A RAKE RECEIVER WITH INTRACELL INTERFERENCE CANCELLATION FOR A DS-CDMA SYNCHRONOUS DOWNLINK WITH ORTHOGONAL CODES

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Abstract - The conventional receiver for DS-CDMA communications is the RAKE receiver. The RAKE receiver is a matched filter, matched to the operations of spreading, pulse shape filtering and channel filtering. Such a matched filter maximizes the Signal-to-Interference-plus-Noise Ratio (SINR) at its output if the interference plus noise is white. This may be approximately the case if user-dependent scrambling (aperiodic spreading) is used. This is one option for the uplink in 3rd generation systems. However, this is not the case in the synchronous downlink with cell-dependent scrambling, orthogonal codes and a common channel. In this paper we propose a restricted class of linear receivers for the downlink, exhibiting a limited or no complexity increase with respect to the RAKE receiver. The linear receivers in this class have the same structure as a RAKE receiver, but the channel matched filter gets replaced by an equalizer filter that is designed to maximize the SINR at the output of the receiver. The complexity of the equalizer filter is variable and can possibly be taken to be as low as in the RAKE receiver (same structure as the channel matched filter), while its adaptation guarantees improved performance with respect to (w.r.t.) the RAKE receiver. Various adaptation strategies are considered and compared in simulations.

I. Introduction

Wireless communications are showing an unpredicted growth and the advent of third generation systems will open up the range of possible services and will significantly increase the available data rates. In the shift from voice services to data services, not only an increase in data rate is required but also a decrease in BER. To achieve such data rates at such BERs, multiple access interference cancellation will be required. Such interference is indeed the major impairment in wireless systems. Third generation systems will use one form or another of Direct Sequence Code Division Multiple Access (DS-CDMA). In such systems (and in contrast to TDMA systems), the interference situation is quite asymmetric between uplink and downlink. This is due to the fact that the transmission in the uplink is typically left asynchronous (mobiles don't get synchronized), whereas the transmission in the downlink is synchronous (from an intracell point of view). This synchronism encourages the use of orthogonal codes in the downlink. Indeed, with orthognal codes, a simple correlator receiver will get rid of all intracell interference (and optimize output SNR in the presence of white noise). This is true, at least, in the absence of delay spread due to multipath propagation. With multipath propagation, the contributions from the different paths can be combined in a maximum ratio combining fashion with a RAKE receiver, to maximize SNR. However, the combined operation of the multipath channel and the RAKE receiver destroys the orthogonality of the intracell user codes, leading to intracell interference at the RAKE output.

In [1], we presented another receiver approach which is based on channel equalization. This approach was introduced independently in [2] and [3]. Whereas the RAKE approach focuses on the noise (and intercell interference), the equalizer approach focuses on the intracell interference. Indeed, if the channel gets equalized as first operation at the receiver, the codes become orthogonal again at the equalizer output. Hence the equalizer can be followed by a correlator to get rid of all intracell interference. By using a fractionally-spaced equalizer, the excess bandwidth can be used to cancel some of the intercell interference also (if multiple antennas are available, then this can be done even better). The problem with this approach is that a zero-forcing equalizer may produce quite a bit of noise enhancement. So between the RAKE receiver and the equalizer receiver, one or the other may be better, depending on whether the intracell interference dominates the intercell interference plus noise or vice versa. A natural solution to improve the equalizer approach would be to replace a zero-forcing design by a MMSE design. Indeed, when cell-dependent scrambling is added to the orthogonal periodic spreading, then the received signal becomes cyclostationary with chip period (or hence stationary if sampled at chip rate) so that a time-invariant MMSE design becomes well-defined. Now, it may not be obvious a priori that such a MMSE equalizer leads to an optimal overall receiver. Due to the unique scrambler for the intracell users, the intracell interference after descrambling exhibits cyclostationarity with symbol period and hence is far from white noise. As a result, 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 matched filter gets replaced by an equalizer filter that is designed to maximize the SINR at the output of the receiver. It turns out that the SINR maximizing receiver uses a MMSE equalizer. 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. In this paper, we consider a wide variety of symbol rate and chip rate adaptation strategies and compare them in simulations.

II. Multiuser Downlink Data Model

Fig. 1 shows the downlink signal model in baseband. The K users are assumed to transmit linearly modulated signals over

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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, $w_{k,l} = c_{k,l \bmod L}\, s_l$:

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

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.

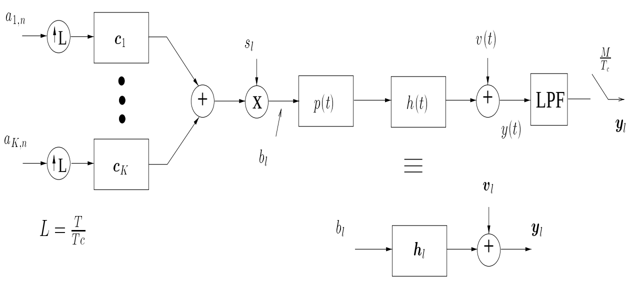

Figure 1. Downlink signal model

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

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

where

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

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.

III. 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.

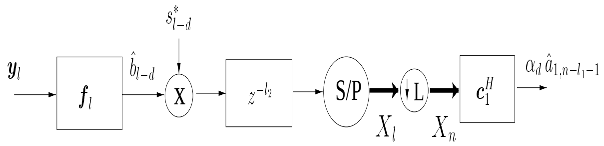

Figure 2. The downlink receiver structure

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 $\boldsymbol{f}(z)\boldsymbol{h}(z) = \sum_{l=0}^{P+N-2} \alpha_l z^{-l} = \boldsymbol{\alpha}(z)$. For a zero-forcing (ZF) equalizer with a delay of d chips, we get $\boldsymbol{f}(z)\boldsymbol{h}(z) = z^{-d}$. After ZF equalization, the spreading codes are again orthogonal (even with a unit magnitude scrambler in place) so that descrambling and correlation suffices to pick out the user of interest symbol and remove all intersymbol and intracell interference. The symbol estimate gets produced with a certain delay of $l_1{+}1$ symbol periods where $d = l_1 L + l_2$ ($l_1 = \lfloor \frac{d}{L} \rfloor, l_2 = d \bmod L$). More precisely, the receiver outputs

$$\alpha_d\, \widehat{a}_{1,n-l_1-1} \ = \ \boldsymbol{c}_1^H X_n \tag{4}$$

where X_n is a vector of descrambled filter outputs,

$$X_n \ = \ S^H_{n-l_1-1} Z_n\ , \ \ Z_n = \mathcal{T}(\boldsymbol{f})\boldsymbol{Y}_n\ , \tag{5}$$

Z_n is a vector of 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

$$\boldsymbol{Y}_n = \mathcal{T}(\boldsymbol{h}')\boldsymbol{S}_n \sum_{k=1}^{K} \boldsymbol{C}_k \boldsymbol{A}_{k,n} \ + \boldsymbol{V}_n \tag{6}$$

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 $\boldsymbol{c}_k$'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

$$\mathcal{T}(\boldsymbol{f})\mathcal{T}(\boldsymbol{h}) = \mathcal{T}(\boldsymbol{\alpha}) = \mathcal{T}(\boldsymbol{\alpha}_d) + \mathcal{T}(\overline{\boldsymbol{\alpha}}_d) \tag{7}$$

where

$$\begin{aligned} \boldsymbol{\alpha} &= [\alpha_0 \cdots \alpha_{P+N-2}]\,,\ \boldsymbol{\alpha}_d = [0 \cdots 0\ \alpha_d\ 0 \cdots 0] \\ \overline{\boldsymbol{\alpha}}_d &= [\alpha_0 \cdots \alpha_{d-1}\ 0\ \alpha_{d+1} \cdots \alpha_{P+N-2}]\,. \end{aligned} \tag{8}$$

In the noiseless case (and no intercell interference), the use of a ZF equalizer leads to $\overline{\boldsymbol{\alpha}}_d = [0 \cdots 0]$ 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 $\boldsymbol{h} = [\boldsymbol{h}_{N-1}^T \cdots \boldsymbol{h}_0^T]^T$.

IV. Maximum SINR Receiver

Let $n' = n-l_1-1$ and $\sigma_k^2 = \mathrm{E}\,|a_{k,n}|^2$, where the user power is incorporated in the symbol constellation scaling. Due to the orthogonality of the codes, we have

$$\boldsymbol{c}_1^H S_{n'}^H \mathcal{T}(\boldsymbol{\alpha}_d) \boldsymbol{S}_n \boldsymbol{C}_k \boldsymbol{A}_{k,n} = \alpha_d\, \delta_{1k}\, a_{k,n'} \tag{9}$$

where δ_{ik} is the Kronecker delta. As a result, we can decompose the receiver output $\alpha_d\,\widehat{a}_{1,n'}$ as

$$\alpha_d\, a_{1,n'} + \sum_{k=1}^{K} \boldsymbol{c}_1^H S_{n'}^H \mathcal{T}(\overline{\boldsymbol{\alpha}}_d) \boldsymbol{S}_n \boldsymbol{C}_k \boldsymbol{A}_{k,n} + \boldsymbol{c}_1^H S_{n'}^H \mathcal{T}(\boldsymbol{f}) \boldsymbol{V}_n\,. \tag{10}$$

We get for the MSE $= \mathrm{E}\,|\alpha_d a_{1,n'} - \alpha_d \widehat{a}_{1,n'}|^2 = |\alpha_d|^2\, \mathrm{E}\,|a_{1,n'} - \widehat{a}_{1,n'}|^2$

$$\begin{aligned} \text{MSE} &= \boldsymbol{c}_1^H \mathrm{E}\left\{S_{n'}^H \mathcal{T}(\boldsymbol{f}) R_{VV} \mathcal{T}^H(\boldsymbol{f}) S_{n'}\right\} \boldsymbol{c}_1 + \\ &\sum_{k=1}^{K} \sigma_k^2\, \mathrm{E}\Big\{\boldsymbol{c}_1^H S_{n'}^H \mathcal{T}(\overline{\boldsymbol{\alpha}}_d) \boldsymbol{S}_n \boldsymbol{C}_k \boldsymbol{C}_k^H \boldsymbol{S}_n^H \mathcal{T}^H(\overline{\boldsymbol{\alpha}}_d) S_{n'} \boldsymbol{c}_1\Big\} \end{aligned} \tag{11}$$

where $R_{VV} = \mathrm{E}\, V_n V_n^H$ is the noise covariance matrix and the remaining expectation is over the random scrambling sequence. Due to its i.i.d. character, we get for the noise contribution $\boldsymbol{c}_1^H\, \overline{\text{diag}}\,\left\{\mathcal{T}(\boldsymbol{f}) R_{VV} \mathcal{T}^H(\boldsymbol{f})\right\}\, \boldsymbol{c}_1 = \boldsymbol{f}\, R_{VV} \boldsymbol{f}^H$ where $\overline{\text{diag}}\,\{A\}$ is a diagonal matrix containing the diagonal part of matrix A, and the block Toeplitz R_{VV} (chip rate cyclostationarity) changes dimensions as appropriate. The interference contribution can be shown to reduce to $\sigma_{tot}^2 \|\overline{\boldsymbol{\alpha}}_d\|^2$ where $\sigma_{tot}^2 = \frac{1}{L}\sum_{k=1}^{K} \sigma_k^2$. Hence, we get for the SINR at the receiver output, $\Gamma = \sigma_1^2 |\alpha_d|^2/\text{MSE}$,

$$\Gamma = \frac{\sigma_1^2 |\alpha_d|^2}{\boldsymbol{f} R_{VV} \boldsymbol{f}^H + \sigma_{tot}^2 \|\overline{\boldsymbol{\alpha}}_d\|^2} = \frac{\sigma_1^2 |\alpha_d|^2}{\boldsymbol{f} A \boldsymbol{f}^H - \sigma_{tot}^2 |\alpha_d|^2} \tag{12}$$

where $A = R_{VV} + \sigma_{tot}^2 \mathcal{T}(\boldsymbol{h}')\mathcal{T}^H(\boldsymbol{h}')$. In the case of the RAKE receiver, $\boldsymbol{f} = \boldsymbol{h}^H$ and the SINR becomes

$$\Gamma_{RAKE} = \frac{\sigma_1^2\, \|\boldsymbol{h}\|^4}{\boldsymbol{h}^H R_{VV} \boldsymbol{h} + \sigma_{tot}^2 \|\overline{\boldsymbol{\alpha}}_d^{RAKE}\|^2} \tag{13}$$

where $\boldsymbol{\alpha}^{RAKE} = \boldsymbol{h}^H \mathcal{T}(\boldsymbol{h}')$. In the ZF equalizer case $\alpha_d^{ZF} = 1$ and $\overline{\boldsymbol{\alpha}}_d^{ZF} = 0$, so that

$$\Gamma_{ZF} = \frac{\sigma_1^2}{\boldsymbol{f}_{ZF} \mathcal{R}_{VV} \boldsymbol{f}_{ZF}^H} \tag{14}$$

where $\boldsymbol{f}_{ZF} = \boldsymbol{\alpha}_d \left(\mathcal{T}^H(\boldsymbol{h}')\mathcal{T}(\boldsymbol{h}')\right)^{-1} \mathcal{T}^H(\boldsymbol{h}')$ in the oversampling case. In what follows we shall assume that filter and channel lengths are equal, $P = N$, and $d = N-1$ (for ease of notation, any other case can equally well be handled). Note that the RAKE receiver uses $P = N$ and $d = N-1$ also. In that case $\alpha_d = \boldsymbol{f}\boldsymbol{h}$. 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

$$\boldsymbol{f}_{MAX} = \arg \max_{\boldsymbol{f}:\boldsymbol{f}\boldsymbol{h}=1} \Gamma = \arg \min_{\boldsymbol{f}:\boldsymbol{f}\boldsymbol{h}=1} \boldsymbol{f} A \boldsymbol{f}^H\,. \tag{15}$$

The solution is

$$\boldsymbol{f}_{MAX} = \left(\boldsymbol{h}^H A^{-1} \boldsymbol{h}\right)^{-1} \boldsymbol{h}^H A^{-1} \tag{16}$$

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

$$\Gamma_{MAX} = \frac{\sigma_1^2}{\left(\boldsymbol{h}^H A^{-1} \boldsymbol{h}\right)^{-1} - \sigma_{tot}^2} \tag{17}$$

Some interpretation is in order here. Due to the stochastic model for the scrambling sequence, the users' chip sequences are independent i.i.d. sequences. As a result, $\boldsymbol{y}_l$ is stationary. Using (6) we get for R_{YY}

$$\begin{aligned} R_{YY} &= R_{VV} + \sum_{k=1}^{K} \sigma_k^2 \mathcal{T}(\boldsymbol{h}')\, \mathrm{E}\left\{\boldsymbol{S}_n \boldsymbol{C}_k \boldsymbol{C}_k^H \boldsymbol{S}_n^H\right\} \mathcal{T}^H(\boldsymbol{h}') \\ &= R_{VV} + \sum_{k=1}^{K} \sigma_k^2 \mathcal{T}(\boldsymbol{h}')\, \overline{\text{diag}}\left\{\boldsymbol{C}_k \boldsymbol{C}_k^H\right\} \mathcal{T}^H(\boldsymbol{h}') \\ &= R_{VV} + \sum_{k=1}^{K} \sigma_k^2 \mathcal{T}(\boldsymbol{h}')\, \frac{1}{L} I\, \mathcal{T}^H(\boldsymbol{h}') \\ &= R_{VV} + \sigma_{tot}^2 \mathcal{T}(\boldsymbol{h}')\mathcal{T}^H(\boldsymbol{h}') = A\ \ ! \end{aligned} \tag{18}$$

Consider now the MMSE receiver to estimate linearly the desired user chip sequence $b_{1,l-d}$ from the data Y_l. That MMSE receiver is

$$\widehat{b}_{1,l-d} = R_{b_{1,l-d} Y_l} R_{YY}^{-1}\, Y_l = \frac{\sigma_1^2}{L}\, \boldsymbol{h}^H R_{YY}^{-1}\, Y_l \tag{19}$$

which is hence proportional to the filter for the max SINR receiver (see (16)). In fact, $\boldsymbol{f}_{MAX}$ is the unbiased MMSE receiver (coefficient of $b_{1,l-d}$ in $\widehat{b}_{1,l-d}$ is 1). Furthermore, the filter $\boldsymbol{f}_{MAX}$ also leads to the unbiased MMSE estimate $\widehat{a}_{1,n-l_1-1}$ at the output of the receiver we are considering. So, 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.

V. Max SINR Receiver: Estimation strategies

V.1. Construction from Theoretical Expression and h

Assume that a training signal is available to estimate the channel. In that case we can consider $\boldsymbol{h}$ to be known. Now also assume that the noise plus intercell interference is white so that R_{VV} is of the form $R_{VV} = \sigma_v^2 I$. Then the only two parameters we need to estimate (further) in order to be able to construct $\boldsymbol{f}_{MAX}$ are σ_v^2 and σ_{tot}^2. In the case of oversampling we can obtain σ_v^2 and σ_{tot}^2 from

$$\begin{aligned} \sigma_v^2 &= \lambda_{min}(R_{YY}) \\ M\sigma_v^2 + \sigma_{tot}^2 \|\boldsymbol{h}\|^2 &= \mathrm{E}\,\|\boldsymbol{y}_l\|^2 \end{aligned} \tag{20}$$

where $\lambda_{min}(.)$ denotes the minimal eigenvalue and note that the signal part of R_{YY} is singular. In practice we estimate the quantities on the right of the equations in (20) from data. Although we believe that $R_{VV} = \sigma_v^2 I$ is a good approximation, work on more general noise covariance models is ongoing.

V.2. Construction from $\widehat{R}_{YY}$ and h

If we again assume that $\boldsymbol{h}$ is available via training data, and we estimate R_{YY}, then we can simply construct $\widehat{\boldsymbol{f}}_{MAX} \sim \boldsymbol{h}^H \widehat{R}_{YY}^{-1}$ or this quantity can also be estimated directly (without forming $\widehat{R}_{YY}^{-1}$) via an LMS-like algorithm. Note that Y_l at all chip periods is available for estimating R_{YY} (chip rate cyclostationarity).

V.3. Linearly Constrained MOE Approach

Again we assume that $\boldsymbol{h}$ is available via training data. From (10), (12) and (18), we get for the variance of the receiver output (often mistakenly called output energy)

$$E|\boldsymbol{c}_1^H X_n|^2 = |\alpha_d|^2(\sigma_1^2 - \sigma_{tot}^2) + \boldsymbol{f} R_{YY} \boldsymbol{f}^H \ . \tag{21}$$

Hence we can obtain the max SINR receiver filter from the following linearly constrained minimum OE (MOE) criterion

$$\boldsymbol{f}_{MAX} = \arg \min_{\boldsymbol{f}:\boldsymbol{f}\boldsymbol{h}=1} E|\boldsymbol{c}_1^H X_n|^2 \tag{22}$$

in which we replace the statistical average by a temporal average. To obtain a better estimation quality, it would be desirable to be able to perform the temporal averaging at the chip rate. To that end, consider the receiver output at an intermediate chip period

$$\boldsymbol{c}_1^H X_{n,m} = \boldsymbol{c}_1^H S_{n',m}^H \mathcal{T}(\boldsymbol{f}) \boldsymbol{Y}_{n,m} \ . \tag{23}$$

$S_{n',m} = \text{diag}\{s_{n'+1,m-1}, \ldots, s_{n'+1,0}, s_{n',L-1}, \ldots, s_{n',m}\}$ and $X_{n,m}$ are like $S_{n'}$ and X_n but shifted m chips into the future so that $X_n = X_{n,0}$ and $S_{n'} = S_{n',0}$, and $\boldsymbol{Y}_{n,m} = [\boldsymbol{y}_{nL+l_2-1+m}^T \, \boldsymbol{y}_{nL+l_2-2+m}^T \cdots \boldsymbol{y}_{nL+l_2-P-L+1+m}^T]^T$. With $\mathcal{T}(\boldsymbol{f})\boldsymbol{Y}_{n,m} = \mathcal{Y}_{n,m}\boldsymbol{f}^T$ for a certain structured matrix $\mathcal{Y}_{n,m}$, we can write

$$E|\boldsymbol{c}_1^H X_{n,m}|^2 = \boldsymbol{f} \, R^* \, \boldsymbol{f}^H, R = \mathrm{E}\left\{\mathcal{Y}_{n,m}^H S_{n,m} \boldsymbol{c}_1 \boldsymbol{c}_1^H S_{n,m}^H \mathcal{Y}_{n,m}\right\}$$

so that $\boldsymbol{f} = \left(\boldsymbol{h}^H R^{-*} \boldsymbol{h}\right)^{-1} \boldsymbol{h}^H R^{-*}$. It can be shown that $|\boldsymbol{c}_1^H X_{n,m}|^2 = \boldsymbol{f} R_{VV} \boldsymbol{f}^H + \sigma_{tot}^2 \|\overline{\boldsymbol{\alpha}}_d\|^2 + |\alpha_d|^2 f_m$ where $f_m = \sum_{k=1}^K \sigma_k^2 \|\boldsymbol{c}_1^H \mathcal{T}(\boldsymbol{\alpha}_d) \boldsymbol{C}_{k,m}\|^2 / |\alpha_d|^2$ and $\boldsymbol{C}_{k,m}$ is like $\boldsymbol{C}_k$ but considered m chips later. So the part of $|\boldsymbol{c}_1^H X_{n,m}|^2$ that varies with m only depends on $\boldsymbol{f}$ through α_d, which is fixed by the linear constraint. So when temporal averaging is performed over all chip periods, the filter estimate $\boldsymbol{f}$ will still converge to $\boldsymbol{f}_{MAX}$. As far as the estimation quality is concerned however, the MSE on the filter estimate is proportional to the excess OE, which in turn is proportional to the MOE. To lower the MOE at intermediate chip instants, it may be advantageous to consider instead the receiver output $\boldsymbol{c}_{1,m}^H X_{n,m}$ where $\boldsymbol{c}_{1,m} = [c_{1,m-1} \cdots c_{1,0} \, c_{1,L-1} \cdots c_{1,m}]^T$ is $\boldsymbol{c}_1$ shifted cyclically over m chips. In this case $|\alpha_d|^2 f_m = \sum_{k=1}^K \sigma_k^2 \|\boldsymbol{c}_{1,m}^H \mathcal{T}(\boldsymbol{\alpha}_d) \boldsymbol{C}_{k,m}\|^2$ whereas the rest of the OE is unchanged. The receiver output in this case is produced by implementing the despreading like the descrambling (multiplication) followed by a sliding summation over L chips.

V.4. Blind Approach Exploiting Unused Codes [4]

Assume that the receiver knows that the codes $[\boldsymbol{c}_{K+1} \cdots \boldsymbol{c}_L] = \boldsymbol{C}^\perp$ are not used. Then we can introduce the following blind (channel knowledge not required) criterion

$$\boldsymbol{f}_B = \arg \min_{\boldsymbol{f}:\|\boldsymbol{f}\|=1} \frac{\mathrm{E}\,\|\boldsymbol{C}^{\perp H} X_n\|^2}{\mathrm{E}\,\|X_n\|^2} \ . \tag{24}$$

We show in [4] that the $\boldsymbol{f}_B = \boldsymbol{f}_{MAX}$. In the simulations below, we actually consider the following variation $\boldsymbol{f}_B = \arg\min \frac{\mathrm{E}\,\|\boldsymbol{C}^{\perp H} X_n\|^2}{\mathrm{E}\,\|\boldsymbol{c}_1^H X_n\|^2}$.

VI. Simulations

Various simulations with different sets of parameters have been performed. All the K users are considered synchronous and use the same spreading factor. The FIR channel is the convolution of a sparse Vehicular A UMTS channel and a pulse shape (root-raised cosine with roll-off factor of 0.22). The UMTS chip rate (3.84 Mchips/sec) is assumed, leading to the channel length of $N = 19$ chips. An oversampling factor of $M = 2$ is used in these simulations. Two cases of user power distribution are considered: all interferers have the same power and the user of interest has either the same power also or 15dB less power (near-far situation).

The spreading factor (SF) is indicated at the top of the figures, which show the performance of various receiver instances in terms of the output signal-to-interference-and-noise (SINR) ratios versus the SNR at the receiver when the length of all the FIR filters, in chips, is the same ($P = N$). In Figures 3-5, we compare the theoretical RAKE (dashed), ZF equalizer (dash-dot) and max SINR (solid) with the estimated max SINR (dotted) obtained via the method of section V.1 (using only 5 symbol periods of data!).

Fig. 3 refers to a highly loaded system with equal distribution of average power between users, while Fig. 4 refers to the same system but with a near-far problem for the user of interest. We can note how the max SINR receiver suffers much less from the near-far effect than the RAKE. In the SNR range of interest, the max SINR receiver performs between 2dB and 10dB better then the RAKE, while the ZF equalizer suffers a lot from the noise enhancement in that SNR range. It can be noted that the adapted max SINR receiver has close to optimal performance. Fig. 5 presents the same near-far situation but with $L = 128$ and $K = 100$ users; performances are similar.

In the next set of three figures we compare the performances of the three theoretical receivers (RAKE, ZF Equalizer and max SINR receiver) with three estimated versions of the max-SINR structure: the method of section V.3 (dense dots), the method of section V.2 (sparse dots), and the blind method of section V.4 (x's).

VII. References

[1] I. Ghauri and D. T. M. Slock, "Linear receivers for the DS-CDMA downlink exploiting orthogonality of spreading sequences," in *Proc. 32nd Asilomar Conf. on Signals, Systems & Computers*, (Pacific Grove, CA), November 1998.

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[4] I. Ghauri and D. T. M. Slock, "Blind Maximum SINR for the DS-CDMA Downlink," in *Proc. ICASSP 2000*, (Istambul, Turkey), May 2000.

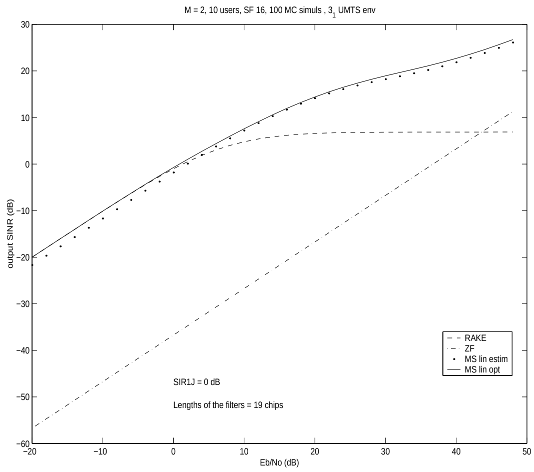

Figure 3. Theoretical output SINR versus SNR, high loaded system, spreading factor 16 and equal power distribution

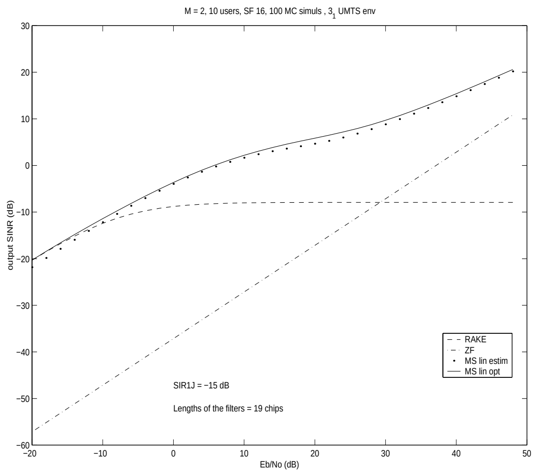

Figure 4. Theoretical output SINR versus SNR, high loaded system, spreading factor 16 and near-far situation

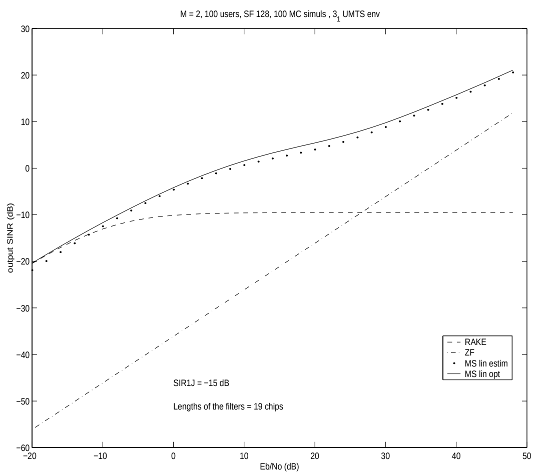

Figure 5. Theoretical output SINR versus SNR, high loaded system, spreading factor 128 and near-far situation

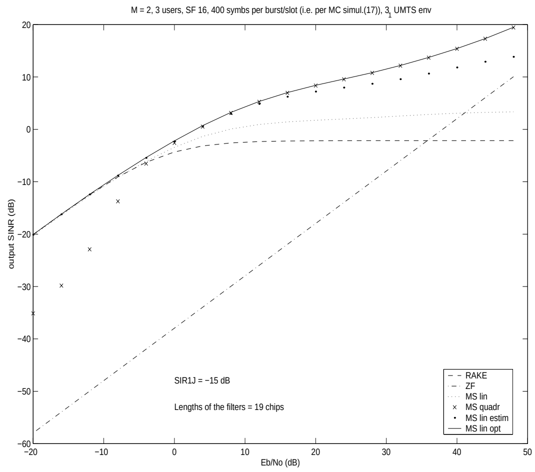

Figure 6. Theoretical output SINR versus SNR, high loaded system, spreading factor 16 and equal power distribution

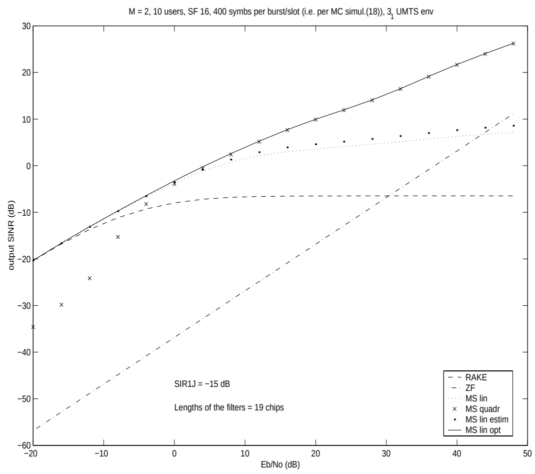

Figure 7. Theoretical output SINR versus SNR, high loaded system, spreading factor 16 and near-far situation

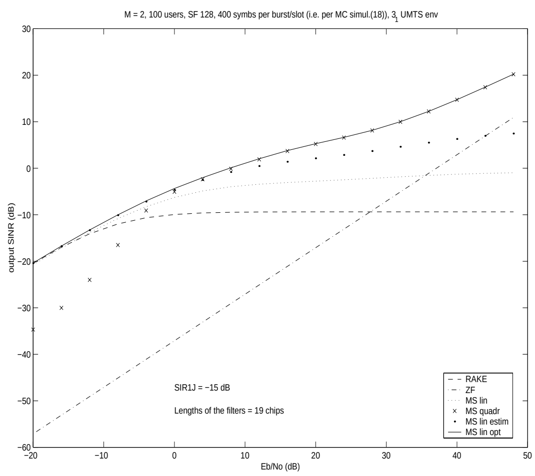

Figure 8. Theoretical output SINR versus SNR, high loaded system, spreading factor 128 and near-far situation