Single Antenna Interference Cancellation using Prefiltering and Multiuser Joint Detection based on the M-Algorithm

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***Abstract*— Cellular mobile radio systems like GSM/ EDGE (Enhanced Data Rates for GSM Evolution) using pulse amplitude modulation (PAM) for transmission over frequency-selective channels are confronted with increasing demands on spectral efficiency. By using advanced signal processing techniques, receivers are able to cope with higher interference power levels from e.g. cochannel cells, therefore allowing a smaller frequency reuse. In mobile phones, only a single antenna can be applied. We assume 8–ary phase–shift keying (8PSK) modulation and consider a joint detection approach for single antenna receivers in the downlink using the M-algorithm and a novel prefiltering approach. The dominant interferer signal is treated together with the desired user signal by joint detection in order to increase robustness to cochannel or adjacent channel interference. Simulations exhibit high performance gains compared to conventional receiver techniques.**

I. INTRODUCTION

Second generation mobile radio networks are interference limited systems. Using conventional equalization approaches, time–division multiple access (TDMA) schemes like the GSM/ EDGE system need to be designed in a way, that cochannel interference from neighboring cells is small enough that a single–input single–output (SISO) receiver ignoring the disturbing signal can be employed. Due to spectrum limitations, recent developments aim at significantly increasing GSM capacity [1]. A promising way for this are interference suppression and multiuser joint detection. Using these methods, receivers are able to cope with much higher interference power levels than conventional receivers. This property can be exploited in network layout, employing a smaller frequency reuse factor.

Cost and size limitations of mobile terminals currently only allow a single receive antenna in the downlink. To increase capacity, GSM/ EDGE employs 8–ary phase–shift keying (8PSK). Thus, maximum-likelihood sequence estimation (MLSE) receivers would require a vast complexity and therefore cannot be applied, in particular for joint detection. Reduced–complexity methods have to be adopted for joint detection in frequency–selective channels. In [2], performance of delayed decision–feedback sequence estimation (DDFSE) for two–user joint demodulation is shown, employing the prefilter of [3]. In this paper, we concentrate on prefilter design and investigate different multiuser detection algorithms like joint DDFSE and joint reduced–state sequence estimation (JRSSE) [4] for a single receive antenna. The M-algorithm [5] is considered as a second approach for improved multiuser joint detection. For multiple receive antennas, filter design can be found e.g. in [6], [7].

In single user reduced–complexity receivers, allpass prefilters are applied to convert the channel into its minimum–phase equivalent representation. Since this cannot be done simultaneously with a single filter for both signals of user and interferer, the prefiltering approach of [3] is examined. It can be observed that for the two–user case, this method can be significantly improved by using a generalized approach. The novel prefilter is based on a possible decision delay difference between desired user and interferer, which can easily be introduced for the used detection methods. The delay difference is utilized within prefilter design as additional degree of freedom in order to get a corresponding optimum overall channel solution for desired user and interferer.

This paper is structured as follows. First the system model is introduced. Then, the prefiltering method and the M-algorithm are presented. Finally, numerical results from simulations are shown.

II. SYSTEM MODEL

Synchronized networks combined with the implementation of single antenna interference cancellation (SAIC) promise a substantial gain in network capacity [1] and are assumed in the following. The considered scenario for derivation of the receiver algorithm comprises two users, namely the *desired user* and the *interferer*, that transmit via independent channels to the receiver, as shown in Fig. 1. The desired user signal $a_0[k]$ and the interferer signal $a_1[k]$ are considered as i.i.d. and mutually independent with variances σ_a^2. The output signal of the channel, cf. Fig. 1, is $r[k]$. We collect the transmit signals in a vector $\mathbf{a}[k] = [a_0[k]\ a_1[k]]^{\mathrm{T}}$ ($(\cdot)^{\mathrm{T}}$: transposition). The input–output relation of the channel describing a GSM/ EDGE

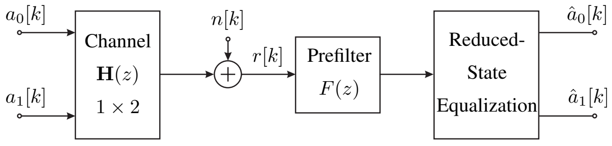

Fig. 1. System model for single antenna interference cancellation with user signal $a_0[k]$ and interferer signal $a_1[k]$.

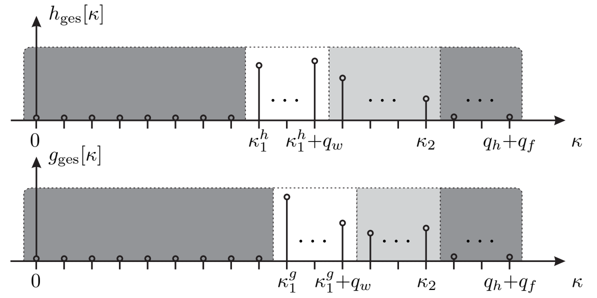

Fig. 2. Overall channel and prefilter impulse response of desired user, $h_{\mathrm{ges}}[\kappa]$, and interferer, $g_{\mathrm{ges}}[\kappa]$, respectively.

transmission with interference is

$$r[k] = \sum_{\kappa=0}^{q_h} \mathbf{H}[\kappa]\, \mathbf{a}[k-\kappa] + n[k]. \qquad (1)$$

Here, $\mathbf{H}[\cdot]$ denotes the causal vector–valued FIR impulse response of order q_h of the overall channel including continuous–time transmit and receive filtering, $\mathbf{H}[\kappa] = [h[\kappa]\ g[\kappa]]$. $h[\cdot]$ and $g[\cdot]$ are the overall impulse responses of the desired user and the interferer, respectively. $n[k]$ is the additive white Gaussian noise (AWGN) process of the receive sequence with variance σ_n^2. Ideal channel knowledge at the receiver is assumed throughout this paper in order to show general performance limits of the algorithms.

III. PREFILTER DESIGN

A reduced–state trellis–based detection is considered in this paper. For this purpose, it is advantageous to concentrate the impulse responses within a range smaller than the original time span. For this, an FIR prefilter with transfer function $F(z) = \sum_{\kappa=0}^{q_f} f[\kappa]\, z^{-\kappa}$, cf. Fig. 1, is applied in the following. The overall channel and prefilter impulse response of the desired user is given by $h_{\mathrm{ges}}[\kappa] = h[\kappa] * f[\kappa]$ and of the interferer by $g_{\mathrm{ges}}[\kappa] = g[\kappa] * f[\kappa]$, respectively. For prefilter design we use the degree of freedom of joint detection, that detection delay of user (κ_1^h) and interferer (κ_1^g) may differ by $q_\Delta = \kappa_1^h - \kappa_1^g$, $|q_\Delta| \le q_b$. q_b denotes the memory length used for detection. By including the optimization of the detection delay difference q_Δ in prefilter design, a better concentration can be achieved than with equal detection delays ($q_\Delta = 0$). The applied reduced–state trellis–based detection can handle postcursors of the overall channel within the range from $\kappa_1 + 1$ to $\kappa_2 = \kappa_1 + q_b$, where $\kappa_1 = \min\{\kappa_1^h, \kappa_1^g\}$.

To concentrate the energy of the overall impulse responses of user and interferer, we introduce a filter design method, which is a generalization of [3]. The novel algorithm optimizes the energy ratios of the impulse responses $h_{\mathrm{ges}}[\kappa]$ and $g_{\mathrm{ges}}[\kappa]$ corresponding to, in general, different ranges. This is demanded by our strategy to allow delays between the two detected signals. More precisely, the energy of the white emphasized regions, $\kappa_a^{h,g} \in \{\kappa_1^{h,g}, \ldots, \kappa_1^{h,g} + q_w\}$, is considered as useful energy and the energy of the regions $\kappa_b^{h,g} \in \{0, \ldots, \kappa_1^{h,g} - 1, \kappa_2 + 1, \ldots, q_h + q_f\}$, marked by dark gray boxes in Fig. 2 as distortion, respectively, to achieve good performance of reduced–state detection, which results if high energy is contained in the first few taps. The energy within the light gray marked regions, $\kappa_c^{h,g} \in \{\kappa_1^{h,g} + q_w + 1, \ldots, \kappa_2\}$, is not considered in filter design.

We can express the cascade of prefilter and channel of the desired user by a channel matrix $\mathbf{H}$, a filter vector $\mathbf{f}$ and an overall impulse response vector $\mathbf{h}_{\mathrm{ges}}$:

$$\begin{aligned}
\mathbf{h}_{\mathrm{ges}} &= [h_{\mathrm{ges}}[0]\, h_{\mathrm{ges}}[1]\, \ldots\, h_{\mathrm{ges}}[q_h+q_f]]^{\mathrm{T}} \\
&= \mathbf{H}\,\mathbf{f}, && (2) \\
[\mathbf{H}]_{mn} &= \begin{cases} h[m-n] & \text{for } 0 \le m-n \le q_h \\ 0 & \text{elsewhere} \end{cases}, && (3) \\
& \quad m \in \{0, \ldots, q_h+q_f\}, \quad n \in \{0, \ldots, q_f\}, \\
\mathbf{f} &= [f[0]\, f[1]\, \ldots\, f[q_f]]^{\mathrm{T}}\,. && (4)
\end{aligned}$$

The overall impulse response vector of the interferer is written in the same way, where the matrix $\mathbf{G}$ is set up as in (3) using $g[\cdot]$ instead of $h[\cdot]$:

$$\mathbf{g}_{\mathrm{ges}} = \mathbf{G}\,\mathbf{f}. \qquad (5)$$

We split up the overall channel and prefilter impulse response $\mathbf{h}_{\mathrm{ges}}$ into the useful part $\mathbf{h}_a = \mathbf{H}_a \mathbf{f}$ and the undesired part $\mathbf{h}_b = \mathbf{H}_b \mathbf{f}$ with

$$\begin{aligned}
\mathbf{H}_a &= \mathbf{H}_{[\kappa_1^h, \ldots, \kappa_1^h + q_w]} \quad \text{and} \\
\mathbf{H}_b &= \begin{bmatrix} \mathbf{H}_{[0, \ldots, \kappa_1^h - 1]} \\ \mathbf{H}_{[\kappa_2+1, \ldots, q_h+q_f]} \end{bmatrix},
\end{aligned}$$

where $\mathbf{H}_{[m_1, \ldots, m_2]}$ denotes the matrix that is built by stacking the rows from index m_1 to m_2 of matrix $\mathbf{H}$. For the interferer we obtain $\mathbf{g}_a = \mathbf{G}_a \mathbf{f}$ and $\mathbf{g}_b = \mathbf{G}_b \mathbf{f}$ with the matrices

$$\begin{aligned}
\mathbf{G}_a &= \mathbf{G}_{[\kappa_1^g, \ldots, \kappa_1^g + q_w]} \quad \text{and} \\
\mathbf{G}_b &= \begin{bmatrix} \mathbf{G}_{[0, \ldots, \kappa_1^g - 1]} \\ \mathbf{G}_{[\kappa_2+1, \ldots, q_h+q_f]} \end{bmatrix},
\end{aligned}$$

respectively.

The useful energy of desired user and interferer is now given by $E_a = \mathbf{h}_a^{\mathrm{H}} \mathbf{h}_a + \mathbf{g}_a^{\mathrm{H}} \mathbf{g}_a = \mathbf{f}^{\mathrm{H}} (\mathbf{H}_a^{\mathrm{H}} \mathbf{H}_a + \mathbf{G}_a^{\mathrm{H}} \mathbf{G}_a)\,\mathbf{f}$ and the energy of the undesired part by $E_b = \mathbf{h}_b^{\mathrm{H}} \mathbf{h}_b + \mathbf{g}_b^{\mathrm{H}} \mathbf{g}_b = \mathbf{f}^{\mathrm{H}} (\mathbf{H}_b^{\mathrm{H}} \mathbf{H}_b + \mathbf{G}_b^{\mathrm{H}} \mathbf{G}_b)\,\mathbf{f}$. The noise energy at the filter output is $E_n = \sigma_n^2\, \mathbf{f}^{\mathrm{H}} \mathbf{f}$. The prefilter is computed for maximization of

$$J = \frac{E_a}{E_b + E_n} = \frac{\mathbf{f}^{\mathrm{H}} (\mathbf{H}_a^{\mathrm{H}} \mathbf{H}_a + \mathbf{G}_a^{\mathrm{H}} \mathbf{G}_a)\,\mathbf{f}}{\mathbf{f}^{\mathrm{H}} (\mathbf{H}_b^{\mathrm{H}} \mathbf{H}_b + \mathbf{G}_b^{\mathrm{H}} \mathbf{G}_b + \sigma_n^2\, \mathbf{I})\,\mathbf{f}}, \qquad (6)$$

where $\mathbf{I}$ denotes the identity matrix of size $(q_f+1)\times(q_f+1)$. For given parameters κ_1^h, κ_1^g and q_w, the solution is calculated via the derivative of J obtained by the Wirtinger calculus [8]:

$$\frac{\delta J}{\delta \mathbf{f}^*} = \frac{\delta}{\delta \mathbf{f}^*}\frac{\mathbf{f}^{\mathrm{H}}\mathbf{A}\,\mathbf{f}}{\mathbf{f}^{\mathrm{H}}\mathbf{B}\,\mathbf{f}} = \frac{\mathbf{A}\,\mathbf{f}\cdot\mathbf{f}^{\mathrm{H}}\mathbf{B}\,\mathbf{f}-\mathbf{f}^{\mathrm{H}}\mathbf{A}\,\mathbf{f}\cdot\mathbf{B}\,\mathbf{f}}{(\mathbf{f}^{\mathrm{H}}\mathbf{B}\,\mathbf{f})^2}, \quad (7)$$

where $\mathbf{A}=\mathbf{H}_a^{\mathrm{H}}\mathbf{H}_a+\mathbf{G}_a^{\mathrm{H}}\mathbf{G}_a$ and $\mathbf{B}=\mathbf{H}_b^{\mathrm{H}}\mathbf{H}_b+\mathbf{G}_b^{\mathrm{H}}\mathbf{G}_b+\sigma_n^2\,\mathbf{I}$ is used. Setting (7) to zero as condition for maximization, we get

$$\mathbf{A}\,\mathbf{f}\cdot(\mathbf{f}^{\mathrm{H}}\mathbf{B}\,\mathbf{f}) \;=\; (\mathbf{f}^{\mathrm{H}}\mathbf{A}\,\mathbf{f})\cdot\mathbf{B}\,\mathbf{f}. \quad (8)$$

With $\lambda=\frac{\mathbf{f}^{\mathrm{H}}\mathbf{A}\,\mathbf{f}}{\mathbf{f}^{\mathrm{H}}\mathbf{B}\,\mathbf{f}}$, the generalized eigenvalue problem $\mathbf{A}\,\mathbf{f}=\lambda\,\mathbf{B}\,\mathbf{f}$ results. Eigenvalue decomposition provides the maximum eigenvalue $J_{\max}=\lambda_{\max}$ and the appropriate eigenvector $\mathbf{f}_{\max}$.

Finally, we use different sets of κ_1^h, κ_1^g and q_w and compute the according optimum energy ratio

$$J_{\mathrm{opt}} = \max_{\kappa_1^h,\kappa_1^g,q_w}\left(J_{\max}\right). \quad (9)$$

The corresponding vector $\mathbf{f}_{\mathrm{opt}}$ is then employed for prefiltering. To reduce the number of computations in simulations, q_w is optimized once for the used detection algorithms and fixed to $q_w=2$ for our application. The set of pairs (κ_1^h, κ_1^g) is limited to $\{(\kappa_1,\kappa_1),(\kappa_1,\kappa_1+1),(\kappa_1,\kappa_1+2),(\kappa_1,\kappa_1+3),(\kappa_1+1,\kappa_1),(\kappa_1+2,\kappa_1),(\kappa_1+3,\kappa_1)\}$. The parameter κ_1 is set to q_f, resulting effectively in an anticausal prefilter, which performes best. By setting $\kappa_1=\kappa_1^h=\kappa_1^g$ and $\kappa_2=\kappa_1+q_w$ the special case of [3] is obtained. Since this choice does not yield reliable results for our application, we did not take it into account further.

The filter design can be easily extended to detection of arbitrary number of users by including additional terms in (6).

IV. M-ALGORITHM

The M-algorithm [5] belongs to the class of sequential decoding schemes and is a purely breadth-first algorithm, extending all paths of a certain depth at once and then selecting the M paths with the best metrics before proceeding forward. Here, we consider the M-algorithm for multiuser joint detection. The tree to be searched by the algorithm is given by all possible combinations of transmit sequences of user and interferer. Since all paths to be compared have the same number of symbols at each processing step, the same metric as for JRSSE [4] can be applied. In terms of complexity, the M-algorithm is similar to JRSSE with M states for moderate M, if sorting of the paths according to their metrics in each detection step is neglected. However, although feasible in principle, obtaining soft information is more involved for the M-algorithm than for trellis-based algorithms, since structure is not as regular as for JRSSE and JDDFSE. In simulations, we compare the M-algorithm to JRSSE and JDDFSE.

V. NUMERICAL RESULTS

For simulations we use the GSM/EDGE typical urban (TU) channel profile. The channel is assumed to be constant within each burst but varies independently from burst to burst. Furthermore, perfect channel knowledge at the receiver side is assumed. The strength of interference is described by the carrier–to–interference ratio (CIR), which is the ratio of the average received symbol energy of desired user and interferers. The dominant interference ratio (DIR) is used to characterize the interference scenario adopted for numerical results consisting of multiple interferers, one dominant and four remaining with equal average powers, and is defined as the ratio of the energy of the dominant interferer, considered in joint detection, to the sum energy of the remaining other interferers not taken into account in detection [1]. E_b/N_0 is fixed during simulations at $30\,\mathrm{dB}$ (E_b: average received bit energy of desired user; N_0: single–sided power spectral density of the underlying passband noise process). This is a typical value for 8PSK receivers. Performance is mainly dependent on the interference situation, and the influence of the noise is quite small in comparison to the distortion created by disregarded interferers. The performance of state-of-the-art receivers corresponds to the curves for a dominant interference ratio of $10\log_{10}(\mathrm{DIR}) \rightarrow -\infty$ in the diagrams of joint detection receivers. In these cases, interference is completely given by unconsidered interferers within detection. The gain with respect to network capacity depends on the interferer constellation and the network setup and has to be determined by separate considerations [9].

We compare two different receiver strategies, denoted by A and B. For detection, we consider JRSSE, JDDFSE and the M-algorithm with about comparable complexity in terms of allocated states/paths. The M-algorithm (64) uses 64 shared states for user and interferer, whereas in JDDFSE $(8,8)$, 8 states are used for the first channel tap of desired user and interferer, respectively. Our investigations have shown, that JRSSE with uniform set partitioning [4] of desired user and interferer using Ungerboeck set partitioning performs poor, if e.g. only 2 subsets are used for each of the three leading channel taps of desired user and interferer. Therefore, JRSSE is not recommendable here.

Receiver strategy A is based on a detection, that ignores precursors of the channel impulse responses of desired user $h[\kappa]$ and interferer $g[\kappa]$, respectively, if their cumulative energy is smaller than a given percentage ξ of the user (resp. interferer) impulse response energy. This value has been optimized by simulations and an optimum for $10\log_{10}(\mathrm{DIR}) \rightarrow \infty$ (only one interferer) and $10\log_{10}(\mathrm{CIR}) = 0\,\mathrm{dB}$ was found at $\xi = 1.2\%$. Precursors of desired user and interferer are considered separately. Therefore, different detection delays for user and interferer may be necessary, if a different number of precursors are ignored. One can observe, that strategy A is highly dependent on the value ξ and the interferer power

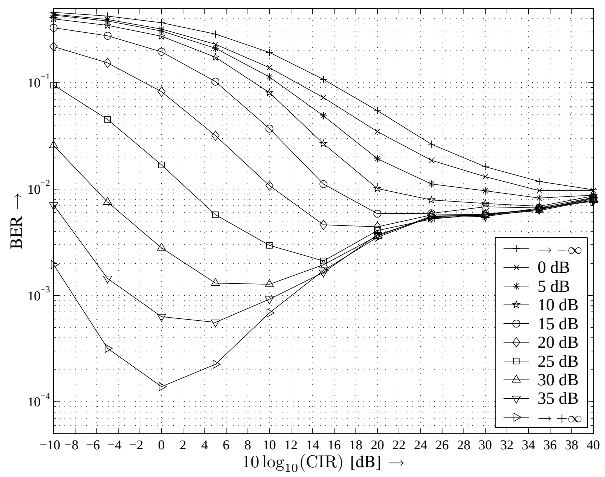

Fig. 3. M-Algorithm (64) and receiver strategy A ($\xi = 1.2\%$): BER vs. $10\log_{10}(\mathrm{CIR})$ for different parameters $10\log_{10}(\mathrm{DIR})$ and 8PSK, $10\log_{10}(E_b/N_0) = 30\,\mathrm{dB}$.

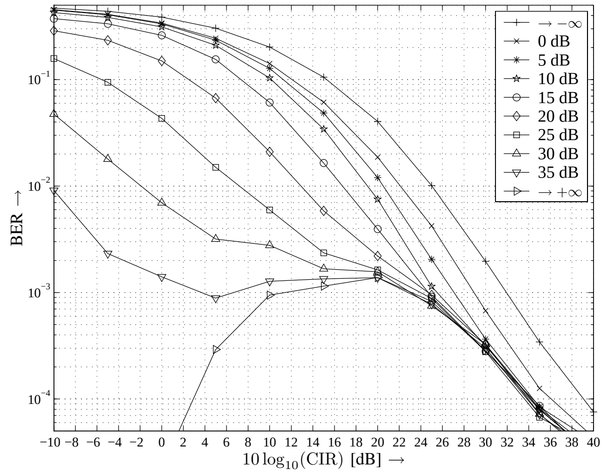

Fig. 4. M-Algorithm (64) and receiver strategy B: BER vs. $10\log_{10}(\mathrm{CIR})$ for different parameters $10\log_{10}(\mathrm{DIR})$ and 8PSK, $10\log_{10}(E_b/N_0) = 30\,\mathrm{dB}$.

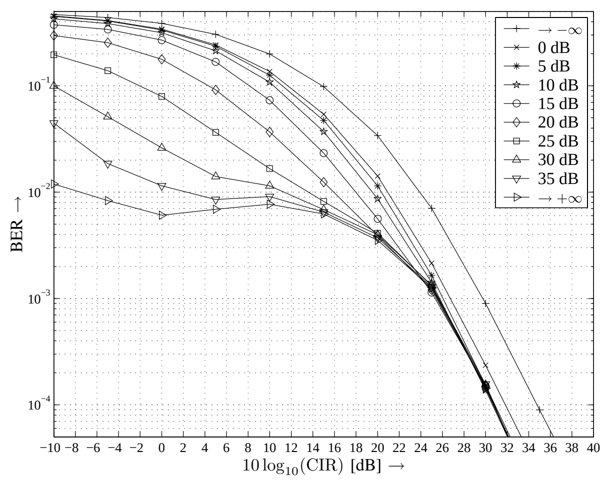

Fig. 5. DDFSE (8,8) and receiver strategy B: BER vs. $10\log_{10}(\mathrm{CIR})$ for different parameters $10\log_{10}(\mathrm{DIR})$ and 8PSK, $10\log_{10}(E_b/N_0) = 30\,\mathrm{dB}$.

levels. The receiver according to strategy B employs the prefilter described in Section III, which is adapted to the given scenario. With this filter, always a good trade-off between impulse concentration and remaining interference is obtained.

Simulation results for the M-algorithm using receiver strategy A are given in Fig. 3, showing good performance for moderate–to–high interference power levels, but worse performance for low interference power levels since the energy of the ignored precursors of the desired user evokes an irreducible error floor. Results for the M-algorithm using strategy B are shown in Fig. 4. High performance is revealed for a wide range of CIR and DIR values. Fig. 5 shows the results for JDDFSE $(8, 8)$ and strategy B. JDDFSE exhibits good performance for low interference power levels, but looses performance in region of high interference power.

An additional gain in performance (results not shown here) of about 1.5–2 dB can be obtained using separate forward and backward detection and choosing the detected sequence with lowest path metric in the end.

VI. CONCLUSIONS

Joint detection reveals a performance advantage compared to traditional single user detection methods at the expense of some additional complexity, if applied to GSM/ EDGE. Our numerical results for perfect channel state information demonstrate a better performance of the M-algorithm for multiuser joint detection in comparison to JDDFSE. By introducing a novel prefiltering method, detection outperforms receiver strategies without prefiltering or with conventional prefiltering, e.g. [3]. In future work, joint detection with channel estimation will be addressed.

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