NONCOHERENT MMSE INTERFERENCE SUPPRESSION FOR DS–CDMA

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***Abstract* — In this paper, a novel robust noncoherent receiver for MMSE interference suppression for DS–CDMA is proposed. The receiver consists of an MMSE filter and a decision–feedback differential detector (DF–DD). The performance of the proposed scheme is investigated analytically and by computer simulations. It is shown that the loss compared to coherent MMSE interference suppression is limited and can be made arbitrarily small by increasing the observation window used for calculation of the reference symbol of the DF–DD. For adjustment of the MMSE filter coefficients a noncoherent normalized least–mean–square (NC–NLMS) algorithm is proposed which is roboust against channel phase variations.**

1. Introduction

Due to its flexibility, direct–sequence code–division multiple–access (DS–CDMA) has gained high attention. Since the conventional matched filter receiver suffers from the near–far problem various optimum and suboptimum coherent multiuser detection schemes have been proposed (cf. [1] and references therein).

Coherent receivers require exact knowledge of the channel phase for optimum performance. Thus, complex carrier phase and frequency synchronization circuits (e.g. PLLs) have to be implemented in the receiver [2]. However, in fading environments or when low–cost local oscillators are employed acquisition and tracking of the carrier phase may be difficult or even impossible. Here, noncoherent receivers are a favorable choice. Various noncoherent receivers for DS–CDMA have been considered in the literature [3, 4, 5, 6, 7]. Although these detectors employ noncoherent detection, this is not exploited for adaptation of the receiver filters. The adopted conventional adaptive algorithms diverge even for relatively small phase variations and thus, despite the noncoherent detection, the resulting receiver is sensitive to frequency offset and phase noise.

Blind adaptive algorithms (e.g. [8, 9]) do not have this problem since they do not attempt to track the phase of the desired user. However, blind algorithms converge more slowly and cause a higher steady–state error than trained adaptive algorithms [8, 10].

Trained minimum mean–squared error (MMSE) receivers with noncoherent detection and *noncoherent adaptation* are discussed in [11, 10, 12]. These receivers employ conventional differential detectors (DDs) and for adaptation the signal autocorrelation matrix has to be estimated.

The proposed noncoherent MMSE receiver for DS–CDMA is related to the noncoherent linear equalizer reported in [13], i.e., the first stage of the receiver is a linear filter for interference suppression and the second stage is a decision–feedback differential detector (DF–DD) [14, 15], which uses a certain observation window to generate a noncoherent decision variable.

2. Channel Model and Receiver Structure

2.1. Channel Model

The received signal of an asynchronous DS–CDMA system with K users can be written as

$$r(t) = \sum_{\mu=1}^{K} r_\mu(t) + n(t), \tag{1}$$

where $n(t)$ denotes additive white Gaussian noise (AWGN) with double–sided power spectral density N_0. Assuming a non–dispersive channel, the received signal from user μ (equivalent complex baseband representation) is modeled as

$$r_\mu(t) = \sqrt{\frac{E_\mu}{T_c}}\, \mathrm{e}^{j\Theta_\mu(t)} \sum_{k=-\infty}^{\infty} b_\mu[k] s_\mu(t - kT - \tau_\mu), \tag{2}$$

where E_μ, $\Theta_\mu(t)$, and $\tau_\mu \geq 0$ are the energy, the phase, and the delay of the μth user, respectively. The kth transmitted M–ary phase–shift keying (MPSK) symbol of user μ is denoted by $b_\mu[k] \in \mathcal{A} = \{\mathrm{e}^{j2\pi\nu/M} | \nu \in \{0, 1, \ldots, M-1\}\}$. T_c and T are the chip and the symbol duration, respectively. The spreading waveform is given by $s_\mu(t) = \sum_{i=0}^{L-1} s_\mu^i \psi(t - iT_c)$, where $L = T/T_c$ is referred to as the spreading factor. Here, s_μ^i, $0 \leq i \leq L-1$, are the elements of the spreading sequence of user μ. In addition, we assume that the chip waveform $\psi(t)$ ($1/T_c \cdot \int_0^{T_c} |\psi(t)|^2 \mathrm{d}t = 1$) has duration T_c and that the spreading sequences are normalized to $\sum_{i=0}^{L-1} |s_\mu^i|^2 = 1$.

Since a noncoherent receiver is employed, differential encoding at the transmitter is necessary, i.e., the MPSK symbols $b_\mu[k]$ are generated from differential (MDPSK) symbols $a_\mu[k]$ via $b_\mu[k] = a_\mu[k] \cdot b_\mu[k-1]$ [2]. Here, we assume that the MDPSK sequences of all users are mutually independent, independent identically distributed (i.i.d) sequences.

Without loss of generality we assume that user 1 is the desired user. Furthermore, perfect chip and symbol synchronization is presumed for this user, i.e., $\tau_1 = 0$ is valid. Thus, the ith sample at the output of the chip matched filter $\frac{1}{T_c}\psi^*(-t)$ in the kth symbol interval can be written as

$$r^i[k] = \sum_{m=1}^{2K-1} \mathrm{e}^{j\Theta^i_m[k]} \bar{b}_m[k] p^i_m + n^i[k]. \qquad (3)$$

Here, we model the K asynchronous users as $2K-1$ synchronous virtual users with equivalent spreading sequences p^i_m, $0 \leq i \leq L-1$, $1 \leq m \leq 2K-1$, as proposed in [16]. The MPSK symbol $\bar{b}_m[k]$ of the mth synchronous virtual user in the kth symbol interval is given by

$$\bar{b}_m[k] = \begin{cases} b_\mu[k], & m = 2\mu - 1,\ 1 \leq \mu \leq K \\ b_\mu[k-1], & m = 2\mu - 2,\ 2 \leq \mu \leq K \end{cases}. \qquad (4)$$

The equivalent spreading sequence of the desired user is given by $p^i_1 = \sqrt{E_1/T_c}\, s^i_1$, $0 \leq i \leq L-1$. The equivalent spreading sequences p^i_m, $0 \leq i \leq L-1$, $2 \leq m \leq 2K-1$, of the remaining $2K-2$ synchronous virtual users depend on the spreading sequences s^i_μ, $0 \leq i \leq L-1$, the energies E_μ, and the delays τ_μ, $2 \leq \mu \leq K$, of the remaining $K-1$ asynchronous users and the chip waveform. For the exact mathematical definition of p^i_m, $0 \leq i \leq L-1$, $2 \leq m \leq 2K-1$, we refer to [16, Section II]. $n^i[k]$ and $\Theta^i_m[k]$ are the discrete–time channel noise and phase of the ith sample of the matched filter output in the kth symbol interval.

For derivation of the noncoherent MMSE receiver we assume that the phases of all users are unknown but constant, i.e., $\Theta^i_m[k] = \Theta_m$. For constant phases, (3) may be rewritten as

$$\boldsymbol{r}[k] = \sum_{m=1}^{2K-1} \mathrm{e}^{j\Theta_m} \bar{b}_m[k] \boldsymbol{p}_m + \boldsymbol{n}[k], \qquad (5)$$

where the definitions

$$\boldsymbol{r}[k] \stackrel{\triangle}{=} [r^0[k]\ r^1[k]\ \ldots\ r^{L-1}[k]]^T, \qquad (6)$$

$$\boldsymbol{p}_m \stackrel{\triangle}{=} [p^0_m\ p^1_m\ \ldots\ p^{L-1}_m]^T, \qquad (7)$$

$$\boldsymbol{n}[k] \stackrel{\triangle}{=} [n^0[k]\ n^1[k]\ \ldots\ n^{L-1}[k]]^T \qquad (8)$$

($[\cdot]^T$ denotes transposition) are used.

2.2. *Receiver Structure*

The receiver structure is depicted in Fig. 1. For simplicity, we restrict our attention to receiver filters of length L. The (complex–conjugated) coefficients $c^i[k]$, $0 \leq i \leq L-1$, of the receiver filter are collected in a vector $\boldsymbol{c}[k]$. The filter output signal $q[k] = \boldsymbol{c}^H[k]\boldsymbol{r}[k]$ can be written as

$$q[k] = \sum_{m=1}^{2K-1} \mathrm{e}^{j\Theta_m} g_m[k] \bar{b}_m[k] + \boldsymbol{c}^H[k]\boldsymbol{n}[k], \qquad (9)$$

($[\cdot]^H$ denotes Hermitian transposition) with

$$g_m[k] \stackrel{\triangle}{=} \boldsymbol{c}^H[k]\boldsymbol{p}_m. \qquad (10)$$

For estimation of the transmitted MDPSK symbol $a_1[k]$ of the desired user a DF–DD is employed, i.e., the complex plane is divided into M sectors corresponding to the possible values of $a_1[k]$ and the estimate $\hat{a}_1[k]$ is determined by the sector into which the decision variable

$$d[k] = q[k] \cdot (q^{N,\alpha}_{\mathrm{ref}}[k-1])^* \qquad (11)$$

falls ($(\cdot)^*$ denotes complex conjugation). Here, $q^{N,\alpha}_{\mathrm{ref}}[k-1]$ denotes the reference symbol, which is defined as

$$q^{N,\alpha}_{\mathrm{ref}}[k-1] \stackrel{\triangle}{=} \beta_0 \sum_{\nu=1}^{N-1} \alpha^{\nu-1} q[k-\nu] \prod_{\mu=1}^{\nu-1} \hat{a}_1[k-\mu], \qquad (12)$$

with the normalization constant

$$\beta_0 \stackrel{\triangle}{=} \begin{cases} 1/(N-1), & N \geq 2,\ \alpha = 1 \\ \frac{1-\alpha}{1-\alpha^{N-1}}, & N \geq 2,\ 0 \leq \alpha < 1 \end{cases}. \qquad (13)$$

Here, N and α are the number of received symbols used for calculation of $d[k]$ and a forgetting factor, respectively. $q^{N,\alpha}_{\mathrm{ref}}[k-1]$ is a generalization of the reference symbols proposed in [14, 15]. The two special cases $\alpha = 1$ ($N \geq 2$) and $N \to \infty$ ($0 \leq \alpha < 1$) will be referred to as *nonrecursive*[1] [14] and *recursive* [15] reference symbols, respectively. From a practical point of view, the recursive case is most interesting since the corresponding reference symbol can be calculated in a simple manner: $q^{\infty,\alpha}_{\mathrm{ref}}[k-1] = (1-\alpha)q[k-1] + \alpha\hat{a}_1[k-1]q^{\infty,\alpha}_{\mathrm{ref}}[k-2]$. Last, it is worth mentioning that for $N = 2$ and/or $\alpha = 0$ a conventional DD [2] results.

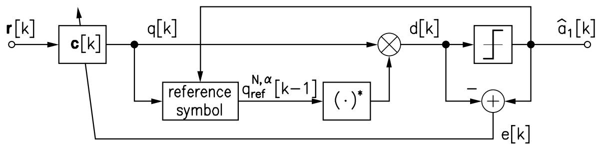

Figure 1: Block diagram of the considered noncoherent MMSE receiver.

3. Noncoherent MMSE Interference Suppression

3.1. *Optimum Filter Coefficients*

The optimum filter coefficients can be obtained by minimizing the variance $\sigma^2_e(\boldsymbol{c})$ of the error $e[k] = a_1[k] - d[k]$ (cf. Fig. 1), where $\hat{a}_1[k] = a_1[k]$ is assumed.

It is shown in [17] that $\sigma^2_e(\boldsymbol{c})$ (in this section $\boldsymbol{c}[k] = \boldsymbol{c}$, $\forall\, k$, is assumed) can be expressed as

$$\sigma^2_e(\boldsymbol{c}) \stackrel{\triangle}{=} 1 - \boldsymbol{c}^H \boldsymbol{p}_1 \boldsymbol{p}^H_1 \boldsymbol{c} \left(2 - (1-\beta_1)\boldsymbol{c}^H \boldsymbol{\Phi} \boldsymbol{c}\right) + \beta_1 (\boldsymbol{c}^H \boldsymbol{\Phi} \boldsymbol{c})^2 + \beta_0^2 |\boldsymbol{g}^H_o \boldsymbol{g}_e|^2, \qquad (14)$$

[1] The reference symbol according to (12) is generated in a nonrecursive way for arbitrary α, of course. However, for the sake of compatibility with the literature and due to the lack of a better denotation, in this paper, *nonrecursive reference symbol* only refers to the case $\alpha = 1$.

where the definitions

$$\boldsymbol{\Phi} \stackrel{\triangle}{=} \sum_{m=1}^{2K-1} \boldsymbol{p}_m \boldsymbol{p}_m^H + \sigma_n^2 \boldsymbol{I}_{L\times L}, \quad (15)$$

$$\boldsymbol{g}_o \stackrel{\triangle}{=} [g_3 \;\; g_5 \;\; \dots \;\; g_{2K-1}]^H, \quad (16)$$

$$\boldsymbol{g}_e \stackrel{\triangle}{=} [g_2 \;\; g_4 \;\; \dots \;\; g_{2K-2}]^H, \quad (17)$$

$$\beta_1 \stackrel{\triangle}{=} \begin{cases} 1/(N-1), & N \geq 2,\, \alpha = 1, \\ \frac{(1-\alpha)(1+\alpha^{N-1})}{(1+\alpha)(1-\alpha^{N-1})}, & N \geq 2,\, 0 \leq \alpha < 1 \end{cases} \quad (18)$$

($\boldsymbol{I}_{L\times L}$ denotes the $L \times L$ identity matrix) are used and $\sigma_n^2 = N_0/T_c$ denotes the noise variance. Notice that $\beta_1 \in (0,\, 1]$ increases with decreasing N and/or α. $\sigma_e^2(\boldsymbol{c})$ may be regarded as a *noncoherent* cost function because it does not depend on the phase of the desired user. First, we consider the case when the last term in (14) is absent which is true for $\beta_0 = 0$ ($N \to \infty$, $\alpha = 1$). It is also true for a synchronous CDMA system since then $\boldsymbol{g}_o = \boldsymbol{g}_e = \boldsymbol{0}_{K-1}$ ($\boldsymbol{0}_{K-1}$ is the all zero vector with $K-1$ rows) is valid. Minimizing (14) it can be shown that the optimum filter setting is given by

$$\boldsymbol{c}_{\text{opt}} \stackrel{\triangle}{=} \gamma \cdot \mathrm{e}^{j\phi} \cdot \boldsymbol{\Phi}^{-1} \boldsymbol{p}_1, \quad (19)$$

where γ is defined as

$$\gamma \stackrel{\triangle}{=} 1/\sqrt{\beta_1 + (1-\beta_1)\boldsymbol{p}_1^H \boldsymbol{\Phi}^{-1} \boldsymbol{p}_1}. \quad (20)$$

The phase ϕ is arbitrary which underlines the noncoherent character of the receiver.

For asynchronous CDMA systems with $\beta_0 > 0$ the term $\beta_0^2 |\boldsymbol{g}_o^H \boldsymbol{g}_e|^2$ appears in (14) because of the correlatedness of $q[k]$ and $q[k-1]$ and a closed–form solution for the optimum coefficient vector cannot be given in general. However, the solution given by (19) is still a very good approximation for the optimum solution since the vectors $\boldsymbol{g}_o$ and $\boldsymbol{g}_e$ only contain gain coefficients g_m, $2 \leq m \leq 2K-1$, which do not belong to the desired user and thus, will be suppressed by the MMSE filter in any case ($K \leq L$ is assumed).

3.2. *Signal–to–Interference Ratio*

For the following, we restrict ourselves to the case where the last term in (14) is absent. Thus, all results obtained are exact for synchronous CDMA and asynchronous CDMA with $\beta_0 = 0$ ($N \to \infty$ and $\alpha = 1$), whereas they are approximations for asynchronous CDMA with $\beta_0 > 0$.

It can be shown easily that the decision variable $d[k]$ (cf. (11)) may be decomposed into

$$d[k] = |g_1|^2 a_1[k] + z[k], \quad (21)$$

where $z[k]$ consists of noise and interference. Like in the coherent case [16], we may define the SIR as the ratio of the desired signal power to the power of interference and noise. This leads to [17]

$$\mathrm{SIR_{NC}} \stackrel{\triangle}{=} \frac{\mathcal{E}\{||g_1|^2 a_1[k]|^2\}}{\mathcal{E}\{|z[k]|^2\}}$$

$$= \frac{(\boldsymbol{p}_1^H \boldsymbol{\Phi}^{-1} \boldsymbol{p}_1)^2}{\beta_1 + (1-\beta_1)\boldsymbol{p}_1^H \boldsymbol{\Phi}^{-1} \boldsymbol{p}_1 - (\boldsymbol{p}_1^H \boldsymbol{\Phi}^{-1} \boldsymbol{p}_1)^2} \quad (22)$$

($\mathcal{E}\{\cdot\}$ denotes expectation). The SIR for coherent MMSE interference suppression is given by [16]

$$\mathrm{SIR_C} = \frac{\boldsymbol{p}_1^H \boldsymbol{\Phi}^{-1} \boldsymbol{p}_1}{1 - \boldsymbol{p}_1^H \boldsymbol{\Phi}^{-1} \boldsymbol{p}_1}. \quad (23)$$

Using this in (22), the SIR for noncoherent MMSE interference suppression can be expressed as

$$\mathrm{SIR_{NC}} = \frac{\mathrm{SIR_C^2}}{\beta_1 + (1+\beta_1)\mathrm{SIR_C}}. \quad (24)$$

(24) shows that $\mathrm{SIR_{NC}}$ decreases monotonically with increasing β_1, i.e., for the maximum value $\beta_1 = 1$, which results if a simple conventional DD ($N = 2$ and/or $\alpha = 0$) is used, the proposed noncoherent receiver has the smallest SIR. For $\beta_1 \to 0$, i.e., if an infinite observation window for calculation of $q_{\text{ref}}^{N,\alpha}[k-1]$ ($N \to \infty$ and $\alpha = 1$) is used, $\mathrm{SIR_{NC}} = \mathrm{SIR_C}$ results. Thus, we can expect that in this asymptotic case the performance of the noncoherent MMSE receiver approaches that of its coherent counterpart.

For $\mathrm{SIR_C} \gg 1$, (24) can be approximated by

$$\mathrm{SIR_{NC}} \approx \frac{1}{1+\beta_1} \mathrm{SIR_C}, \quad (25)$$

i.e., on a logarithmic scale the noncoherent receiver causes a loss of $10\log_{10}(1+\beta_1)$ dB. The maximum loss of 3 dB results for $N = 2$ and/or $\alpha = 0$ (conventional DD).

From (24) and (25) it is obvious that the maximum loss of the noncoherent receiver compared to the coherent one is limited. Since it is well known that the coherent MMSE receiver is near–far resistant [16], we can conclude that also the noncoherent MMSE receiver is near–far resistant.

3.3. *Bit Error Rate (BER) for BDPSK and QDPSK*

Approximating the filter output symbols $q[k]$ by Gaussian random variables and using the same approach as in [14], it can be shown that the BER of the genie–aided receiver (i.e., $\hat{a}_1[k-\nu] = a_1[k-\nu]$, $1 \leq \nu \leq N-1$, in (12)) is given by

$$P_b^{\text{genie}} = \mathrm{Q_M}(a,b) - \frac{1}{2} \cdot \mathrm{I_0}(ab) \cdot \exp\left(-\frac{a^2+b^2}{2}\right), \quad (26)$$

where $\mathrm{Q_M}(\cdot,\cdot)$ and $\mathrm{I_0}(\cdot)$ denote the Marcum Q–function and the zeroth order modified Bessel function of the first kind, respectively [2]. Furthermore, a and b are given by

$$\begin{Bmatrix} a \\ b \end{Bmatrix} = \sqrt{\left(\frac{1+\beta_1}{2\beta_1} \mp \frac{1}{\sqrt{\beta_1}} \Re\{C\}\right) \mathrm{SIR_C}} \quad (27)$$

with $C = 1$ and $C = \mathrm{e}^{j\pi/4}$ for BDPSK and QDPSK, respectively.

With a realizable receiver, P_b^{genie} cannot be achieved because of errors in the feedback symbols, of course. However, it is well known that, in general, feedback errors increase BER of DF–DD by approximately a factor of two (cf. [14]). Therefore, we can approximate the BER of a realizable noncoherent MMSE receiver by

$$P_b \approx \begin{cases} P_b^{\mathrm{genie}}, & \beta_1 = 1 \\ 2 \cdot P_b^{\mathrm{genie}}, & 0 < \beta_1 < 1 \end{cases}. \tag{28}$$

Note that for $\beta_1 = 1$ ($N = 2$ and/or $\alpha = 0$) no feedback is necessary and thus, the BER has not to be doubled.

4. Filter Adaptation

For adaptive adjustment of the filter coefficients, we propose a noncoherent normalized least–mean–square (NC–NLMS) algorithm.

4.1. NC–NLMS Algorithm

For derivation of the NC–NLMS algorithm a similar technique as in [13] can be used. The resulting NC–NLMS algorithm is given by

$$\boldsymbol{c}[k+1] = \boldsymbol{c}[k] + \delta_{\mathrm{LMS}}[k] e^*[k] \boldsymbol{u}[k], \tag{29}$$

with $\boldsymbol{u}[k] \overset{\triangle}{=} \boldsymbol{r}[k](q_{\mathrm{ref}}^{N,\alpha}[k-1])^*$, $e[k] = \hat{a}_1[k] - d[k]$, and time–varying step size

$$\delta_{\mathrm{LMS}}[k] = \frac{\tilde{\delta}_{\mathrm{LMS}}}{a_{\mathrm{LMS}} + \boldsymbol{u}^H[k]\boldsymbol{u}[k]}, \tag{30}$$

where $\tilde{\delta}_{\mathrm{LMS}}$ is the adaptation constant and $a_{\mathrm{LMS}} > 0$ avoids division by zero.
For the NC–NLMS algorithm the initialization of $\boldsymbol{c}[k]$ is an important issue. The initial setting $\boldsymbol{c}[0] = \boldsymbol{0}_L$ cannot be used since in this case $\boldsymbol{c}[k] = \boldsymbol{0}_L$, $\forall\, k$, results. Therefore, we propose to adopt $c^{\lfloor L/2 \rfloor}[0] = 0.05$ ($\lfloor x \rfloor$ denotes the largest natural number equal or smaller than x) and $c^i[0] = 0$, $i \neq \lfloor L/2 \rfloor$.

4.2. Speed of Convergence

A synchronous CDMA system with $K = 15$ equal power users is considered. For spreading Gold sequences of length $L = 15$ are adopted. Furthermore, $10\log_{10}(E_1/N_0) = 15$ dB is valid. Figs. 2a), b), and c) show the SIR (averaged over 500 adaptation processes) vs. the number of iterations for the conventional NLMS and the NC–NLMS algorithm (recursive reference symbol) for constant phase, a normalized frequency offset of $\Delta fT = 0.005$, and phase noise (Wiener process) with variance $\sigma_S^2 = 0.005$ (per symbol interval T), respectively. $\tilde{\delta}_{\mathrm{LMS}} = 0.5$ and $a_{\mathrm{LMS}} = 1.0$ are adopted for both conventional NLMS and NC–NLMS algorithm. The dashed lines in Fig. 2 correspond to the maximum SIR for a fixed MMSE filter and constant phase (cf. (22), (23)). For constant phase the conventional NLMS algorithm achieves the highest SIR. The speed of convergence of the NC–NLMS algorithm with $\alpha = 0.4$ and $\alpha = 0.8$ is similar to that of the conventional NLMS algorithm, whereas the NC–NLMS algorithm with $\alpha = 0.0$ converges more slowly since the reference symbol $q_{\mathrm{ref}}^{\infty,0.0}[k-1]$ is very noisy in this case. For frequency offset and phase noise, the conventional NLMS algorithm does not converge since it is not able to track the phase variations. For frequency offset the SIR shows a periodic behavior with period 200 ($200 \cdot \Delta fT = 1$). The NC–NLMS algorithm, however, also converges if the phase is time–variant.

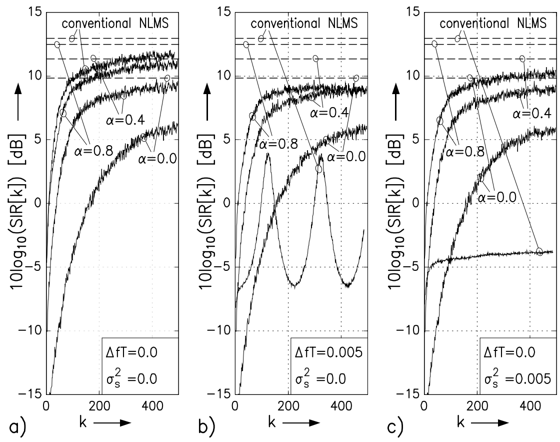

Figure 2: SIR vs. number of iterations k for a) constant phase, b) frequency offset, and c) phase noise.

5. Simulation Results

In our simulations, again Gold sequences of length $L = 15$ are used for spreading and the proposed NC–NLMS algorithm is employed for filter adjustment. All active users transmit with equal energy.
Figs. 3a) and b) show BER vs. $10\log_{10}(E_b/N_0)$ for asynchronous[2] QDPSK transmission (energy per bit $E_b = E_1/2$, $K = 10$) with nonrecursive and recursive reference symbol, respectively. Although in this case the MMSE filter obtained from (19) and used for calculation of the theoretical BER curves is only an approximation, theory and simulation coincide very well. The performance of the noncoherent receiver improves as N (or α) increase, while the loss compared to the respective single user bound is almost independent of N (α).
In Fig. 4, the performance of the proposed receiver in the presence of phase noise (Wiener process) is shown for asynchronous QDPSK transmission ($K = 10$, $10\log_{10}(E_b/N_0) = 10$ dB). It can be observed that the sensitivity to phase variations increases with increasing N (α). On the other hand, in the absence of phase variations power efficiency increases with increasing N (α). Thus, there is a trade–off between

[2] For all simulations presented here, a chip–synchronous system with $\tau_\nu = v_\nu \cdot T_c$, $0 \leq v_\nu \leq L-1$, $2 \leq \nu \leq K$, is used. In particular, $v_2 = 5$, $v_3 = 2$, $v_4 = 3$, $v_5 = 8$, $v_6 = 1$, $v_7 = 6$, $v_8 = 7$, $v_9 = 4$, $v_{10} = 9$, is valid.

power efficiency for constant phase and robustness against phase variations, i.e., in practice, N and α have to be chosen carefully taking into account the maximum expected phase variations of the channel.

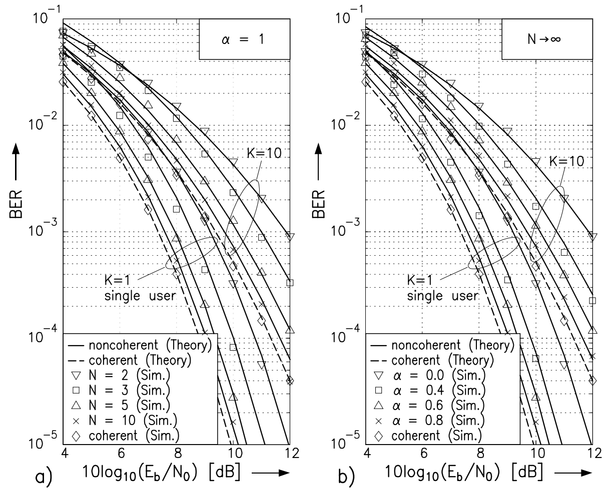

Figure 3: BER for QDPSK vs. $10\log_{10}(E_b/N_0)$ for a) nonrecursive and b) recursive reference symbol.

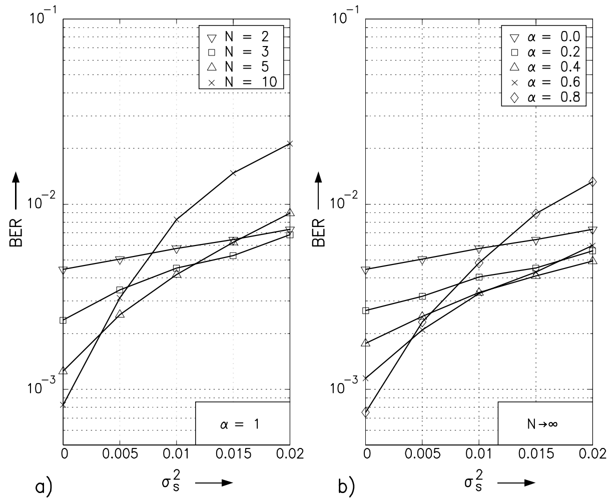

Figure 4: BER for QDPSK vs. phase noise variance σ_S^2 (per symbol interval T) for a) nonrecursive and b) recursive reference symbol.

6. Conclusions

In this paper, a novel noncoherent MMSE receiver for DS–CDMA has been proposed. It is shown that the loss compared to coherent MMSE interference suppression is limited and can be adjusted via the observation window used for generation of the reference symbol for the DF–DD. For long observation windows the performance of the coherent receiver is approached. On the other hand, for smaller observation windows the receiver is more robust against phase noise and frequency offset. The proposed NC–NLMS algorithm also converges to the desired solution in the presence of phase variations, whereas its conventional (coherent) counterpart diverges.

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