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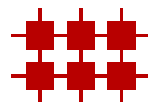
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# Downlink Processing for Mitigation of Intracell Interference in DS-CDMA Systems

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*Abstract* — We present a downlink processing algorithm which provides the necessary signal level at the output of the demodulator at the mobile station (MS). Both, the base station (BS) and MS are equipped with only one antenna. Furthermore, we presume a time division duplex (TDD) system, therefore, the complete channel is known to the BS. With this knowledge an exact representation of the downlink transmission is possible at the BS.

The conventional way to create the signal for transmission in a direct sequence code division multiple access (DS-CDMA) system is to sum the weighted outputs of spread QPSK signals dedicated to the different mobiles. Contrary, we optimize the transmission signal itself to provide the demanded outputs at the mobiles. Simulation results show the superiority of the new approach.

## I. INTRODUCTION

DS-CDMA will be a major part of the 3G systems. In January 1998, the *European Telecommunications Standards Institute* (ETSI) agreed to base the FDD transmission on *Wideband CDMA* (WCDMA). In this paper, we examine a TDD system with a modulation scheme similar to WCDMA.

The assumption of a TDD system implies that the channel is known to the base station. This knowledge can be used to improve the transmission and, thus, increase system capacity. Furthermore, the transmission signal of a TDD system has *burst structure*, hence, we can apply burst-wise downlink processing. This paper presumes perfect channel estimation at the BS and MS, therefore, its results can be viewed as a *lower bound* of raw BER achievable with downlink processing.

After explaining the system model in Section II, we review two recently published downlink processing approaches and present a downlink processing algorithm in Section III, and Section IV contains BER simulation results, where we compare the different downlink processing schemes.

## II. SYSTEM MODEL

### A. DS-CDMA Modulation

We examine a DS-CDMA system which is similar to WCDMA [1]. The symbols  $s_k^{(m)}$  of user  $k \in \{1, \dots, K\}$  are spread by a spreading code  $c_k^{(m)}(t)$ . The resulting base band DS-CDMA signal for one TDD burst reads as

$$s_k(t) = \sum_{m=0}^{M_k-1} s_k^{(m)} c_k^{(m)}(t - mT_k), \quad (1)$$

where  $T_k = SF_k \cdot T_c$ ,  $SF_k$ , and  $1/T_c$  are the symbol time, the spreading factor of user  $k$ , and the chiprate, respectively. The number of symbols per slot  $M_k$  is the ratio between the number of chips per slot  $N_c$  and the spreading factor  $SF_k$  of the respective mobile. The symbols  $s_k^{(m)}$  are QPSK modulated, therefore,  $s_k^{(m)} \in \{e^{j\pi/4}, e^{j3\pi/4}, e^{-j3\pi/4}, e^{-j\pi/4}\}$ . The spreading code of symbol  $m$

$$c_k^{(m)}(t) = \sum_{l=0}^{SF_k-1} d_k^{(m)}[l] p_{\text{rrc}}(t - lT_c) \quad (2)$$

is composed of  $SF_k$  chips  $d_k^{(m)}[l] \in \{-1, 1\}$  weighted by the chip waveform  $p_{\text{rrc}}(t)$  which is characterized by unit energy and a square-root raised cosine spectrum with a roll-off factor  $\alpha = 0.22$ .

The chips  $d_k^{(m)}[l]$  are a combination of the *Orthogonal Variable Spreading Factor* (OVSF, [2]) code  $o_k[l]$  and the pseudo-noise Gold sequence  $g[l]$  of length  $N_g$  used for scrambling [2]:

$$d_k^{(m)}[l] = o_k[l] g[(m SF_k + l) \bmod N_g], \quad (3)$$

where ‘mod’ denotes modulo division.

### B. Channel Model

At the BS, the signals  $s_k(t)$  dedicated to the different MS are weighted and added. Therefore, every MS receives portions of all sent signals  $s_{k'}(t)$ :

$$r_k(t) = h_k(t) * \sum_{k'=1}^K w_{k'} s_{k'}(t) + n_k(t), \quad (4)$$

where ‘\*’,  $K$ , and  $n_k(t)$  denote linear convolution, the number of mobiles, and *intercell interference* combined with *noise* at mobile  $k$  which are both assumed to be white Gaussian random processes, respectively. In WCDMA, the default value for the weight  $w_{k'}$  is set to  $1/\sqrt{SF_{k'}}$ .

We presume a channel with *discrete multipaths*. Hence, the channel impulse response  $h_k(t)$  representing the transmission to mobile  $k$  is a FIR filter:

$$h_k(t) = \sum_{q=1}^Q h_{k,q} \delta(t - \tau_{k,q}), \quad (5)$$

with the Dirac delta function  $\delta(t)$ . The  $Q$  paths leading to mobile  $k$  are represented by the the transmission factors  $h_{k,q} \in \mathbb{C}$  (the reciprocal of the respective path attenuations) and the path delay times  $\tau_{k,q}$  which are assumed to be uniformly distributed within the interval  $[0, \tau_{\text{max}}]$ .

If we combine Equation (1), (4), and (5) and neglect noise and intercell interference, the received base band signal at mobile  $k$  can be written as

$$r_k(t) = \sum_{q=1}^Q h_{k,q} y_{\text{rrc}}(t - \tau_{k,q}) \quad (6)$$

where

$$y_{\text{rrc}}(t) = \sum_{k'=1}^K \sum_{m=0}^{M_k-1} s_{k'}^{(m)} w_{k'}^{(m)} c_{k'}^{(m)}(t - mT_{k'}). \quad (7)$$

Note that  $y_{\text{rrc}}(t)$  is the signal which is transmitted over the antenna and that we introduced a dependence of  $w_{k'}^{(m)}$  on the symbol index  $m$  as in [3] to stress that  $w_{k'}^{(m)}$  can change from *symbol to symbol*.

The purpose of downlink processing is to choose appropriate  $w_{k'}^{(m)}$  to ensure that the symbols  $s_{k'}^{(m)}$  are detected correctly at mobile  $k$ . In the case of symbol-wise downlink processing [3] the weights  $w_{k'}^{(m)}$  change from symbol to symbol. However, splitting up the problem into the computation of  $K$  weights  $w_{k'}^{(m)}$  is not a must. We only have to find a transmission signal  $y_{\text{rrc}}(t)$  which is a weighted sum of time shifted square root raised cosine pulses

$$y_{\text{rrc}}(t) = \sum_{l=0}^{N_c-1} y[l] p_{\text{rrc}}(t - lT_c) \quad (8)$$

to fulfill the requirement of correct detection at all MS. Note that the step from a sum of *weighted discrete* value sequences to a simple sequence of *continuous* values is the main contribution of this work.

### C. Reciprocity of the Channel Parameters

Our downlink processing algorithm is based on the channel parameters estimated in the uplink. This can be done by simply correlating the received uplink signal with the pilot signal or by more sophisticated channel estimation algorithms (e. g. [4]). However, we assume *perfect channel estimation* in the uplink. Hence, we do not make a distinction between the real and the estimated channel parameters.

In TDD systems, a major criterion for the validity of the downlink channel parameters estimated in the uplink is the coherence time [5, 6]. We assume that the burst duration time is small compared to the coherence time of the channel. Thus, not only the channel is constant over one burst, but also the channel parameters estimated in the *uplink* are *valid* for the *downlink*.

### D. Correlation Demodulator

After matched filtering and sampling the received signal  $r_k(t)$  at chiprate, therefore,  $r_k^{\text{mf}}[n] = r_k^{\text{mf}}(nT_c)$ , and performing *coherent demodulation*, the detection signal for the  $m$ -th symbol can be written as

$$u_k^{(m)} = \sum_{l=0}^{SF_k-1} d_k^{(m)}[l] h_{k,1}^* r_k^{\text{mf}}[l + mSF_k + l_{k,1}], \quad (9)$$

where  $r_k^{\text{mf}}[n]$  is the matched filtered and sampled received signal. Note that we assume without loss of generality that  $|h_{k,1}| > |h_{k,q}|$ ,  $1 < q \leq Q$ . Additionally, we assume a chip rate  $1/T_c$  so that the multipath delays  $\tau_{k,q}$  are integer multiples  $l_{k,q} \in \{0, l_{\max}\}$  of the chip time  $T_c$ .

The combination of Equation (9) with Equation (6) yields

$$\mathbf{u}_k = h_{k,1}^* \mathbf{D}_k^T \mathbf{H}_k \mathbf{y}, \quad (10)$$

where we collected the demodulator outputs of one TDD burst in the  $\mathbb{C}^{M_k \times 1}$  vector

$$\mathbf{u}_k = [u_k^{(0)}, \dots, u_k^{(M_k-1)}]^T. \quad (11)$$

The scrambled spreading sequences of one burst were put into the  $\{-1, +1\}^{N_c \times M_k}$  matrix

$$\mathbf{D}_k = \mathbf{G}(\mathbf{1}_{M_k} \otimes \mathbf{o}_k) \quad (12)$$

with the  $M_k \times M_k$  identity matrix  $\mathbf{1}_{M_k}$  and the Kronecker product denoted by ' $\otimes$ '. The elements of the  $N_c \times N_c$  diagonal matrix  $\mathbf{G}$  are the values of the pseudo-noise Gold sequence  $g[l]$  and the OVSF code of mobile  $k$  can be found in the  $\{-1, +1\}^{SF_k \times 1}$  vector

$$\mathbf{o}_k = [o_k[0], \dots, o_k[SF_k - 1]]^T. \quad (13)$$

By introducing the  $M \times M$  nilpotent shift matrix

$$\mathbf{\Gamma}_M = \begin{bmatrix} \mathbf{0}_{M-1}^T & 0 \\ \mathbf{1}_{M-1} & \mathbf{0}_{M-1} \end{bmatrix} \quad (14)$$

with the  $M \times 1$  zero vector  $\mathbf{0}_M$  and defining  $\mathbf{\Gamma}_M^0 = \mathbf{1}_M$  and the inverse  $(\mathbf{\Gamma}_M)^{-1} = (\mathbf{\Gamma}_M)^T$ , the channel can be represented by the  $N_c \times N_c$  Toeplitz matrix

$$\mathbf{H}_k = \sum_{q=1}^Q h_{k,q} \mathbf{\Gamma}_{N_c}^{l_{k,q} - l_{k,1}}. \quad (15)$$

The remaining quantity is the transmitted signal

$$\mathbf{y} = [y[0], \dots, y[N_c - 1]]^T \in \mathbb{C}^{N_c \times 1}. \quad (16)$$

Note that we moved the matched filtering to the transmitter, thus, the transmitted signal sampled at chiprate is simply the sequence  $y[l]$ .

### E. Rake Demodulator

After matched filtering and sampling the received signal at chiprate and performing *maximum ratio combining* (MRC), the rake detection signal for the  $m$ -th symbol can be written as

$$u_k^{(m)} = \sum_{l=0}^{SF_k-1} d_k^{(m)}[l] \sum_{f=1}^F v_{k,f} r_{k,q}^{\text{mf}}[l + mSF_k + l_{k,f}], \quad (17)$$

where  $F$  is the number of rake fingers. The rake finger weights  $v_{k,f}$  are set to the conjugate complex of the  $F$  strongest channel impulse response taps  $h_{k,f}$ . Note that we again assume without loss of generality that  $|h_{k,f}| > |h_{k,q}|$ ,  $1 \leq f \leq F$ ,  $F + 1 \leq q \leq Q$ .

Similar to Equation (10), we yield after transforming to matrix vector notation

$$\mathbf{u}_k = \mathbf{D}_k^T \mathbf{V}_k \mathbf{H}'_k \mathbf{y}, \quad (18)$$

where  $\mathbf{D}_k$  and  $\mathbf{y}$  contain the scrambled spreading codes (cf. Equation 16) and the transmitted signal (cf. Equation 12), respectively.

With the  $M \times M + N$  selection matrix

$$\mathbf{S}_{(L,M,N)} = [\mathbf{0}_{M \times L}, \mathbf{1}_M, \mathbf{0}_{M \times N-L}], \quad (19)$$

where  $\mathbf{0}_{M \times L}$  denotes the  $M \times L$  zero matrix, the channel reads as

$$\mathbf{H}'_k = \sum_{q=1}^Q h_{k,q} S_{(l_{k,q}, N_c, l_{\max})}^T \in \mathbb{C}^{N_c + l_{\max} \times N_c} \quad (20)$$

and the rake demodulator is represented by the  $N_c \times N_c + l_{\max}$  Toeplitz matrix

$$\mathbf{V}_k = \sum_{f=1}^F v_{k,f} S_{(l_{k,f}, N_c, l_{\max})}, \quad (21)$$

where  $l_{\max}$  is the number of samples according to  $\tau_{\max}$ .

### III. DOWNLINK PROCESSING

#### A. Pre-Rake Downlink Processing

In [7], Esmailzadeh and Nakagawa presented a processing scheme for downlink transmission in TDD systems called Pre-Rake. They showed for the single user case that it is possible to move the maximum ratio combining rake from the MS to the BS without changing the BER performance. The remaining receiver at the mobile is a simple correlation demodulator. Thus, the complexity of the MS is decreased while system capacity is not effected.

Esmailzadeh et. al. also investigated the generalization to the multi-user case [8]. The *maximum ratio combining* of each mobile is *moved* to the BS. Therefore, the transmitted signal is the sum of the Pre-Rake filtered signals dedicated to the different mobiles:

$$y_{\text{pre}}(t) = \sum_{k=1}^K \frac{1}{\sqrt{\sum_{q=1}^Q |h_{k,q}|^2}} \sum_{q=1}^Q h_{k,q}^* s_k(t + \tau_{k,q}), \quad (22)$$

where the normalization with  $1/\sqrt{\bullet}$  is introduced to provide equal transmission power for every mobile.

The Pre-Rake system is *superior* to the conventional rake system for *non-orthogonal* spreading codes. However, if *orthogonal spreading* codes are utilized, *Pre-Rake* is *worse* than *conventional rake*. Moreover, using *orthogonal* codes leads to *better* BER performance than using *non-orthogonal* codes. Thus, there has to be found a trade-off between simplicity of the receiver at the MS and BER performance, when orthogonal spreading codes are used.

#### B. Post-Rake Downlink Processing

Recently, Barreto and Fettweis [9] extended the Pre-Rake to a Post-Rake. Their idea is to maximize the *signal to noise ratio* (SNR) at the MS in a Pre-Rake system. To this end, they suggest to replace the correlation demodulator by a rake receiver which is matched to the combination of Pre-Rake and channel. Therefore, the impulse response of the Post-Rake reads as

$$\begin{aligned} v_{\text{post},k}(t) &= v_{\text{post},k}^*(-t) = \sum_{f=1}^{F_k} v_{k,f} \delta(t + \tau'_{k,f}) = \\ &= \sum_{q=1}^Q \sum_{p=1}^Q h_{k,q} h_{k,p}^* \delta(t - \tau_{k,q} + \tau_{k,p}). \end{aligned} \quad (23)$$

Note that the number of necessary rake fingers at the receiver  $F_k \geq 2Q - 1$ . Hence, the receiver complexity is significantly increased compared to a Pre-Rake system and also compared to a conventional rake system.

Barreto et. al. showed that Post-Rake processing improves the BER performance for a small number of users  $K$  compared to the spreading factor  $SF$ . However, if the number of mobiles lies in the vicinity of the spreading factor, Pre-Rake outperforms Post-Rake processing although the latter one has a higher receiver complexity. Moreover, Barreto et. al. only investigated *non-orthogonal* codes. Because [8] showed that orthogonal codes lead to better system performance than non-orthogonal codes for Pre-Rake processing, our feeling is that Post-Rake processing has to be further investigated.

#### C. New Downlink Processing Algorithm

In the previous section we found a representation of the demodulator output at mobile  $k$  depending on the transmitted signal (cf. Equation 10 and 18). In this section we use the abbreviation

$$\mathbf{u}_k = \mathbf{T}_k \mathbf{y}. \quad (24)$$

If we combine the equations for all mobiles  $k \in \{1, \dots, K\}$ , then we get

$$\mathbf{u} = \mathbf{T} \mathbf{y}, \quad (25)$$

where  $\mathbf{u} = [\mathbf{u}_1^T, \dots, \mathbf{u}_K^T]^T$  and  $\mathbf{T} = [\mathbf{T}_1^T, \dots, \mathbf{T}_K^T]^T$  are the stacked versions of the demodulator outputs and transmission matrices, respectively.

The transmitted signal  $y_{\text{rrc}}(t)$  has to be chosen to ensure that the demodulated symbols  $u_k^{(m)}$  are equal to the demanded values  $s_k^{(m)}$ . Moreover, we can assume that most of the time the number of mobiles  $K$  is below the maximum number of mobiles in a CDMA system, i. e.  $\sum_{k=1}^K 1/SF_k < 1$ . Therefore, we have unused degrees of freedom which we can use to minimize transmission power. Consequently, we can write:

$$\min \|\mathbf{y}\|_2^2 \quad (26)$$

$$\text{s.t.: } \mathbf{T} \mathbf{y} = \mathbf{s},$$

where  $\mathbf{s}$  is a vector of the demanded outputs at the mobiles  $s_k^{(m)}$  defined in the same way as  $\mathbf{u}$ .

The solution of Equation (26) is the multiplication with the pseudo-inverse of  $\mathbf{T}$ , i. e.

$$\mathbf{y} = \mathbf{T}^H (\mathbf{T} \mathbf{T}^H)^{-1} \mathbf{s} = \mathbf{T}^\dagger \mathbf{s}. \quad (27)$$

Note that the values of  $\mathbf{s}$  can be scaled arbitrarily. Thus, *transmission power control* (TPC) can be implemented quite easily, i. e. if mobile  $k$  needs more signal power, the respective values of  $\mathbf{s}$  can be increased.

### IV. SIMULATION RESULTS

We compare our new downlink processing approach to the conventional rake demodulator, the Pre-Rake approach, and the Post-Rake receiver. We assume  $Q = 5$  paths connecting the BS and each MS. Each path has a phase which is uniformly distributed within the interval  $[0, 2\pi]$  and an amplitude which is Rayleigh distributed with unit power, therefore,  $E[|h_{k,q}|^2] = 1$ . The maximum delay spread  $\tau_{\max}$  is set to  $10\mu\text{s}$  which is approximately  $40T_c$ , if a chip rate similar to WCDMA is assumed. The channel is computed for every simulation run and the results are the mean of 1000 simulation runs. The channel estimation is assumed to be ideal and the channel stays constant over the whole burst. Each burst consists of  $N_c = 800$  chips.

Our investigated scenario includes  $K = 4$  high-rate mobiles with a spreading factor of  $SF_k = 4, k \in \{1, \dots, K\}$ , thus,

$M_k = 200$  symbols,  $k = 1, \dots, K$ . The maximum number of mobiles for this spreading factor is reached, therefore, the examined scenario is interference limited. The shown results are the mean of the results for all  $K = 4$  mobiles.

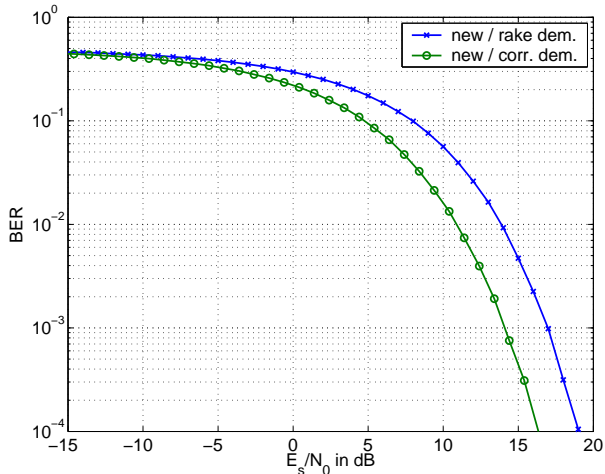


Figure 1: New approach: BER vs. SNR

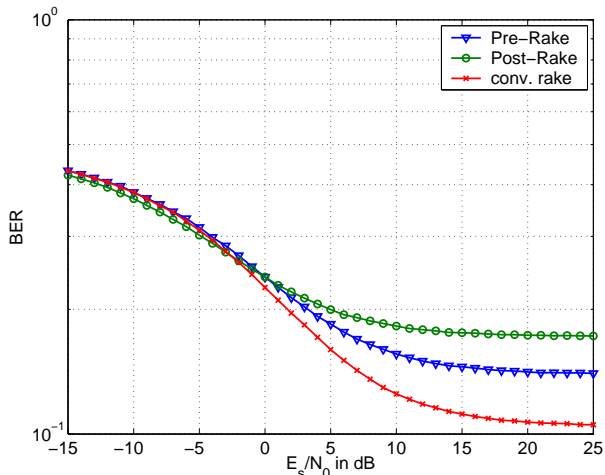


Figure 2: Pre-Rake, Post-Rake, and conventional rake: BER vs. SNR

Figure 1 shows the BER performance of the new downlink processing scheme. First, we observe that the BER converges to zero for increasing SNR. This result is not surprising, since our method was designed to remove the effects of intracell interference. Second, the system with a correlation receiver outperforms the system with a rake receiver at the mobile. The reason for this difference can be found in the way the rake combines the received signal. Our algorithm provides the same signal energy at the correlator output for both receiver types. However, maximum ratio combining leads to noise enhancement which is overcompensated in a conventional rake system. Contrary, in the case of our new downlink processing scheme signal energy is not enhanced, but interference is removed.

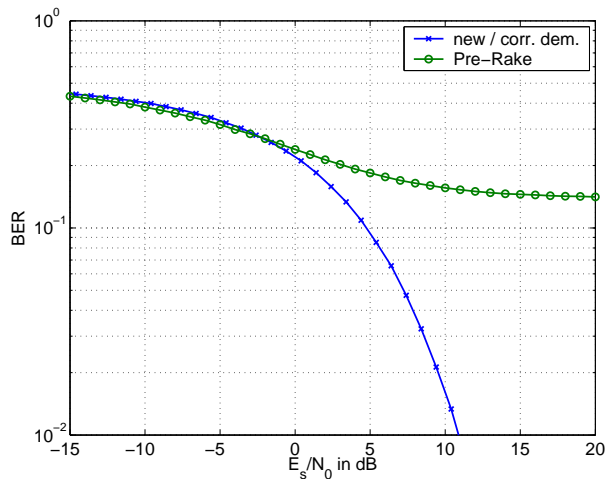


Figure 3: New approach / correlation demodulator and Pre-Rake: BER vs. SNR

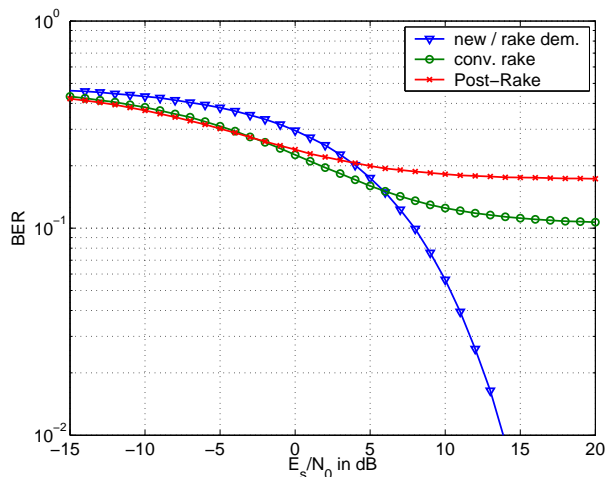


Figure 4: New approach / rake demodulator, Post-Rake, and conventional rake: BER vs. SNR

Pre-Rake processing, Post-Rake processing, and conventional rake receiver are compared in Figure 2. As expected using Post-Rake leads to a reduction of BER for the high noise case, but the difference to the other two approaches is very small. For large SNR values, the conventional rake receiver outperforms Pre-Rake and Post-Rake, because these two lead to interference enhancement. Figure 2 also suggests that the conventional rake system outperforms the Pre-Rake approach for orthogonal spreading codes as has been shown in [8]. Moreover, employing Post-Rake seems to further increase interference, hence, leading to worse BER than using Pre-Rake only.

For the comparison of our new downlink processing approach to Pre-Rake processing we have to use correlation receivers at the mobile stations. This comparison is shown in Figure 3. If the system is noise limited the two schemes exhibit nearly equal performance, but the Pre-Rake approach suffers from interference for high SNR values, whereas our

new method removes interference completely.

In Figure 4, our new approach with rake demodulators at the MS is compared to a Post-Rake system and a conventional rake system. Again, we can see a slight degradation of our new downlink processing scheme for low SNR compared to the other two methods and the superiority of the new algorithm for the interference limited case.

The previous plots showed the results for the ideal case, where full knowledge was available at the base station. Figure 5 gives an impression how sensitive downlink processing is to a-priori known system parameters. As an example we see the degradation for large SNR values, if the number of rake fingers which are used at the mobile stations differs from the number of rake fingers which are included in the optimization at the BS.

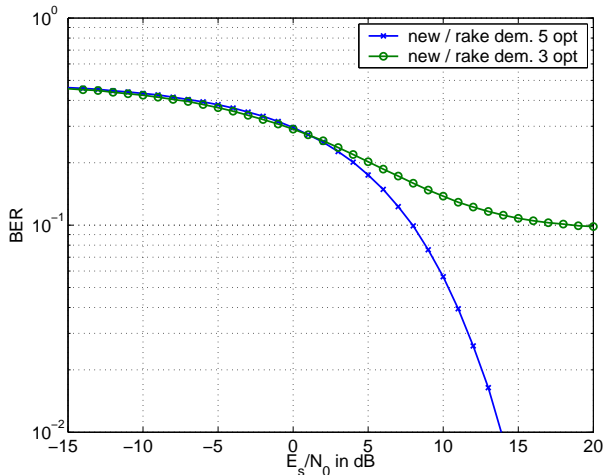


Figure 5: New approach, dependency on system information: BER vs. SNR

## CONCLUSION

We presented a new downlink processing scheme for DS-CDMA which reduces the effects of interference due to multipath propagation. Contrary to the conventional approaches, we do not optimize the combination of spread signals dedicated to the different mobiles, but optimize the transmitted signal itself. The made assumptions are the best case, since we assumed that the channel does not change over two successive TDD bursts and we also assumed perfect channel estimation at the BS and MS. Therefore, the results can be seen as a lower bound of achievable BER at least for high SNR values. The optimization only takes into account interference and not noise, therefore, the performance for the noise limited case can be improved, if noise is not neglected.

This work is the first step to a downlink processing scheme which is also applicable for a *frequency division duplex* (FDD) system. Unfortunately, the transmitted signal has no burst structure in FDD systems. Hence, an iterative version of our algorithm has to be developed. Also, the assumption that the channel impulse responses are known at the base station does not hold anymore for a FDD system, because the channels in the uplink fade independently from the channels in the downlink. Thus, feedback has to be considered, but for moving mobiles the channel changes over time leading to increasing

feedback rates. Therefore, we try to avoid feedback and suggest to use adaptive antennas which are able to estimate the *directions of arrival* (DOA) of the impinging wavefronts. Because the *directions of departure* (DOD) are nearly the same as the DOAs although there is a frequency shift between uplink and downlink, we will focus on DOA based algorithms in future work.

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