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SINGLE ANTENNA INTERFERENCE CANCELLATION (SAIC) FOR CELLULAR TDMA NETWORKS BY MEANS OF DECOUPLED LINEAR FILTERING/NONLINEAR DETECTION

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ABSTRACT

In this paper, a receiver structure suitable for single-antenna co-channel interference cancellation in cellular TDMA networks is studied. The receiver under investigation consists of a novel linear prefilter followed by a nonlinear detector. The task of the prefilter is to suppress non-Gaussian interference and—simultaneously—to shorten the overall impulse response, whereas the task of the nonlinear detector is to cancel intersymbol interference. The prefilter is designed according to the principle of minimum mean square error estimation, without solving an eigenvalue problem. Optionally the prefilter can be designed utilizing available knowledge of all co-channels, knowledge of the channel of a desired user only, or with no channel knowledge at all. The nonlinear detector may be an arbitrary nonlinear equalizer, in the simplest case a memoryless detector. The receiver structure under investigation is compatible with conventional TDMA receivers ignoring co-channel interference. Performance results are presented for synchronous and asynchronous GSM/GPRS networks.

I. INTRODUCTION

Currently, single antenna co-channel interference cancellation (SAIC) is a hot topic, especially for the GSM/EDGE downlink. In field trails, tremendous capacity gains have recently been demonstrated when co-channel interference (CCI) is mitigated, particularly for synchronized networks in urban areas. As a consequence, the upcoming GSM/EDGE release will be tightened with respect to CCI. One of the most challenging tasks is the design of the interference canceller, especially if (due to cost, volume, power consumption, and design aspects) only one receive antenna is available. In general, linear interference canceller fail if the number of receive antennas does not exceed the number of co-channels [11, 6].

In this paper focus is on a receiver structure, where CCI reduction is separated from intersymbol interference (ISI) suppression. CCI reduction is done by means of a time-domain linear filter, followed by a nonlinear detector (such as a trellis-based, tree-based, graph-based equalizer or a decision-feedback equalizer) in order to perform ISI suppression of the desired users channel. Besides interference suppression, the linear filter is able to truncate the overall channel impulse response seen by the subsequent nonlinear detector. A block diagram of the corresponding receiver structure is shown in Fig. 1. Note that this receiver structure is compatible with conventional TDMA receivers ignoring CCI.

The receiver structure is motivated as follows:

• A linear co-channel interference canceller is typically less sensitive with respect to model mismatch compared to a multiuser-based CCI canceller. Model mismatch may be due to asynchronous interference, nonperfect channel estimation, EDGE interferers in a GSM environment or vice versa, etc.

A similar receiver structure (with linear filtering and decoupled nonlinear detection) has been studied in [2, 3] using multiple receive antennas, whereas our focus is on single-antenna interference cancellation, which is more demanding than space-time processing. Still, multiple antennas are optional in our design. The special case of linear filtering in conjunction with a memoryless detector has been investigated in [6, 9, 10] for the case of multiple receive antennas. The novel prefilter under investigation in this paper may be interpreted as a generalization of Kammeyer’s prefilter [5], which is designed in order to truncate a single ISI channel, but which does not take CCI into account. A similar prefilter has also been proposed by Miller and Taylor for MIMO channels [8] and by Al-Dhahir and Cioffi for the case of Gaussian interference [1]. In this paper, however, focus is on a few dominant interferers (which is the typical case in cellular networks), i.e., on non-Gaussian interference. An adaptive MMSE-DFE filter design by means of the LMS algorithm and using channel estimation of a desired user is proposed in [7], where a prefilter with real-valued processing is designed according to the principles established in [10].

The outline of the remainder of this paper is as follows. In Section II, the channel model under consideration is presented. In Section III, the design of the decoupled filter with consideration of multiple receive signals is described. A nice feature of the prefilter under investigation is its suitability for the case that knowledge of all co-channels are available (case A), only knowledge of the channel of the desired user is available (case B), or no channel knowledge at all is available (case C). A comparison of different receiver designs is presented in the numerical results in Section IV, and finally conclusions are drawn in Section V.
Throughout the paper the complex baseband notation is used. Vectors and matrices are written in bold face. Hypotheses are identified by a tilde, final estimates are characterized by a hat, and $(\cdot)^*$ denotes complex conjugate.

II. EQUIVALENT DISCRETE-TIME CHANNEL MODEL

In the presence of $J$ active co-channel interferers, the equivalent discrete-time channel model considered in this paper is given by

$$y^{(i)}[k] = \sum_{j=0}^{J} \sum_{l=0}^{L} h^{(i)}_{j,L}[a_j[k - l]] + n^{(i)}[k], \quad (i)$$

where $y^{(i)}[k] \in \mathbb{C}$ is the $k$th baud-rate output sample of the $i$th polyphase channel, $k$ is the time index $(0 \leq k \leq K - 1)$, $K$ is the number of $M$-ary data symbols per burst, $L$ is the effective memory length of the discrete-time ISI channel model, $h^{(i)}_{j,L}[k] \in \mathbb{C}$ is the $i$th channel coefficient of the $i$th polyphase channel of the $j$th user, $a_j[k]$ is the $k$th data symbol of the $j$th user, randomly drawn from an $M$-ary alphabet $(E(a_j[k]) = 0$, $E\{a_j[k]\}^2 = 1)$, and $n^{(i)}[k] \in \mathbb{C}$ is the $k$th sample of a Gaussian noise process $(E\{n^{(i)}[k]\} = 0, E\{|n^{(i)}[k]|^2\} = N_0/E_s$, where $N_0$ is the one-sided spectral noise density and $E_s$ is the average symbol energy). All random processes are assumed to be mutually independent. If the signal bandwidth does not exceed twice the Nyquist bandwidth, two samples per symbol are sufficient. This scenario is assumed in the following. The corresponding polyphase channels are labeled $i = 1$ and $i = 2$, respectively. The channel coefficients $h^{(i)}_{j,L}[k] := [h^{(i)}_{j,L}[0],...,h^{(i)}_{j,L}[L]]^T \quad (E\{|h^{(i)}_{j,L}[k]|^2\} = 1)$, $(\sum_{j=1}^{J} E\{|h^{(i)}_{j,L}[k]|^2\} = 1)/(C/I), \quad (C/I)$ is the carrier-to-interference ratio) comprise pulse shaping at the transmitter side, the physical channel, analog and digital receive filtering, the sampling phase, and the sampling frequency. Without loss of generality, the effective memory length, $L$, is assumed to be the same for all co-channels; eventually, some coefficients are zero. In case of square-root Nyquist receive filtering and baud-rate sampling, the Gaussian noise processes of all polyphase channels are white. This case is assumed in the following. The equivalent discrete-time channel model is suitable both for synchronous as well as asynchronous TDMA networks.

III. PROPOSED DESIGN OF THE DECOUPLED LINEAR FILTER

For simplicity, a synchronous TDMA network and low vehicular speeds are assumed in the following for the purpose of filter design. In conjunction with frequency hopping, the second assumption corresponds to a so-called block fading channel model. Correspondingly, the time index $k$ can be dropped for convenience. (In case of asynchronous bursts, the channel coefficients are only piece-wise constant.)

As a possible design criterion for the linear filter, the signal to interference plus noise ratio (SINR) at the output of the linear filter can be maximized:

$$\text{SINR}_{\text{max}} := \max_{\mathbf{w}, \mathbf{h}_w} \frac{E\{|\mathbf{w}^H y|^2\}}{E\{|\mathbf{w}^H y - \mathbf{h}_w^H a|^2\}} \quad (2)$$

where the maximization is done jointly with respect to the filter coefficients, $\mathbf{w}$, and the overall channel coefficients (including filtering) of the desired user, $\mathbf{h}_w$.

- The filter coefficients can be written in vector form as $\mathbf{w} = [\mathbf{w}^{(1)}]^T, \mathbf{w}^{(2)}]^T$, where the filter coefficients of each polyphase channel are given by $\mathbf{w}^{(i)} = [w^{(i)}_0, w^{(i)}_1, ..., w^{(i)}_{L_w}]^T$. $L_w$ is the filter order of each polyphase channel. Without loss of generality, oversampling by a factor of two is assumed, i.e., $i \in \{1,2\}$. Note that further oversampling does not lead to any performance improvement (in contrast to increasing the number of receive antennas). Specifically, the degree of diversity is not further enhanced.
- The output samples of the analog receive filter can be written in vector form as $y = [y^{(1)}]^T, y^{(2)}]^T$, where $y^{(i)} = [y^{(i)}[0], y^{(i)}[1], ..., y^{(i)}[k - L_w]]^T$.
- Hypotheses of the overall channel coefficients (including filtering) of the desired user are denoted as $\mathbf{h}_w = [h_{0,w}, ..., h_{L',w}]^T$, where $L'$ is the effective memory length. If the filter order $L_w$ is sufficiently large, the filter truncates the overall impulse response, i.e., $L' \leq L$. For a well designed filter in conjunction with a suitable nonlinear detector (such as a trellis-based, tree-based, or graph-based equalizer), $L'$ is a design parameter. It can be interpreted as the desired memory length of the overall impulse response.
- The corresponding data vector is given by $a = [a[k - k_0], a[k - k_0 - 1], ..., a[k - k_0 - L]]^T$, where $k_0$ is the decision delay of the filter.

Recall that the coefficients of the linear filter, $\mathbf{w}$, and the channel coefficients, $\mathbf{h}_w$, are jointly optimized. The optimum filter and channel coefficients with respect to the maximum SINR are denoted as $\mathbf{w}^*$ and $\mathbf{h}_w^*$, respectively. A maximization of (2) corresponds to a minimization of the cost function

$$C := E\{|\mathbf{w}^*^H y - \mathbf{h}_w^H a|^2\}, \quad (3)$$

that is

$$C_{\text{min}} = \min_{\mathbf{w}, \mathbf{h}_w} \left\{ E\left\{ (\mathbf{w}^H y - \mathbf{h}_w^H a) (\mathbf{w}^H y - \mathbf{h}_w^H a)^H \right\} \right\}. \quad (4)$$

By using the Wirtinger derivative of the cost function $C$ and after some substitutions a solution by solving an eigenvalue problem is obtained.

The main problem of the filter design introduced above is the computational complexity and the robustness [5]. It appears to be difficult to solve the eigenvalue problem, even by means of a Cholesky factorization as proposed by several authors.

An alternative method for calculating the coefficients of the linear filter is based on an MMSE-DFE equalizer. It is shown in the following that the feedforward filter of a fractionally-spaced MMSE-DFE equalizer is equivalent to the desired linear filter. The solution can be written in closed form. The computational complexity is less than solving the corresponding eigenvalue problem. The filter design gives insight with respect to impulse response truncation. Based on (3) and choosing the overall impulse response of the feedback filter of the MMSE-DFE
equalizer to be monic, a modified cost function can be obtained as
\[
C := E \{ |\mathbf{w}^H \mathbf{y} - \hat{h}_w^H \mathbf{a} - a[k - k_0]|^2 \}. \tag{5}
\]
The data sequence in (5) is defined as \( \mathbf{a} := [a[k-k_0-1], \ldots, a[k-k_0-L_0^d]]^T \), and the overall channel coefficients of the desired user are defined as \( \hat{h}_w := [\hat{h}_{1,w}, \hat{h}_{2,w}, \ldots, \hat{h}_{L_w,w}]^T \). The parameter \( k_0 \) is the decision delay and \( L' \) is the effective memory length of the overall channel impulse response of the desired user. \( k_0 \) and \( L' \) are design parameters. The proposed solution generalizes the receiver published in [5], which does not consider interference cancellation and which applies a symbol-spaced feedforward filter. The Wirtinger partial derivatives, which minimize the cost function \( C \), are given as follows:
\[
\frac{\partial C}{\partial \mathbf{w}} = 0^T, \quad \frac{\partial C}{\partial \hat{h}_w} = 0^T. \tag{6}
\]
Assuming i.i.d. data, after some calculation the following relations are obtained:
\[
\mathbf{w}^H \mathbf{R}_{yy} = \hat{h}_w^H \mathbf{R}_{ya} + \mathbf{r}_{ya}^H \quad \hat{h}_w^H = \mathbf{w}^H \mathbf{R}_{ya}. \tag{7}
\]
As defined before, \( \mathbf{R}_{yy} = E\{\mathbf{y}\mathbf{y}^H\} \) denotes the \( 2(L_w + 1) \times 2(L_w + 1) \) autocorrelation matrix of the received samples, \( \mathbf{R}_{ya} = E\{\mathbf{a}\mathbf{y}^H\} \) denotes the \( L' \times 2(L_w + 1) \) cross-correlation matrix between the data and the received samples, and the \( 2(L_w + 1) \) vector is defined as \( \mathbf{r}_{ya} := [r_{ya}[k_0], \ldots, r_{ya}[k_0 - 2L_w]]^T := E\{*\mathbf{[k - k_0]}\} \). Upon insertion of (8) in (7) the optimum filter coefficients are obtained as
\[
\mathbf{w}^H = \mathbf{r}_{ya}^H \left( \mathbf{R}_{yy} - \mathbf{R}_{ya} \mathbf{R}_{ya}^H \right)^{-1}. \tag{9}
\]
The corresponding channel coefficient vector \( \hat{h}_w \) is obtained by inserting (9) into (8). Hence, in the proposed solution the filter coefficients are computed prior to the channel coefficients. The symbol-spaced channel coefficients \( \hat{h}_w \) are provided to the decoupled nonlinear detector. According to (8), and (11), the overall channel coefficients \( \hat{h}_w \) can be written as a convolution between the filter coefficients, \( \mathbf{w} \), and the channel coefficients \( \mathbf{h}_0 \).

The solution discussed so far is suitable for complex symbol alphabets (such as 8-PSK). For one-dimensional symbol alphabets (such as M-PAM or derotated GMSK), the performance can be improved by applying real-valued processing as proposed in [10].

A. Calculation of Correlation Matrices in Case of Available Channel Estimates (Case A)

In the proposed solution in particular the autocorrelation and crosscorrelation matrices \( \mathbf{R}_{yy} = E\{\mathbf{y}\mathbf{y}^H\} \) and \( \mathbf{R}_{ya} = E\{\mathbf{a}\mathbf{y}^H\} \) have to be computed for each burst. This computation can be done in a straightforward manner by taking the expected values \( E\{\mathbf{y}\mathbf{y}^H\} \) and \( E\{\mathbf{a}\mathbf{y}^H\} \) without applying any channel knowledge. However, in case of i.i.d. data we observed a significant performance improvement (compared to using a short training sequence) if channel estimates are used.

In this subsection, we assume that channel estimates for the desired user as well as for the interferer are available. This assumption is suitable for synchronous TDMA networks. In the next subsection, we assume that channel estimates are only available for the desired user, a scenario which is more suitable for asynchronous TDMA networks. In the third subsection, we finally assume that no channel estimates are available for the filter design.

In case of i.i.d data, the elements \( r_{yy}(i,j) \) of the \( 2(L_w + 1) \times 2(L_w + 1) \) autocorrelation matrix \( \mathbf{R}_{yy} = E\{\mathbf{y}\mathbf{y}^H\} \) can be calculated as follows:
\[
r_{yy}(i,j) = \begin{cases} 
0 & i,j \leq L_w \\
\sum_{m=0}^{L_w} h_{1,m}^{(1)}(h_{i+j-m,1})^* + \sigma_n^2 \delta_{i-j}, & 0 \leq i,j \leq L_w \\
\sum_{m=0}^{L_w} h_{1,m}^{(2)}(h_{i+j-m,2})^* + \sigma_n^2 \delta_{i-j}, & L_w + 1 \leq i,j \leq 2L_w + 1 \\
\sum_{m=0}^{L_w} h_{2,m}^{(1)}(h_{i+j-m,2})^* + \sigma_n^2 \delta_{i-j}, & L_w + 1 \leq i,j \leq 2L_w + 1.
\end{cases} \tag{10}
\]

where \( \sigma_n^2 \) is the noise variance. Correspondingly, the elements \( r_{ya}(i,j) \) of the \( L' \times 2(L_w + 1) \) crosscorrelation matrix \( \mathbf{R}_{ya} = E\{\mathbf{a}\mathbf{y}^H\} \) can be calculated as follows:
\[
r_{ya}(i,j) = \begin{cases} 
(h_{k_0,i-j,0})^*, & 0 \leq i \leq L_w, 0 \leq j \leq L_w \\
(h_{k_0,i-j,0})^*, & 0 \leq i \leq L', L_w + 1 \leq j \leq 2L_w + 1.
\end{cases} \tag{11}
\]

Finally, the elements of the crosscorrelation vector \( \mathbf{r}_{ya} \) can be computed as
\[
r_{ya}(i) = \begin{cases} 
(h_{k_0,i,0}), & 0 \leq i \leq L_w \\
(h_{k_0,i,0}), & L_w + 1 \leq i \leq 2L_w + 1.
\end{cases} \tag{12}
\]

Note that for each burst estimates of \( h_1 \), \( 0 \leq j \leq J \), and an estimate of the noise variance \( \sigma_n^2 \) have to be known at the receiver. The degradation is small, however, if a constant noise variance (let us say, \( 0.05 \leq \sigma_n^2 \leq 0.001 \)) is assumed. Though an explicit knowledge of the data is not necessary, training sequences are helpful in order to compute the channel estimates.

B. Calculation of Correlation Matrices in Case that Channel Estimates are not Available for the Interferer (Case B)

Particularly in asynchronous TDMA networks, an estimation of the channel coefficients of the interferer is difficult. In case that estimates of the channel coefficients of the desired user are available, but estimates of the channel coefficients of the interferer are not available, the autocorrelation matrix \( \mathbf{R}_{yy} \) can be approximated as follows: The main idea is to partition \( \mathbf{R}_{yy} \) in the form
\[
\mathbf{R}_{yy} = E\{\mathbf{y}_d \mathbf{y}_d^H\} + E\{\mathbf{y}_{i+n} \mathbf{y}_{i+n}^H\} = \mathbf{R}_{yy}^d + \mathbf{R}_{yy}^{i+n}, \tag{13}
\]
where
\[
y_{d}^{(i)}[k] = \sum_{l=0}^{L} h_{i,0}^{(i)} a_0[k - l], \quad y_{i+n}^{(i)}[k] = \sum_{l=0}^{L} h_{i+n,0}^{(i)} a_0[k - l] \tag{14}
\]
The elements of the autocorrelation matrix $\mathbf{R}_{yy}^{d}$ can be written as

$$
r_{yy}(i, j) = \begin{cases} 
\sum_{l=0}^{L} h_{i,0}^{(1)} (h_{i+1-l,0})^*, & 0 \leq i, j \leq L_w \\
\sum_{l=0}^{L} h_{i,0}^{(2)} (h_{i+1-l,0})^*, & 0 \leq i \leq L_w, \\
L_w + 1 \leq j \leq 2L_w + 1, \\
\sum_{l=0}^{L} h_{i,0}^{(2)} (h_{i+1-l,0})^*, & L_w + 1 \leq i \leq 2L_w + 1, \\
\sum_{l=0}^{L} h_{i,0}^{(2)} (h_{i+1-l,0})^*, & 0 \leq j \leq L_w, \\
L_w + 1 \leq i \leq 2L_w + 1. 
\end{cases}
$$

(16)

The elements of the autocorrelation matrix $\mathbf{R}_{yy}^{i+n}$ can be approximated as

$$
r_{yy}^{i+n}(i, j) \approx \begin{cases} 
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} y_{i+n}^{(1)}[k - i] (y_{i+n}^{(1)}[k - j])^*, & 0 \leq i, j \leq L_w \\
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} y_{i+n}^{(2)}[k - i] (y_{i+n}^{(2)}[k - j])^*, & 0 \leq i \leq L_w, L_w + 1 \leq j \leq 2L_w + 1, \\
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} y_{i+n}^{(2)}[k - i] (y_{i+n}^{(2)}[k - j])^*, & L_w + 1 \leq i \leq 2L_w + 1. 
\end{cases}
$$

(17)

where $y_{i+n}^{(i)}[k] := y^{(i)}[k] - y_{d}^{(i)}[k]$. The parameter $K'$ is the length of the training sequence if training symbols are used in order to calculate $\mathbf{R}_{yy}^{i+n}$. If tentative decisions are used in order to calculate $\mathbf{R}_{yy}^{i+n}$, the parameter $K'$ is the burst length. Note that no additional noise variance estimation is needed.

C. Calculation of Correlation Matrices in Case that no Channel Estimates are Available (Case C)

In the case that co-channel interference is strong, channel estimates for the desired user are poor in synchronous and asynchronous networks. Also, channel estimates for the interferer are difficult to obtain in asynchronous networks. For these reasons, the following novel solution is proposed, where no channel estimation and no estimation of the noise variance is necessary for the design of the linear filter.

In a first step, we compute the filter coefficients

$$
\hat{\mathbf{w}}^H = \mathbf{r}_{ya}^H \mathbf{R}_{yy}^{-1},
$$

(18)

The elements of the autocorrelation matrix $\mathbf{R}_{yy}$ can be approximated as

$$
r_{yy}(i, j) \approx \begin{cases} 
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} y^{(1)}[k - i] (y^{(1)}[k - j])^*, & 0 \leq i, j \leq L_w \\
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} y^{(2)}[k - i] (y^{(2)}[k - j])^*, & 0 \leq i \leq L_w, L_w + 1 \leq j \leq 2L_w + 1, \\
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} y^{(2)}[k - i] (y^{(2)}[k - j])^*, & L_w + 1 \leq i \leq 2L_w + 1. 
\end{cases}
$$

(19)

The parameter $K'$ is the length of the training sequence if training symbols are used in order to calculate $\mathbf{R}_{ya}$. If tentative decisions are used in order to calculate $\mathbf{R}_{ya}$, the parameter $K'$ is the burst length. The elements of the autocorrelation matrix $\mathbf{R}_{ya}$ can be approximated as

$$
r_{ya}(i, j) \approx \begin{cases} 
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} a[k - k_0 - i] (y^{(1)}[k - j])^*, & 1 \leq i \leq L', 0 \leq j \leq L_w \\
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} a[k - k_0 - i] (y^{(2)}[k - j])^*, & 1 \leq i \leq L', L_w + 1 \leq j \leq 2L_w + 1 
\end{cases}
$$

(20)

Finally, the elements of the crosscorrelation vector $\mathbf{r}_{ya}$ can be approximated as

$$
r_{ya}(i) \approx \begin{cases} 
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} a[k - k_0] (y^{(1)}[k - i])^*, & 0 \leq i \leq L_w \\
\frac{1}{K'} \sum_{k=k_0}^{K' + k_0 - 1} a[k - k_0] (y^{(2)}[k - i])^*, & L_w + 1 \leq i \leq 2L_w + 1. 
\end{cases}
$$

(21)

In a second step, the convolution

$$
y_w := \mathbf{w}^H \mathbf{y}
$$

(22)

is performed. In a third step, conventional LS channel estimation is done given $y_w$ and $a$. Due to the interference suppression, the channel estimation error is reasonably small. Finally, MLSE of the desired user is done given $y_w$ and the estimated channel coefficients $\mathbf{h}_w$.

IV. NUMERICAL RESULTS

For the numerical results presented in Fig. 2- Fig. 5, the following scenarios are assumed:

- Synchronous and asynchronous GSM network
- $J = 1$ dominant interferer
- Uniformly distributed training sequence code (TSC) for desired user and interferer
- TSC of dominant interferer differs from TSC of desired user
- TU50 channel model, $L = 3$
- Proposed decoupled MMSE filter/Viterbi detector
• Real-valued processing in MMSE filter
• $L_w = 4$ (corresponding to 20 real-valued filter coefficients), $k_0 = (L_w + 1)/2 - 1$
• $L' = L = 3$ (no truncation, i.e., $M^{L'} = 8$ state Viterbi detector)

As a reference, performance results for the conventional receiver are included as well. The performance/complexity trade-off is quite remarkable.

Figure 2: Raw BER (of desired user) versus $C/I$, TU50 channel model, joint LS channel estimation, synchronous GSM network, case A.

Figure 3: Raw BER (of desired user) versus $C/I$, TU50 channel model, LS channel estimation for desired user, synchronous GSM network, case B.

V. CONCLUSIONS

In this paper, a decoupled linear filter/nonlinear detector has been investigated. The purpose of the fractionally-spaced linear filter is to suppress CCI and to shorten the overall impulse response.

An efficient algorithm for the design of the linear filter is introduced. The proposed solution generalizes the receiver published in [5]. Particular emphasis is on adaptation techniques for the linear filter taking channel knowledge of the desired user and, optionally, of the interferers into account. Also, the case of no channel knowledge for the purpose of designing the linear filter is considered. The numerical results are shown for the popular GSM system and demonstrate significant performance improvements for the three cases under investigation.

REFERENCES


