SINR MAXIMIZING EQUALIZER RECEIVER FOR DS-CDMA

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ABSTRACT

The conventional receiver for DS-CDMA communications is the RAKE receiver which is a (linear) matched filter (MF), matched to the operations of spreading, pulse shape filtering and channel filtering. Such a MF maximizes the Signal-to-Interference-plus-Noise Ratio (SINR) at its output if the interference plus noise is white noise. This may be approximately the case if user-dependent scrambling (aperiodic spreading) is used. However, if no scrambling (hence the spreading is periodic) or only cell-dependent scrambling is used, then the interference exhibits cyclostationarity with symbol period and hence is far from white noise. In that case, the SINR at the output of a RAKE receiver can be far from optimal in the sense that other linear receivers may perform much better. In this paper we propose a restricted class of linear receivers that have the same structure as a RAKE receiver, but the channel MF and the pulse shape MF get replaced by equalizer filters that are designed to maximize the SINR at the output of the receiver. The complexity of the equalizer filters is variable and can possibly be taken to be as low as in the RAKE receiver, while its adaptation(s) guarantees improved performance with respect to the RAKE receiver. The adaptation of the SINR maximizing equalizer receiver can be done in a semi-blind fashion at symbol rate, while requiring the same information (channel estimate) as the RAKE receiver.

1 INTRODUCTION

In the Wideband CDMA (WCDMA) option of the FDD mode of the 3GPP UMTS proposal for cellular wireless communications, both uplink and downlink use DS-CDMA communications. This paper focuses on the downlink, where a set of orthogonal periodic spreading sequences are used, to take advantage of the synchronicity (between users) of the downlink. To limit interference between cells though, a cell-dependent scrambling gets added which does not destroy the orthogonality between the intracell users. Even with the scrambling present (if considered as a stationary chip rate sequence), the received signal is cyclostationary at symbol rate and linear multiuser detectors (MUD) that are time invariant at symbol rate can be applied in a meaningful way. The complexity of linear MUDs is relatively high when applied to WCDMA since they require multiplications of signals with coefficients at chip rate. From this point of view, nonlinear IC such as Parallel IC (PIC) is more interesting since it consists of a cascade of RAKE reception and refiltering by the channel. However, nonlinear approaches require a good initialization (by a linear receiver) for proper operation. The RAKE receiver is a restricted linear optimal receiver in the sense that it would be optimal if only the additive white noise (and not the interference) would be present). When considering the downlink, we presented another restricted optimal receiver in [1] which would be optimal if only the intracell interference (and not the intercell interference and noise) would be present. Intracell interference is indeed a problem because the delayspread of multipath propagation detroys the orthogonality of the downlink spreading sequences. However, if the receiver would start with an equalizer which eliminates the delay spread of the multipath propagation, then the spreading sequences would be orthogonal again at the equalizer output and it would suffice to follow the equalizer by a correlator to pick out only the contribution of the user of interest (among the intracell users). Such a receiver is also suboptimal though since the zero-forcing equalizer enhances the noise and intercell interference. In [2] we proposed a generalized linear receiver, the max-SINR receiver, which encompasses the RAKE and the equalizer-plus-correlator receivers. The structure is the same of the RAKE receiver, but the channel and pulse shape matched filters are replaced by an equalizer filter that is designed to maximize the SINR at the output of the receiver. In this paper we study different implementation and adaptation of the max-SINR receiver, analizing their performances with respect to its theoretical

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expression and to the RAKE receiver.

2 MULTIUSER DOWNLINK SIGNAL MODEL

Fig. 1 shows the downlink signal model in baseband. The K users are assumed to transmit linearly modulated signals over the same linear multipath channel with additive noise and intercell interference. The symbol and chip periods T and T_c are related through the spreading factor L: $T = LT_c$, which is assumed here to be common for all the users. The total chip sequence b_l is the sum of the chip sequences of all the users, each one given by the product between the nth symbol of the kth user and an aperiodic spreading sequence $w_{k,l}$ which is itself the product of a periodic Walsh-Hadamard (with unit energy) spreading sequence $\boldsymbol{c}_k = [\boldsymbol{c}_{k,0}\ \boldsymbol{c}_{k,1} \cdots \boldsymbol{c}_{k,L-1}]^T$, and a base-station specific unit magnitude complex scrambling sequence s_l with variance 1, $w_{k,l} = c_{k,l\ mod\ L}\, s_l$:

$$b_l = \sum_{k=1}^{K} b_{k,l} = \sum_{k=1}^{K} a_{k,\lfloor \frac{l}{L} \rfloor} w_{k,l} \ . \qquad (1)$$

The scrambling operation is a multiplication of chip rate sequences. The spreading operation could be represented similarly, or alternatively as a filtering of an upsampled symbol sequence with the spreading sequence as impulse response, as indicated in the figure. The chip sequence b_l gets transformed into a continuous-time signal by filtering it with the pulse shape $p(t)$ and then passes through the multipath propagation channel $h(t)$ to yield the received signal $y(t)$. The receiver samples M times per chip the lowpass filtered received signal.

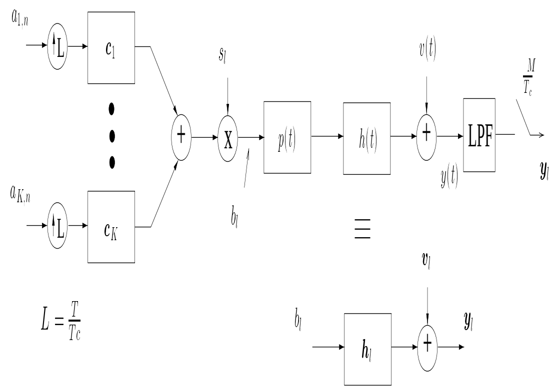

Figure 1: Downlink signal model

Stacking the M samples per chip period in vectors, we get for the sampled received signal

$$\boldsymbol{y}_l = \sum_{k=1}^{K} \sum_{i=0}^{N-1} \boldsymbol{h}_i b_{k,l-i} + \boldsymbol{v}_l \ , \qquad (2)$$

where

$$\boldsymbol{y}_l = \begin{bmatrix} y_{1,l} \\ \vdots \\ y_{M,l} \end{bmatrix}, \boldsymbol{h}_l = \begin{bmatrix} h_{1,l} \\ \vdots \\ h_{M,l} \end{bmatrix}, \boldsymbol{v}_l = \begin{bmatrix} v_{1,l} \\ \vdots \\ v_{M,l} \end{bmatrix} . \qquad (3)$$

Here $\boldsymbol{h}_l$ represents the vectorized samples of the overall channel, including pulse shape, propagation channel and receiver filter. The overall channel is assumed to have a delay spread of N chips. If we model the scrambling sequence and the symbol sequences as independent i.i.d. sequences, then the chip sequence b_l is a sum of K independent white noises (chip rate i.i.d. sequences, hence stationary). The intracell contribution to $\boldsymbol{y}_l$ then is a stationary (vector) process (the continuous-time counterpart is cyclostationary with chip period). The intercell interference is a sum of contributions that are of the same form as the intracell contribution. The remaining noise is assumed to be white stationary noise. Hence the sum of intercell interference and noise, $\boldsymbol{v}_l$, is stationary.

3 MAX-SINR RECEIVER STRUCTURE

As shown in Fig. 2, the receiver is constrained to be a chip rate filter $\boldsymbol{f}$ followed by a descrambler and a correlator with the spreading code of the user of interest, which is here assumed to be user 1. So the receiver has the same structure as a RAKE receiver, except that the channel matched filter gets replaced by a general filter $\boldsymbol{f}$. If a sparse (path-wise) representation is used for the channel, then the channel matched filter leads to a RAKE structure with one finger per path. The channel matched filter is anticausal in principle, if the channel is causal. We shall assume the filter $\boldsymbol{f}$ to be causal so that the receiver outputs symbol estimates for the user of interest with a certain delay. In Fig. 2, the operation "S/P" denotes a serial to parallel conversion which stacks the L most recent inputs into a vector. The correlator can also be viewed as a matched filter, matched to the spreading code filter, but here it is simply depicted as an inner product on a downsampled vectorized signal.

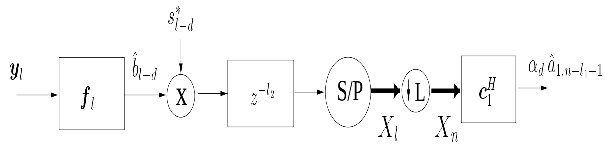

Figure 2: The downlink receiver structure

While the RAKE is one particular instance of the proposed receiver structure, another special case is the equalizer receiver. To describe this case more precisely, let $\boldsymbol{h}(z) = \sum_{l=0}^{N-1} \boldsymbol{h}_l z^{-l}$ be the $M{\times}1$ FIR channel transfer function and $\boldsymbol{f}(z) = \sum_{l=0}^{P-1} \boldsymbol{f}_l z^{-l}$ the $1{\times}M$ FIR filter transfer function of length P chips. The cascade of channel and filter gives $\boldsymbol{f}(z)\boldsymbol{h}(z) = \sum_{l=0}^{P+N-2} \alpha_l z^{-l} = \alpha(z)$. In particular, for a zero-forcing (ZF) equalizer with a delay of d chips, we get $\boldsymbol{f}(z)\boldsymbol{h}(z) = z^{-d}$, but generally the symbol estimate gets produced with a certain delay of $l_1{+}1$ symbol periods where $d = l_1 L + l_2$ $(l_1 = \lfloor \frac{d}{L} \rfloor$, $l_2 = d \; mod \; L)$. More precisely, the receiver outputs

$$\alpha_d \, \widehat{a}_{1,n-l_1-1} = \boldsymbol{c}_1^H X_n \; , X_n = S^H_{n-l_1-1} \mathcal{T}(\boldsymbol{f}) \boldsymbol{Y}_n \quad (4)$$

where X_n is a vector of descrambled filter outputs, $S_n = \text{diag}\{s_{n,L-1}, \ldots, s_{n,1}, s_{n,0}\}$ is a diagonal matrix of scrambling code coefficients $s_{n,l} = s_{nL+l}$, $\mathcal{T}(\boldsymbol{f})$ is the block Toeplitz filtering matrix with $\boldsymbol{f} = [\boldsymbol{f}_0 \cdots \boldsymbol{f}_{P-1}]$ (padded with zeros) as first block row, and $\boldsymbol{Y}_n = \left[\underline{Y}^T_{n,l_2} \, Y^T_{n-1} \cdots Y^T_{n-l_3} \, \overline{Y}^T_{n-l_3-1,l_4}\right]^T$ where $P{+}L{-}1{-}l_2 = l_3 L{+}l_4$, $Y_n = \left[\boldsymbol{y}^T_{n,L-1} \cdots \boldsymbol{y}^T_{n,0}\right]^T$, $\underline{Y}_{n,l} = \left[\boldsymbol{y}^T_{n,l-1} \cdots \boldsymbol{y}^T_{n,0}\right]^T$, $\overline{Y}_{n,l} = \left[\boldsymbol{y}^T_{n,L-1} \cdots \boldsymbol{y}^T_{n,L-l}\right]^T$, and $\boldsymbol{y}_{n,l} = \boldsymbol{y}_{nL+l}$. The structure of the vector $\boldsymbol{Y}_n$ of received data that contribute to the estimate $\widehat{a}_{1,n-l_1-1}$ is

$$\boldsymbol{Y}_n = \mathcal{T}(\boldsymbol{h}')\boldsymbol{S}_n \sum_{k=1}^{K} \boldsymbol{C}_k \boldsymbol{A}_{k,n} + \boldsymbol{V}_n \quad (5)$$

where $\mathcal{T}(\boldsymbol{h}')$ is again a block Toeplitz filtering matrix with the zero padded $\boldsymbol{h}' = [\boldsymbol{h}_0 \cdots \boldsymbol{h}_{N-1}]$ as first block row, $\boldsymbol{S}_n = \text{blockdiag}\left\{\underline{S}_{n,l_2}, S_{n-1}, \ldots, S_{n-l_5}, \overline{S}_{n-l_5-1,l_6}\right\}$, $\boldsymbol{C}_k = \text{blockdiag}\{\underline{\boldsymbol{c}}_{k,l_2}, \boldsymbol{c}_k, \ldots, \boldsymbol{c}_k, \overline{\boldsymbol{c}}_{k,l_6}\}$ $(l_5 \quad \boldsymbol{c}_k\text{'s})$, $\boldsymbol{A}_{k,n} = [a_{k,n} \cdots a_{k,n-l_5-1}]^T$, $\boldsymbol{V}_n$ is defined like $\boldsymbol{Y}_n$, and $\underline{S}_{n,l}$, $\overline{S}_{n,l}$, $\underline{\boldsymbol{c}}_{k,l}$ and $\overline{\boldsymbol{c}}_{k,l}$ are defined similarly to $\underline{Y}_{n,l}$ and $\overline{Y}_{n,l}$ except that $\underline{S}_{n,l}$ and $\overline{S}_{n,l}$ are diagonal matrices, and $P{+}L{+}N{-}2{-}l_2 = l_5 L{+}l_6$. We have for the filter-channel cascade

$$\mathcal{T}(\boldsymbol{f})\mathcal{T}(\boldsymbol{h}) = \mathcal{T}(\alpha) = \mathcal{T}(\alpha_d) + \mathcal{T}(\overline{\alpha_d}) \quad (6)$$

where

$$\begin{array}{l} \alpha = [\alpha_0 \cdots \alpha_{P+N-2}] \, , \; \alpha_d = [0 \cdots 0 \; \alpha_d \; 0 \cdots 0] \\ \overline{\alpha_d} = [\alpha_0 \cdots \alpha_{d-1} \; 0 \; \alpha_{d+1} \cdots \alpha_{P+N-2}] \, . \end{array} \quad (7)$$

In the noiseless case (and no intercell interference), the use of a ZF equalizer leads to $\overline{\alpha_d} = [0 \cdots 0]$ and $\widehat{a}_{1,n-l_1-1} = a_{1,n-l_1-1}$ $(\alpha_d = 1)$. A RAKE receiver corresponds to $\boldsymbol{f} = \boldsymbol{h}^H$, $\alpha_d = ||\boldsymbol{h}||^2$, $P = N$, where $\boldsymbol{h} = [\boldsymbol{h}^T_{N-1} \cdots \boldsymbol{h}^T_0]^T$.

The analysis done in [2] shows that, due to the orthogonality of the spreading codes and to the i.i.d. character of the noise, the SINR at the receiver output, , , is

$$, = \frac{\sigma_1^2 |\alpha_d|^2}{\boldsymbol{f} R_{VV} \boldsymbol{f}^H + \sigma^2_{tot} ||\overline{\alpha_d}||^2} = \frac{\sigma_1^2 |\alpha_d|^2}{\boldsymbol{f} R_{YY} \boldsymbol{f}^H - \sigma^2_{tot} |\alpha_d|^2} \quad (8)$$

where $\sigma_k^2 = \text{E}\,|a_{k,n}|^2$, $\sigma^2_{tot} = \frac{1}{L}\sum_{k=1}^{K} \sigma_k^2$ and $R_{YY} = R_{VV} + \sigma^2_{tot} \mathcal{T}(\boldsymbol{h}')\mathcal{T}^H(\boldsymbol{h}')$. The choice for the filter $\boldsymbol{f}$ that leads to maximum receiver output SINR is unique up to a scale factor and can be found as the solution to the following problem

$$\begin{array}{rcl} \boldsymbol{f}_{MAX} & = & \arg \max\limits_{\boldsymbol{f}:\boldsymbol{f}\boldsymbol{h}=1} \, , \; = \arg \min\limits_{\boldsymbol{f}:\boldsymbol{f}\boldsymbol{h}=1} \boldsymbol{f} R_{YY} \boldsymbol{f}^H \\ \Rightarrow \boldsymbol{f}_{MAX} & = & \left(\boldsymbol{h}^H R^{-1}_{YY} \boldsymbol{h}\right)^{-1} \boldsymbol{h}^H R^{-1}_{YY} \end{array} \quad (9)$$

The maximum SINR becomes $(\alpha_d^{MAX} = 1)$

$$,_{MAX} = \frac{\sigma_1^2}{\left(\boldsymbol{h}^H R^{-1}_{YY} \boldsymbol{h}\right)^{-1} - \sigma^2_{tot}} \quad (10)$$

As pointed out in [2], this receiver corresponds to the cascade of an (unbiased if $\alpha_d = 1$) MMSE receiver for the desired user's chip sequence, followed by a descrambler and a correlator. In the noiseless case, the MMSE receiver $\boldsymbol{f}_{MAX}$ becomes a ZF equalizer.

4 ADAPTATION STRATEGIES

Assuming that $\boldsymbol{h}$ is known (via training signal), noise plus intercell interference is white $(R_{VV} = \sigma_v^2 I)$ with a known variance σ_v^2 and σ^2_{tot} is known by construction, we can analyze different implementations of the max-SINR equalizer. The equalizer filter $\boldsymbol{f}_{MAX}$ presented in section 3 replaces at the same time the pulse shape and the channel matched filters, leaving complete freedom to the optimization process. Other possibilities rise when we want to impose a particular structure to the receiver.

4.1 Root Raised Cosine MF

The pulse shape adopted by the 3G UMTS norm is the root raised cosine (RRC) with roll-off 0.22. If we want to impose match filtering with this pulse shape, what is left to be optimized to maximize the output SINR are the coefficients of the matched filter for the propagation channel. In fact, we can write the overall channel $\boldsymbol{h}$ as

$$\boldsymbol{h} = P\boldsymbol{h}_{prop} = P_{sp}\boldsymbol{h}_{sp} \quad (11)$$

where P is the convolution matrix of the root raised cosine $p(t)$ and $\boldsymbol{h}_{prop}$ is the vector of samples of the (sparse) multipath propagation channel (MPC). Due to the sparseness of the MPC (train of pulses), P can be reduced to P_{sp} (selected columns) and $\boldsymbol{h}_{prop}$ to $\boldsymbol{h}_{sp}$ (non-zero coefficients). The receiver filter is then factored into a RRC matched filter (represented by a convolution matrix as P or P_{sp}) and an optimized (sparse) filter:

$$\boldsymbol{f} = \boldsymbol{f}_{prop} P^H = \boldsymbol{f}_{sp} P^H_{sp} \quad (12)$$

The max-SINR optimization problem in (9) gives in this case the following solution:

$$\begin{array}{rcl} \boldsymbol{f}_{RRC} & = & \left(\boldsymbol{h}^H B \left(B^H R_{YY} B\right)^{-1} B^H \boldsymbol{h}\right)^{-1} \times \\ & & \boldsymbol{h}^H B \left(B^H R_{YY} B\right)^{-1} B^H \end{array} \quad (13)$$

where B can be P or P_{sp}. In the latter case, just a number of coefficients corresponding to the number of paths in the MPC get optimized. As we will note in section 5, optimizing only the non-zero coefficients in the sparse MPC gives only modest gains w.r.t. the RAKE receiver. A better strategy could be the one in which we optimize those coefficients that corresponds to the biggest taps in the MPC, instead of taking as fixed the delays as in 13. The number of coefficients can also be augmented w.r.t. the number of paths in the MPC itself. The equation 13 is still applicable, but B is replaced by a different convolution matrix P_{opt}.

4.2 Propagation channel MF

Another choice of optimization is the one in which we impose to match filter with the propagation (sparse) channel, but we optimize the pulse shape MF coefficients in order to maximize the output SINR. In this case, the receiver filter is factored into an optimized pulse shape matched filter and a propagation channel matched filter:

$$\boldsymbol{f} = \boldsymbol{h}_{prop} G^H = \boldsymbol{g}^H T^H(\boldsymbol{h}_{prop}) \tag{14}$$

being g the optimized pulse shape (from (9)) and $T(\boldsymbol{h}_{prop})$ the convolution matrix of the (sparse) MPC. The max-SINR filter is:

$$\begin{aligned}\boldsymbol{f}_G &= \left(\boldsymbol{h}^H T(\boldsymbol{h}_{prop}) B T^H(\boldsymbol{h}_{prop})\boldsymbol{h}\right)^{-1} \times \\ & \boldsymbol{h}^H T(\boldsymbol{h}_{prop}) B T^H(\boldsymbol{h}_{prop}) \end{aligned} \tag{15}$$

where $B = \left(T^H(\boldsymbol{h}_{prop}) R_{YY} T(\boldsymbol{h}_{prop})\right)^{-1}$.

5 NUMERICAL EXAMPLES

To evaluate the loss of the max-SINR receivers in section 4 with respect to its theoretical version of section 3, we performed various simulations, with different set of parameters. All the K users are considered synchronous and use the same spreading factor SF. The UMTS chip rate is assumed (3.84 Mchips/sec) and an oversampling factor of $M = 2$ is used in the simulations. A near-far situation for the user of interest (10 dB less power than all other users) is also taken into account. In the figures below, “RAKE” refers to the RAKE receiver, “ZF” refers to the ZF equalizer (for the overall channel), and for the max-SINR curves, “theor” refers to the fully optimized max-SINR, “RRC-taps” refers to the RAKE but with the sparse channel coefficients optimized, “G-sp” refers to the RAKE buth with the pulse shape MF replaced by an optimized filter, and “RRC-nx” refer to a receiver filter consisting of the pulse shape MF and a sparse filter with n times more coeffients than the number of paths, in positions corresponding to the positions of the largest coefficients in the overall channel MF. When optimizing filters w.r.t. the RAKE, the lengths of the filters are kept.

In Fig. 3, the environment is UMTS Indoor B, spreading factor is 16, with 10 users. In Fig. 4, the environment is UMTS Vehicular A, spreading factor is 64, with 10 users. We see that the constrained optimized receivers show saturation, due to the fact that their number of degrees of freedom is not large enough to perform zero-forcing equalization. Nevertheless, in the SNR region of interest (0-20dB), the reduced complexity techniques show useful gains w.r.t. the RAKE receivers.

References

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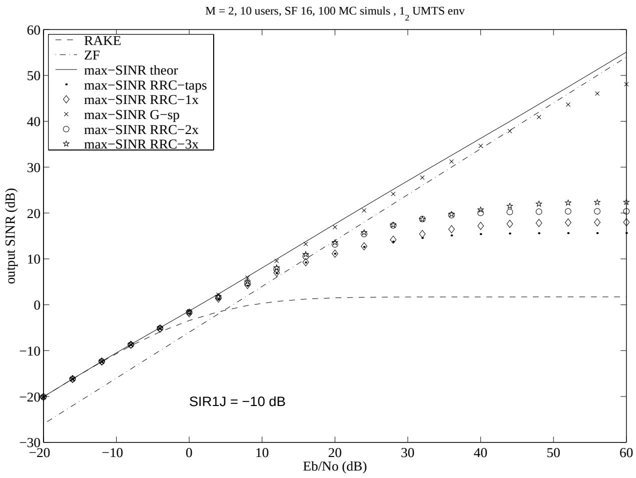

Figure 3: Output SINR versus SNR: Ind B enviroment, high loaded system, spreading factor 16 and near-far situation

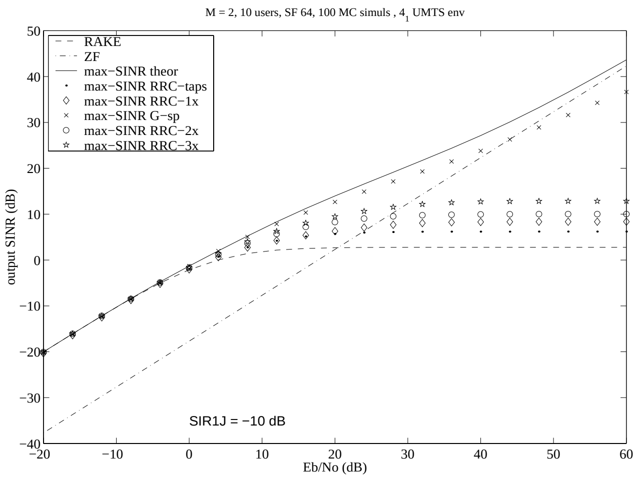

Figure 4: Output SINR versus SNR: Veh enviroment, medium loaded system, spreading factor 64 and near-far situation