Training Sequence Aided Multichannel Identification in the Presence of Interference and Noise

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ABSTRACT

In wireless communications, spatial (via antenna arrays) and temporal (excess bandwidth) diversity may be exploited to simultaneously equalize a user of interest while canceling or reducing (cochannel) interfering users (CCI). This can be done using the Interference Canceling Matched Filter (ICMF) which we introduced previously. The ICMF depends on the channel for the user of interest, to be estimated with a training sequence, and contains a blind interference cancellation part. The simulations show that significant improvements may result from the exploitation of the prior knowledge of the transmission pulse shape.

I. PROBLEM FORMULATION

Consider the linearized version of the GMSK modulation transmitted over a linear channel with additive noise. The cyclostationary received signal can be written as

$$y'(t)=\sum_k h'(t-kT)b_k+v'(t)=\sum_k h'(t-kT)j^k d_k+v'(t) \quad (1)$$

where $h'(t)$ is the combined impulse response of the modulation $f'(t)$ [1] and the channel $c'(t)$, ie.,$h'(t) = f'(t) * c'(t)$ and d_k are the BPSK symbols. The channel impulse response $h'(t)$ is assumed to be FIR with duration NT. If K sensors are used and each sensor waveform is oversampled at the rate $\frac{p}{T}$, the discrete-time input-output relationship at the symbol rate can be written as:

$$\mathbf{y}'_k = \sum_{i=0}^{N-1}\mathbf{h}'_i b_{k-i} + \mathbf{v}'_k = \mathbf{H}'_N B_N(k) + \mathbf{v}'_k,$$
$$\mathbf{y}'_k = \begin{bmatrix} y'_{1,k} \\ \vdots \\ y'_{m,k}\end{bmatrix}, \mathbf{v}'_k = \begin{bmatrix} v'_{1,k} \\ \vdots \\ v'_{m,k}\end{bmatrix}, \mathbf{h}'_k = \begin{bmatrix} h'_{1,k} \\ \vdots \\ h'_{m,k}\end{bmatrix}$$
$$\mathbf{H}'_N = \left[\mathbf{h}'_{N-1}\cdots\mathbf{h}'_0\right], B_N(k) = \left[b^H_{k-N+1}\cdots b^H_k\right]^H \quad (2)$$

where the first subscript i denotes the i^{th} channel, $m = pK$, and superscript ^{H} denotes Hermitian transpose. We have introduced the p phases of the K oversampled antenna signals: ${y'}_{(n-1)p+l,k} = {y'}_n(t_0+(k+\frac{l}{p})T)$, $n = 1,\ldots,K$, $l = 1,\ldots,p$ where ${y'}_n(t)$ is the signal received by antenna n.

The propagation environment is described by a channel $c'(t) = \left[{c'}^H_1(t)\cdots{c'}^H_K(t)\right]^H$. We consider GSM channel models which are taken as specular multipath channels with L_c paths of the form ${c'}_n(t) = \sum_{r=1}^{L_c} a_{r,n}\delta(t-\tau_{r,n})$ for the n^{th} antenna. $a_{r,n}$ and $\tau_{r,n}$ are the amplitude and the delay of path r. The distribution of the amplitudes and the values of the delays depend on the propagation environment (urban, rural, hilly terrain). We consider independent channel realizations for the K antennas. The received signal for antenna n can be written as

$${y'}_n(t)=\sum_{k=-\infty}^{+\infty} {h'}_n(t-kT)b_k\,, \quad {h'}_n(t)=\sum_{r=1}^{L_c} a_{r,n}f'(t-\tau_{r,n}). \quad (3)$$

The continuous-time channel ${h'}_n(t)$ for the n^{th} antenna when sampled at the instant $t_0+(k+\frac{l}{p})T$ yields the $((n-1)p+l)^{th}$ component of the vector ${\mathbf{h}'}_k$.

The constellation for the symbols b_k, the inputs to the discrete-time multichannel, is complex whereas the constellation for the symbols d_k is real. It will be advantageous to express everything in terms of real quantities and in this way double the number of (fictitious) channels. To that end we demodulate the received signal by j^{-k} [2]:

$$j^{-k}\mathbf{y}'_k=\sum_{i=0}^{N-1} j^{-k}\mathbf{h}'_i b_{k-i} + j^{-k}\mathbf{v}'_k = \sum_{i=0}^{N-1}(j^{-i}\mathbf{h}'_i)d_{k-i} + j^{-k}\mathbf{v}'_k \quad (4)$$

and then we decompose the complex quantities into their real and imaginary parts like

$$\begin{cases} \mathbf{y}^R_k = Re(j^{-k}\mathbf{y}'_k) = & \mathbf{H}^R(q)d_k + \mathbf{v}^R_k \\ \mathbf{y}^I_k = Im(j^{-k}\mathbf{y}'_k) = & \mathbf{H}^I(q)d_k + \mathbf{v}^I_k \end{cases} \quad (5)$$

where q^{-1} is the delay operator: $q^{-1}\,\mathbf{y}_k = \mathbf{y}_{k-1}$ and $\mathbf{H}^R(q) = \sum_{i=0}^{N-1}\mathbf{h}^R_i q^{-i} = \sum_{i=0}^{N-1} Re(j^{-i}\mathbf{h}'_i)q^{-i}$ and similarly for $\mathbf{H}^I(q)$. We can represent this system more conveniently in the following obvious notation

$$\mathbf{y}_k = \begin{bmatrix} \mathbf{y}_k^R \\ \mathbf{y}_k^I \end{bmatrix} = \begin{bmatrix} \mathbf{H}^R(q) \\ \mathbf{H}^I(q) \end{bmatrix} d_k + \begin{bmatrix} \mathbf{v}_k^R \\ \mathbf{v}_k^I \end{bmatrix} = \mathbf{H}(q) d_k + \mathbf{v}_k \ . \tag{6}$$

The term $\mathbf{v}_k$ will be considered here to consist of both spatially and temporally correlated additive zero mean noise. In simulations, we often assume $\mathbf{v}_k$ to consist of temporally and spatially i.i.d. noise plus co-channel multi-user interference. This would mean that the additive noise $v'(t)$ is a combination of stationary and cyclostationary components with period T. When the noise consists of multiuser interference plus Gaussian noise, the optimal receiver performs joint detection of all users. However, the estimation of the matrix transfer function from all users to all antennas (and/or sampling phases) is a formidable and often prohibitive task. Furthermore, the complexity of MLSE can be enormous in this case. Therefore we shall concentrate on the detection of one user of interest, ignore the discrete distribution of the interferers and approximate them with a Gaussian distribution. We shall assume that the channel transfer functions for the interferers are also FIR and that their symbol sequences are uncorrelated. Hence we assume that $\mathbf{v}_k$ is a multivariate MA(N'–1) process.

In the blind estimation problem on the basis of a burst of received data $\mathbf{Y}_M(M) = \mathcal{T}_M(\mathbf{H}) D_{M+N-1}(M) + \mathbf{V}_M(M)$ or $\mathbf{Y} = \mathcal{T}(\mathbf{H}) D + \mathbf{V}$, where $\mathbf{Y} = [\mathbf{y}_1^H \cdots \mathbf{y}_M^H]^H$ and similarly for $\mathbf{V}$, and $\mathcal{T}(\mathbf{H})$ is a block Toeplitz matrix with M block rows and $[\mathbf{h}_{N-1} \cdots \mathbf{h}_0 \;\; 0_{2m \times (M-1)}]$ as first block row, the unknown parameters are the channel $\mathbf{H}$, the transmitted symbols D and the noise correlation sequence $r_{\mathbf{VV}}(0 \!:\! N'\!-\!1)$. However, this ensemble of unknown parameters is unidentifiable from the received data $\mathbf{Y}$. A training sequence, i.e. a subset of known transmitted symbols, has to be available to enable estimation of all unknown parameters. In [3], which builds upon previous work as discussed in [3], a two-step procedure was proposed in which the training sequence was used to estimate the channel $\mathbf{H}$ via least-squares (as is usually done for training-sequence based channel estimation). A set of parameters equivalent to $r_{\mathbf{VV}}(0 \!:\! N'\!-\!1)$ in a filtering structure called the Interference Canceling Matched Filter (ICMF) was then estimated blindly. The remaining symbols can then be estimated using any of the existing receiver techniques that are based on known channel and noise statistics.

II. TX/RX FILTER KNOWLEDGE

The prior information of the transmit and receive filter used in [4] to perform a training sequence based structured channel estimation is shown to give good performance with the Viterbi Algorithm. Alternatively, one can review, and then adapt, the derivation of [5] to the GSM case. Consider a certain oversampling factor p, and let the oversampled transfer function $H'_n(z) = C'_n(z) F'(z)$ of the overall channel for the n^{th} antenna be the cascade of the actual oversampled anti-aliasing filtered channel $C'_n(z)$ and the oversampled combined TX/RX filter $F'(z)$ (the oversampling factor should satisfy the Nyquist criterion for the TX/RX filter). Each of these transfer functions can be decomposed into its polyphase components at the symbol rate, e.g. $H'_n(z) = \sum_{i=0}^{p-1} z^{-i} H'_{n,i}(z^p)$. These components can also be represented in the SIMO form, $\mathbf{F}'(z) = [F'^H_1(z) \cdots F'^H_p(z)]^H = \sum_{k=0}^{L_f-1} \mathbf{f}'(k) z^{-k}$ and $\mathbf{C}'_n(z) = [C'^H_{n,1}(z) \cdots C'^H_{n,p}(z)]^H = \sum_{k=0}^{L-1} \mathbf{c}'(k) z^{-k}$ with $L_f + L - 1 = N$. The relations between the polyphase components can be obtained from

$$\sum_{i=0}^{p-1} z^{-i} H'_{n,i}(z^p) = \sum_{k=0}^{p-1} z^{-k} F'_k(z^p) \sum_{l=0}^{p-1} z^{-l} C'_{n,l}(z^p) \tag{7}$$

In particular for $p = 2$ we get

$$\begin{bmatrix} H'_{n,0}(z) \\ H'_{n,1}(z) \end{bmatrix} = \begin{bmatrix} F'_0(z) & z^{-1} F'_1(z) \\ F'_1(z) & F'_0(z) \end{bmatrix} \begin{bmatrix} C'_{n,0}(z) \\ C'_{n,1}(z) \end{bmatrix} = \begin{bmatrix} C'_{n,0}(z) & z^{-1} C'_{n,1}(z) \\ C'_{n,1}(z) & C'_{n,0}(z) \end{bmatrix} \begin{bmatrix} F'_0(z) \\ F'_1(z) \end{bmatrix} \tag{8}$$

or $\mathbf{H}'_n(z) = \underline{\mathbf{F}}'(z) \mathbf{C}'_n(z) = \underline{\mathbf{C}}'_n(z) \mathbf{F}'(z)$. In the time domain, we get

$$\mathcal{T}_M(\mathbf{H}'_n) = \mathcal{T}_M(\underline{\mathbf{F}}') \mathcal{T}_{M+L_f-1}(\mathbf{C}'_n) \tag{9}$$

where $\mathcal{T}_M(\mathbf{X})$ is a block Toeplitz matrix with M block rows and $[\mathbf{X} \;\; 0_{p \times (M-1)q}]$ as first block row, $\mathbf{X}$ being considered as a block row vector with $p \times q$ blocks, $\mathbf{C}'_n$ is similar to $\mathbf{H}'_n$ and

$$\underline{\mathbf{F}}' = \left[\underline{f}'(L_f - 1) \cdots \underline{f}'(0)\right], \; \underline{f}'(k) = \begin{bmatrix} f'_0(k) & f'_1(k-1) \\ f'_1(k) & f'_0(k) \end{bmatrix} \tag{10}$$

and we assume $f'_{p-1}(L_f - 1) = 0$.

Now from (4), one should write for the GSM case $\mathbf{H}'_n(jz) = \underline{\mathbf{F}}'(jz) \mathbf{C}'_n(jz)$ and then we decompose the complex quantities into their real and imaginary parts like

$$\begin{bmatrix} \mathbf{H}_n^R(z) \\ \mathbf{H}_n^I(z) \end{bmatrix} = \begin{bmatrix} Re(\underline{\mathbf{F}}'(jz)) & -Im(\underline{\mathbf{F}}'(jz)) \\ Im(\underline{\mathbf{F}}'(jz)) & Re(\underline{\mathbf{F}}'(jz)) \end{bmatrix} \begin{bmatrix} \mathbf{C}_n^R(z) \\ \mathbf{C}_n^I(z) \end{bmatrix} \tag{11}$$

Then, for the n^{th} antenna, $\mathbf{H}_n(z) = \underline{\underline{\mathbf{F}}}'(z) \mathbf{C}_n(z)$, where $\underline{\underline{\mathbf{F}}}'(z) = \sum_{k=0}^{L_f-1} \underline{\underline{f}}'(k) z^{-k}$, $\underline{\underline{f}}'(k) = J^k \otimes \underline{f}'(k)$ and

$J = \begin{bmatrix} 0 & 1 \\ -1 & 0 \end{bmatrix}$.

In the case of an array of K antennas, $\mathbf{H}(z) = \underline{\underline{\mathbf{F}}}(z)\mathbf{C}(z)$ and $\underline{\underline{\mathbf{F}}}(z) = \underline{\underline{\mathbf{F}}}'(z) \otimes I_K$ where I_K is the K by K identity matrix.

III. PARTIAL EQUALIZATION OF THE TX FILTER

Since the overall channel is the cascade of the TX filter and the propagation channel, one can equalize the TX filter which is known by the receiver. This yields a reduced-order channel model that is suitable for the implementation in the Viterbi (or any linear) receiver. However, it is known that the GMSK filter is too hard to be inverted by a Zero Forcing (ZF) or a Minimum Mean-Square-Error (MMSE) Linear Equalizer (LE). The resulting equalizer is too long to be implemented in a real burst processing application as it is the case for Mobile Communications and, more importantly, the noise enhancement is considerable. One solution to overcome this problem is to equalize the GMSK filter to a shorter filter and not to a single tap as it is usually done in an equalization framework. Hence, the name "partial equalization"(PE) [6]. Let $\mathbf{f}' = [f'_0(0) \cdots f'_{p-1}(0) \cdots f'_0(L_f-1) \cdots f'_{p-1}(L_f-1)]$ be the 1 by pL_f GMSK filter, $\mathbf{g} = [g_0(0) \cdots g_{p-1}(0) \cdots g_0(L_g-1) \cdots g_{p-1}(L_g-1)]$ the 1 by pL_g equalizer and $\mathbf{b} = [b(0) \cdots b(L_b-1)]$ the 1 by L_b desired, in the mean square sense, impulse response of the combined TX filter-equalizer impulse response. Now, we define by $\mathcal{F}' = \mathcal{T}_{pL_g}(\mathbf{f}')$ a Toeplitz matrix with pL_g rows and $[\mathbf{f}' \;\; 0_{1\times(pL_g-1)}]$ as first row that can be partitionned into $\overline{\mathcal{F}}' = \mathcal{F}'\overline{\mathcal{I}}$ and $\overline{\overline{\mathcal{F}}}' = \mathcal{F}'\overline{\overline{\mathcal{I}}}$ where $\overline{\mathcal{I}} = \begin{bmatrix} 0_{\Delta,L_b} \\ I_{L_b} \\ 0_{s,L_b} \end{bmatrix}$, $\overline{\overline{\mathcal{I}}} = \overline{\mathcal{I}}^{\perp} = \begin{bmatrix} I_\Delta & 0_{\Delta,s} \\ 0_{L_b,\Delta} & 0_{L_b,s} \\ 0_{s,\Delta} & I_s \end{bmatrix}$ and $s = pL_g + pL_f - 1 - \Delta - L_b$. Δ accounts for a desirable delay.

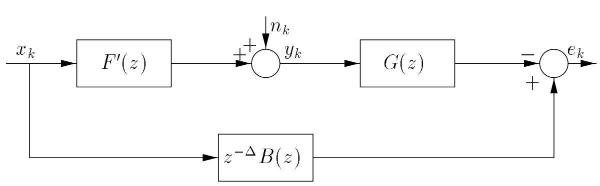

Figure 1. Block diagram for a MMSE Linear Partial Equalizer (LPE).

Referring to Fig. 1, the coefficients of the equalizer and the desired impulse response (assumed to be obtained with no error) are chosen to minimize w.r.t. $\mathbf{g}$ the variance of the noise e_k: $\sigma_n^2\|\mathbf{g}\|^2 + \sigma_x^2\|\mathbf{g}\overline{\overline{\mathcal{F}}}'\|^2$ under the constraint $\|\mathbf{g}\overline{\mathcal{F}}'\|^2 = 1$ and uncorrelated input symbols (x_k).

Then the equalizer $\mathbf{g}$ is obtained as the Hermitian transpose of the generalized eigenvector that is associated to the minimal eigenvalue of the matrices $\sigma_n^2 I_{pL_g} + \sigma_x^2 \overline{\overline{\mathcal{F}}}'\overline{\overline{\mathcal{F}}}'^H$ and $\overline{\mathcal{F}}'\overline{\mathcal{F}}'^H$.

In [6], the authors minimize the mean square of the error sequence e_k. Under the assumptions of a unit-energy constraint of $\mathbf{b}$ and uncorrelated input symbols, they have shown that the optimum desired impulse response is equal to the unit-norm eigenvector that corresponds to the minimum eigenvalue of the matrix $\mathbf{R}_\Delta = \overline{\mathcal{I}}^H \left(\frac{I_{pL_g+pL_f-1}}{\sigma_x^2} + \frac{\mathcal{F}'^H\mathcal{F}'}{\sigma_n^2}\right)^{-1}\overline{\mathcal{I}}$.

Once $\mathbf{b}$ is determined, the minimum mean square error unit energy constrained equalizer coefficients are calculated from $\mathbf{g} = [0_{1,\Delta}\; \mathbf{b}\; 0_{1,s}]\, \mathcal{F}'^H \left(\mathcal{F}'\mathcal{F}'^H + \frac{\sigma_n^2}{\sigma_x^2} I_{pL_g}\right)^{-1}$.

In our simulations ($p = 2$ and $L_f = 4$), the two approaches are equivalent but in general we claim that the first one should work slightly better.

Back to the time-domain, we shall filter the received burst by $\underline{\underline{\mathbf{G}}}(z) = \underline{\underline{\mathbf{G}}}'(z) \otimes I_K$ where $\underline{\underline{\mathbf{G}}}'(z) = \sum_k \underline{\underline{g}}'(k) z^{-k}$, $\underline{\underline{g}}'(k) = J^k \otimes \underline{g}'(k)$ and $\underline{g}'(k)$ is similarly defined as $\underline{f}'(k)$. For $p = 2$, we get $\underline{g}'(k) = \begin{bmatrix} g_0(k) & g_1(k-1) \\ g_1(k) & g_0(k) \end{bmatrix}$.

The reduced-order model is then

$$\mathbf{z}_k = z^{\Delta}\underline{\underline{\mathbf{G}}}(q)\mathbf{y}_k = \mathbf{Q}(q)d_k + \boldsymbol{v}_k, \tag{12}$$

where $\mathbf{Q}(z) = \underline{\underline{\mathbf{B}}}(z)\mathbf{C}(z)$, $\underline{\underline{\mathbf{B}}}(z) = z^{\Delta}\underline{\underline{\mathbf{G}}}(z)\underline{\underline{\mathbf{F}}}(z)$ and $\boldsymbol{v}_k = z^{\Delta}\underline{\underline{\mathbf{G}}}(q)\mathbf{v}_k$.

IV. IMPLEMENTATION ISSUES

As far as the design of the various filters is concerned, the channel transfer function $\mathbf{Q}(z)$ can be estimated with the training sequence for the user of interest. From $\mathbf{Q}(z)$, one can determine the whitened matched filter and the blocking equalizers $\mathbf{Q}^{\perp\dagger}(z)$. The theoretical expression for $W(z) = S_{x_1\mathbf{X}_2} S^{-1}_{\mathbf{X}_2\mathbf{X}_2}$ is

$$W(z) = \mathbf{Q}^{\dagger}(z) S_{\mathbf{ZZ}}(z)\mathbf{Q}^{\perp}(z)\left(\mathbf{Q}^{\perp\dagger}(z) S_{\mathbf{ZZ}}(z)\mathbf{Q}^{\perp}(z)\right)^{-1}. \tag{13}$$

In the noiseless case, $W(z)$ satisfies,

$$W(z)\mathbf{Q}^{\perp\dagger}(z)\mathbf{Q}_d(z) = \mathbf{Q}^{\dagger}(z)\mathbf{Q}_d(z), \tag{14}$$

where $\mathbf{Q}_d(z)$ ($2m \times d$) regroups the channel transfer functions of the $d \le 2m-1$ interferers.

This system of equations allows an FIR solution for $W(z)$ if the number of interferers is limited to $d \le 2m-2$. The optimal length of $W(z)$ is

$$L_w \ge \left\lceil \frac{2(N_q-1)d}{2m-1-d} \right\rceil \tag{15}$$

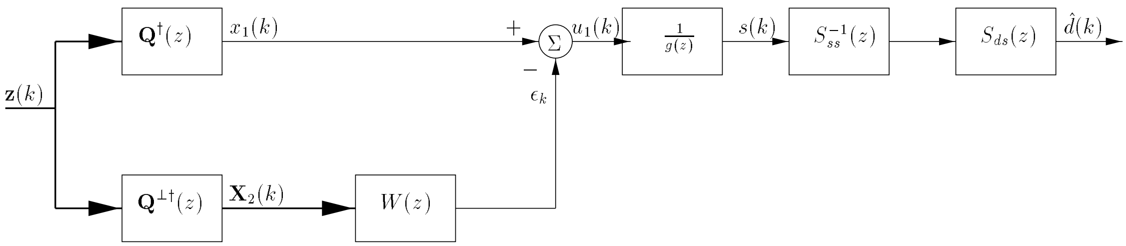

Figure 2. Optimal receiver for one user in the presence of colored noise.

where it is assumed that the maximal duration of the channel impulse responses $[\mathbf{Q}(z)\mathbf{Q}_d(z)]$ is $N_qT \geq (N+1-L_f)T$. In general, $W(z)$ is IIR and will be approximated by an FIR filter. The $1 \times (2m-1)$ Wiener filter $W(z)$ can be estimated from the signal $x_1(k)$ and $\mathbf{X}_2(k)$. When $W(z)$ is singular, one can use the Moore-Penrose Pseudoinverse which corresponds to a reduced-rank filter [7]. In practice, $W(z)$ contains only a few coefficients that can be estimated from the training sequence after removing the signal of interest $\mathbf{Q}(z)d_k$ from $x_1(k)$. In a second step (after a first detection of the symbols of the user of interest), one can use the samples of the whole time slot for the estimation of $W(z)$.

It will be convenient to process $u_1(k)$ further by a whitening filter $1/g(z)$ (see Fig. 2) (this filter can be combined with any other filter that may follow): $g(z) = (\mathbf{Q}^{\dagger}(z)\mathbf{Q}(z))^{\frac{1}{2}}$. We get for the resulting signal s_k:

$$s_k \;=\; g^{\dagger}(z)\, d_k + \mathrm{n}_k \tag{16}$$

where $z^{-(N_q-1)}g^{\dagger}(z)$ is a maximum-phase FIR filter of length N_q.

For implementing an actual receiver, we need to estimate $S_{ss}(z)$ which can be done from the signal s_k observed over the time slot. For MMSE equalizers, we consider the transfer function (Wiener filter)

$$S_{ds}(z)S_{ss}^{-1}(z) \;=\; \sigma_d^2\, g(z)\, S_{ss}^{-\frac{\dagger}{2}}(z)S_{ss}^{-\frac{1}{2}}(z)\ . \tag{17}$$

This is the transfer function of the MMSE linear equalizer (LE). For the MMSE DFE, we consider the last expression in which the first two factors correspond to the feedforward filter while the last factor, the feedback filter, gets implemented in decision feedback form. Note that $S_{ss}^{-\frac{1}{2}}(z)$ is proportional to the prediction filter for the psd $S_{ss}(z)$.

V. ADAPTIVE INTERFERENCE CANCELLATION

In a real scenario, the CCI is non-stationnary because the different users can not be perfectly synchronized at the frame level (even if the base stations are synchronous). One may track the filter $W(z)$ for each new detected symbol by the RLS algorithm and update the survivor path of each state of the Viterbi Algorithm (VA) [8]. The noise n_k at the ICMF output is assumed to be white (which corresponds to an approximation). Thus we minimize the branch metric

$$\mathcal{M}(k) = \left\| s_k - \left[g_0^* \cdots g_{N_q-1}^*\right] \begin{bmatrix} d_{k-N_q+1} \\ \vdots \\ d_k \end{bmatrix} \right\|^2 \tag{18}$$

w.r.t. the information sequence d_k where the superscript * denotes complex conjugate.

VI. SIMULATION RESULTS

We consider a three-cluster size cellular system and a typical urban propagation environment (TU). For this environment, we consider the 6-tap ($L_c = 6$) statistical channel impulse responses as specified in the ETSI standard [9] and this for both the user of interest and the interferers. The channels for multiple antennas are taken independently. We show bit error rates (BER) curves as a function of SIR for a fixed SNR = 20dB. The BER curves are averaged over 1000 realizations of the symbols and the channel impulse responses of the user of interest and the interferer(s), according to the statistical channel model, with the channel response of the interferers being rescaled to have a desired SIR. To have an idea about the distribution of BERs, one can plot the cumulative probability of the SIR for a uniform distribution of the users in the cells. Apart from the optimal receiver (ICMF followed by the Viterbi Algorithm), we also consider the performance of MMSE-LEs after an optimal interference cancellation or just after a matched filter. In a first step the filter $W(z)$ is estimated by the training sequence. We make soft decisions (hyperbolic tangent) if the equalizer output is far from 1 or -1 and hard decisions in the other case. In the second step, the channel estimate is improved if new sequences of ± 1 longer than N_q are obtained. The filter $W(z)$ is reestimated over almost the whole time slot. We compare our results with a spatio-temporal approach [10] where the temporal effect is reduced to one symbol period. The Lower Bound corresponds to the optimal receiver where the true channels are considered. We consider the uplink (2 antennas) and the downlink (1 antenna) transmission, the former for either one or two interferers.

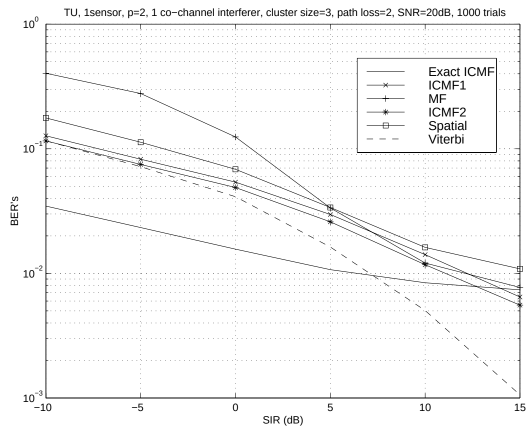

Figure 3. Bit Error Rates for one antenna and one interferer

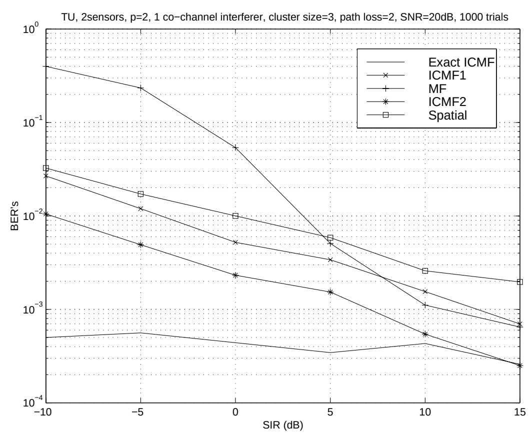

Figure 4. Bit Error Rates for two antennas and one interferer

Note that the lower bound (Exact ICMF in the plots) can be approached in the case of small radius cells (1 or 2 kilometers) served by synchronous base stations. In this case, the burst of the user of interest and those of the interferers are almost synchronous. Since in GSM only 8 training sequences are considered, one can correlate the received burst by each sequence (the prior knowledge of the TX/RX filter can also be incorporated) and then detect the active training sequences. Then, we perform a structured joint estimation of all the channels which works very well even with 5 or 6 users since the number of the parameters is reduced due to the LPE. Now, the ICMF is parameterized by all the accurately estimated channels.

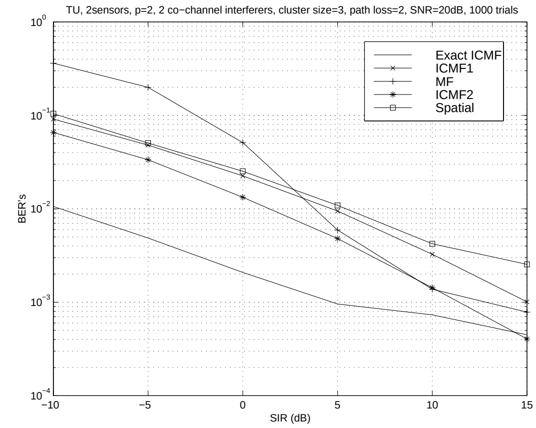

Figure 5. Bit Error Rates for two antennas and two interferers

VII. CONCLUSIONS

The proposed receiver structure is appropriate for the downlink at the mobile unit (where only the training sequence for the user of interest is assumed known). For instance in the GSM system, using multiple antennas at the mobile unit may not be realistic, but oversampling with a factor of $p = 2$ can be applied in a meaningful fashion. This would imply that if only one (dominant) interferer is present, it could be perfectly canceled with the ICMF, whose implementation requires no changes to the GSM standard. The ICMF could also be used as a suboptimal receiver structure for treating the users separately in the uplink at the base station. In this case, the ICMF yields to comparable performances if it is followed by the VA or a MMSE-LE as was shown previously [11].

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