Linear Receivers for the DS-CDMA Downlink Exploiting Orthogonality of Spreading Sequences[*]

Irfan Ghauri   Dirk T. M. Slock

Institut Eurécom, 2229 Route des Crêtes, B.P. 193
06904 Sophia Antipolis Cedex, France
{ghauri,slock} @eurecom.fr

ABSTRACT

We address the problem of downlink interference rejection in a DS-CDMA system. Periodic orthogonal Walsh-Hadamard sequences spread different users' symbols followed by scrambling by a symbol aperiodic base-station specific overlay sequence. This corresponds to the downlink of the European UMTS wideband CDMA proposal. The point to point propagation channel from the cell-site to a certain mobile station is the same for all downlink signals (desired user as well as the interference). The composite channel is shorter than a symbol period for some user signals, while other users can have significant ISI owing to a faster transmission rate. In any case, orthogonality of the underlying Walsh-Hadamard sequences is destroyed by multipath propagation, resulting in multiuser interference if a coherent combiner (the RAKE receiver) is employed. We propose linear zero-forcing (ZF) and minimum mean-squared-error (MMSE) receivers which equalize for the estimated channel, thus rendering the user signals orthogonal again. A simple code matched filter subsequently suffices to cancel the multiple access interference (MAI) from intracell users.

1. INTRODUCTION AND PREVIOUS WORK

While complex joint multiuser detection techniques are being pursued for the uplink of the third generation wireless systems mostly due to the inability of the RAKE receiver in combatting the *near-far* effect, downlink situations are considered to be much too deficient in terms of information and processing power to implement multiuser techniques. However, the capacity of the overall communication system can only be increased if both links can support similar transmission rates.

In situations where a relatively small number of users are active ($\approx 20\%$ of the processing gain), the RAKE receiver [1] might perform in an adequate manner and more complex signal processing might be unnecessary. Increasing the number of users to approach the spreading gain, however, has a catastrophic effect on the performance since small contributions of multipath signals of of each interferer captured by the matched-filter bank add up to large values, even in the power controlled case. This effect is simply due to the suboptimal treatment of the MAI as uncorrelated noise by the RAKE receiver.

As an alternative to the RAKE, linear receiver techniques based upon single user *a priori* information have lately been an active topic of research [2] (and the references therein). These receiver algorithms are based upon symbol rate wide sense cyclostationarity and have been shown to converge to the MMSE solution. The application of these techniques in existing systems, however, is not straightforward, since symbol rate cyclostationarity no longer exists when aperiodic overlay sequences spread/randomize the orthogonal user sequences.

It is to be noted that in the structure of the downlink problem, the only entity fixed over the processing interval is the propagation channel. Burst processing techniques can thus be applied once the channel has been estimated, and single user information (symbol spreading code of the user of interest) and cell specific randomizing codes of active base stations are available.

In the IS-95 CDMA standard, a perpetually present known downlink pilot signal is used to estimate the downlink channel. This pilot is much stronger than other user signals and a correlation based searcher constantly searches the best fingers for building the RAKE. Channel estimation on the downlink using the known pilot symbols was presented in [3], where it was assumed that all downlink codes were known. Such an assumption is reasonable in the case of the UMTS WCDMA norm [4], where a fixed number of downlink codes can be used at one time, and all common cell information is constantly broadcast over the entire cell. Some blind algorithms exploiting the *i.i.d.* nature of the spreading sequences and symbols have been presented [5] [6]. However, these algorithms are statistically inefficient.

In this paper, we introduce linear zero-forcing (ZF) and MMSE receivers for the DS-CDMA downlink which equalize for the propagation channel, once the latter has been estimated, thus rendering the user signals orthogonal again. Oversampling/multiple sensors are used at the mobile station to facilitate equalization. Multichannels can also be created by treating real and imaginary parts of a signal when the input constellation is one dimensional.

2. DOWNLINK SIGNAL MODEL

Fig. 1 illustrates the downlink channel model. The J intracell users are assumed to transmit linearly modulated signals over a linear multipath channel with additive Gaussian noise. It is assumed that the signal is received at the mobile station through multiple (diversity) discrete-time channels, obtained from oversampling the received signal multiple times per chip or through multiple sensors (or a combination of the two schemes). We shall

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consider the signal to be received through precisely M channels where, M = no. of sensors $\times$ oversampling factor. The signal received through the mth channel can be written in baseband notation as

$$y_m(t) = \sum_{j=1}^{J} \sum_{k} b_j(k) h_{mj}(t \ominus kT) + v_m(t), \qquad (1)$$

where the subscript j denotes the user index; T is the chip period; the chip sequences $\{b_j(k)\}_{j=1}^{J}$ are assumed to be independent of the additive noise $\{v_m(t)\}$; and $h_{mj}(t)$ characterizes the channel impulse response between the jth user signal and the mth sensor or phase of the received signal. Let us
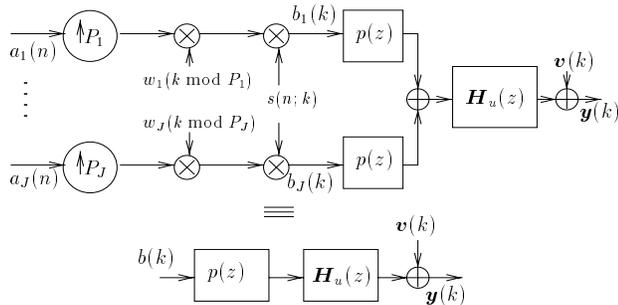
Figure 1. The Downlink Model

denote by $c_j(n) = [c_j(n;0)\ c_j(n;1)\cdots c_j(n; P_j \ominus 1)]^T$, the structured aperiodic spreading sequence vector for the nth symbol of the jth user. The aperiodic spreading sequences consist of a periodic Walsh-Hadamard spreading sequence $w_j = [w_j(0)\cdots w_j(P_j \ominus 1)]$, overlaid by a base-station specific scrambling sequence $s(n;k)$. Then, $c_j(n;k) = w_j(k)s(n;k)$, $k \in \{1, \cdots, P_j\}$. We can write the chip sequence $\{b_j(k)\}$ corresponding to the data symbol, $a_j(n)$, for the jth user as

$$b_j(k) = a_j(n)c_j\{n; k \ominus (n \ominus 1)P_j\}. \qquad (2)$$

The chip period T is a constant, while the symbol period $T_j, \forall j$, is a function of the transmission rate of the jth user. The symbol and chip periods are related through the processing gain P_j: $T_j = P_j T$.

In the DS-CDMA downlink, the propagation channel for the desired user's signal and the interference is the same as seen by the mobile station. Therefore, we shall suppress the subscript j from $h_{mj}(t)$ in the following development. We consider that there are no intercell interferers (or are considerably weaker) and their effect is momentarily ignored. We assume that the timing information has already been acquired from the initial synchronization procedure. Due to the synchronous nature of the channel, user signals will be considered to be perfectly aligned. The oversampled cyclostationary received signal at M times the chip rate can be stacked together to obtain the $M \times 1$ stationary vector signal $y(k)$ at the chip rate, which can be expressed as

$$y(k) = \sum_{j=1}^{J} \sum_{i=0}^{N-1} h(i)b_j(k \ominus i) + v(k)$$
$$= \sum_{j=1}^{J} H_N B_{j,N}(k) + v(k) = H_N B_N(k) + v(k), \quad (3)$$

where,

$$y(k) = \begin{bmatrix} y_1(k) \\ \vdots \\ y_M(k) \end{bmatrix}, \ h(k) = \begin{bmatrix} h_1(k) \\ \vdots \\ h_M(k) \end{bmatrix}, \ v(k) = \begin{bmatrix} v_1(k) \\ \vdots \\ v_M(k) \end{bmatrix}$$

$$H_N = [h(N \ominus 1) \cdots h(0)]$$
$$h = \left[h^H(N \ominus 1) \cdots h^H(0) \right]^H$$
$$B_j(k) = C_j(k) A_{n_j}(k)$$
$$B_N(k) = [C_1(k) \dots C_J(k)] A_n(k)$$
$$A_n(k) = \left[A_{n_1}^H(k) \cdots A_{n_J}^H(k) \right]^H$$
$$A_{n_j}(k) = \left[a_j^H(k \ominus n_j + 1) \cdots a_j^H(k) \right]^H, \qquad (4)$$

$n = \sum_j n_j$; $n_j = \lceil (N + P_j \ominus 1)/P_j \rceil$, and the superscript ^{H} denotes Hermitian transpose. The matrices $C_j(k)$ are $N \times n_j$ matrices with n_j giving the degree of ISI for the jth user's signal, while the index k accounts for the aperiodicity of spreading sequences. Let us make the following assumptions:

(**A1**). The spreading codes are binary *i.i.d.*
(**A2**). The user data symbols are also *i.i.d.*

Based upon these assumptions, and due to the linearity of the modulation and the channel model, the chip-rate vector signal as described in (3) can be thought of as a single input multiple output (SIMO) model with a modified input alphabet given by the elements of the composite chip-sequence $b(k)$. See fig. 1.
Let us stack L successive vectors $y(k)$ in a supervector

$$Y_L(k) = \mathcal{T}(h)B_{N+L-1}(k) + V_L(k), \qquad (5)$$

where $\mathcal{T}_L(h)$ is a block Toeplitz convolution matrix filled up with the channel coefficients grouped in h, and has in general full column rank.

3. DOWNLINK RECEIVERS

3.1. The Zero-Forcing Receiver

In the CDMA downlink problem, there are several kinds of zero-forcing criteria that can be pursued. We shall consider the following two special cases:

3.1.1. Zero-Forcing conditions for ISI and MAI

In a deterministic framework, the overall channel including the code filter $C_j(k)$, as seen by the jth user symbol sequence is a time-varying filter expressed in symol rate co-efficients, $g_j(k) = (C_j(k) \otimes I_M)h$. Let us denote by $\mathcal{G}_j(k)$, the composite channel convolution matrix for the jth user. Then we can write the received signal (5) as: $Y_L(k) = \sum_{j=1}^{J} \mathcal{G}_j(k) A_{L+n_j-1}(k) = \mathcal{G}(k)A_{L+N-1}$, with $\mathcal{G}(k) = [\mathcal{G}_1(k)\cdots\mathcal{G}_J(k)]$ and $N = \sum_{j=1}^{J} n_j$ is the channel length in symbol periods. Consider now, a fractionally spaced $(\frac{T}{M})$ receiver vector $F_L(k)$ of length L symbol periods, which is a time-varying filter. Then, the condition for the receiver to be zero-forcing at the kth instant is that

$$F_L(k)\ \mathcal{G}_L(k) = [0 \cdots 0\ 1\ 0 \cdots 0], \qquad (6)$$

where, $\mathcal{G}_L(k)$ is now a tall matrix assumed to be of full column rank, and the 1 lies in a position so as to correctly select the desired column of $\mathcal{G}_j(k)$. To enable ZF solution, we need to choose the receiver length such that the system of equations (6) is exactly or underdetermined. Hence,

$$L \geq \underline{L} = \left\lceil \frac{N \Leftrightarrow J}{(MP_j)_{\text{eff}} \Leftrightarrow J} \right\rceil, \qquad (7)$$

where, $(MP_j)_{\text{eff}} = \min\{MP_j, N\}$ is the effective number of channels.

The receiver in (6) results in perfect ISI and MAI rejection, in the noiseless case, with the inconvenience that it is a time-varying filter from symbol to symbol and that the $\underline{L}$ is fairly long for saturated systems.

3.1.2. Zero-Forcing conditions for ISI only

Alternatively, given the *i.i.d.* nature of the input chip sequence, which results in a modified constellation of the chip sequences, and the knowledge that the downlink channel is the same for all signals, we can treat the problem as a single-input multiple-output (SIMO) vector channel transfer function, obtained from the single input $b(k) = \sum_{j=1}^{J} a_j(n)c_j\{n; k \Leftrightarrow (n \Leftrightarrow 1)P_j\}$ to the multiple outputs: $\boldsymbol{H}(z) = \sum_{i=0}^{N-1} \boldsymbol{h}(i)z^{-i}$, so that we can write the vector received signal: $\boldsymbol{y}(k) = \boldsymbol{H}(q)b(k) + \boldsymbol{v}(k)$, where, $q^{-1}b(k) = b(k \Leftrightarrow 1)$. This gives us a reduced set of constraints

$$\boldsymbol{F}\ \mathcal{T}_L(\boldsymbol{h}) = [0 \cdots 0\ 1\ 0 \cdots 0], \qquad (8)$$

with 1 in the $\delta + 1$st position from the end. To be able to satisfy all the constraints (8), we need to choose the filter length such that the system of equations is exactly or underdetermined. Hence,

$$L \geq \bar{L} = \left\lceil \frac{N \Leftrightarrow 1}{M_{\text{eff}} \Leftrightarrow 1} \right\rceil, \qquad (9)$$

where, $M_{\text{eff}} = \text{rank}\{\boldsymbol{H}_N\}$ is the effective number of channels, and L is measured in chip periods as opposed to (7).

At the output of the equalizer, we obtain the chip-sequence $\hat{b}_i(k \Leftrightarrow \delta) = \boldsymbol{F}_L \boldsymbol{Y}_L(k)$, where δ is the equalizer delay expressed in chip periods. After filtering and downsampling, regenerated PN sequences are applied for first descrambling, $\hat{d}(n; k) = \hat{b}(n; k)s(n; k)$ and then despreading

$$\hat{a}_{j,n-\tau} = \text{dec}\left[\boldsymbol{w}_j \hat{D}(k)\right], \qquad (10)$$

where, $\hat{D}(k) = \left[\hat{d}^H(n; k \Leftrightarrow \delta \Leftrightarrow 1) \cdots \hat{d}^H(n; k \Leftrightarrow \delta \Leftrightarrow P_j)\right]^H$ is the descrambled chip sequence, dec is the decision operator, and τ is the equalizer delay in symbol periods.

As an alternative scheme, we can first despread $y(k)$ at L chip spaced delays in a successive fashion:

$$\boldsymbol{X}_L(n) = [\boldsymbol{Y}_L(1) \cdots \boldsymbol{Y}_L(P_j)]\,\boldsymbol{c}_j(n), \qquad (11)$$

where $\boldsymbol{c}_j(n)$ is the jth user spreading sequence during the kth interval. We can further write $\boldsymbol{X}_L(n)$ as

$$\boldsymbol{X}_L(n) = \mathcal{T}_L(\boldsymbol{h}) \underbrace{\begin{bmatrix} b_1 & b_2 & \cdots & b_{P_j} \\ b_2 & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ b_{N+L-1} & \cdots & \cdots & b_{P_j+N+L-2} \end{bmatrix}}_{\text{B}} \overbrace{\begin{bmatrix} c_j(n; 1) \\ \vdots \\ \vdots \\ c_j(n; P_j) \end{bmatrix}}^{\text{c}(n;j)},$$

where, $\text{B} = \sum_{i=1}^{J} \text{B}_i$. Let us denote by $\boldsymbol{R}$, the matrix consisting of the product $\mathcal{T}_L(\boldsymbol{h})\text{Bc}(n; j)\text{c}^H(n; j)\text{B}^H \mathcal{T}^H(\boldsymbol{h})$ and by $\boldsymbol{R}_j$, the matrix consisting of the diagonal of the matrix $\mathcal{T}_L(\boldsymbol{h})\text{B}_j\text{c}(n; j)\text{c}^H(n; j)\text{B}_j^H \mathcal{T}^H(\boldsymbol{h})$. Then the instantaneous SINR for the jth user is given as

$$\text{SINR}_j(n) = \frac{\sigma_a^2 \boldsymbol{F}_L \boldsymbol{R}_j \boldsymbol{F}_L^H}{\sigma_a^2 \boldsymbol{F}_L (\boldsymbol{R} \Leftrightarrow \boldsymbol{R}_j)\, \boldsymbol{F}_L^H + \sigma_v^2 \boldsymbol{F}_L \boldsymbol{F}_L^H}, \qquad (12)$$

and, the desired symbol estimate is given by

$$\hat{a}_{j,n-\tau} = \text{dec}\left[\boldsymbol{F}_L \boldsymbol{X}_L(n)\right], \qquad (13)$$

and $\boldsymbol{F}_L$ is the one that satisfies (8), and τ is the same as in (10). Hence, there is no essential difference in the order of the despreading and filtering operations [5], once the receiver has been determined. The subtle difference lies in the observation that the former approach suggests some kind of decisions (soft if no quantization is applied) at the chip level and then removing the effect of the MAI by the despreading operation. The other method, on the contrary, relies on first despreading the received signal and then applying further processing, which is more in the spirit of traditional spread-spectrum systems.

3.2. Linear MMSE Receiver

From the standpoint of noise enhancement, the MMSE criterion gives better performance than the ZF receiver. The MMSE receiver can be obtained as: $\min_{\boldsymbol{F}_M} E\|b_k \Leftrightarrow \boldsymbol{F}_M Y(k)\|^2$ which gives the MMSE receiver: $\boldsymbol{F}_M = \sigma_b^2 \tilde{\boldsymbol{h}}_j^H \boldsymbol{R}_{YY}^{-1}$, and $\tilde{\boldsymbol{h}}_j$ is the corresponding column of the matrix $\mathcal{T}(\boldsymbol{h})$ in (5). The MMSE receiver might incur some performance loss due to residual non-orthogonality resulting from non-perfect zero-forcing but is more robust to noise enhancement.

3.3. Discussion

It is meaningful to suppress both the ISI and the MAI as in (6) in order to enhance the SINR at the receiver output. However, doing so means imposing a large number of constraints (equal to $N + J(L \Leftrightarrow 1)$ for a given L). This corresponds to a small number of degrees of freedom resulting in significant noise enhancement, and eventually lowering the output SINR. It is advantageous, therefore, to exploit inherent properties of the data/channel model (here the orthogonality of the underlying spreading sequences) thus reducing the number of constaints to the ones in ISI ZF-equalization problem (8) with $N \Leftrightarrow J + J(L \Leftrightarrow 1)$ constraints. Equation (8) when written in the same way as (6) corresponds to keeping one column for each user's composite channel matrix $\mathcal{G}_j(k)$, $\forall j$ scaled by an arbitrary scalar, while forcing all others to zero. Hence we have more degrees of freedom and a better performance is obtained. The resultant MAI is automatically suppressed due to the orthogonality of user codes.

It is clear from (8) that in the single user case, the ZF solution would be sub-optimal since all the energy is concentrated in one tap and due to the resulting noise enhancement. Such is not the case for the RAKE receiver which collects the energy from all paths to maximize the SNR at the output (being a matched filter). However, when interferers are present, constraining all the energy in one tap is still suboptimal for the user of interest *vis-á-vis* the noise enhancement, but a better SINR is achieved since the interference is perfectly cancelled.

4. CHANNEL ESTIMATION

Consider a downlink synchronous slot corresponding to the transmitted data vector of length $A_{j,L+n_j-1}, \forall j$, which can be written as a sequence of transmitted chips $B_{j,L+N-1}, \forall j$. The corresponding observations are $Y = [y^H(0) \cdots y^H(L \Leftrightarrow 1)]^H$, and the noise is denoted by V. The relationship can be written in a concise fashion as $Y = \mathcal{T}B + V$.

4.1. Training-data Based Channel Estimation

We suppose here that the training symbols are known over the common training period of the frame/slot. In the UMTS WCDMA downlink, the control/data slot contains a fixed number of training symbols for all users which are time-aligned due to the synchronous nature of the downlink. As in the current GSM standard, the training symbols $A_N(k)$ can be pre-selected quantities. In the case of intracell users, the number of active users and their rate information is broadcast on a common downlink channel, thus making the spreading code information available to all mobile stations managed by the cell site. Then, the estimation of the channel is the one obtained by the least-squares criterion. By exploiting the commutativity of the convolution, we get: $\mathcal{T}B = \mathcal{B}h$,

$$\mathcal{B} = B_{L,N} \otimes I_M \qquad (\otimes = \text{tensor product})$$

where,

$$B_{L,N} = \begin{bmatrix} b_{-N+1} & b_{-N+2} & \cdots & b_0 \\ b_{-N+2} & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ b_{L-N} & \cdots & \cdots & b_{L-1} \end{bmatrix}.$$

Then, solving the maximum likelihood criterion

$$\min_h \|Y \Leftrightarrow \mathcal{T}B\|^2 \Leftrightarrow \min_h \|Y \Leftrightarrow \mathcal{B}h\|^2, \qquad (14)$$

for the channel, admits as solution: $h = (\mathcal{B}^H\mathcal{B})^{-1}\mathcal{B}^H Y$. The inputs $B(k)$ are from a modified discrete constellation alphabet which comes about from the linear combinations of the individual user chip alphabets (which are the same for all users).

4.2. Blind and Semi-Blind Methods

The SIMO model can be solved for the channel coefficients from the second-order statistics of the received signal (blind methods) or by optimizing jointly for the blind and training sequence information (semi-blind methods). We shall refer to [7] (and the references therein) for the details of these methods.

5. MULTICELLULAR ENVIRONMENT

In the presence of signals from U base stations, the L received vectors, $y(k) = \sum_{u=1}^{U} H_{N_u} B_{N_u}(k) + v(k)$, can be stacked together to give

$$Y_L(k) = \sum_{u=1}^{U} \mathcal{T}(h_u) B_{N_u+L-1}(k) + V_L(k). \qquad (15)$$

We also assume, further to (A1) and (A2), that given fairly long *i.i.d.* binary sequences, their linear combination is an *i.i.d.* non-binary sequence.
Once again, the channel can be estimated jointly from the least squares criterion, and a zero-forcing equalizer satisfying (8) can be determined, given a certain smoothing factor.

5.1. Dimensional Requirement

Consider the noiseless case ($v(t) \equiv 0$). Then we can write (15) as

$$Y_L(k) = \mathcal{T}(h)B_{N+L-1}$$
$$= [\mathcal{T}(h_1) \cdots \mathcal{T}(h_U)][b_{1,N_1+L-1}^H \cdots b_{U,N_U+L-1}^H]^H. \qquad (16)$$

Now, $\mathcal{T}(h)$ is of dimension $ML \times N + U(L \Leftrightarrow 1)$, with $N = \sum_{u=1}^{U} N_u$. Then in order to be zero-forcing in the noiseless case, L has to be such that $\mathcal{T}(h)$ is a tall matrix of full column rank in general. Then, $L \geq \underline{L} = \left\lceil \frac{N-U}{M_{\text{eff}}-U} \right\rceil$, where, U is the number of active base stations for the mobile user. Note that $M_{\text{eff}} > U$ is a condition that is easily satisfied for worst-case scenarios, i.e., $U = 3$, in the hexagonal cell geometry.

5.2. Cyclostationary nature of intercell interferers

It is interesting to observe the behavior of intercellular interference in terms of its statistical properties. Due to the aperiodic overlay sequences, the out-of-cell interferers add up as cyclostationary noise at the chip rate, $\sigma_b^2 \sum_{u=2}^{U} \mathcal{T}(h_u)\mathcal{T}^H(h_u) \to \sigma_b^2 R_h$, under the assumptions (A1) and (A2). R_h is a banded toeplitz matrix with a strong diagonal element (considering chip-rate sampling is performed) and $R_h \to I$ as the delay spread of the channel reduces. If these interferers are weak, then their effect can be ignored due to the relatively small terms on the bands of R_h.

6. NUMERICAL EXAMPLES

The simulation framework modelizes the downlink of a UMTS wideband CDMA type system with orthogonal channelization codes overlaid by a cell-site specific scrambler randomizing the periodic user code sequences. A spreading gain of 16 is assumed. We consider a channel shorter than the symbol period (about 40% of T_s, the symbol period, assumed to the same for all users). There is no change in the model if users with different rates are present, since the basic signature waveforms are orthogonal. However, we consider shorter channels since it is pointless to employ a RAKE receiver in the case where ISI spans quite a few symbol periods. The input signal constellation is QPSK with the primary spreading
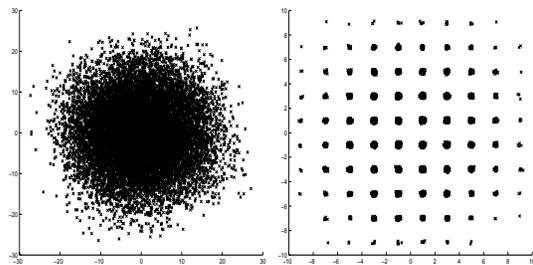
Figure 2. Eye of the received and equalized signals with as SNR=30 dB. and 9 intracell users, P=16, with an input constellation of QPSK

sequences from the binary Walsh-Hadamard set, followed by the randomly selected scrambler with an alphabet $s(n) \in \{+1, \Leftrightarrow 1\}$. The eye of the received and equalized signals are shown in fig. 2. It is seen that the equalized signal vectors are combinations of the input alphabets. A root-raised cosine pulse with a roll-off factor of 0.22 is used in these simulations conform with the UMTS

WCDMA norm [4]. We choose a relatively long (64 chip periods) equalizer in these simulations in order to satisfy $\underline{L}$ in all cases. Furthermore, it is a well-known result that longer equalizers give better results. Fig. 3 compares the output signal-to-interference-and-
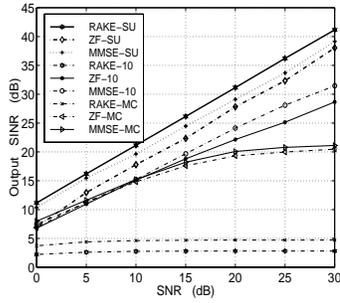
Figure 3. SINR comparison of RAKE, ZF and MMSE

noise ratio (SINR) performance of the ZF and the MMSE receivers with the RAKE receiver. It is seen that the performance of the RAKE receiver is effected by finite cross-correlations among delayed versions of spreading sequences. A flooring effect is hence noticeable. As for the ZF and MMSE receivers, once the channel is equalized, the effect of other users can be perfectly removed owing to the underlying orthogonality. At low SNRs, some performance loss for the ZF receiver is incurred due to the noise enhancement. For reference, the SINR of single user case are also shown. The coherent RAKE is a code-channel matched filter in that case. The figure also shows the performance degradation as intercellular interference starts to creep in. We have 14 users in a code space of 16, of which 8 are orthogonal users sharing the same downlink channel. The other 7 issue from the neighboring cell site and are all 10 dB weaker than the user of interest. The figure depicts performance loss incurred by ignoring such interference. "SU" stands for single user, and "MC" for the multicellular case.

Fig. 4 shows the performance of the ZF and MMSE receivers as a function of the signal-to-interference ratio (SIR) corresponding to the 5 out-of-cell interferers at 5 dB SNR with 5 in-cell users. The ZF receiver suffers some performance penalty in the high noise case. Finally, fig. 5 shows the degradation of the training sequence based channel estimate as 7 intercellular interfereres (10 dB weaker) from a different cell sites are introduced.
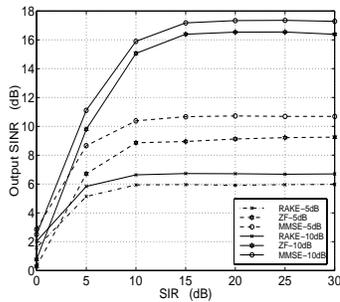
Figure 4. SINR vs. SIR in multicellular case

7. CONCLUSIONS

We presented linear ZF and MMSE receivers for the downlink of a DS-CDMA system. It is seen that given an estimate of the common downlink channel, perfect zero-forcing equalization is possi-

ble in the noiseless case, irrespective of the number of users, as long as their inner spreading sequences are orthogonal (which is the case in various existing CDMA norms), and if sufficient spatio-temporal diversity is made available. Performance comparison
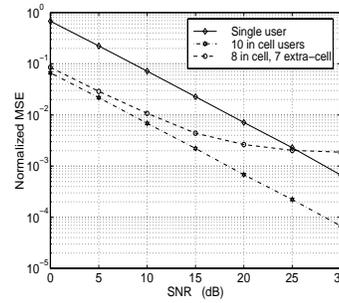
Figure 5. Channel estimation perf. for T.S. method, P=16

with the RAKE receiver shows that in the absence of intercellular interference, these receivers are near far resistant and provide promising gains. Extension to multicellular environments is also possible if all downlink channels can be estimated. However, more diversity channels will be needed to zero-force in this case. Towards the end of this work, we became aware of [8] which proposes block processing schemes for downlink orthogonal transmission. However, the processing involves complex matrix operations over blocks of data and the *being tall* condition of the channel matrix is satisfied by considering zeros transmitted before and after the burst instead of multiple channels in our formulation. Furthermore, in our approach, performance of the receivers is expected to be much better due to the multichannel aspect.

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