is large with regard to spreading factor G. Nevertheless, equalization for small delay spreads is also beneficial if the system load increases. This is shown in Fig. 4.

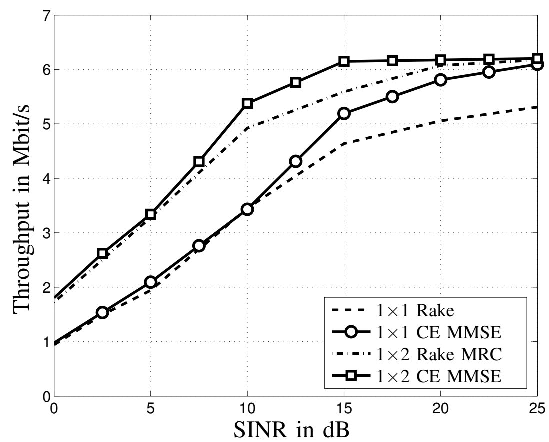

Fig. 4. Throughput results for the ITU Pedestrian A channel, single transmit antenna, 3 kph UE speed and load $2/3$ ($K = 10$)

With the utilization of 10 spreading codes, the equalizer is superior to Rake reception. The same observation holds for 1×2 systems with Maximum Ratio Combining (MRC) of all receive antennas. Although almost the same throughput can be achieved at SINR values less than 5 dB, the equalizer reaches the maximum possible throughput with a benefit of around 10 dB compared to Rake reception.

As mentioned before, the knowledge of the channel at the transmitter and the receiver is essential for the throughput performance. In this regard, Fig. 5 denotes the relation between SINR and the relative throughput decrease using different receiver types with respect to the throughput that can be achieved with perfect channel estimation. Here, the ITU Pedestrian A channel is considered again. Due to imperfect channel estimation the MMSE receiver using Rake reception shows highly sensitive behavior concerning channel estimation errors with up to 30% less throughput compared to perfect channel estimation. In contrast, the equalizers are more robust with only a maximum of 10% in throughput decrease and have the general tendency to converge to the perfect channel knowledge case for increasing SINR.

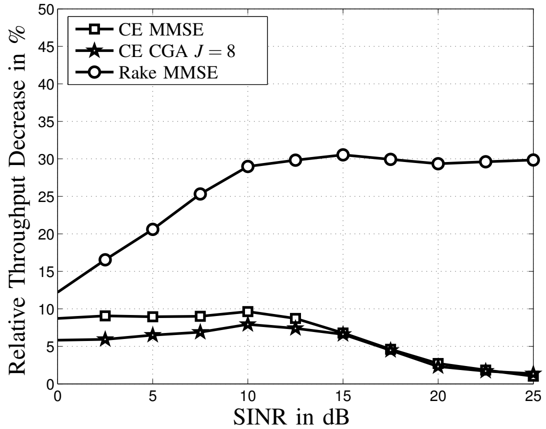

Fig. 5. SINR vs. the relative throughput decrease due to imperfect channel estimation for the ITU Pedestrian A channel, 2×2 PARC, 3 kph UE speed and 5 spreading codes using different receivers

The same observation holds for an increased feedback bit error rate as stated in Fig. 6 at 5 dB SINR. With an increase in feedback errors in the uplink direction, the MMSE Rake receiver always shows worse performance than the equalizers and has a higher throughput decrease especially for more feedback errors.

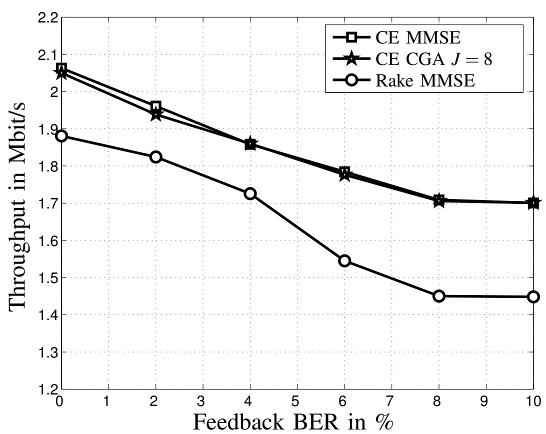

Fig. 6. Throughput vs. feeback bit error rate at 5 dB SINR for the ITU Pedestrian A channel, 2×2 PARC, 3 kph UE speed and 5 spreading codes using different receivers

V. Summary and Conclusions

In this paper, we showed some throughput results of chip-level equalization techniques concerning a MIMO enhancement for the 3GPP Wideband CDMA FDD Downlink with PARC. These results for HSDPA indicate a superior performance of advanced receivers especially at high SINR ranges, particularly in scenarios with more extensive multicode transmission and appearance of large delay spreads. There, the maximum available throughput can be doubled due to a second transmit antenna. For low SINR values, the application of a single transmit antenna using equalization in combination with receive diversity by multiple receive antennas shows best results. Concerning the effect of decreasing throughput due to imperfect channel estimation and erroneous feedback information a less sensitive performance of the equalization techniques was observed.

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