AMOUR - GENERALIZED MULTICARRIER CDMA IRRESPECTIVE OF MULTIPATH

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ABSTRACT

Suppression of multiuser interference (MUI) and mitigation of multipath effects constitute major challenges in the design of third-generation wireless mobile systems. Most wideband and multicarrier uplink CDMA schemes suppress MUI statistically in the presence of unknown multipath and impose restrictive and difficult to check conditions on the FIR channel nulls. Relying on symbol blocking, we design A Mutually-Orthogonal Usercode-Receiver (AMOUR) system for quasi-synchronous blind CDMA that eliminates MUI deterministically and mitigates fading irrespective of the unknown multipath and the adopted signal constellation. AMOUR converts a multiuser CDMA system into parallel single-user systems irrespective of multipath and guarantees identifiability of users' symbols. Simulations reveal the generality, flexibility, and superior performance of AMOUR over competing alternatives.

1. INTRODUCTION

Multiuser interference (MUI) and multipath-induced interchip interference (ICI) are critical performance limiting factors in the design of third-generation wireless systems because they define their capabilities in handling high data rates and interactive multimedia services. MUI and ICI suppression is thus of paramount importance in mobile wideband CDMA standards such as UMTS and IMT-2000. Multipath causes frequency-selective fading, destroys orthogonality of user codes, and when unknown, it precludes usage of linear zero-forcing (ZF), minimum mean-square error (MMSE) [4], or nonlinear (decision-feedback (DF) and maximum-likelihood) multiuser detectors for MUI suppression. But even when multipath channel estimates are available (e.g., using bandwidth consuming training sequences) it is well known that especially for multichannel uplink CDMA systems multiuser equalization is only possible under certain rank conditions on channel matrices that are difficult to check at the receiver [7, 2].

Relying on symbol blocking, we develop in this paper A Mutually-Orthogonal Usercode-Receiver structure (AMOUR) for quasi-synchronous blind *uplink* CDMA that eliminates MUI deterministically and mitigates fading irrespective of the unknown multipath. The system encompasses LV-CDMA [5] and MC-CDMA [1] systems as special cases, can have low FFT-based complexity, and appears to offer considerable design flexibility. Although our design targets the uplink CDMA channel, it can be also applied to the downlink channel as well as to other wire-line applications such as DSL (Digital Subscriber Line). Our focus will be on the blind scenario but AMOUR is also attractive when the multipath channel has been estimated (e.g., via pilot signals). Based on the multirate block model of Section 2, we develop the AMOUR-CDMA system in Section 3. Section 4 is devoted to simulations and comparisons with competing schemes.

2. AMOUR SYSTEM MODELING

Generalizing the filterbank CDMA model proposed in [7], the block diagram in Fig. 1 represents the uplink channel of a CDMA system, described in terms of its discrete-time equivalent baseband model, where signals, codes, and channels are represented by samples of their complex envelopes taken at the chip rate (only transmitter and receiver filters for one, the mth, user are shown). Advance elements and down/up-samplers serve the purpose of blocking and inserting zeros, so that each of the M users maps successive blocks of K symbols of the information sequence $s_m(n)$ to blocks of length $P > K$, where K and P are design parameters to be specified in Section 3. First, the mth user's information symbol stream $s_m(n)$ is parsed into consecutive K-long blocks, $\mathbf{s}_m(i) := [s_m(iK), \ldots, s_m(iK+K-1)]^T$. Then each symbol $s_m(iK+k)$ in this ith block is spread by a code $c_{m,k}(n)$ of length P and the transmitted sequence $u_m(n)$ is the sum of the resulting chip sequences:

$$u_m(n) = \sum_{i=-\infty}^{\infty} \sum_{k=0}^{K-1} s_m(iK+k) c_{m,k}(n-iP) \,. \quad (1)$$

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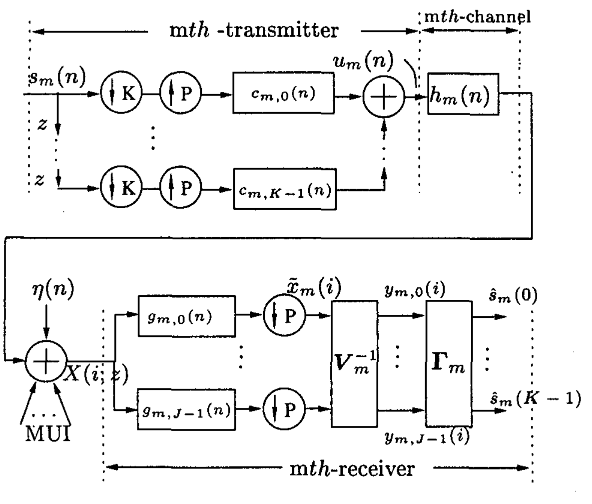

Figure 1: Discrete-time baseband AMOUR system

The coded chip sequence $u_m(n)$ then passes through the discrete-time equivalent baseband channel, denoted by its impulse response $h_m(n)$. The discrete-time received sequence is $x(n) = \sum_{m=0}^{M-1} x_m(n) + \eta(n)$, where: $x_m(n) = \sum_{j=-\infty}^{\infty} u_m(j)h_m(n-j)$, and $\eta(n)$ is the AGN.

The channels $h_m(n)$ or their $\mathcal{Z}$-transforms $H_m(z)$, $m = 0, \ldots, M-1$, are assumed to be of order $\leq L$. The downlink scenario is a special case of our setup and corresponds to having $h_m(l) = h(l)$, $\forall m \in [0, M-1]$. In addition to multipath and the transmit-receive filters, $h_m(n)$ includes the mth user's asynchronism in the form of delay factors. In quasi-synchronous (QS) CDMA systems, mobile users attempt to synchronize with the base-station's pilot waveform (see e.g., [1]). But exact synchronization is difficult to implement in the reverse channel due to multipath and Doppler effects arising because of relative motion, especially when the chip period is small. In general, mobile users' timing maybe off by $2-3$ chips. To avoid inter-block interference (IBI), we add a guard time interval of length L to each transmitted block by setting the last L samples of $c_{m,k}(n)$ equal to zero.

In Fig. 1, the ith received block $\boldsymbol{x}(i) := [x(iP) \cdots x(iP+P-1)]^T$ (^{T} denotes transpose) consists of chips from the mth user of interest along with MUI chips from other users and AGN $\boldsymbol{\eta}(i) := [\eta(iP) \cdots \eta(iP+P-1)]^T$. The receive-filterbank consists of J parallel filters each of length P and will be described by the $J \times P$ matrix $\boldsymbol{G}_m$ whose (j,p)th entry is $g_{m,j}(p)$. The columns of $\boldsymbol{G}_m$, denoted by $\boldsymbol{g}_m(p) := [g_{m,0}(p) \cdots g_{m,L+K-1}(p)]^T$, perform block filtering of $\boldsymbol{x}(i)$. Through $\boldsymbol{G}_m$, the block $\boldsymbol{x}(i)$ is mapped to a transformed domain where user m is separated from its MUI. Downsamplers bring the transformed MUI-free block back to the symbol rate block $\tilde{\boldsymbol{x}}(i)$, while the $J \times (L+K)$ matrix $\mathbf{V}_m^{-1}$ performs the inverse transform on $\tilde{\boldsymbol{x}}(i)$, to yield the $(L+K) \times 1$ vector $\boldsymbol{y}_m(i) := [y_{m,0}(i) \cdots y_{m,L+K-1}(i)]^T$. Finally, symbol estimates $\hat{\boldsymbol{s}}_m(i)$ are obtained by multiplying $\boldsymbol{y}_m(i)$ with the equalizing matrix $\boldsymbol{\Gamma}_m$ (matrices $\boldsymbol{G}_m$, $\mathbf{V}_m^{-1}$, and $\boldsymbol{\Gamma}_m$ will be specified in Section 3).

The structure in Fig. 1 generalizes the one in [3] and provides a framework for CDMA signal processing. Let us now define the $\mathcal{Z}$-transforms $C_{m,k}(z) := \sum_{n=0}^{P-1} c_{m,k}(n)z^{-n}$, $X(i;z) := \sum_{p=0}^{P-1} x(iP+p)z^{-p}$, $N(i;z) := \sum_{p=0}^{P-1} \eta(iP+p)z^{-p}$, and the $D \times 1$ Vandermonde vector $\boldsymbol{v}(\rho, D)$ built from the complex constant ρ as: $\boldsymbol{v}(\rho, D) := [1\, \rho^{-1} \cdots \rho^{-(D-1)}]^T$. With z replacing ρ and P instead of D, this Vandermonde vector describes the $\mathcal{Z}$-transform operation in the sense that $X(i;z) := \boldsymbol{v}^T(z,P)\boldsymbol{x}(i)$. Thanks to the L trailing zeros, no IBI is present in our P-long blocks. Therefore, despite the presence of MUI, and ICI that is allowed in our QS setup, one can focus on each received block separately, and express it in the $\mathcal{Z}$-domain as [c.f. (1)]:

$$X(i;z) = \sum_{\mu=0}^{M-1} H_\mu(z) \sum_{k=0}^{K-1} s_\mu(iK+k)C_{\mu,k}(z) + N(i;z). \tag{2}$$

Our goal is to design codes $\{C_{m,k}(z)\}_{m,k=0,0}^{M-1,K-1}$ and receive-filters $\{G_{m,j}(z)\}_{j=0}^{L+K-1}$, where $G_{m,j}(z) := \sum_{n=0}^{P-1} g_{m,j}(n)z^{-n}$, that guarantee: i) deterministic MUI cancellation with a simple linear receiver; ii) blind channel equalization; and iii) symbol recovery, irrespective of the channels, the adopted signal constellation and with minimum transmit-redundancy.

3. DETERMINISTIC MUI ELIMINATION

Our basic idea is to achieve MUI elimination in the frequency ($\mathcal{Z}$-) domain. That is, we seek code polynomials $C_{m,k}(z)$, $m = 0, \ldots M-1$, so that for each m there exist J points $\{\rho_{m,j}\}_{j=0}^{J-1}$ on which $X(i;z)$ contains the mth user's contribution but MUI from the remaining $M-1$ users is eliminated, irrespective of $H_\mu(z)$, $\mu = 0, \ldots, M-1$ (J is a design parameter to be specified). If such MUI-eliminating points $\rho_{m,j}$ exist and we choose

$$\boldsymbol{G}_m = [\boldsymbol{v}(\rho_{m,0}, P) \cdots \boldsymbol{v}(\rho_{m,J-1}, P)]^T, \tag{3}$$

then $\tilde{\boldsymbol{x}}_m(i) := \boldsymbol{G}_m\boldsymbol{x}(i) = [X(i;\rho_{m,0}) \cdots X(i;\rho_{m,J-1})]^T$ will be free of MUI. Note that the mth user separating Vandermonde matrix $\boldsymbol{G}_m$ maps $\boldsymbol{x}(i)$ to preselected values of its $\mathcal{Z}$-transform $\tilde{\boldsymbol{x}}(i)$. We call these points $\{\rho_{m,j}\}_{j=0}^{J-1}$ signature points of the mth user.

It follows from (2) that the desired channel independent separation of the users in the $\mathcal{Z}$-domain can be achieved if and only if we design $\{C_{\mu,k}(z)\}_{\mu=0}^{M-1}$ such that $\forall k \in [0, K-1]$, $j \in [0, J-1]$, $\mu, m \in [0, M-1]$, it holds that:

$$C_{\mu,k}(\rho_{m,j}) = \theta_m(j,k)\delta(\mu - m)\,, \tag{4}$$

where $\theta_m(j,k)$ are arbitrary complex constants, and δ denotes Kronecker's delta. Condition (4) implies that $\{\rho_{m,j}\}_{j=0}^{J-1}$ are roots common to all $C_{\mu,k}(z)$ except $C_{m,k}(z)$ – an observation that plays a key role in our MUI-free design. Plugging $z = \rho_{m,j}$ into (2) and using (3), we can write $\tilde{\boldsymbol{x}}_m(i)$ under (4) as:

$$\tilde{\boldsymbol{x}}_m(i) = \boldsymbol{D}_{H_m}\boldsymbol{\Theta}_m \boldsymbol{s}_m(i) + \tilde{\boldsymbol{\eta}}_m(i), \tag{5}$$

where $\boldsymbol{D}_{H_m}$ denotes a $J \times J$ diagonal matrix with diagonal entries $\{H_m(\rho_m, j)\}_{j=0}^{J-1}$, $\boldsymbol{\Theta}_m$ is a $J \times K$ matrix with its (j,k)th entry $\theta_m(j,k)$, and the noise term $\tilde{\boldsymbol{\eta}}_m(i) := [N(i;\rho_{m,0}),\ldots,N(i;\rho_{m,J-1})]^T$. Choosing $J \geq K$, we deduce from (5) that a necessary and sufficient condition to guarantee identifiability of symbols $\boldsymbol{s}_m(i)$ from $\tilde{\boldsymbol{x}}_m(i)$ irrespective of the input symbol constellation is:

$$\text{rank}(\boldsymbol{D}_{H_m}\boldsymbol{\Theta}_m) \geq K\,. \tag{6}$$

If we select our user codes to satisfy (4) and (6), then MUI- and ICI-free symbol recovery becomes possible irrespective of the channels $H_m(z)$. Two possibilities arise:

Case 1) if the mth transmitter has access to channel status information (CSI), we can always choose $J = K$ different points $\rho_{m,j}$ so that $H_m(\rho_{m,j}) \neq 0\ \forall j \in [0, K-1]$, and with any non-singular $K \times K$ matrix $\boldsymbol{\Theta}_m$, we can fulfill (4) and (6).

Case 2) if CSI is not available at the transmitters, since $H_m(z)$ has order $\leq L$, at most L of $\{H_m(\rho_{m,j})\}_{j=0}^{J-1}$ can be zero. Thus, in order to satisfy (6), we need: 2a) $J \geq L + K$; and 2b) any $J - L$ rows of $\boldsymbol{\Theta}_m$ to span the vector space of complex K-tuples $\mathbb{C}^K$.

Because our ultimate goal is channel irrespective MUI elimination and blind symbol recovery, we focus on Case 2 and choose $J = L + K$, which will also turn out to achieve maximum bandwidth efficiency. With $J = L + K$, condition 2b) translates to requiring any K rows of $\boldsymbol{\Theta}_m$ to be linearly independent. One possible choice for $\boldsymbol{\Theta}_m$ that will prove useful and flexible enough for our purposes is

$$\boldsymbol{\Theta}_m = \mathcal{K}_m[\boldsymbol{v}(\rho_{m,0}, K)\cdots \boldsymbol{v}(\rho_{m,L+K-1}, K)]^T, \tag{7}$$

where $\mathcal{K}_m$ is a constant introduced to control transmitted power, and the points $\{\rho_{m,j}\}_{j=0}^{L+K-1}$ are distinct so that the

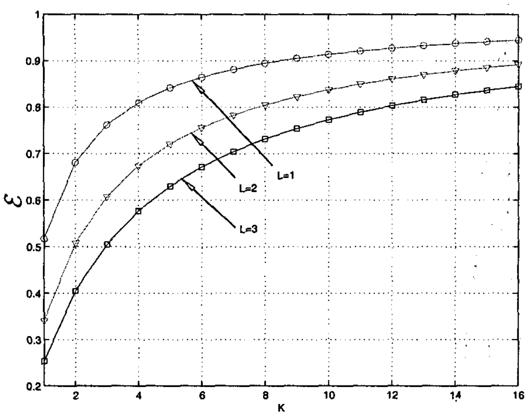

Figure 2: Bandwidth efficiency for an AMOUR-CDMA system with 16 users

Vandermonde matrix $\boldsymbol{\Theta}_m$ in (7) is full rank. With this choice for $\boldsymbol{\Theta}_m$, the MUI-free code specification (4) becomes: $\forall$ $k \in [0, K-1]$, $j \in [0, L+K-1]$, $\forall \mu, m \in [0, M-1]$,

$$C_{\mu,k}(\rho_{m,j}) = \mathcal{K}_m \rho_{m,j}^{-k}\delta(\mu - m)\,. \tag{8}$$

For fixed μ and k, (8) prescribes $C_{\mu,k}(z)$ at $M(L+K) = MJ$ points $\rho_{m,j}$. Thus, $C_{\mu,k}(z)$ polynomials that satisfy (8) in general should have degree $\deg[C_{\mu,k}(z)] \geq MJ - 1$. When $\deg[C_{\mu,k}(z)] = MJ - 1$, such code polynomials can be uniquely determined by Lagrange interpolation through the points $\rho_{m,j}$ [c.f. (8)]

$$C_{\mu,k}(z) = \mathcal{K}_\mu \sum_{\lambda=0}^{L+K-1} \rho_{\mu,\lambda}^{-k} \prod_{m=0}^{M-1} \prod_{\substack{j=0 \\ (m,j)\neq(\mu,\lambda)}}^{L+K-1} \frac{1 - \rho_{m,j}z^{-1}}{1 - \rho_{m,j}\rho_{\mu,\lambda}^{-1}}. \tag{9}$$

From (9) and taking into account the L trailing zeros, the code length in our system is $P = M(L+K) + L$, and since we deal with transmissions of K-long blocks from M users, the bandwidth efficiency of our system is

$$\mathcal{E} := \frac{MK}{P} = \frac{MK}{M(L+K)+L}\,. \tag{10}$$

For sufficiently large $K \gg L$, we have $\mathcal{E} \approx 1$; hence, bandwidth is not over expanded (see also Fig. 2).

In steps, AMOUR's design procedure can be summarized as follows:

d1) Choose the symbol block length $K \gg L$;

d2) Select $M(L+K)$ distinct points $\rho_{m,j}$ on the complex plane and assign $L+K$ of them $\{\rho_{m,j}\}_{j=0}^{L+K-1}$ to be signatures of the mth user;

d3) Compute the codes according to (9), add L trailing zeros, and set the receiver filters according to (3).

Using (7) and (8), eq. (5) can be written component-wise as

$$X(i;\rho_{m,j})=S_m(i;\rho_{m,j})H_m(\rho_{m,j})+N(i;\rho_{m,j}), \quad (11)$$

which confirms that we have achieved user separation at certain $\mathcal{Z}$-transform values of the ith received block. The separability of the entire $\mathcal{Z}$-transform $X(i;z)$, and hence $\boldsymbol{x}(i)$, can be similarly established as in [3] via Vandermonde matrices $\mathbf{V}_m := [\boldsymbol{v}(\rho_{m,0}, L+K)) \cdots \boldsymbol{v}(\rho_{m,L+K-1}, L+K)]^T$; namely, it follows that [3]: $\boldsymbol{y}_m(i) = \mathbf{V}_m^{-1}\tilde{\boldsymbol{x}}_m(i) = \mathbf{V}_m^{-1}\boldsymbol{G}_m\boldsymbol{x}(i)$.

In vector form, (11) can be written as $\boldsymbol{y}_m(i) = \mathbf{H}_m\boldsymbol{s}_m(i) + \boldsymbol{\eta}_m(i)$, where $\boldsymbol{\eta}_m(i) := \mathbf{V}_m^{-1}\boldsymbol{G}_m\boldsymbol{\eta}(i)$, and $\mathbf{H}_m$ is the Toeplitz convolution matrix with the first row $[h_m(i;0)\,0\ldots0]$ and the first column $[h_m(i;0)\ldots h_m(i;K-1)\,0\ldots0]^T$. Due to the trailing zeros introduced at the transmitter [c.f. d3)], $\mathbf{H}_m$ is always full rank.

After the MUI elimination, blind [6, 3] or non blind single user equalizers $\boldsymbol{\Gamma}_m$ can be applied to $\boldsymbol{y}_m(i)$ to recover $\boldsymbol{s}_m(i)$. Typical choices include:

- RAKE receiver $\boldsymbol{\Gamma}_m^{(rake)} = \mathbf{H}_m^{\mathcal{H}}$ (Hermitian),
- ZF equalizer $\boldsymbol{\Gamma}_m^{(zf)} = \mathbf{H}_m^{\dagger}$ (pseudo-inverse),
- MMSE equalizer $\boldsymbol{\Gamma}_m^{(mmse)} = \boldsymbol{R}_{s_m}\mathbf{H}_m^{\mathcal{H}}(\boldsymbol{R}_{\eta_m} + \mathbf{H}_m\boldsymbol{R}_{s_m}\mathbf{H}_m^{\mathcal{H}})^{-1}$, where $\boldsymbol{R}_{s_m}$ and $\boldsymbol{R}_{\eta_m}$ are autocorrelation matrices of $\boldsymbol{s}_m(i)$ and $\boldsymbol{\eta}(i)$, respectively.

We remark that the AMOUR system can have low (FFT based) complexity if we choose $\rho_{m,l}$ to be regularly spaced around the unit circle:

$$\rho_{m,l} = e^{j\frac{2\pi(m+lM)}{M(L+K)}}, \; \forall l \in [0, L+K-1]. \quad (12)$$

User code polynomials with such signature points have coefficients given by [c.f. (9)]

$$c_{m,k}(n) = \begin{cases} K_m e^{j\frac{2\pi m}{M}q} & \text{if } n \bmod (L+K) = k, \\ 0 & \text{otherwise.} \end{cases} \quad (13)$$

Observe that for every user m, the code $c_{m,k}(n) = c_{m,0}(n-k)\ \forall k = 0,\ldots,K-1$, which means that at the transmitter, entails a simple shift and scaling only. At the receiver, matrix multiplications (by $\boldsymbol{G}_m$) and inversion ($\mathbf{V}_m^{-1}$) can be replaced by FFTs. Algorithmically, the receiver computes a P-point (Zoom-) FFT of the received block $\boldsymbol{x}(i)$ at the $M(L+K)$ frequencies given in (12). Then each user's receiver computes an $(L+K)$-point IFFT at the $L+K$ frequency samples corresponding to the user's signature points to obtain $\boldsymbol{y}_m(i)$. In addition to computational simplicity, such a selection in (12) can be thought as a generalized MC-CDMA scheme.

4. SIMULATIONS

In this section, we test the performance of the proposed AMOUR system via simulations.

Test case 1 For a AMOUR system with parameters $(M, K, L) = (16, 16, 2)$, we compared the performance of ZF AMOUR receiver $\boldsymbol{\Gamma}_m^{(zf)}$ *and* MMSE AMOUR receiver $\boldsymbol{\Gamma}_m^{(mmse)}$ with that of a RAKE receiver $\boldsymbol{\Gamma}_m^{(rake)}$. In Fig. 3, we show the BER for the three receivers averaged over 1000 random Rayleigh fading channels. The AMOUR receivers outperform the RAKE receiver by one or two orders of magnitude of BER at high SNRs.

Test case 2 We compared performance of AMOUR with codes as in (13) and $M = 8$, $L = 2$, $K = 14$, against an MC-CDMA system with MRC at the receiver. At the MC-CDMA transmitter, each user spreads K consecutive information symbols with Walsh-Hadamard codes of length $Q = M(L+K) = 128$, and then uses OFDM with cyclic prefix of length $L = 2$ before transmission. Both systems assume the channel to be perfectly known at the receiver end. Fig. 4 shows the corresponding BERs averaged over 200 independent Rayleigh fading channels. AMOUR outperforms MC-CDMA especially at high SNR. We repeated this test case with $M = 32$, $L = 2$, $K = 8$ (corresponds to bandwidth efficiency $\mathcal{E} \approx 0.8$) against an MC-CDMA system with MRC, $Q = 32$, and $\mathcal{E}$ varying from 0.2 to 0.8. We varied the number of users M to meet the different $\mathcal{E} = M/(Q+L)$ levels. Fig. 5 confirms that in order to outperform AMOUR (solid line) at 80% efficiency, the MC-CDMA system in the uplink should operate at 20% efficiency.

Test case 3 To test AMOUR's performance for blind equalization in uplink CDMA systems, we used the indirect blind method proposed in [6, 3], which first estimates the channels and then applies ZF or MMSE equalization. We simulated 100 Monte-Carlo realizations of Rayleigh channels of order 1, and tested performance using 34 blocks of received data for equalization. Blind performance is compared with the theoretical BER formula derived in [3] with CSI available (see Fig. 6). Interestingly, the blind method suffers only a small (2dB) penalty compared with the theoretical bound that assumes perfect channel knowledge.

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5. REFERENCES

[1] V. M. DaSilva and E. S. Sousa, "Multicarrier orthogonal CDMA signals for quasi-synchronous communication

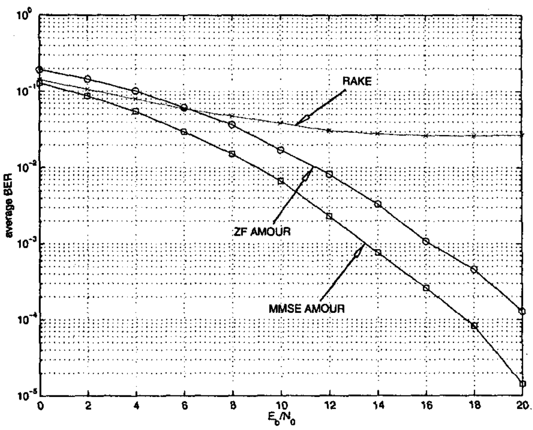

Figure 3: AMOUR vs. RAKE receivers

systems," *IEEE Journal on Selected Areas in Comm.*, pp. 842–852, 1994.

[2] D. Gesbert, J. Sorelius, and A. Paulraj, "Blind multi-user MMSE detection of CDMA signals," in *Proc. of ICASSP*, 1998, pp. 3161–3164.

[3] G. B. Giannakis, Z. Wang, A. Scaglione, and S. Barbarossa, "Mutually orthogonal transceivers for blind uplink CDMA irrespective of multipath channel nulls," in *Proc. of ICASSP*, Phoenix, AZ, Mar. 1999.

[4] M. Honig, U. Madhow, and S. Verdu, "Blind adaptive multiuser detection," *IEEE Trans. on Info. Theo.* , vol. 41, pp. 944–960, 1995.

[5] A. Scaglione, and G. B. Giannakis, "Design of user codes in QS-CDMA systems for MUI elimination in unknown multipath," *IEEE Communication Letters*, February 1999.

[6] A. Scaglione, G. B. Giannakis, and S. Barbarossa, "Self-recovering multirate equalizers using redundant filterbank precoders," in *Proc. of ICASSP*, 1998, pp. 3501–3504.

[7] M. K. Tsatsanis, "Inverse filtering criteria for CDMA systems," *IEEE Trans. on Sig. Processing*, vol. 45, no. 1, pp. 102–112, Jan. 1997.

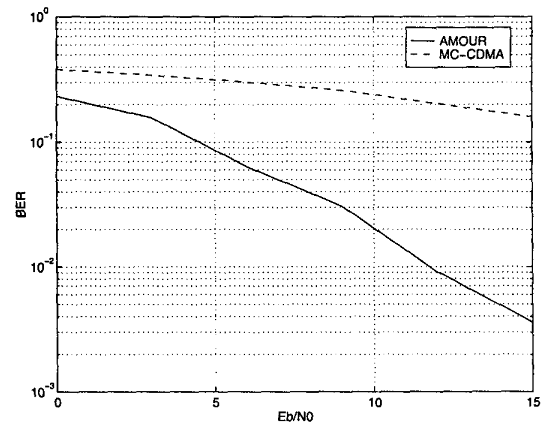

Figure 4: ZF AMOUR vs. MC-CDMA with MRC

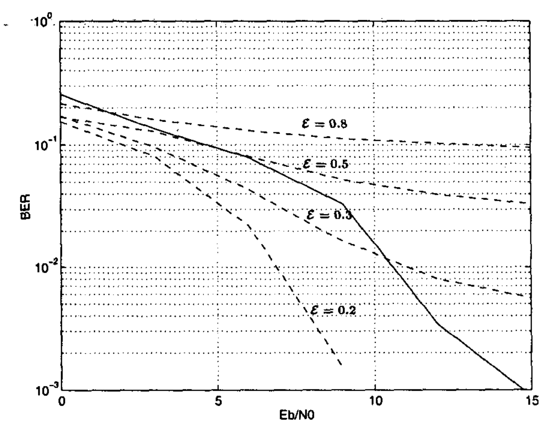

Figure 5: ZF AMOUR vs. MC-CDMA with MRC

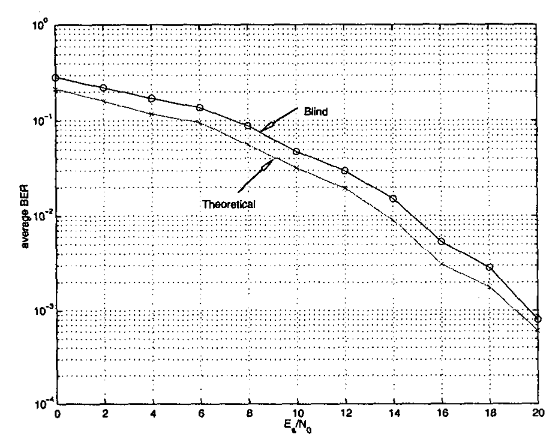

Figure 6: Performance of blind AMOUR equalizer