

Reduced and Differential Parallel Interference Cancellation for CDMA Systems

Achim Nahler, *Associate Member, IEEE*, Ralf Irmer, *Student Member, IEEE*, and
Gerhard Fettweis, *Senior Member, IEEE*

Abstract—Since code division multiple access systems in multipath environments suffer from multiple access interference (MAI), multiuser detection schemes should be used in the receivers. Parallel interference cancellation (PIC) is a promising method to combat MAI due to its relatively low computational complexity and good performance. It is shown in this paper that the complexity of PIC is still high for realistic scenarios in terms of the symbol rate, the number of users, spreading gain, and multipath components. However, two novel methods are introduced to reduce significantly the complexity without sacrificing performance. The first approach, called reduced PIC, takes advantage of the composition of the interference to concentrate interference cancellation only on significant terms. The second approach, called differential PIC, exploits the multistage character of PIC to avoid unnecessary double calculations of certain terms in consecutive stages. It is shown that a combination of both approaches leads to a performance very close to the single-user bound whereas the complexity can be kept on the order of the conventional RAKE receiver.

Index Terms—Code division multiple access, multicarrier spread spectrum, multiuser detection, parallel interference cancellation.

I. INTRODUCTION

BECAUSE of its wide range of advantages compared to other access schemes, code division multiple access (CDMA) is an access scheme for many future applications [1]. However, in CDMA systems multiple access interference (MAI) arises because of the nonideal crosscorrelation properties of the spreading signals and multipath propagation. The detection of such CDMA signals using the conventional RAKE receiver leads to unacceptable performance degradations in terms of the bit error rate (BER). This presents the need for more sophisticated detection strategies, such as multiuser detection (MUD) to overcome these performance degradations. An overview of different MUD methods can be found in [2]. All of them have to be measured by two figures of merit: low SNR degradation compared to the single-user bound for a

specified BER and the affordable computational complexity. In terms of the BER, the MLSD detector proposed in [3] is optimum. Unfortunately, its complexity grows exponentially with the number of users, multipath components, and symbol alphabet [4].

One suboptimum, but less complex method is the parallel interference cancellation (PIC) scheme introduced in [5] that can be repeated in multiple stages. The concept of one such stage is to regenerate the transmitted signals based on the tentative estimated data from the previous stage, emulate the distortions occurring from the multipath channels and, finally, subtract all regenerated interfering signals from the received signal to obtain more reliable estimated data for the user of interest. The performance of the PIC is determined by the number of stages and how the tentatively estimated data are weighted in each stage. Much research work has been devoted to the choice of the weighting function [6]–[9] and [10].

From an implementation point of view the PIC is investigated in [11]–[13]. Implementation aspects are also in the scope of this paper. A detailed analysis of the computational effort of two performance-equivalent implementation variants, the so-called wideband PIC (WB-PIC) and the narrowband PIC (NB-PIC) will be given. It will be shown that the complexity of both variants is too high for low-cost solutions in realistic scenarios. Hence, it is of interest to investigate how the complexity could be reduced to a feasible amount. In this paper, two novel approaches are given to reduce the computational complexity significantly. One approach is based on the idea of cancelling only a few strong interference terms instead of all interference terms. For that purpose, the interference scenario is evaluated for each received burst and the terms with the highest interference power are selected and cancelled. The second complexity reduction approach, termed differential PIC (D-PIC), exploits the multistage character of PIC. If the estimated data does not change in two consecutive stages, it will not change in all following stages. Therefore, redundant interference calculations can be avoided in later stages by this method.

The remainder of this paper is organized as follows. In Section II, the system model including multicarrier spread-spectrum frequency code division multiple access (FCDMA) is introduced. Section III-A–1 explains the RAKE receiver. In Section III, the concept of PIC is explained and its performance and computational complexity are investigated for a frequency-selective radio channel. In Sections IV and V, both novel approaches for the complexity reduction are introduced, and in Section VI the concatenation of both schemes is investigated. Finally, Section VII concludes the paper.

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A. Nahler was with Dresden University of Technology, D-01062 Dresden, Germany. He is now with Systemonic AG, D-01099 Dresden, Germany (e-mail: achim.nahler@systemonic.de).

R. Irmer is with Dresden University of Technology, D-01062 Dresden, Germany (e-mail: irmer@ifn.et.tu-dresden.de).

G. Fettweis is with Dresden University of Technology, D-01062 Dresden, Germany, and also with Systemonic AG, D-01099 Dresden, Germany (e-mail: fettweis@ifn.et.tu-dresden.de).

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TABLE I
REAL AND IMAGINARY PART OF THE COEFFICIENTS $C[k]$ USED IN THIS PAPER

k	0	1	2	3	4	5	6	7
$\text{Re}(C[k])$	0.2057	0.2139	0.0568	-0.1232	-0.2228	0.0466	-0.0210	-0.2500
$\text{Im}(C[k])$	-0.1421	0.1294	0.2435	0.2175	0.1134	-0.2456	0.2491	-0.0041
k	8	9	10	11	12	13	14	15
$\text{Re}(C[k])$	0.1985	-0.2437	0.1093	-0.0925	-0.1904	-0.1602	0.0408	0.2425
$\text{Im}(C[k])$	-0.1520	0.0556	0.2248	-0.2323	0.1620	0.1920	0.2467	0.0608

II. SYSTEM MODEL

A. Transmission Model

The uplink CDMA system model is based on the discrete-time equivalent baseband model. Assuming burstwise transmission of N BPSK symbols $d_u[n] \in \{+1, -1\}$, $n = 0, \dots, N-1$ for user $u = 0, \dots, U-1$ users, the signal at the output of the transmitter of user u is

$$s_u[\ell] = \sum_{n=0}^{N-1} c_u[\ell - nKL] \cdot d_u[n], \quad \ell = 0, \dots, NKL - 1 \quad (1)$$

where K is the spreading gain of the short code spreading sequence c_u and L is the oversampling factor. The propagation channel is modeled as a multipath frequency-selective channel. It is considered as time invariant for one burst of symbols. The chip-spaced channel impulse response (CIR) of the uplink channel for user u is

$$h_u[\ell] = \sum_{p=0}^{P-1} h_u[p] \delta[\ell - pL] \quad (2)$$

where $h_u[p]$, $p = 0, \dots, P-1$ are the channel coefficients. The signal at the receiver is

$$r[\ell] = \sum_{u=0}^{U-1} \sum_{p=0}^{P-1} h_u[p] \sum_{n=0}^{N-1} c_u[\ell - (nK + p)L] \cdot d_u[n] + w[\ell] \quad (3)$$

with the added white Gaussian noise (AWGN) $w[\ell]$ with variance $\mathcal{N}_0$. The CIRs for all users are assumed to be perfectly known to the receiver.

B. Modulation Scheme: Multicarrier Spread Spectrum

In the early 1990s, multicarrier spread spectrum (MC-SS) was introduced [14], [15]. A detailed state-of-the-art overview about MC-SS is found in [16] and [17]. Because of MC-SSs suitability for low-cost terminals [18], [19], it is used as a spread-spectrum technique in this paper.

The spreading signal is given by

$$c[\ell] = \sum_{k=0}^{K-1} C[k] \cdot e^{j2\pi k \frac{\ell}{KL}} \quad \text{with} \quad |C[k]| = 1/\sqrt{K} \quad \text{and} \quad 0 \leq \ell < KL. \quad (4)$$

Hence, the spreading signal is determined by the coefficients $C[k]$, $k = 0, \dots, K-1$. In this paper, coefficients with a low peak-to-average power ratio are used according to Table I. They were generated by a time-frequency swapping algorithm [20].

In [21], the RAKE receiver is compared with fast-Fourier-transform (FFT)-based solutions for MC-SS. Because of the lower complexity of RAKE compared to the frequency-domain equalizer [21], the RAKE receiver is used as single-user detector in this paper.

Because of implementation aspects it is useful to apply an FCDMA scheme [18]. This means that all users have the same spreading signal but user distinction is achieved by a slight user-specific frequency offset, so that the cyclic spreading signal of the u th user is given by

$$c_u[\ell] = e^{j2\pi u \frac{\ell}{KL}} \cdot c[\ell] = \sum_{k=0}^{K-1} C[k] \cdot e^{j2\pi(k+u) \frac{\ell}{KL}} \quad \text{with} \quad 0 \leq \ell < KL. \quad (5)$$

All numerical evaluations for analysis of the following algorithms are done for MC-SS FCDMA. However, it should be noted that the presented algorithms are stated in a more general manner and are also valid for direct sequence spread spectrum (DS-SS) CDMA since MC-SS is, in fact, very similar to DS-SS [18].

III. PIC

In this section, the PIC concept is revised and two implementation versions are analyzed regarding computational complexity.

A. Concept

The basic concept of parallel interference cancellation is depicted in Fig. 1. In an initial stage, the tentative symbols for each user are detected. In this paper, independent single-user RAKE detectors are used. After that, the most probable transmitted signals of all users are regenerated using the symbol estimates from the previous stage. Based on that, the MAI estimates are composed for each user and are subtracted from the received signal. After this cancellation step a new symbol detection is done based on the interference reduced signal leading to more reliable symbol estimates. This procedure can be repeated in further stages as shown in Fig. 2.

The input of PIC stage (i) is the tentative symbol $\tilde{d}_v^{(i-1)}[n]$ from the output of the RAKE receiver of the previous stage $(i-1)$, weighted by the tentative decision function $f_v^{(i)}$

$$\hat{d}_v^{(i)}[n] = f_v^{(i)} \left(\tilde{d}_v^{(i-1)}[n] \right). \quad (6)$$

The decision function in (6) has a large impact on the PIC performance. In-depth analysis of the decision function can be found

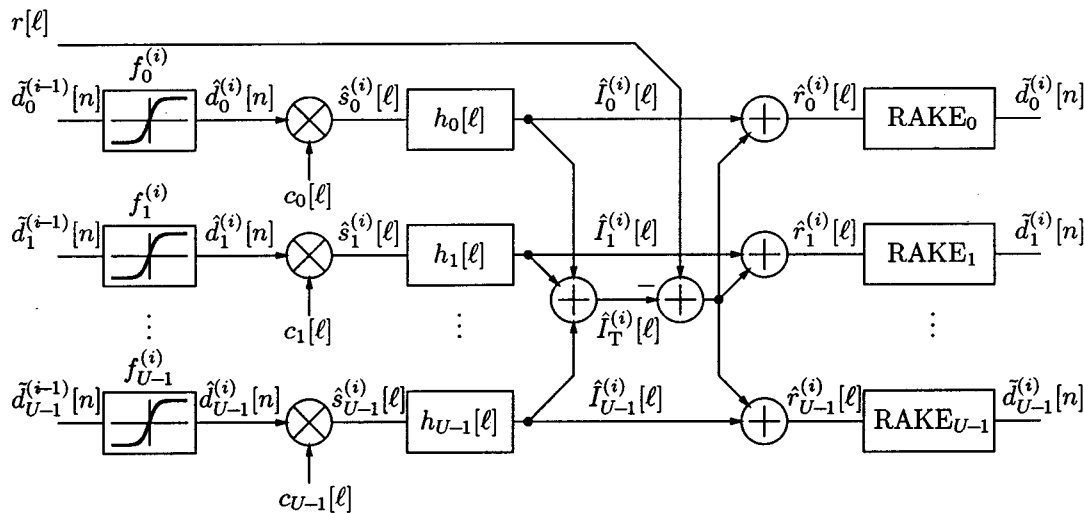


Fig. 1. Concept of the i th PIC stage.

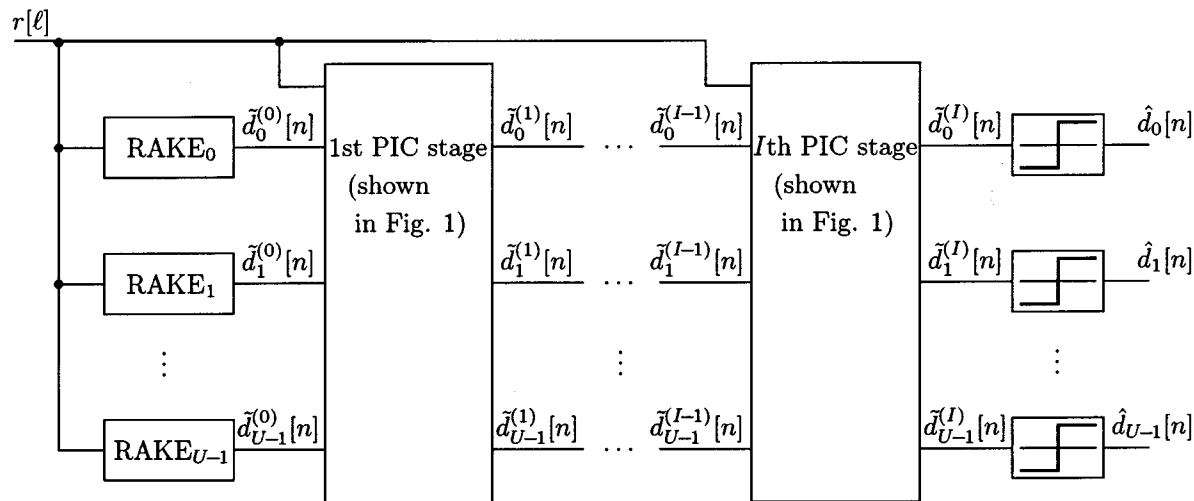


Fig. 2. Multistage PIC.

in [6], [7], [9], [10], and [22]. In the following, only hard decisions $f_v(x) = \text{sgn}(\text{Re}(x))$ are assumed.

1) *Initial PIC Stage: RAKE*: Following, the decision variables and the interference terms at the output of the RAKE receiver in the initial PIC stage are given. Only the F most powerful paths ($|h[p_{0,v}]|^2 > |h[p_{1,v}]|^2 > \dots > |h[p_{F-1,v}]|^2$) are considered, where $p_{f,v}$ denotes the delay of the f th most powerful path of the v th user. The data symbols at the output of the RAKE combiner are

$$\tilde{d}_v[n] = \frac{1}{LK} \sum_{f=0}^{F-1} h_v^*[p_{f,v}] \sum_{m=0}^{LK-1} r[m + (nK + p_f)L] \cdot c_v^*[m] \quad (7)$$

$$= \sum_{f=0}^{F-1} |h_v[p_{f,v}]|^2 \cdot d[n] + I_{A_v}[n] + I_{C_v}[n] + \eta_v[n] \quad (8)$$

with the v th user's self-interference term $I_{A_v}[n]$ resulting from nonideal autocorrelation, and the multiuser interference $I_{C_v}[n]$ being (9), as shown at the bottom of the next page, resulting from nonideal crosscorrelation of the codes and the noise term $\eta_v[n]$.

The partial discrete-time crosscorrelation between the users u and v is

$$\varphi_{u,v}^{(1)}[\ell] = \frac{1}{LK} \sum_{m=0}^{\ell L-1} c_u[m + L(K - \ell)] \cdot c_v^*[m] \quad (10)$$

$$\varphi_{u,v}^{(2)}[\ell] = \frac{1}{LK} \sum_{m=\ell L}^{LK-1} c_u[m - \ell L] \cdot c_v^*[m]. \quad (11)$$

For codes with good even and odd autocorrelation function, the self-interference term I_{A_v} is negligible compared to the crosscorrelation term I_{C_v} and will not be taken into account in further sections of this paper. It is assumed that the channel impulse length is less than one symbol ($P \leq K$). Thus, one symbol $d_v[n]$ is only influenced by the preceding or following symbols. The following concepts and analyses could be extended to longer channel impulse responses but this would lead to a more complicated description of the system and reduce the readability of this paper.

2) *WB-PIC*: In the WB-PIC, all operations are based on the respread signals. The symbol hypothesis from the previous PIC

TABLE II
DATA-DEPENDENT CROSSCORRELATION $\varphi_{u,v}^{(\text{data})}(d_1, d_2, \ell)$ (FOR BPSK, HARD DECISION)

d_1	$+1$	$+1$	-1	-1
d_2	$+1$	-1	$+1$	-1
$\varphi_{u,v}^{(\text{data})}(d_1, d_2, \ell)$	$+\varphi_{u,v}^{(\text{even})}[\ell]$	$-\varphi_{u,v}^{(\text{odd})}[\ell]$	$+\varphi_{u,v}^{(\text{odd})}[\ell]$	$-\varphi_{u,v}^{(\text{even})}[\ell]$

stage $\hat{d}_v^{(i)}[n]$ is used to respread the signal with the code c_v and to filter it by a finite-impulse response (FIR) filter with the estimated CIR coefficients for user v to obtain the interference hypothesis at the receiver caused by user v

$$\hat{I}_v^{(i)}[\ell] = \sum_{p=0}^{P-1} h_v[p] \sum_{n=0}^{N-1} c_v[\ell - (nK + p)L] \cdot \hat{d}_v^{(i)}[n]. \quad (12)$$

Throughout this paper, CIR estimates in the receiver are assumed to be perfect. For efficient implementation, the total interference was introduced in [12]. This is the sum of the interference hypotheses of all users, hence,

$$\hat{I}_T^{(i)}[\ell] = \sum_{u=0}^{U-1} \hat{I}_u^{(i)}[\ell]. \quad (13)$$

It is not user specific, i.e., it has only to be calculated once for all users. The signal

$$\begin{aligned} \hat{r}_v^{(i)}[\ell] &= r[\ell] - \hat{I}_T^{(i)}[\ell] + \hat{I}_v^{(i)}[\ell] \\ &= \sum_{p=0}^{P-1} h_v[p] \sum_{n=0}^{N-1} c_v[\ell - (nK + p)L] \cdot d_v[n] \\ &\quad + \sum_{\substack{u=0 \\ u \neq v}}^{U-1} \sum_{p=0}^{P-1} h_u[p] \sum_{n=0}^{N-1} c_u[\ell - (nK + p)L] \\ &\quad \cdot \Delta^{(i)} d_u[n] + w[\ell] \end{aligned} \quad (14)$$

is the received signal for user v freed from the interference caused by all other users $u, u \neq v$. The first term in (15) is the transmitted signal of the user of interest v and the second term contains the remaining interference and noise. In (15), $\Delta^{(i)} d_u[n] = d_u[n] - \hat{d}_u^{(i)}[n]$ is the difference between the transmitted symbol $d_u[n]$ and the symbol hypothesis from the output

of the previous stage $\hat{d}_u^{(i)}[n]$. The signal $\hat{r}_v^{(i)}[\ell]$ is the input signal to the RAKE receiver of PIC stage (i) . At the output of each RAKE receiver, one obtains the new symbol hypothesis

$$\begin{aligned} \hat{d}_v^{(i)}[n] &= \frac{1}{LK} \sum_{f=0}^{F-1} h_v^*[p_{f,v}] \\ &\quad \times \sum_{m=0}^{LK-1} \hat{r}_v^{(i)}[m + (nK - (p - p_{f,v}))L] \cdot c_v^*[m] \end{aligned} \quad (16)$$

which can be processed further in the next PIC stage or is detected in the last PIC stage.

3) **NB-PIC**: Another implementation version of the PIC algorithm is the NB-PIC. This algorithm is equivalent in performance to the WB-PIC, but operates at symbol level, in contrast to the chip-level operation of the WB-PIC. In the NB-PIC, the most probable transmitted signals are not recalculated, but the interferer estimates occurring from all P paths of all other $U - 1$ users are directly subtracted from the symbol estimates from the previous PIC stage. Assuming BPSK with hard decision function, in the i th PIC stage one obtains a decision variable $\hat{d}_v^{(i)}[n]$ for the interference reduced symbol of the v th user as (17), as shown at the bottom of the page. In (17), $\hat{I}_{v,f,u}^{(i)}[n]$ denotes the interference estimate from user u affecting the f th strongest path of user v . The data-dependent crosscorrelation $\varphi_{u,v}^{(\text{data})}(d_1, d_2, \ell)$ between the user of interest v and the interferer u with its symbols d_1 and d_2 is formed according to Table II. The even and odd crosscorrelation functions of the codes of the user v and the interferer u

$$\varphi_{u,v}^{(\text{even})}[\ell] = \varphi_{u,v}^{(1)}[\ell] + \varphi_{u,v}^{(2)}[\ell] \quad (18)$$

$$\varphi_{u,v}^{(\text{odd})}[\ell] = -\varphi_{u,v}^{(1)}[\ell] + \varphi_{u,v}^{(2)}[\ell] \quad (19)$$

are based on the partial crosscorrelations in (10) and (11).

$$\begin{aligned} I_{C_v}[n] &= \sum_{f=0}^{F-1} h_v^*[p_{f,v}] \sum_{\substack{u=0 \\ u \neq v}}^{U-1} \left(\sum_{p=p_{f,v}}^{P-1} h_u[p] \left(\varphi_{u,v}^{(1)}[p - p_{f,v}] \cdot d_u[n - 1] + \varphi_{u,v}^{(2)}[p - p_{f,v}] \cdot d_u[n] \right) \right. \\ &\quad \left. + \sum_{p=0}^{p_{f,v}-1} h_u[p] \left(\varphi_{u,v}^{(1)}[K - (p_{f,v} - p)] \cdot d_u[n] + \varphi_{u,v}^{(2)}[K - (p_{f,v} - p)] \cdot d_u[n + 1] \right) \right) \end{aligned} \quad (9)$$

$$\begin{aligned} \hat{d}_v^{(i)}[n] &= \hat{d}_v[n] - \sum_{f=0}^{F-1} h_v^*[p_{f,v}] \sum_{\substack{u=0 \\ u \neq v}}^{U-1} \underbrace{\left(\sum_{p=p_{f,v}}^{P-1} h_u[p] \varphi_{u,v}^{(\text{data})} \left(\hat{d}_u^{(i)}[n - 1], \hat{d}_u^{(i)}[n], p - p_{f,v} \right) \right.} \\ &\quad \left. + \sum_{p=0}^{p_{f,v}-1} h_u[p] \varphi_{u,v}^{(\text{data})} \left(\hat{d}_u^{(i)}[n], \hat{d}_u^{(i)}[n + 1], K - (p_{f,v} - p) \right) \right) \end{aligned} \quad (17)$$

$\hat{I}_{v,f,u}^{(i)}[n]$

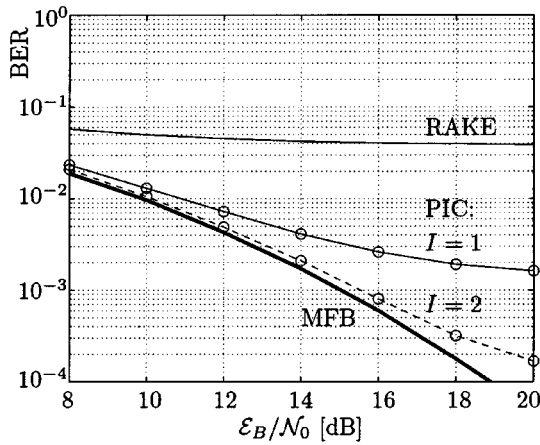


Fig. 3. BER versus $\mathcal{E}_B/\mathcal{N}_0$ for conventional PIC compared to the RAKE receiver and the MFB, $U = 16$, spreading factor $K = 16$.

B. Performance

The performance of the PIC was evaluated using Monte Carlo simulations for an MC-SS FCDMA system described in Section II-B. The data rate is 500 kb/s, the spreading gain is $K = 16$ and the symbol duration is $T_s = 2 \mu\text{s}$. Assuming the carrier frequency $f_c = 2.4$ GHz and a maximum velocity of $v = 3$ m/s for indoor applications, the coherence time becomes [23] $(\Delta t)_c = 9c/(16\pi v f_c) = 7.5$ ms, so that for a burst length of $N = 100$ symbols the channel can be seen as time invariant.

The channel model is based on measurements for frequencies around 2.2 GHz [24], [25]. The nonline-of-sight indoor model Pic2 is used which is characterized by a maximum delay of 500 ns and a coherence bandwidth of 2.5 MHz.

Fig. 3 shows for $U = 16$ users the uncoded BER of PIC with $I = 1$ and $I = 2$ stages compared to the performance of the RAKE receiver and the matched filter bound (MFB). It can be seen that the use of PIC leads to significant performance gains. A PIC scheme consisting of two stages can almost reach the MFB. The SNR degradation $D(P_B, U, I)$ gives the ratio of $\mathcal{E}_B/\mathcal{N}_0$ required to achieve a BER P_B for U users, referred to that which would be required to achieve the same P_B in the case of single-user communication [6]. In Fig. 4, the SNR degradation $D(P_B, U, I)$ depending on the number of users is depicted for $P_B = 10^{-3}$, $I = 1$ and $I = 2$, respectively. Even for a heavily loaded system, the two-stage PIC shows only a moderate SNR degradation of less than 1 dB.

C. Computational Complexity

The computational complexity is one of the most critical aspects in implementation. First, it determines whether an algorithm can be implemented at all in a real-time power-limited system. Second, it determines the costs in terms of number of necessary resources like operations per second in a digital signal processor (DSP) or number of gates and chip size in an ASIC. Third, it determines the power consumption which is critical in mobile devices.

In this paper, DSP-based implementations are the main focus. The scope of this section is to express the computational complexity for comparison of different implementation versions,

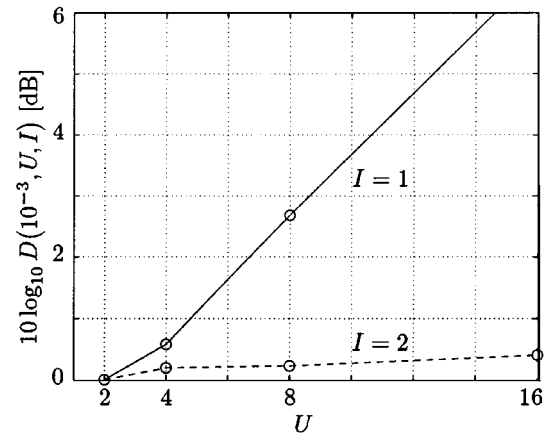


Fig. 4. SNR degradation $D(P_B, U, I)$ for PIC at $P_B = 10^{-3}$ for $I = 1$ and $I = 2$ iterations.

not to find an absolute complexity measure for a specific hardware. Therefore, a virtual processor is introduced that can perform one floating-point operation in one cycle. Operations are summation (ADD), multiplication (MUL), and multiply and add (MAC). Hence, the complex number operations are $\text{CADD} = 2 \text{ ADD} = 2 \text{ ops}$, $\text{CMUL} = 4 \text{ MUL} = 4 \text{ ops}$, and $\text{CMAC} = 4 \text{ MAC} = 4 \text{ ops}$. The loop overhead is assumed to be zero. The complexity O measured in number of operations (ops) is calculated for the previously introduced receiver structures. Because of the base station point of view in the uplink, O is the number of operations to detect one symbol for each of the U users.

1) *RAKE Receiver*: For the RAKE receiver in (7) with F RAKE fingers, spreading factor K , and oversampling factor L the complexity is given by

$$O_{\text{RAKE}}(U, F, K, L) = U \cdot F \cdot (KL \text{ CMAC} + 1 \text{ CMAC}) \quad (20)$$

$$= U \cdot F \cdot (4KL + 4) \text{ ops.} \quad (21)$$

2) *WB-PIC*: The complexity O is composed of the number of operations for resampling, channel emulation, calculation of the total interference, interference subtraction, and for the RAKE receiver. The effort for the tentative decision is neglected here. For BPSK modulation and a tentative hard decision device, the effort for resampling can be neglected as well because the possible signal constellation could be stored in a lookup table. For one WB-PIC stage in (12)–(16), the complexity is

$$O_{\text{WB,BPSK}}(U, F, P, K, L) = U \left(\overbrace{KL/UCADD}^{\text{Eq. (13)}} + \overbrace{PKL\text{CMAC}}^{\text{Eq. (12)}} + \overbrace{KL \text{ CADD} + KL \text{ CADD}}^{\text{Eq. (14)}} + \overbrace{F(KL \text{ CMAC} + 1 \text{ CMAC})}^{\text{Eq. (16)}} \right) \quad (22)$$

$$= 4U \left(KL \left(P + F + 1 + \frac{1}{2U} \right) + F \right) \text{ ops.} \quad (23)$$

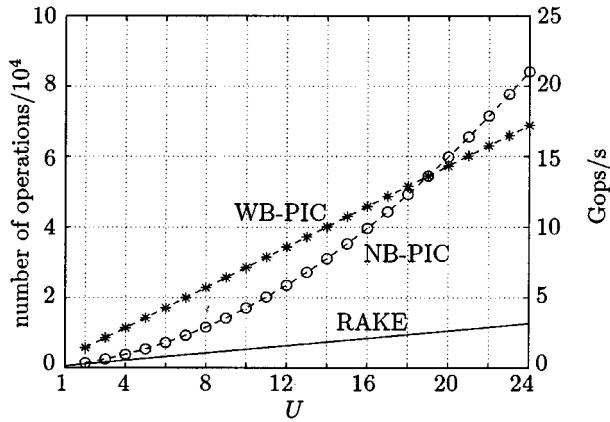


Fig. 5. Computational complexity of NB-PIC and WB-PIC for $K = 16$, $L = 2$, $F = P = 4$, $I = 2$ compared to RAKE.

The complexity grows linearly with the number of users U , the spreading gain K , the number of paths and fingers P , and F , and the oversampling factor L .

3) *NB-PIC*: The effort of one stage of the NB-PIC in (17) is

$$O_{\text{NB,BPSK}}(U, F, P) = UF((U-1)P \text{ CMAC} + 1 \text{ CMAC}) \quad (24)$$

$$= 4UF((U-1)P + 1) \text{ ops.} \quad (25)$$

It is assumed that the data-dependent crosscorrelations are stored in a lookup table. Thus, they do not have to be calculated from the partial crosscorrelations and from the data estimates. However, for soft decision and higher order modulation the calculation of the data-dependent crosscorrelations would be necessary. The complexity of the NB-PIC grows quadratically with the number of users U and number of considered paths/fingers P , F .

4) *Complexity Comparison*: Although the performance in terms of BER of both the NB-PIC and the WB-PIC is identical, their complexity behavior is different. The complexity of the WB-PIC depends linearly on U and F , P in (23), whereas the NB-PIC has quadratic complexity growth in (25). The computational complexity O_{RAKE} for the RAKE receiver and two stages of WB-PIC and NB-PIC is shown in Fig. 5 for different numbers of users U . It can be seen that the complexity of the NB-PIC is much smaller compared to that of the WB-PIC for low user load and low diversity. The complexity crossing point in Fig. 5 is given by

$$O_{\text{NB1,BPSK}}(U, F, P) < O_{\text{WB1,BPSK}}(U, F, P, K, L) \quad (26)$$

for (27), shown at the bottom of the page. Equation (27) gives a design rule for PIC implementation, specifying which algorithm should be selected for a specific system from a complexity point of view. However, it is obvious from Fig. 5 that the complexity of both methods is much higher than that for the RAKE

receiver, especially for high system loads. For example, a RAKE followed by two NB-PIC stages requires about 10 Gops/s for an uplink data rate of 250 kb/s as shown in Fig. 5. The question arises: "Can the complexity be reduced by a modification of the PIC algorithm without sacrificing the strong performance of PIC?" The next three sections provide answers to this question.

IV. REDUCED PIC (R-PIC)

In this section, one modification of the PIC is presented to reduce the computational complexity significantly. This method is the so-called R-PIC, which was first described by the authors in [26]. It works only for the NB-PIC as will be clear later. The NB-PIC in (17) tries to eliminate all interference terms occurring from all paths of all users according to the CDMA scheme. Instead of cancellation of all of these $(U-1) \cdot F \cdot P$ interference terms, the idea of the R-PIC is to cancel only a few, but strong interference terms. This leads to the elimination of a sizeable part of the interference power while the numerical effort is decreased to a significantly lower amount.

At first, an interference analysis is made to show how the interference term affecting user v , I_{C_v} in (9) is composed of and how efficient the R-PIC could be. It is assumed that the data symbols d_u are uniformly distributed and uncorrelated, hence, the crosscorrelations $\varphi_{u,v}^{(\text{even})}$, $-\varphi_{u,v}^{(\text{even})}$, $\varphi_{u,v}^{(\text{odd})}$, $-\varphi_{u,v}^{(\text{odd})}$ are uniformly distributed as well. With $v = 0, \dots, U-1$; $f = 0, \dots, F-1$; $p = 0, \dots, P-1$; $u = 0, \dots, U-2$ and the $U-1 \times U$ matrix

$$\nu = (\nu_0, \dots, \nu_{U-1}) \quad \text{with} \quad \nu_v = (0, 1, \dots, v-1, v+1, \dots, U-1)^T \quad (28)$$

the complete interference scenario is described by the $(U-1) \cdot F \cdot P \times U$ matrices (29)–(32), shown at the bottom of the next page. For the analysis of this interference scenario the elements of $\mathbf{S}$ are sorted row-wise in descending order. This leads to the new $(U-1) \cdot F \cdot P \times U$ matrices

$$\mathbf{S}^{(s)} = (\mathbf{S}_{\ell,v}^{(s)}) \quad (33)$$

$$\mathbf{I} = (\mathbf{I}_{\ell,v}) \quad (34)$$

with

$$S_{\ell,v}^{(s)} = \max_{\ell_1 \in \mathbb{I}_S(\ell,v)} S_{\ell_1,v} \quad (35)$$

$$I_{\ell,v} = \arg \max_{\ell_1 \in \mathbb{I}_S(\ell,v)} S_{\ell_1,v}$$

with the index set

$$\mathbb{I}_S(\ell,v) = \{0, \dots, (U-1) \cdot F \cdot P - 1\} \setminus \{I_{0,v}, \dots, I_{\ell-1,v}\}. \quad (36)$$

Hence, the element $S_{0,v}^{(s)}$ contains the value of the strongest mean interference term influencing the v th user and the element

$$U < \frac{P + \frac{KL}{F}(P + F + 1) + \sqrt{(P + \frac{KL}{F}(P + F + 1))^2 + \frac{2KL}{F}}}{2P} \quad (27)$$

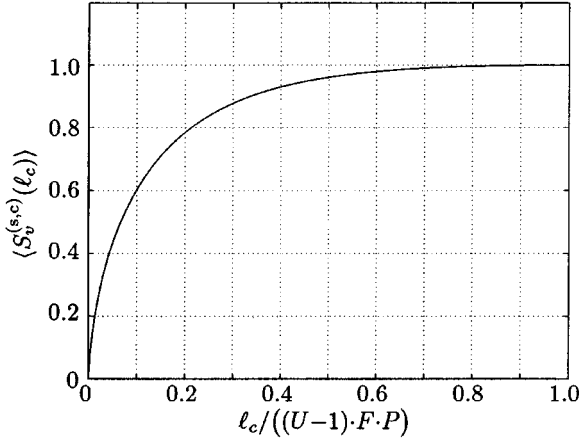


Fig. 6. Mean normalized cumulative interference $\langle S_v^{(s,c)}(\ell_c) \rangle$ versus number of considered interference terms for $U = 16, F = P = 4$.

$I_{0,v}^{(s)}$ contains the value of the index for S, F, P , and U belonging to $S_{0,v}^{(s)}$. The element $S_{1,v}^{(s)}$ contains the value of the second strongest mean interference term, etc.

For a better numerical evaluation of this interference analysis the cumulative mean interference power normalized to the total interference power is introduced by

$$S_v^{(s,c)}(\ell_c) = \frac{\sum_{\ell=0}^{\ell_c-1} S_{\ell,v}^{(s)}}{\sum_{\ell=0}^{(U-1)FP-1} S_{\ell,v}^{(s)}}. \quad (37)$$

Fig. 6 shows $\langle S_v^{(s,c)}(\ell_c) \rangle$ versus the number ℓ_c of considered interference terms for the CODIT channel Pic2 ($F = P = 4$), $U = 16$ and $v = 0$. It can be seen that a few strong interference terms comprise a large part of the total interference power, e.g., the strongest 20% interference terms accumulate about 70%–80% of the total interference power.

This interference analysis above leads to the following modification of the NB-PIC. Instead of the cancellation of all interference terms only the strongest ℓ_c interference terms are elim-

TABLE III
 $n_1(\ell), n_2(\ell)$, and $p(\ell)$ AS FUNCTIONS OF $P_{I_{\ell,v,v}} - F_{I_{\ell,v,v}}$

	$n_1(\ell)$	$n_2(\ell)$	$p(\ell)$
$P_{I_{\ell,v,v}} \geq F_{I_{\ell,v,v}}$	$n-1$	n	$P_{I_{\ell,v,v}} - F_{I_{\ell,v,v}}$
$P_{I_{\ell,v,v}} < F_{I_{\ell,v,v}}$	n	$n+1$	$K - (F_{I_{\ell,v,v}} - P_{I_{\ell,v,v}})$

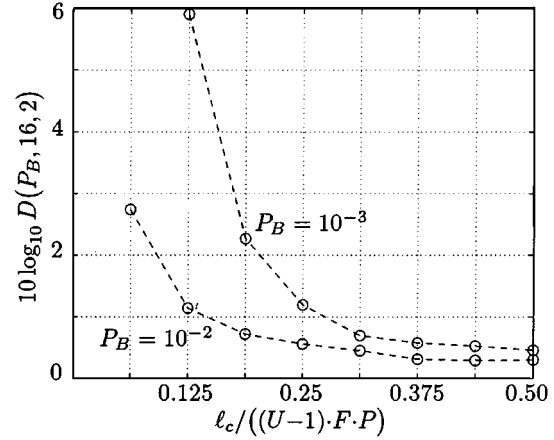


Fig. 7. SNR degradation of R-PIC versus number of considered interference terms for $U = 16, F = P = 4, I = 2$.

inated, hence, (38), as shown at the bottom of the page with $n_1(\ell), n_2(\ell)$, and $p(\ell)$ according to Table III.

For the R-PIC, Fig. 7 depicts the SNR degradation versus the number ℓ_c of considered interference terms for the specified BER $P_B = 10^{-2}, 10^{-3}$, $U = 16$ users and $I = 2$ R-PIC stages. It can be seen that the desired BERs can be reached for a low number of considered interference terms. The SNR degradation D referred to the single-user system depends on the target BERs and the number of cancelled interference terms and converges for a higher number of cancelled interference terms to the SNR degradation of the full-complexity PIC.

The complexity of the R-PIC is composed of the following parts.

$$S = (S_{\ell,v}) \quad \text{with } S_{f(U-1)P+uP+p,v} \\ = \begin{cases} \left| \text{Re} \left(h_v^*[p_{f,v}] \cdot h_{\nu_{u,v}}[p] \frac{\varphi_{\nu_{u,v},v}^{(\text{even})}[p-p_{f,v}] + \varphi_{\nu_{u,v},v}^{(\text{odd})}[p-p_{f,v}]}{2} \right) \right| & \text{for } p \geq p_{f,v} \\ \left| \text{Re} \left(h_v^*[p_{f,v}] \cdot h_{\nu_{u,v}}[p] \frac{\varphi_{\nu_{u,v},v}^{(\text{even})}[K-(p_{f,v}-p)] + \varphi_{\nu_{u,v},v}^{(\text{odd})}[K-(p_{f,v}-p)]}{2} \right) \right| & \text{for } p < p_{f,v} \end{cases} \quad (29)$$

$$F = (F_{\ell,v}) \quad \text{with } F_{f(U-1)P+uP+p,v} = p_{f,v} \quad (30)$$

$$P = (P_{\ell,v}) \quad \text{with } P_{f(U-1)P+uP+p,v} = p \quad (31)$$

$$U = (U_{\ell,v}) \quad \text{with } U_{f(U-1)P+uP+p,v} = \nu_{u,v} \quad (32)$$

$$\tilde{d}_v^{(i)}[n] = \tilde{d}_v[n] - \sum_{f=0}^{F-1} h_v^*[p_{f,v}] \sum_{\substack{\ell=0 \\ f(U-1)P \leq I_{\ell,v} < (f+1)(U-1)P}}^{\ell_c-1} h_{U_{I_{\ell,v},v}}[P_{I_{\ell,v},v}] \cdot \varphi_{U_{I_{\ell,v},v}}^{(\text{data})} \left(\tilde{d}_{U_{I_{\ell,v},v}}^{(i)}[n_1(\ell)], \tilde{d}_{U_{I_{\ell,v},v}}^{(i)}[n_2(\ell)], p(\ell) \right) \quad (38)$$

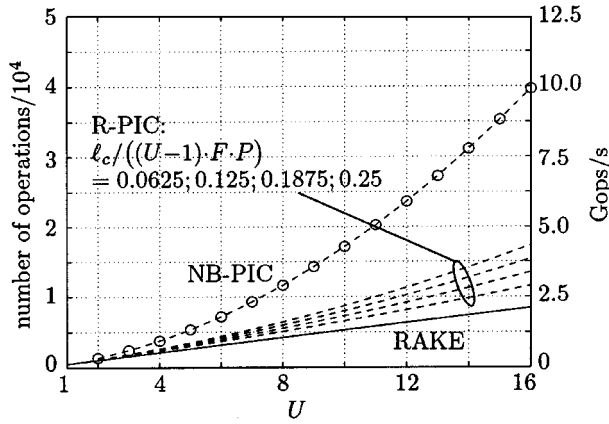


Fig. 8. Computational complexity of R-PIC for $K = 16$, $L = 2$, $F = P = 4$, $N = 100$, $I = 2$ compared to RAKE and NB-PIC.

- 1) The interference scenario evaluation in (29) requires per burst

$$O_{IS}(U, F, P) = U((U-1) \cdot F \cdot P(\text{CADD} + 2 \text{ CMUL})) \quad (39)$$

$$= 10U(U-1) \cdot F \cdot P \text{ ops.} \quad (40)$$

- 2) The determination of the index matrix I needs per burst

$$O_{IM}(U, F, P) = U \cdot (U-1) \cdot F \cdot P \log_2((U-1) \cdot F \cdot P) \text{ ops} \quad (41)$$

because $N \log_2 N$ operations are needed for indexing of a table with N entries [27].

- 3) The cancellation of ℓ_c interference terms in (38) requires per symbol

$$O_{RC21}(U, F, \ell_c) = U(\ell_c \cdot \text{CMAC} + F \cdot \text{CMAC}) \quad (42)$$

$$= 4U(\ell_c + F) \text{ ops.} \quad (43)$$

The first two steps have to be done only once per burst, before the first R-PIC stage is performed. This does not contribute an important amount to the total computational complexity of the R-PIC. Fig. 8 shows the complexity versus the number of users for an R-PIC with two stages and a different number of considered interference terms. Compared to the NB-PIC, the R-PIC leads to a considerable lower amount of operations. As an example, the additional PIC complexity can be reduced by 90% if only 6.25% of the interference terms are cancelled.

V. D-PIC

A. Concept

In the multistage PIC, all stages are identical time-invariant linear systems, considering only the part after the decision function $f_v^{(i)}$ in (6). If the symbol estimates at the input of two different PIC stages are identical for all users v , the output of both PIC stages would be the same. Therefore, all signal processing of stage (i) could be omitted, saving costly computing power. This is exploited in the D-PIC that is introduced in this paper. It will be shown that the complexity can be reduced significantly by omitting all unnecessary operations. After the conventional RAKE receiver stage and the first PIC stage (which must be

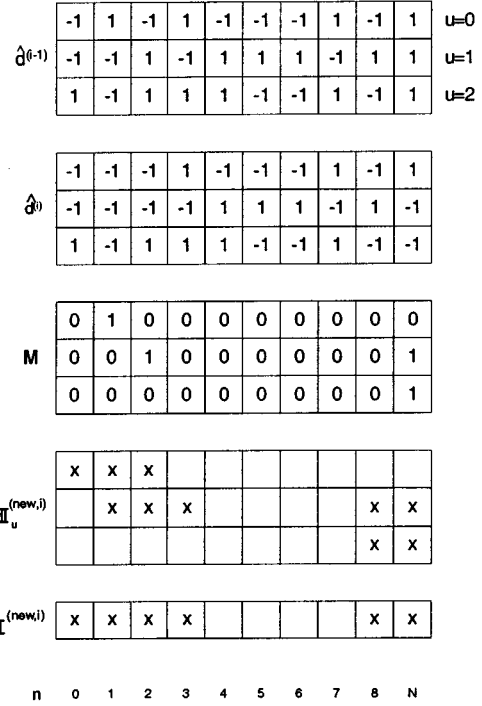


Fig. 9. Input symbols, masking matrix, and index sets of D-PIC.

a conventional or R-PIC stage), D-PIC stages can be applied. Following, necessary new operations of the i th D-PIC stage are identified by reusing symbol estimates and intermediate results from the previous PIC stage for all other values.

The alteration of a tentative symbol $\tilde{d}_v^{(i-1)}[n]$ at the input of stage $(i-1)$ to symbol $\tilde{d}_v^{(i)}[n]$ at the input of stage (i) has a certain influence on the interference estimates and symbol outputs of all users. This is pointed out in Fig. 9 analyzed using a masking matrix and sets of indexes. The element $M_{u,n}^{(i)}$ of the masking matrix $M^{(i)}$ indicates if symbol n of user u has changed from stage $(i-1)$ to stage (i) :

$$M^{(i)} = \begin{pmatrix} M_{u,n}^{(i)} \end{pmatrix} = \begin{cases} 1, & \text{for } \tilde{d}_u^{(i-1)}[n] \neq \tilde{d}_u^{(i)}[n] \\ 0, & \text{otherwise} \end{cases} \quad (44)$$

for $u = 0, \dots, U-1$ and $n = 0, \dots, N-1$. In Fig. 9, the input symbols $\tilde{d}$ for three users and a symbol burst length $N = 10$ are shown for the actual D-PIC stage (i) and the previous PIC stage $(i-1)$. Four symbols are changed, resulting in $M^{(i)}$.

For each user $u = 0, \dots, U-1$, the set of indexes with all symbol alterations from stage $(i-1)$ to stage (i) is defined

$$\mathbb{I}_u^{(\text{alt}, i)} = \bigcap_{n=0}^{N-1} \mathbb{I}_{u,n} \quad \text{with} \quad \mathbb{I}_{u,n} = \begin{cases} \{n\}, & \text{for } M_{u,n}^{(i)} = 1 \\ \emptyset, & \text{for } M_{u,n}^{(i)} = 0 \end{cases} \quad (45)$$

Each symbol alteration has an influence to other symbols of the same user and of other users, depending on the channel impulse length. For the system as defined in Section II with the channel impulse length being smaller than the symbol duration new sets

of indexes can be defined with all symbols in stage (i) which are affected by the input symbol alteration, thus resulting in (46) and (47), shown at the bottom of the page, which points to all indexes of the symbols which could be affected in stage (i). In Fig. 9, all elements of the respective index sets are marked.

B. Implementation of Differential WB-PIC

New calculation steps have only to be performed for the elements the according index set points to. For all other terms, the unchanged values from the previous stage can be taken.

Respreading and channel emulation to obtain the interference contribution of user v , $\hat{I}_v^{(i)}[l]$ in (12) is only necessary for all samples¹ $[(\ell/KL)] \in \mathbb{I}_v^{(\text{new},i)}$, otherwise the interference contribution from the previous stage $\hat{I}_v^{(i-1)}[l]$ can be used. The total interference calculation and subtraction from the received signal in (14) to obtain $\hat{r}_v^{(i)}[l]$ is only necessary for the samples $[(\ell/KL)] \in \mathbb{I}^{(\text{new},i)}$. The RAKE filter bank in (16) which results in $\tilde{d}_v^{(i)}[n]$ for all users has only to be activated for $n \in \mathbb{I}^{(\text{new},i)}$.

D-PIC can be extended to partial PIC as introduced in [6] by scaling of all terms from the previous stage by the quotient of the partial cancellation factors of stages ($i-1$) and (i).

C. Implementation of Differential NB-PIC

In (17), the interference estimate from user u affecting the path f of user v , $\hat{I}_{v,f,u}^{(i)}[n]$ has only to be calculated for $n \in \mathbb{I}_u^{(\text{new},i)}$, otherwise $\hat{I}_{v,f,u}^{(i-1)}[n]$ is still valid in stage (i). The calculation of $\tilde{d}_v^{(i)}[n]$ must be done for $n \in \mathbb{I}^{(\text{new},i)}$. Partial differential NB-PIC is also possible as described previously for the partial differential WB-PIC.

D. Complexity

The amount of complexity reduction is determined by the relative number of symbol alterations from one PIC stage to the next. Furthermore, the channel impulse length relative to the symbol duration and the number of active users affect the possible complexity reduction by the D-PIC since symbol changes of one user influence the symbols of all other users for the channel impulse length.

The computational effort of stage (i) depends mainly on the number of elements of the index sets defined in (45)–(47). Fol-

lowing, this computational effort is approximated in a worst case manner. It is assumed that $P_{B,D-PIC}$ is the sum of the BERs of stage ($i-2$) and stage ($i-1$). Considering a symbol burst length N , the index set $\mathbb{I}_u^{(\text{alt},i)}$ for $u = 0, \dots, U-1$ has approximately $P_{B,D-PIC} \cdot N$ elements. For a channel impulse length smaller than the duration of one symbol, three symbols of all users could be affected by one symbol change. Hence, $\mathbb{I}_u^{(\text{new},i)}$ has approximately $\min(3P_{B,D-PIC}, 1) \cdot N$ entries and $\mathbb{I}^{(\text{new},i)}$ has approximately $\min(3UP_{B,D-PIC}, 1) \cdot N$ entries.

The computational complexity of one D-PIC stage for the WB-PIC is

$$\begin{aligned} O_{D-WB,BPSK}(U, F, P, K, L) &= 4U \left(PKL \min(3P_{B,D-PIC}, 1) \right. \\ &\quad \left. + \left(KL + F \left(KL + \frac{1}{2U} \right) \right) \right) \\ &\quad \times \min(3UP_{B,D-PIC}, 1) \text{ ops} \end{aligned} \quad (48)$$

and for the differential NB-PIC

$$\begin{aligned} O_{D-NB,BPSK}(U, F, P, K, L) &= U(PFU \text{CMAC} \min(3P_{B,D-PIC}, 1) \\ &\quad + F((U-1)\text{CADD} + \text{CMAC}) \min(3UP_{B,D-PIC}, 1)) \end{aligned} \quad (49)$$

$$\begin{aligned} &= 4UF(UP \min(3P_{B,D-PIC}, 1) \\ &\quad + ((U-1)/2 + 1) \min(3UP_{B,D-PIC}, 1)) \text{ ops.} \end{aligned} \quad (50)$$

It is shown in (48)–(50) that the computational complexity of a D-PIC stage depends on the BERs $P_{B,D-PIC}$. For practical systems, uncoded BERs in the range of 10^{-1} – 10^{-3} (and lower) are strived for often. Therefore, the complexity of a D-PIC stage could be significantly small compared to a full PIC stage, especially for low $P_{B,D-PIC}$. Fig. 10 shows the complexity of one D-PIC stage depending on $P_{B,D-PIC}$.

VI. CONCATENATED R-PIC AND D-PIC

In the two previous sections, two different modifications of the PIC are given to reduce the computational complexity to a feasible amount. In this section, the concatenation of the R-PIC and D-PIC is proposed. The detector architecture is as follows: the initial bank of RAKE receivers is followed by two R-PIC

$$\mathbb{I}_u^{(\text{new},i)} = \bigcap_{n=0}^{N-1} \mathbb{I}_{u,n} \quad \text{with} \quad \mathbb{I}_{u,n} = \begin{cases} \{n, n+1\} & \text{for } M_{u,n}^{(i)} = 1 \quad \text{and} \quad n = 0 \\ \{n-1, n, n+1\} & \text{for } M_{u,n}^{(i)} = 1 \\ \{n-1, n\} & \text{for } M_{u,n}^{(i)} = 1 \quad \text{and} \quad n = N-1 \\ \emptyset & \text{for } M_{u,n}^{(i)} = 0 \end{cases} \quad (46)$$

$$\mathbb{I}^{(\text{new},i)} = \bigcap_{u=0}^{U-1} \mathbb{I}_u^{(\text{new},i)} \quad (47)$$

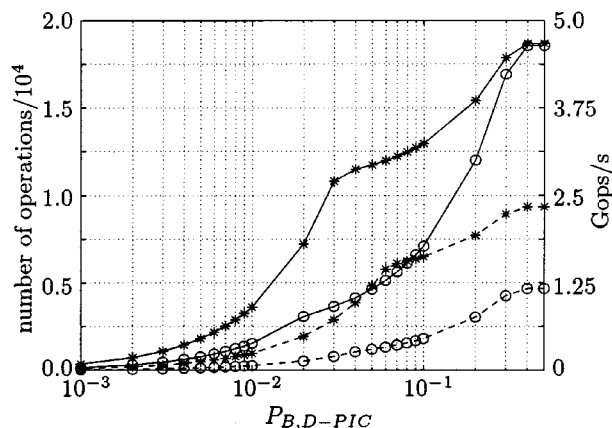


Fig. 10. Computational complexity of one D-PIC stage versus $P_{B,D-PIC}$ for $F = P = 4, K = 16, L = 2$: D-WB-PIC ($U = 8$: —*, $U = 16$: —*—). D-NB-PIC ($U = 8$: —o—, $U = 16$: —o—).

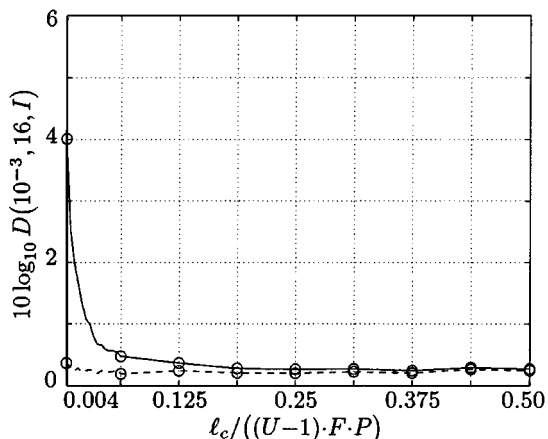


Fig. 11. SNR degradation of RD-PIC versus number of considered interference terms, $P_B = 10^{-3}, U = 16$: $I = 3$ (—o—), and $I = 4$ (—o—).

stages and two D-PIC stages. In the two R-PIC stages, the BER is improved in such a way that the D-PIC stages work very efficiently because only a few data symbols have to be estimated again, and the D-PIC stages are able to overcome the SNR degradations occurring from the R-PIC. Therefore, this concatenation leads to a performance very close to the MFB even if only a few cancelled interference terms are cancelled in the two R-PIC stages as shown in Fig. 11. It is worth noting, that for four stages the SNR degradation for $\text{BER} = 10^{-3}$ is almost independent from the number of cancelled interference terms. In Fig. 12, the computational complexity of that architecture is plotted for $\mathcal{E}_B/N_0 = 16$ dB. It can be seen that there exists a local minimum for the complexity depending on the number of considered interference terms. Hence, a configuration with near-optimum performance and minimum computational complexity can be found depending on the number of cancelled interference terms.

VII. CONCLUSION

A detailed complexity analysis shows the high computational effort for parallel interference cancellation in realistic scenarios in terms of data rate, spreading gain, number of users, and multipath components. To lower the complexity and, hence, the cost,

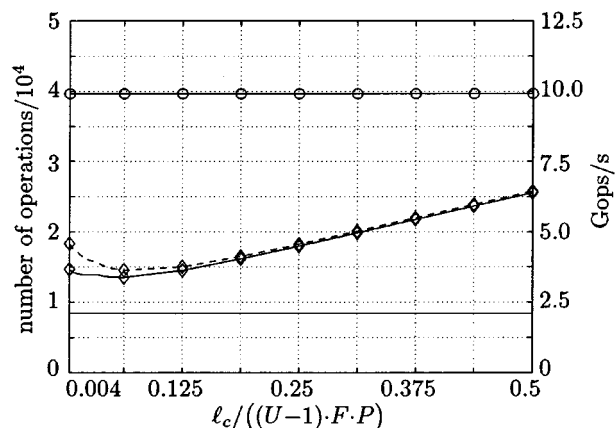


Fig. 12. Computational complexity of RD-PIC versus number of considered interference terms, $\mathcal{E}_B/N_0 = 16$ dB, $U = 16, F = P = 4, K = 16, L = 2, N = 100$: RD-PIC ($I = 3$: —o—, $I = 4$: —o—) compared to RAKE (—) and NB-PIC ($I = 2$: —o—).

two novel modifications of the parallel interference cancellation have been introduced in this paper.

The reduced PIC eliminates only a few strong interference terms, which comprise almost the whole interference power. The more that the total interference power is concentrated in a few interference terms, the better this idea works. The relative number of the considered interference terms is the parameter that influences the complexity reduction and the SNR degradation. For example, the additional PIC complexity to the single-user receiver can be decreased by up to 90% if only 6.25% of the interference terms are considered. However, a small performance degradation has to be taken into account, depending on the target BER.

D-PIC avoids unnecessary calculations by exploiting the multistage character of the PIC. This principle is very efficient if the BERs of the previous stages are still low ($5 \cdot 10^{-2}, \dots, 5 \cdot 10^{-3}$), so that only very few symbols have to be estimated again.

The drawbacks of R-PIC and D-PIC are SNR degradation and BER-depending efficiency, respectively. However, the concatenation of both algorithms offers further performance and complexity optimization. The RD-PIC architecture consisting of an initial RAKE bank, followed by two R-PIC stages and two D-PIC stages, leads to a performance very close to the single-user bound, whereas the computational complexity can be kept in the order of the conventional RAKE receiver.

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Achim Nahler (S'96–A'00) received the Dipl.-Ing. and Ph.D. degrees from Dresden University of Technology, Dresden, Germany, in 1995 and 2001, respectively.

The focus of his Ph.D. work was on the design of multicarrier-based spread-spectrum systems, in particular, the design of low-cost/low-complexity receiver architectures. In May 2001, he joined Systemonic AG, Dresden, Germany.



Ralf Irmer (S'98) received the M.Sc./Dipl.-Ing. degree, in 2000, from Dresden University of Technology, Dresden, Germany, where he is currently working toward the Ph.D. degree.

During 1995–1996, he studied at the University of Edinburgh, Edinburgh, U.K. He was with the Centre for Wireless Communications, Singapore, for his internship during 1998–1999. His research interests include multiuser detection and transmission, space-time signal processing, and efficient implementation of algorithms.



Gerhard Fettweis (S'84–M'90–SM'98) received the M.Sc./Dipl.-Ing. and Ph.D. degrees in electrical engineering from Aachen University of Technology (RWTH), Aachen, Germany, in 1986 and 1990, respectively.

From 1990 to 1991, he was a Visiting Scientist at the IBM Almaden Research Center, San Jose, CA, working on signal processing for disk drives. From 1991 to 1994, he was a Scientist with TCSI, Berkeley, CA, responsible for signal processor developments for mobile phones. Since September

1994, he has held the Mannesmann Mobilfunk Chair for Mobile Communications Systems at Dresden University of Technology, Dresden, Germany. Since 1999, he has been the CTO of Systemonic AG, Dresden, Germany, a startup spun out of TU Dresden, focusing on broadband wireless chip-sets.

Dr. Fettweis has been an elected member of the SSC Society's Administrative Committee since 1999 and was a member of the IEEE ComSoc Board of Governors from 1998 to 2000. He has been an Associate Editor of the *IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS II* and the *IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS* Wireless Communications Series. Over the years, he has organized and served on the program committees of numerous IEEE workshops and conferences.