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Impulse Response Shortening for Multiple Co-Channels

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Abstract—In this paper, a novel design of a receiver-side FIR prefilter is proposed which is able to jointly shorten the impulse responses of multiple co-channels given just one receive antenna. The prefilter coefficients are computed without explicitly solving an eigenvalue problem. In conjunction with a reduced-complexity multiuser detector, a good error performance can be achieved. As a possible application, co-channel interference cancellation for the EDGE system is studied. The prefilter, however, is also suitable for interference suppression in MIMO systems, for crosstalk suppression, or for reducing the length of the cyclic prefix in multitone systems, among other applications.

Index Terms—Wireless communication, cellular radio, co-channel interference cancellation, joint prefiltering, joint channel estimation, joint equalization, multiuser detection.

I. INTRODUCTION

In urban areas and in hot spots (like train stations and airports), the capacity of TDMA cellular radio networks such as GSM/EDGE is limited by co-channel interference (CCI) rather than noise, unless co-channel interference cancellation is done. Co-channel interference cancellation is particularly challenging in the downlink, where usually (due to cost, volume, power consumption, and design reasons) only one receive antenna is available. Corresponding techniques are called single-antenna (co-channel) interference cancellation (SAIC) techniques [1], [13]. A prominent class of SAIC techniques is based on multiuser detection, where the desired user and the interferer(s) are treated jointly. In case of trellis-based joint maximum-likelihood sequence estimation or joint maximum a posteriori detection [9], [10], the number of states is \(M^{(J+1)L} \), where \(M\) is the cardinality of the symbol alphabet, \(J\) is the number of active interferers, and \(L\) is the effective memory length of the equivalent discrete-time channel model. In case of GSM (\(M = 2\)), one dominant interferer (\(J = 1\)) is about handable in typical urban areas (\(L \approx 3\)). However in case of EDGE (\(M = 8\)), the computational complexity is prohibitive even in the presence of just one interferer. Hence, for the EDGE system reduced-state versions have to be applied, see e.g. [5], [12], [19]. Similar to reduced-state single-user detectors (such as delayed-decision feedback sequence estimation (DDFSE) [6] and reduced-state sequence estimation (RSSE) [7]), a major drawback of reduced-state joint detectors is their sensitivity with respect to the power distribution of the overall channel impulse responses (CIRs) seen by the detector, however. This problem can be solved by using a reduced-state joint detector in conjunction with a prefilter, which shortens the overall impulse response of all co-channels.

The design of a channel-shortening prefilter is a challenging task, if more co-channels than available receive antennas shall be jointly shortened. Related work has been published for different applications:

- **OFDM**: In [3], [4], [15], [17], [23], joint impulse response shortening has been investigated for multitone (OFDM) systems. The aim is to reduce the length of the cyclic prefix.
- **MIMO systems**: Channel shortening for MIMO systems has been studied in [3], [4], [21], [22], [24], for example.
- **CDMA**: A channel-shortening multiuser detector for direct-sequence CDMA systems has been proposed in [16].
- **CCI suppression**: Using an algorithm that is similar\(^1\) to the one in [17], in [11] the problem of co-channel interference suppression is tackled.

In all these previous works, an eigenvalue problem has explicitly to be solved.

In [8], [18], [25], prefiltering for reduced-state equalization of MIMO channels is considered. The prefilter transforms the MIMO channel to its minimum-phase (or maximum-phase) equivalent. This approach is conceptually different from channel shortening.

\(^1\)Unfortunately, no further details concerning the prefilter design are presented in [11].
In this paper, an alternative prefilter design for the purpose of simultaneously shortening the overall CIRs of all co-channels is proposed, which is based on a joint minimum mean squared error (MMSE) decision-feedback equalizer. The novel prefilter generalizes the prefilter proposed in [14], which considers only a single user. As a possible application, joint shortening of the impulse responses of two co-channels (i.e., the channels of the desired user and of the dominant interferer) are considered. The results may, however, easily be extended to more than two co-channels or to related applications, including crosstalk suppression, suppression of interference from multiple transmit antennas, and shortening of the cyclic prefix in multitone systems.

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.

II. CHANNEL MODEL

In the presence of one dominant interferer, the equivalent discrete-time channel model (before prefiltering) considered in this paper is denoted as (cf. Fig. 1)

\[ y[k] = \sum_{l=0}^{L} h_t a[k-l] + \sum_{l=0}^{L} g_t b[k-l] + n[k], \quad 0 \leq k \leq K-1, \]

(1)

where \( y[k] \in \mathcal{C} \) is the \( k \)-th baud-rate output sample of the analog receive filter, \( L \) is the effective memory length of the discrete-time ISI channel model before prefiltering, \( h_t \in \mathcal{C} \) are the channel coefficients of the desired user, \( g_t \in \mathcal{C} \) are the channel coefficients of the dominant interferer, \( a[k] \) and \( b[k] \) are the \( k \)-th i.i.d. data symbols of the desired user and the dominant interferer, respectively, both randomly drawn over an \( M \)-ary alphabet \( E\{a[k]\} = E\{b[k]\} = 0, E\{|a[k]|^2\} = E\{|b[k]|^2\} = 1 \), \( n[k] \in \mathcal{C} \) is the \( k \)-th sample of a Gaussian noise process \( E\{|n[k]|^2\} = 0 \), \( E\{|n[k]|^2\} := \sigma_n^2 = N_0/E_s \), \( k \) is the time index, and \( K \) is the number of \( M \)-ary data symbols per burst. All random processes are assumed to be mutually independent. The channel coefficients \( h := [h_0, \ldots, h_L]^T \) \((E\{|h|^2\} = 1)\) and \( g := [g_0, \ldots, g_L]^T \) \((E\{|g|^2\} = 1/(C/I))\) comprise pulse shaping, the respective physical channel, analog 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 process is white. In order to derive the prefilter, the channel coefficients are assumed to be time-invariant within a burst but may entirely change from burst to burst (block fading assumption), and a synchronous network is assumed. In case of an asynchronous network with frequency hopping, a piece-wise filter design is necessary.

III. RECEIVER STRUCTURE

The receiver structure under investigation is shown in Fig. 2. It consists of the prefilter under investigation, a joint channel estimator (Joint-CE), and a nonlinear joint detector.

The baud-rate output samples \( \{y[k]\} \) of the analog receive filter and estimates \( \hat{h} \) and \( \hat{g} \) of the channel coefficients \( h \) and \( g \) are used to design the discrete-time channel model before prefiltering, \( h_t \in \mathcal{C} \) are the channel coefficients of the desired user, \( g_t \in \mathcal{C} \) are the channel coefficients of the dominant interferer, \( a[k] \) and \( b[k] \) are the \( k \)-th i.i.d. data symbols of the desired user and the dominant interferer, respectively, both randomly drawn over an \( M \)-ary alphabet \( E\{a[k]\} = E\{b[k]\} = 0, E\{|a[k]|^2\} = E\{|b[k]|^2\} = 1 \), \( n[k] \in \mathcal{C} \) is the \( k \)-th sample of a Gaussian noise process \( E\{|n[k]|^2\} = 0 \), \( E\{|n[k]|^2\} := \sigma_n^2 = N_0/E_s \), \( k \) is the time index, and \( K \) is the number of \( M \)-ary data symbols per burst. All random processes are assumed to be mutually independent. The channel coefficients \( h := [h_0, \ldots, h_L]^T \) \((E\{|h|^2\} = 1)\) and \( g := [g_0, \ldots, g_L]^T \) \((E\{|g|^2\} = 1/(C/I))\) comprise pulse shaping, the respective physical channel, analog 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 process is white. In order to derive the prefilter, the channel coefficients are assumed to be time-invariant within a burst but may entirely change from burst to burst (block fading assumption), and a synchronous network is assumed. In case of an asynchronous network with frequency hopping, a piece-wise filter design is necessary.

Fig. 1. Equivalent time discrete channel model in the presence of one dominant interferer.

Fig. 2. Receiver structure under investigation.
IV. PREFILTER DERIVATION

Fig. 3 shows a joint decision-feedback equalizer\(^2\) (JDFE) for the case of one desired user and one interferer\(^3\). The task of the feedforward filter \(f\) of the JDFE is to shorten the channel impulse response of the desired user and the channel impulse response of the dominant interferer jointly. A symbol-spaced FIR feedforward filter is assumed throughout this paper. After suitable filtering, effectively \(L_1 + 1\) and \(L_2 + 1\) channel coefficients remain. The last consecutive \(L_1\) and \(L_2\) channel coefficients are eliminated by the corresponding feedback filters. Hence, the feedforward filter of the JDFE corresponds to the desired prefilter. The hard decisions \(\hat{a}[k]\) and \(\hat{b}[k]\) are obtained as (cf. Fig. 3)

\[
[\hat{a}[k-k_0], \hat{b}[k-k_0]] = \arg \min \left[\hat{a}[k-k_0], \hat{b}[k-k_0]\right] \left| z[k] - \hat{h}_{s,0} \hat{a}[k-k_0] - \sum_{l=1}^{L_1} \hat{h}_{s,l} \hat{a}[k-k_0 - l] - \hat{g}_{s,0} \hat{b}[k-k_0] - \sum_{l=1}^{L_2} \hat{g}_{s,l} \hat{b}[k-k_0 - l]\right|^2,
\]

where \(k_0\) is the decision delay of the JDFE.

A nonlinear joint detector attached to the feedforward filter must be able to handle \(L_1 + 1\) and \(L_2 + 1\) channel coefficients of the desired user channel and the interferer channel, respectively. For example, a joint Viterbi detector with \(M^{L_1+L_2}\) states may be applied, where \(M\) denotes the cardinality of the symbol alphabet.

Assuming i.i.d. data, after some calculation the following relations are obtained:

\[
\hat{f}^H R_{yy} = \hat{h}^H_{s,0} R_{ya} + \hat{g}_{s,0}^H R_{yb} + \hat{h}_{s,0}^H r_{ya} + \hat{g}_{s,0}^H r_{yb}
\]

(6)

\[
\hat{h}_s = \hat{f}^H R_{ya}
\]

(7)

\[
\hat{g}_s = \hat{f}^H R_{yb},
\]

(8)

where \(R_{yy} := E\{yy^H\}\) denotes the \((L_f + 1) \times (L_f + 1)\) autocorrelation matrix of the received samples, \(R_{ya} := E\{ay^H\}\) denotes the \(L_1 \times (L_f + 1)\) cross-correlation matrix between the desired user data and the received samples, \(R_{yb} := E\{by^H\}\) denotes the \(L_2 \times (L_f + 1)\) cross-correlation matrix between the interference user data and the received samples, and \(r_{ya} := E\{a^*[k-k_0]y\}\) and \(r_{yb} := E\{b^*[k-k_0]y\}\) are \((L_f + 1) \times 1\) vectors. For i.i.d. data symbols, the elements of \(R_{yy}, R_{ya}, R_{yb}, r_{ya}\), and \(r_{yb}\) can be calculated as

\[
r_{yy}(i,j) = \sum_{l=0}^{L_1} h_{l}^* h_{l+i-j} + \sum_{l=0}^{L_2} g_{l}^* g_{l+i-j} + \sigma_n^2 \delta_{i-j},
\]

where \(r_{ya}(i,j) = h_{k_0+i-j}^* y_{a}(i,j) = \sqrt{1/(C^H C)} g_{k_0+i-j}^* y_{a}(i,j) = h_{k_0+i-j}^* y_{a}(i,j)\), and \(r_{yb}(i,j) = g_{k_0+i-j}^* y_{a}(i,j)\), respectively. This result justifies the use of a joint\(^4\) channel estimator for calculating the prefilter coefficients.

\(^2\)Note that a JDFE is the simplest special case of a JDDFSE [12].

\(^3\)As opposed to the JDFE proposed in [2], the output of our feedforward filter is a scalar rather than a vector.

\(^4\)The joint channel estimator may be replaced by a bank of conventional (single-user) channel estimators, if the training sequences of all co-channels are mutually orthogonal.
Upon insertion of (7) and (8) in (6) the optimum filter coefficients are obtained as
\[
\hat{f}^H = (\hat{h}_{s,0} r_{ya} + \hat{g}_{s,0} r_{yb}) \left[ R_{yy} - R_{ya} R_{ya} - R_{yb} R_{yb} \right]^{-1}.
\] (9)

Note that (9), which is the key formula here, is highly symmetrical. This symmetry immediately suggests a possible generalization to the case of more than two co-channels. Vice versa, for \( r_{yb} = 0 \) and \( R_{yb} = 0 \) the corresponding prefilter for a single channel as proposed in [14] is obtained.

The shortened channel coefficients \( \hat{h}_s \) and \( \hat{g}_s \) are obtained by inserting (9) into (7) and (8), respectively. Hence, in the proposed solution the filter coefficients are computed prior to the shortened channel coefficients. According to (7) and (8), the shortened channel coefficients \( \hat{h}_s \) and \( \hat{g}_s \) can be written as a convolution between the prefilter coefficients, \( \hat{f} \), and the original channel coefficients, \( h \) and \( g \), respectively:
\[
\hat{h}_s = \hat{f} * \hat{h} \quad (10)
\]
\[
\hat{g}_s = \hat{f} * \hat{g} \quad (11)
\]
The symbol-spaced shortened channel coefficients \( \hat{h}_s \) and \( \hat{g}_s \) (incl. \( \hat{h}_{s,0} \) and \( \hat{g}_{s,0} \)) are finally provided to the nonlinear joint detector, cf. Fig. 2. Note that the feedback filters and the memoryless detector featured in Fig. 3 do not have to be implemented.

V. NUMERICAL RESULTS

For the numerical results presented in Fig. 4 to Fig. 8 the following scenario is assumed:
- Synchronous EDGE network
- 8-PSK modulation (\( M = 8 \))
- \( J = 1 \) dominant EDGE interferer
- Uniformly distributed training sequence code (TSC) for desired user and interferer
- TSC of dominant interferer differs from TSC of desired user
- Typical urban (TU) channel model according to GSM 05.05; square-root Nyquist receive filter; effective memory length before prefiltering: \( L = 3 \); block fading
- Signal-to-noise ratio: \( E_s/N_0 = 30 \) dB (This is a typical value in urban areas. GSM/EDGE is interference limited rather than noise limited.)
- MMSE prefilter of order \( L_f = 19 \) (unless otherwise stated)
- Effective memory length after prefiltering: \( L_1 = 1 \) and \( L_2 = 1 \)
- \( \hat{h}_{s,0} = 1 \) (monic CIR) and \( \hat{g}_{s,0} = \sqrt{I/(C/I)} \)
- Decision delay of JDFE: \( k_0 = (L_f + 1)/2 \)
- \( 5 \cdot 10^4 \) statistically independent bursts.

The GSM 05.05 TU channel model is typically used in the context of SAIC for GSM and EDGE.

The average power of the overall (shortened) channel impulse response of the desired user and the dominant interferer are presented for two different \( C/I \) values in Fig. 4 to Fig. 7.
In Fig. 8, the raw bit error rate versus C/I for joint Viterbi detection with 64 states in conjunction with and without prefiltering is presented.

VI. SUMMARY AND CONCLUSIONS

We conclude from the numerical results that the proposed prefilter performs well, even for C/I = 0 dB. For a reduced-state multiuser detector without prefilter, the raw BER is unacceptable. Compared to the related techniques proposed in [3], [4], [11], [15], [16], [17], [21], [22], [23], [24], no eigenvalue problem has to be solved explicitly. Furthermore, the proposed algorithm is numerically stable. The JDFE-based design generalizes the corresponding single-user prefilter in an elegant way and gives nice insights.

REFERENCES


