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Multiple Antenna Interference Cancellation for M-PSK Signals Using the SAGE Algorithm

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Abstract— **This paper presents a novel iterative receiver for mobile handsets using multiple antennas in the GSM/EDGE system. The receiver is derived based on the space alternating-generalized expectation-maximization (SAGE) method. It performs joint channel estimation, spatial diversity combining and successive co-channel interference (CCI) cancellation. The channel impulse response of all users are jointly estimated, while the users' transmitted symbols are detected successively. The receiver has a tractable complexity that increases linearly with the number of interferers. Two practically relevant applications are investigated by means of Monte Carlo simulations: single antenna interference cancellation (SAIC) in the GSM system and dual antenna interference cancellation (DAIC) in the EDGE system. Simulations using the TU3 channel model show that the new proposed scheme achieves a significant performance gain compared to the conventional receiver at low signal-to-interference ratios (SIRs).**

I. Introduction

Cellular mobile communication systems, like the GSM/EDGE system, exploit efficiently the band-limited spectrum by applying effective frequency reuse patterns. However, nearby cells using the same carrier frequencies introduce co-channel interference (CCI). This interference prevents the application of aggressive frequency-reuse factors, and thus constitutes a major limiting factor for the capacity of the network. In order to mitigate CCI, the receivers are required to implement CCI cancellation. In the base station (BS), large antenna arrays are usually deployed and digital array processing is used to mitigate the CCI. Recently, multiple-element antennas for the mobile receivers have been considered in the standardization [1] to allow for more effective CCI cancellation. However due to the size limitation only a dual antenna system can be considered in practice. Thus the antenna system shall be coupled with a joint demodulation method to efficiently mitigate the interference.

In receivers applying joint demodulation methods the signals of the desired user and the interferers are jointly detected using their individual channel impulse responses (CIRs). The optimum joint demodulator was presented first by Van Etten in [2] for known channel conditions. It was later extended

to the unknown channel case in [3]. In [3] the multiuser detector is coupled with a multichannel estimator, that jointly estimates the channels of all users based on the users' training sequences. The computational complexity of these receivers is too high for practical implementation. Low complexity approaches have been proposed based on a joint reduced-state sequence estimation equalizer combined with a prefilter [4] or successive interference cancellation (SIC) [5].

In this paper, a new low complexity iterative receiver is proposed that performs channel estimation, co-channel interference (CCI) cancellation, spatial diversity processing and single-user sequence detection in multiuser M-level phase shift keying (M-PSK) systems. The receiver is derived within the space-alternating generalized expectation-maximization (SAGE) framework [6]. The EDGE and linearized GSM modulation formats are two variants of M-PSK, which use the linearized Gaussian minimum shift-keying (GMSK) signal pulse shape [7] and an additional incremental phase rotation of each symbol. While the receiver is derived for the unified M-PSK modulation transmission, its performance is evaluated in the context of GSM/EDGE systems.

The remainder of the paper is organized as follows. The system model under investigation is introduced in Section II. In Section III, the SAGE framework for multiuser detection is briefly described first. Then the novel receiver is derived using the SAGE algorithm and its computational complexity is assessed. In Section IV, the performance of the receiver in terms of bit error rate (BER) is presented. Concluding remarks follow in Section V.

The following notation is used in the paper: $[\boldsymbol{x}]_m$ is the m-th element of the vector $\boldsymbol{x}$, and $[\boldsymbol{X}]_{mn}$ denotes the element in the m-th row and n-th column of the matrix $\boldsymbol{X}$. Moreover $\lfloor x \rfloor$ denotes the largest integer smaller than or equal to x, $\Re\{x\}$ is the real part of the complex scalar x, $\mathrm{diag}\{x_1, \cdots, x_N\}$ is the diagonal matrix with diagonal entries $x_1, \cdots, x_N$ and $\mathrm{Tr}\{\boldsymbol{X}\}$ is the trace of the square matrix $\boldsymbol{X}$.

II. System Model

We consider a downlink scenario: each of the $(K+1)$ active transmit BSs is equipped with a single antenna; the mobile receiver uses an antenna array with U elements.

The mobile station receives the desired user's signal, superimposed with K co-channel interfering users' signals, all experiencing frequency-selective slow fading. The sequence of modulation symbols of the k-th user is denoted by $\boldsymbol{d}_k = [d_k[0], \cdots, d_k[L-1]]^{\mathrm{T}}$; the symbols $d_k[l]$ belong to a M-PSK modulation alphabet $\mathbb{D}$ and have unit variance, $E\{|d_k[l]|^2\} = 1$. The sequence $\boldsymbol{d}_k$ consists of three consecutive sub-sequences, namely a user-specific midamble of pilot symbols placed in between two sub-sequences of data symbols, which are assumed to be independent and uniformly distributed. At the u-th receive branch, the receive filter output is sampled at baud rate $1/T_d$, where T_d denotes the symbol duration. This yields

$$z_u[l] = \sum_{k=0}^{K} \sum_{p=0}^{P} d_k[l - \gamma_k^p] h_{uk}[l, p] + \eta_u[l], \qquad (1)$$

with $l = 0, \cdots, L-1$ and $\gamma_k^p = p + \lfloor \tau_k / T_d \rfloor$. By convention, $k = 0$ indexes the desired user. The receiver is assumed to be synchronized with the reception of the desired user's signal. Thus with τ_k denoting the relative delay for the k-th user, τ_0 is known and without loss of generality we assume $\tau_0 = 0$.

The complex coefficients $h_{uk}[l, p]$ denote the p-th tap of the discrete-time composite channel impulse response (CIR) between the k-th user's BS and the u-th receive array element at the signaling time l. The composite CIR comprises the transmit pulse shape $c_0(\tau)$, the time-varying CIR $a_{uk}(t, \tau)$ of the propagation channel of the k-th BS to the u-th antenna element, and the square-root Nyquist receive filter $f_r(\tau)$. In accordance with the GSM/EDGE test conditions [8], the CIRs of the propagation channels of all users have the same duration, such that $h_{uk}[l, p] = 0$, $p \neq 0, \cdots, P$ for any k-th user and u-th branch. The channel memory length P and the symbol-wise delays $\lfloor \tau_k / T_d \rfloor$ of the interferers are assumed to be known to the receiver. The sequence of noise samples $\{\eta_u\}$ is a white circularly-symmetric complex Gaussian process with variance $\sigma_\eta^2 = N_0/E_d$, normalized to the desired user's signal power E_d (given $E\{1/U \sum_{u=1}^{U} \sum_{p=0}^{P} |h_{u0}[l, p]|^2\} = 1$). The U noise sequences in the receive branches are assumed to be uncorrelated.

Assuming the propagation channels are wide-sense-stationary with uncorrelated scattering (WSSUS), the wideband time correlation function of their CIRs $\boldsymbol{a}_k(t, \tau) = [a_{1k}(t, \tau), \cdots, a_{Uk}(t, \tau)]^{\mathrm{T}}$ is expressed using the Kronecker channel model [9], according to

$$\boldsymbol{\Sigma}_{\boldsymbol{a}_k, \boldsymbol{a}_k}(\Delta_t, \tau) = \boldsymbol{\Omega}_k(\Delta_t) \boldsymbol{G} \boldsymbol{\Sigma}(\tau). \qquad (2)$$

In (2), $\boldsymbol{\Sigma}(\tau)$ is a diagonal matrix with equal diagonal entries proportional to the delay power spectrum of the propagation channel. The space-time correlation matrix $\boldsymbol{\Omega}_k(\Delta_t)$ is defined as $\boldsymbol{\Omega}_k(\Delta_t) = \int_{-\pi}^{\pi} e^{j2\pi f_D(\theta)\Delta_t} \boldsymbol{\nu}^*(\theta) \boldsymbol{\nu}(\theta)^{\mathrm{T}} f_k(\theta) d\theta$. Here, $f_k(\theta)$ is the angular distribution of scatterers in the surrounding of the receive antennas with θ denoting the direction of arrival (DoA). Considering isotropic antennas, the steering vector $\boldsymbol{\nu}(\theta)$ is defined as $\boldsymbol{\nu}(\theta) = [1, e^{-j2\pi\mu\cos(\theta)}, \cdots, e^{-j2\pi\mu(U-1)\cos(\theta)}]^{\mathrm{T}}$, with μ

denoting the ratio of the equidistant antenna spacing by the wavelength λ of the transmit signal. The Doppler frequency is $f_D(\theta) = \frac{v}{\lambda} \cos(\theta - \beta)$, where v and β denotes the velocity and the direction of movement of the mobile, respectively. Finally, the diagonal matrix $\boldsymbol{G} = \mathrm{diag}\{|g_1|^2, \cdots, |g_U|^2\}$ embodies the effects of the antenna gain imbalance (AGI) and the shadowing due $e.g.$ to the hand and the head of the handset holder.

In the derivation of the receiver we assume perfect frequency hopping and a block-fading channel. In other words, the composite CIRs of all users remain constant within a burst transmission and are independent from one burst to another. Thus, by concatenating the sampled output of the receive filters in the vector $\boldsymbol{z} = [\boldsymbol{z}[0]^{\mathrm{T}}, \cdots, \boldsymbol{z}[L-1]^{\mathrm{T}}]^{\mathrm{T}}$, with $\boldsymbol{z}[l] = [z_1[l], \cdots, z_U[l]]^{\mathrm{T}}$, the signal model (1) can be rewritten in a vector notation as

$$\boldsymbol{z} = \boldsymbol{D}_0 \boldsymbol{h}_0 + \sum_{k=1}^{K} \boldsymbol{D}_k \boldsymbol{h}_k + \boldsymbol{\eta}, \qquad (3)$$

where $\boldsymbol{D}_0$ is the symbol matrix of the desired user and $\boldsymbol{D}_k$, $k = 1, \cdots, K$, are the symbol matrices of the interferers. Notice that $\boldsymbol{D}_k = \tilde{\boldsymbol{D}}_k \otimes \boldsymbol{I}_U$, where $\otimes$ denotes the Kronecker product, $\boldsymbol{I}_U$ is the $U \times U$ identity matrix and

$$\left[\tilde{\boldsymbol{D}}_k\right]_{mn} = \begin{cases} d_k[m-n] & \text{for } \left\lfloor \frac{\tau_k}{T_d} \right\rfloor \leq m-n \leq \left\lfloor \frac{\tau_k}{T_d} \right\rfloor + P, \\ 0 & \text{elsewhere.} \end{cases} \qquad (4)$$

The column vector $\boldsymbol{h}_k = [\boldsymbol{h}_k[0]^{\mathrm{T}}, \cdots, \boldsymbol{h}_k[P]^{\mathrm{T}}]^{\mathrm{T}}$, with $\boldsymbol{h}_k[p] = [h_{1k}[p] \cdots, h_{Uk}[p]]^{\mathrm{T}}$, is a vector representation of the composite CIR of the k-th user. The column vector $\boldsymbol{\eta} = [\boldsymbol{\eta}[0]^{\mathrm{T}}, \cdots, \boldsymbol{\eta}[L-1]^{\mathrm{T}}]^{\mathrm{T}}$, with $\boldsymbol{\eta}[l] = [\eta_1[l], \cdots, \eta_U[l]]^{\mathrm{T}}$ contains the noise samples. For the subsequent derivation, it is also useful to define the vectors $\boldsymbol{d} = [\boldsymbol{d}_0^{\mathrm{T}}, \cdots, \boldsymbol{d}_K^{\mathrm{T}}]^{\mathrm{T}}$ and $\boldsymbol{h} = [\boldsymbol{h}_0^{\mathrm{T}}, \cdots, \boldsymbol{h}_K^{\mathrm{T}}]^{\mathrm{T}}$ comprising the modulation symbols and the multichannel impulse responses of all users, respectively.

III. THE SAGE RECEIVER

In [10], *Kocian et al.* applied the SAGE algorithm to perform multiuser detection in DS-CDMA operating in flat fading channels. There are, however, some major differences between the DS-CDMA scenario considered in [10] and the scenario considered in this contribution. While in our case the specific CIRs are the only discrimination means of the different users, the DS-CDMA receiver exploits the individual signature sequences in addition to the individual CIRs to separate the superimposed users' signals. Moreover, in our context the channels are narrowband with continuous-time CIRs, so that the approach of estimating only the CIRs of the propagation channels is not suitable, as shown in [11]. The composite CIRs have to be estimated instead. Finally, the receiver is extended to the context of multiple receive antennas in order to take advantage of spatial diversity to mitigate CCI. The resulting receiver structure is presented in Fig. 1. In the sequel, we assume that the reader is familiar with the basic notions of the SAGE framework [6].
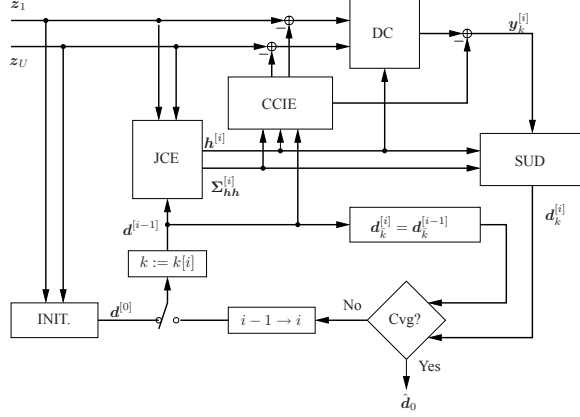
Fig. 1. The proposed receiver consists of a joint multichannel estimator (JCE), a receive diversity combiner (DC), a co-channel interference estimator (CCIE), a single-user detector (SUD), an initialisation scheme (INIT.), a convergence test procedure (Cvg?) and a block that selects the user of interest at iteration i ($k := k[i]$).

A. SAGE Framework for Multiuser Detection

Following an approach similar to that used in [10], we consider the parameters to be estimated to be the symbols of all users d, the incomplete data to be the sampled outputs of the receive filter z, and the admissible hidden data to be $\{z, h\}$ for each iteration. At iteration i, only one user's sequence d_k is updated. Throughout the paper this user is called the user of interest at iteration i. Notice that the user of interest is not necessarily the desired user. The vector $d_{\bar{k}}$ is the vector d where the elements of d_k have been removed.

The initialization procedure (further discussed in Subsection III-E) provides an initial guess of the symbols d that will be used in the first iteration. This guess is denoted by $d^{[0]}$.

Each iteration i of the SAGE algorithm consists of

- *an expectation step (E-step):*

$$\mathcal{Q}\left(d_k|d^{[i-1]}\right) = E\left\{\log p\left(z, h|d_k, d_{\bar{k}}^{[i-1]}\right)|z, d^{[i-1]}\right\} \quad (5)$$

- *and a maximization step (M-step):*

$$d_k^{[i]} = \arg \max_{d_k \in \mathbb{D}^L} \mathcal{Q}\left(d_k|d^{[i-1]}\right),$$
$$d_{\bar{k}}^{[i]} = d_{\bar{k}}^{[i-1]}. \quad (6)$$

Updating each user once, corresponding to $K+1$ iteration, is referred to as one stage. The order of the users' update during each stage is arranged in the ascending order of their receive signal powers. Thus, the symbols of the user received with the weakest power are updated first. This procedure limits the propagation of errors during iterations due to wrong estimates in the initialisation step, as the weakest signals are more likely to be in error; overall the probability of converging to a local maximum is reduced.

The algorithm terminates when convergence is achieved, *i.e.*, when the estimates of the modulation symbols of all users stay unchanged during two successive stages. The estimate of the desired-user sequence $\hat{d}_0$ is then output.

B. Computation of the E-step

The log-likelihood function of d_k and $d_{\bar{k}}^{[i-1]}$ in (5) is decomposed using Bayes' rule and taking into account the

conditional independence of z and h. Discarding the terms independent of d_k, the $\mathcal{Q}$-function can be rewritten as[1]

$$\mathcal{Q}\left(d_k|d^{[i-1]}\right) \propto E\left\{\log p\left(z|d_k, d_{\bar{k}}^{[i-1]}, h\right)|z, d^{[i-1]}\right\}. \quad (7)$$

According to the system model (3) and discarding further summands independent of d_k, the $\mathcal{Q}$-function can be recast as

$$\mathcal{Q}\left(d_k|d^{[i-1]}\right) \propto 2\Re\left\{\text{Tr}\left\{D_k h_k^{[i]} z^{\text{H}}\right\}\right\}$$
$$- \Re\left\{\text{Tr}\left\{D_k \left(h_k h_k^{\text{H}}\right)^{[i]} D_k^{\text{H}}\right\}\right\}$$
$$- 2\Re\left\{\sum_{\substack{k'=0 \\ k' \neq k}}^{K} \text{Tr}\left\{D_k \left(h_k h_{k'}^{\text{H}}\right)^{[i]}\left(D_{k'}^{[i-1]}\right)^{\text{H}}\right\}\right\}. \quad (8)$$

The conditional mean and the conditional correlation matrix of the multichannel impulse response h given the observations z and the estimates of the modulation symbol sequences at the $(i-1)$-th iteration $d^{[i-1]}$ are given by[2]

$$h_k^{[i]} = h_{(k)}^{[i]}, \qquad \left(h_k h_{k'}^{\text{H}}\right)^{[i]} = \left(h h^{\text{H}}\right)_{(k,k')}^{[i]}, \quad (9)$$

where

$$h^{[i]} = E\left\{h|z, d^{[i-1]}\right\}, \qquad \left(h h^{\text{H}}\right)^{[i]} = E\left\{h h^{\text{H}}|z, d^{[i-1]}\right\}. \quad (10)$$

The computations of the channel estimate $h_k^{[i]}$ and of the correlation matrix $\left(h_k h_{k'}^{\text{H}}\right)^{[i]}$ are discussed in Subsection III-D. Notice that $\left(h_k h_{k'}^{\text{H}}\right)^{[i]} = h_k^{[i]}\left(h_{k'}^{[i]}\right)^{\text{H}} + \Sigma_{h_k h_{k'}}^{[i]}$, where $\Sigma_{h_k h_{k'}}^{[i]} = \left(\Sigma_{hh}^{[i]}\right)_{(k,k')}$, with $\Sigma_{hh}^{[i]}$ denoting the conditional error covariance matrix of the multichannel at the i-th iteration. As the receive filter outputs are independent conditioned on the modulation symbols and CIRs, the r.h.s. expression in (8) can be recast according to

$$\mathcal{Q}\left(d_k|d^{[i-1]}\right) \propto \sum_{l=0}^{L-1} \phi_k[l], \quad (11)$$

where

$$\phi_k[l] = \Re\left\{\sum_{p=0}^{P} d_k^*[l-\gamma_k^p]\left(y_k^{[i]}[l,p]\right.\right.$$
$$\left.\left.- \sum_{u=1}^{U}\left(\sum_{\substack{p'=0 \\ p'>p}}^{P} (h_{uk}^*[p]h_{uk}[p'])^{[i]} d_k\left[l-\gamma_k^{p'}\right]\right)\right)\right\}, (12)$$

and

$$y_k^{[i]}[l,p] = \sum_{u=1}^{U}\left(h_{uk}^*[p]^{[i]}\left(z_u[l] - \sum_{\substack{k'=0 \\ k' \neq k}}^{K}\sum_{p'=0}^{P} h_{uk'}^{[i]}[p']d_{k'}^{[i-1]}\left[l-\gamma_{k'}^{p'}\right]\right)\right.$$
$$\left.- \sum_{\substack{k'=0 \\ k' \neq k}}^{K}\sum_{p'=0}^{P}\left[\Sigma_{h_k h_{k'}}^{[i]}\right]_{Up+u\ Up'+u} d_{k'}^{[i-1]}\left[l-\gamma_{k'}^{p'}\right]\right). \quad (13)$$

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[1]We let $x \propto y$ denote $x = \alpha + \beta y$ where α and β are arbitrary constants.

[2]The $(P+1)U$-dimensional vector $x_{(k)}$ contains the elements $[x]_v$, with $v = k(P+1)U, \cdots, (k+1)(P+1)U - 1$. The $(P+1)U \times (P+1)U$ sub-matrix $X_{(k,k')}$ comprises the elements $[X]_{vw}$, where $v = k(P+1)U, \cdots, (k+1)(P+1)U - 1$ and $w = k'(P+1)U, \cdots, (k'+1)(P+1)U - 1$.

The vector $\boldsymbol{y}_k^{[i]}$, with entries $y_k^{[i]}[l,p]$, is the receive-filter output signal for the k-th user cleaned of its CCI and spatially combined. Indeed (13) can be interpreted as follows: at the i-th iteration the CCI signals for the k-th user are regenerated using the composite CIR estimate $\boldsymbol{h}^{[i]}$ and the estimate of the interferers' modulation sequences $\boldsymbol{d}_{\bar{k}}^{[i-1]}$. Then the CCI is subtracted from the receive filter output at each antenna branch. The cleaned observations are then processed by the maximum ratio combiner (MRC) matched to the composite CIR estimates $\boldsymbol{h}^{[i]}$. The last line in (13) is a post-processing algorithm that cancels the residual interference due to channel estimation errors (see Fig. 1).

C. Computation of the M-step

The $\mathcal{Q}$-function defined in (11) is maximized w.r.t. the symbol sequence of the k-th user. The sequence estimate is computed by a single-user sequence detector that provides hard decision on the modulation symbols $\boldsymbol{d}_k$ (see Fig. 1). Due to the additive structure of the $\mathcal{Q}$-function, the Viterbi algorithm can be used to efficiently implement the detector. The metrics of the algorithm are given by $\phi_k[l]$ in (12). The trellis of the sequence detector has M^P states that track the memory state of the composite channel. The equalizer performs ISI cancellation based on the channel weight estimation $\boldsymbol{h}_k^{[i]}$ computed in the E-step. The new estimate $\boldsymbol{d}_k^{[i]}$ is then fed back and used in the following K iterations.

D. Channel Estimation

The conditional distribution of $\boldsymbol{h}$ given $\boldsymbol{z}$ and $\boldsymbol{d}^{[i]}$ is Gaussian with expectation

$$\boldsymbol{h}^{[i]} = \frac{1}{\sigma_\eta^2} \boldsymbol{\Sigma}_{\boldsymbol{hh}}^{[i]} \left(\boldsymbol{D}^{[i-1]} \right)^{\mathrm{H}} \boldsymbol{z} \qquad (14)$$

and covariance matrix

$$\boldsymbol{\Sigma}_{\boldsymbol{hh}}^{[i]} = \left(\frac{1}{\sigma_\eta^2} \left(\boldsymbol{D}^{[i-1]} \right)^{\mathrm{H}} \boldsymbol{D}^{[i-1]} + \boldsymbol{\Sigma}_{\boldsymbol{hh}}^{-1} \right)^{-1}, \qquad (15)$$

where $\boldsymbol{\Sigma}_{\boldsymbol{hh}}$ is the prior covariance matrix of the multichannel. In the E-step at iteration i of the SAGE algorithm, estimates $\boldsymbol{d}^{[i-1]}$ of the modulation symbols are considered as the true transmitted modulation symbols. Under this assumption and the fact that $\boldsymbol{h}$ and $\boldsymbol{z}$ are jointly Gaussian, the channel estimate results in a linear MMSE estimate.

The MMSE estimator incorporates the a-priori knowledge on the composite CIR $\boldsymbol{h}$, i.e. the mean $E\{\boldsymbol{h}\} = \boldsymbol{0}$ and the covariance matrix $\boldsymbol{\Sigma}_{\boldsymbol{hh}}$. The covariance matrix of the multichannel is a block diagonal matrix due to the fact that the CIRs of all users are assumed to be independent: $\boldsymbol{\Sigma}_{\boldsymbol{hh}} = \mathrm{diag}\{\boldsymbol{\Sigma}_{\boldsymbol{h}_0\boldsymbol{h}_0}, \cdots, \boldsymbol{\Sigma}_{\boldsymbol{h}_K\boldsymbol{h}_K}\}$, where $\boldsymbol{\Sigma}_{\boldsymbol{h}_k\boldsymbol{h}_k}$ is the covariance matrix of the k-th user's CIRs. As noted in [11], the covariance matrix of the composite channel $\boldsymbol{\Sigma}_{\boldsymbol{h}_k\boldsymbol{h}_k}$ is computed based on the continuous time function of the receive filter $f_r(\tau)$, the transmit shape filter $c_0(\tau)$ and

$\boldsymbol{\Sigma}_{\boldsymbol{a}_k\boldsymbol{a}_k}(\tau) = \boldsymbol{\Sigma}_{\boldsymbol{a}_k\boldsymbol{a}_k}(\Delta_t, \tau)|_{\Delta_t=0}$. Thus using (2), we obtain[3]

$$(\boldsymbol{\Sigma}_{\boldsymbol{h}_k\boldsymbol{h}_k})_{[p,p']} = E\left\{ \boldsymbol{h}_k[p]\boldsymbol{h}_k[p']^{\mathrm{H}} \right\}$$

$$= \boldsymbol{\Omega}_k \boldsymbol{G} \int_{-\infty}^{+\infty} \boldsymbol{\Sigma}(\tau) f_c(pT_d - \tau) f_c(p'T_d - \tau) d\tau, \quad (16)$$

with $f_c(\tau) = c_0(\tau) * f_r(\tau)$. Since $f_r(\tau)$ and $c_0(\tau)$ are known, the correlation matrix of the composite channel can be precomputed, if $\boldsymbol{\Sigma}(\tau)$, $\boldsymbol{\Omega}_k = \boldsymbol{\Omega}_k(\Delta_t)|_{\Delta_t=0}$ and $\boldsymbol{G}$ are available at the receiver.

E. Initialization Issues

At iteration $i = 0$, the initial estimates of the transmit data symbols $\boldsymbol{d}^{[0]}$ and of the channel vector $\boldsymbol{h}^{[0]}$ are computed. The channel coefficients are estimated using the linear MMSE estimator described in Section III-D based on the pilot symbols of all users. Notice that the training sequences of both the desired and the interfering users are assumed to be known in the receiver. An MMSE diversity combining scheme [12] applied at the receive filter outputs of the U branches computes $\boldsymbol{d}^{[0]}$ based on $\boldsymbol{h}^{[0]}$.

F. Complexity Analysis

The complexity analysis is conducted in terms of floating point operations. Notice that complex operations and real operations are counted in the same way. Remember that K is the number of co-channel interferers, U is the number of receive antenna elements, L is the length of the sequence of transmitted symbols, M is the alphabet size of the modulation, and P is the memory length of the channel.

At first, the complexity of each functional block is analyzed for small K. In the E-step, the complexity of the MMSE channel estimator is $\mathcal{O}\left(L \times (K+1)^2 \times (P+1)^2 \times U^2\right)$, i.e., quadratic in $K+1$, $P+1$ and U. The CCI canceling scheme has complexity $\mathcal{O}\left(L \times K \times U \times (P+1)^2\right)$, i.e. quadratic in $P+1$ and linear in L, U and K. Finally the single-user detector has exponential complexity in the length of the CIR: $\mathcal{O}\left(L \times U \times M^{(P+1)} \times P^2\right)$.

The complexity of one iteration is dominated by the complexity of the single-user equalizer, when the number of co-channel interferers is low. The entire complexity of the SAGE algorithm is $\mathcal{O}\left(S \times L \times U \times M^{(P+1)} \times (K+1) \times P^2\right)$, where S is the number of processing stages. In comparison, the complexity of the joint ML receiver without channel estimator is $\mathcal{O}\left(L \times U \times M^{(P+1)(K+1)} \times (K+1)^2 \times P^2\right)$.

IV. SIMULATION RESULTS

In this section, we evaluate the averaged BER performance of the proposed receiver by means of Monte Carlo simulations. We choose the GSM/EDGE communication system as a particular implementation of an M-PSK system to exemplify the performance of this receiver. The interference originating from only one dominating CCI ($K = 1$) is considered. In the

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[3]The $U \times U$ sub-matrix $\boldsymbol{X}_{[p,p']}$ comprises the elements $[\boldsymbol{X}]_{vw}$ where $v = pU, \cdots, (p+1)U - 1$ and $w = p'U, \cdots, (p'+1)U - 1$.

numerical examples, all users' sequences are transmitted at a carrier frequency of $f = 900$ MHz and the typical urban model with a velocity of 3 km/h (TU3) [8] is considered.

The channel coefficients are generated using the slowly time-variant composite channel model introduced in [13] extended to a multiple-input multiple-output system. In such a geometrical model, the CIR of the propagation channel, $a_{uk}(t, \tau)$, is assumed to be composed of N components with equal power $\sqrt{E_{a_k}/N}$, such that $E_{a_k} = E\left\{\frac{1}{U}\sum_{u=1}^{U}\int |a_{u,k}(t,\tau)|^2 d\tau\right\}$. The n-th plane wave emitted from the k-th BS is scattered by a local scatterer in the surrounding of the mobile phone and impinges the U array elements with a DoA θ_n [14]. Thus, the contribution in the CIR of each plane wave exhibits an individual phase shift ϕ_n and a delay τ_n. The composite channel impulse response $h_{uk}[l,p]$ reads

$$h_{uk}[l,p] = \sqrt{\frac{E_{a_k}}{N}}\, g_u \sum_{n=1}^{N} e^{j\phi_n} e^{j2\pi f_D(\theta_n)lT_d} \times$$
$$\times [\boldsymbol{\nu}(\theta_n)]_u f_c(pT_d - \tau_n). \quad (17)$$

In (17), the continuous random variables ϕ_n, τ_n and θ_n follow the marginal distributions $f_k(\phi)$, $f_k(\tau)$, $f_k(\theta)$ for the k-th user, respectively. Notice that $f_k(\tau)$ is proportional to the delay power spectrum of each link $a_{uk}(t,\tau)$ as proved in [13], and $f_k(\phi)$ is uniform over $[-\pi, \pi]$. Perfect frequency hopping from burst to burst is assumed, i.e. the CIRs in different bursts are uncorrelated. In the simulation, we set the number of inpinging waves to be $N = 64$.

In the receiver, the input filter is an analog root-raised cosine filter with a rolloff-factor $\alpha = 0.3$. For the channel estimation, no prior knowledge of the delay power profile is considered, apart from the memory length of the channel P, and the symbol-wise delay for each user. The antenna gain and space correlation coefficients are assumed to be known if not stated otherwise. A flat delay power spectrum for the propagation channels of all users is assumed to pre-compute (16). In order to reduce the complexity of the single-user detector, the considered composite CIR of each user is pruned in such a way that only the 4 dominant channel weights are kept. This approach is motivated by the fact that the 4 dominant channel weights in the TU profile exhaust all together most of the total channel power [15].

Fig. 2 shows the averaged BER performance of the proposed receiver implemented as a single-antenna interference cancellation scheme ($U = 1$) for the GSM system. A signal-to-noise ratio (SNR) of $E_d/N_0 = 20$ dB is considered and the Jakes channel model is assumed. The averaged BER of the desired user is plotted versus the signal-to-interference ratio (SIR) for three different scenarios: perfect burst and symbol synchronization between the users, i.e. $\tau_0 = \tau_1 = 0$ (*Scenario A*); synchronous network within small cells, i.e. $\lfloor \tau_1/T_d \rfloor \le 1$ (*Scenario B*); and asynchronous network, with $\lfloor \tau_1/T_d \rfloor \le 7$ (*Scenario C*). The top curve in Fig. 2 represents the performance of the conventional receiver. The conventional receiver consists of a single-user MLSE and a pilot-based MMSE channel estimator; the estimator interprets the
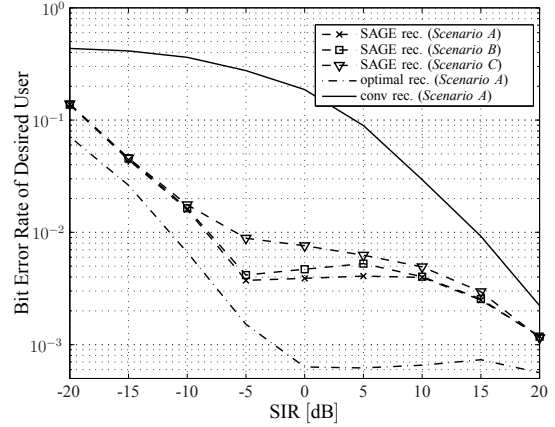

Fig. 2. BER versus SIR for the GSM system operating in the three scenarios with the following setting: TU3 channel profile, $U = 1$, and $E_d/N_0 = 20$ dB.

interferer as part of the white Gaussian noise. The bottom curve shows the performance of the optimal receiver with perfect channel state information (CSI). It represents the lower bound for the multiuser receiver performance. These two curves are generated using *Scenario A*. The proposed receiver outperforms the conventional receiver at low SIR. For example, a gain of $20 - 23$ dB is observed at BER=10^{-2} for all scenarios. The gaps between the curves corresponding to the optimal receiver and the proposed receiver at high and low SNR are due to the channel estimation error. The performance degradation of all receivers at low SIR is due to (i) the remaining ISI due to the pruned channel taps and (ii) the mismatch between the linear model assumed in the receiver and the true non-linear GMSK model.

In Fig. 3 and 4, the performance of the proposed receiver is presented for dual antenna interference cancellation ($U = 2$) in the context of a synchronous EDGE system. A SNR of $E_d/N_0 = 25$ dB is assumed in both simulations. Moreover the DoA θ is chosen to follow the Von Misses distribution as first proposed in [16].

In Fig. 3, the averaged BER for the two test-case scenarios proposed in [17] are investigated: the worst-case scenario with high spatial correlation, i.e. with an envelope correlation $|\varrho|^2 = 0.7$, and with an AGI of 6 dB, and the best-case scenario with uncorrelated antenna gains, i.e. $|\varrho|^2 = 0$ and AGI=0 dB. Several alternative scenarios are also shown. The upper curve shows the performance of the conventional receiver for the best test-case scenario. As similarly observed in Fig. 2, a huge gain is obtained by using the proposed receiver: for example at BER=10^{-2} a gain of more than 25 dB is observed. The performance of the SAGE receiver is also investigated when the spatial statistics of the channel are unknown at the receiver. In this scenario, the pre-computation of (16) makes use of the identity matrix for the spatial correlation matrix and the gain matrix, i.e. $\boldsymbol{\Omega} = \boldsymbol{G} = \boldsymbol{I}_U$; this corresponds to the best-case scenario. The receiver does not suffer any degradation, because the covariance matrix of the channel only slightly affects (16) at high SNR.

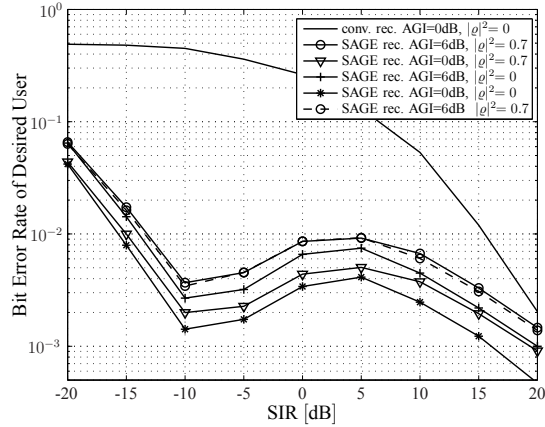Fig. 4 shows the impact of the difference between the

Fig. 3. BER versus SIR for the EDGE system in *Scenario A* with the following setting: TU3 channel profile, $U = 2$, and $E_d/N_0 = 25$ dB; 'solid' the receiver knows the spatial correlation and AGI; 'dashed' the receiver assumes that the channel is uncorrelated and has no knowledge of the AGI.
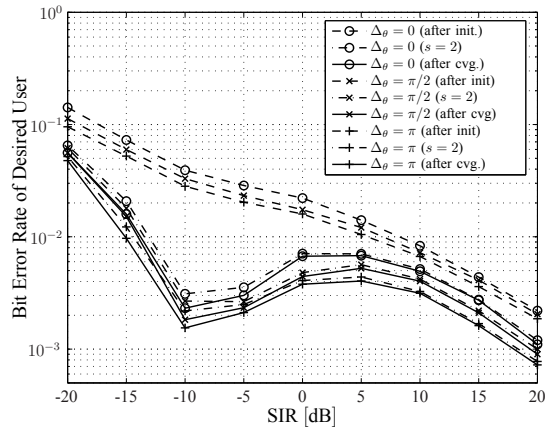

Fig. 4. BER versus SIR for the EDGE system with the angular difference Δ_θ between the mean DoA of the desired and interfering users as a parameter for *Scenario A* with TU3 channel profile, $U = 2$, $E_d/N_0 = 25$ dB and $|\varrho|^2 = 0.7$.

phases of the spatial correlations of the desired and interfering users' signals on the receiver performance. To do so, we vary the difference betwen the nominal DoAs of the two users, which we denote by Δ_θ. Investigations are shown for an average envelope correlation of $|\varrho|^2 = 0.7$ and no AGI. The proposed receiver performs best when the difference is maximized (*i.e.* $\Delta_\theta = \pi$), however the gap between the curve is not significant.

Finally, in all simulated scenarios the proposed receiver converges on average after 6 stages at low SIR. However, only a slight improvement is obtained after the second stage ($s = 2$ in Fig. 4). At high SIR, only one stage is typically necessary to reach convergence, as the initialization scheme is able to cope with low-power CCI.

V. CONCLUSIONS

In this work, the SAGE algorithm has been applied to derive an iterative successive interference cancellation scheme for multiple antenna interference cancellation. The resulting receiver performs multichannel estimation, interference cancellation, spatial diversity combining and single-user detection in an iterative manner. The order of the computational complexity has been analyzed. Monte Carlo simulations for two realistic applications (single antenna interference cancellation for the GSM system and dual antenna interference cancellation for the EDGE system) in TU3 channel profile indicate that a significant gain is obtained compared to the conventional receiver. The proposed receiver also converges after few stages and thus exhibits a low complexity compared to the optimal mutiuser receiver. Furthermore, in the considered scenario the proposed receiver is robust against the mismatch due to the absence of knowledge about the channel statistics (spatial correlation and antenna gain imbalance) and against moderate time misalignements in the reception of the users' signals.

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