GENERALIZED MULTI-CARRIER CDMA FOR MUI/ISI-RESILIENT UPLINK TRANSMISSIONS IRRESPECTIVE OF FREQUENCY-SELECTIVE MULTIPATH

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Abstract Relying on symbol blocking and judicious design of user codes, this paper develops a Generalized Multicarrier (GMC) quasi-synchronous CDMA system capable of multiuser interference (MUI) elimination and inter-symbol interference (ISI) suppression, irrespective of the wireless frequency selective channels encountered in the uplink. As the term reveals, GMC-CDMA provides a unifying framework for multicarrier (MC) CDMA systems and as this paper shows, it offers flexibility in full load (maximum number of users allowed by the available bandwidth) and in reduced load settings. A blind channel estimation algorithm is also derived. Analytic evaluation and simulations illustrate that GMC-CDMA outperforms competing MC-CDMA alternatives especially in the presence of uplink multipath channels.

1. INTRODUCTION

Mitigation of frequency selective multipath and elimination of multiuser interference have received considerable attention as they constitute the main limiting performance factors in wireless CDMA systems. Multicarrier (MC) CDMA systems have been introduced to mitigate both MUI and ISI caused by frequency selective channel effects, but they do not guarantee (blind or not) recovery of the transmitted symbols in the uplink without imposing constraints on the unknown multipath channel nulls [1, 6, 9]. In [7, 8] a spread-spectrum multicarrier multiple access scheme was developed and shown to achieve MUI

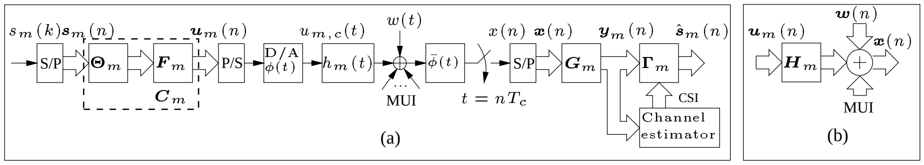

Figure 1 (a) Baseband transceiver model (b) Discrete-time equivalent channel model.

elimination irrespective of the frequency selective uplink channels. Relative to [7, 8], the generalized multicarrier (GMC) CDMA system designed in this paper, offers the following distinct features: (i) it quantifies the minimum redundancy needed for uplink bandwidth efficient transmissions; (ii) without channel coding and symbol interleaving, it establishes conditions that guarantee FIR-channel-irrespective symbol recovery with FIR linear equalizers; (iii) it offers capabilities for blind channel estimation by exploiting the redundant GMC-precoded transmission; (iv) it has low (linear) complexity and does not trade-off bandwidth efficiency in order to lower the exponential complexity of MLSE receivers. Features i)-iv) are present also in the so called AMOUR system of [4, 5, 10], which however, was designed for fully loaded systems. GMC-CDMA retains AMOUR's low complexity and bandwidth efficiency while at the same combines spread-spectrum with multicarrier features to improve performance when the system is not fully loaded.

2. SYSTEM MODEL

The baseband equivalent transmitter and receiver model for the mth user is depicted in Fig. 1(a), where $m \in [0, M_a - 1]$ and M_a is the number of active users out of a maximum M users that can be accommodated by the available bandwidth. The information stream $s_m(k)$ with symbol rate $1/T_s$ is first serial-to-parallel (S/P) converted to blocks[1] $\boldsymbol{s}_m(n)$ of length $K \times 1$ with the kth entry of the nth block denoted as: $s_{m,k}(n) := s_m(k + nK)$, $k \in [0, K-1]$. The $\boldsymbol{s}_m(n)$ blocks are multiplied by a $J \times K$ $(J > K)$ tall matrix $\boldsymbol{\Theta}_m$, which introduces redundancy and spreads the K symbols in $\boldsymbol{s}_m(n)$ by J-long codes. Precoder $\boldsymbol{\Theta}_m$ will facilitate ISI suppression, while the subsequent redundant precoder described by the tall $P \times J$ matrix $\boldsymbol{F}_m$ will accomplish MUI elimination. The precoded $P \times 1$ output vector $\boldsymbol{u}_m(n)$ is first parallel to serial (P/S) and then digital to analog converted using a chip waveform $\phi(t)$ of duration T_c with Nyquist characteristics, before being transmitted through the frequency selective channel $h_m(t)$. Although we focus on the uplink, the downlink scenario is subsumed by our model (it corresponds to having $h_m(t) = h(t)\ \forall m$). The resulting aggregate signal $x(t)$ from all active users is filtered with a receive-filter $\bar{\phi}(t)$ matched to $\phi(t)$ and then sampled at the chip rate $1/T_c$. Next, the sampled signal $x(n)$ is serial to parallel converted and processed by the digital multichannel receiver.

[1] Throughout this paper, k is the symbol index and n will be used to index blocks-of-symbols.

From the multichannel input-output viewpoint depicted in Fig. 1(b), the $K \times 1$ vector $\boldsymbol{u}_m(n)$ propagates through an equivalent channel described by the $P \times P$ lower triangular Toeplitz (convolution) matrix $\boldsymbol{H}_m$ with (i,j)th entry $h_m(i-j)$, where $h_m(l)$, $l \in [0, L]$, $m \in [0, M_a - 1]$ are the taps of the discrete chip-rate sampled FIR channels assumed to have maximum order L. In addition to transmit-receive filters, each channel $h_m(l)$, includes user quasi-synchronism in the form of delay factors (in this case $L = L_d + L_m$, where L_d captures asynchronism ($\ll P$ chips) relative to a reference user, and L_m expresses (in chips) the maximum multipath delay spread). To avoid channel-induced inter-block interference (IBI), we pad our transmitted blocks $\boldsymbol{u}_m(n)$ with L zeros (guard bits). Specifically, we design our $P \times J$ precoders $\boldsymbol{F}_m$ such that:
d1) $P = M_a J + L$ and the $L \times J$ lower submatrix of $\boldsymbol{F}_m$ is set to zero.
Under d1), the $P \times 1$ data vector $\boldsymbol{x}(n)$ received in AGN $\boldsymbol{w}(n)$ is given by:

$$\boldsymbol{x}(n) = \sum_{m=0}^{M_a-1} \boldsymbol{H}_m \boldsymbol{C}_m \boldsymbol{s}_m(n) + \boldsymbol{w}(n), \qquad \boldsymbol{C}_m := \boldsymbol{F}_m \boldsymbol{\Theta}_m. \tag{1}$$

Processing the multichannel data $\boldsymbol{x}(n)$ by the mth user's receiver amounts to multiplying it with the $J \times P$ receiver matrix $\boldsymbol{G}_m$ that yields $\boldsymbol{y}_m(n) = \boldsymbol{G}_m \boldsymbol{x}(n)$. Similar to [4], the precoder/decoder matrices $\{\boldsymbol{F}_m, \boldsymbol{G}_m\}$ will be judiciously designed such that MUI is eliminated from $\boldsymbol{x}(n)$ independent of the channels $\boldsymbol{H}_m$. Channel status information (CSI) acquired from the channel estimator (see Fig. 1) will be used to specify the linear equalizer $\boldsymbol{\Gamma}_m$ which removes ISI from the MUI-free signals $\boldsymbol{y}_m(n)$ to obtain the estimated symbols $\hat{\boldsymbol{s}}_m(n) = \boldsymbol{\Gamma}_m \boldsymbol{G}_m \boldsymbol{x}(n)$ that are passed on to the decision device.

3. MUI/ISI ELIMINATING CODES

We pursue MUI elimination from $\boldsymbol{x}(n)$ in the Z-domain [4]. Let us define $\boldsymbol{v}_P(z) := [1 \; z^{-1} \; ... \; z^{-P+1}]^{\mathrm{T}}$ ($^{\mathrm{T}}$ denotes transpose), and Z-transform the entries of $\boldsymbol{x}(n)$ in (1) to obtain: $X(n;z) := \boldsymbol{v}_P^{\mathrm{T}}(z)\boldsymbol{x}(n)$. Substituting $\boldsymbol{x}(n)$ from (1), we find:

$$X(n;z) = \sum_{m=0}^{M_a-1} H_m(z)[F_{m,0}(z) \; ... \; F_{m,J-1}(z)]\boldsymbol{\Theta}_m \boldsymbol{s}_m(n) + \boldsymbol{v}_P^{\mathrm{T}}(z)\boldsymbol{w}(n), \tag{2}$$

where $H_m(z) := \sum_{l=0}^{L} h_m(l) z^{-l}$ and $F_{m,j}(z)$ is the Z-transform of matrix $\boldsymbol{F}_m$'s jth column. Note that evaluating $X(n;z)$ at $z = \rho_{\mu,i}$ amounts to using a receiver $\boldsymbol{v}_P(\rho_{\mu,i})$ that performs a simple inner product operation $\boldsymbol{v}_P^{\mathrm{T}}(\rho_{\mu,i})\boldsymbol{x}(n)$. Hence, forming $\boldsymbol{y}_\mu := [X(n;\rho_{\mu,0}) \;\; X(n;\rho_{\mu,1}) \;\; \ldots \;\; X(n;\rho_{\mu,J-1})]^{\mathrm{T}}$ requires a receiver $\boldsymbol{G}_\mu := [\boldsymbol{v}_P(\rho_{\mu,0}) \; ... \; \boldsymbol{v}_P(\rho_{\mu,J-1})]^{\mathrm{T}}$ to obtain:

$$\boldsymbol{y}_\mu(n) = \boldsymbol{G}_\mu \boldsymbol{x}(n). \tag{3}$$

The principle behind designing MUI-free precoders $\boldsymbol{F}_m$ is to seek J points $\{\rho_{\mu,i}\}_{i=0}^{J-1}$ for every active user $\mu \in [0, M_a - 1]$ on which $X(n; z = \rho_{\mu,i})$ contains the μth user's signal of interest, while MUI from the remaining $M - 1$ users is eliminated. If in addition to MUI we also want to cancel the inter-chip interference from precoder $\boldsymbol{F}_m$, we must select:

$$F_{m,j}(\rho_{\mu,i}) = \delta(j-i)\delta(m-\mu), \quad \forall m,\mu \in [0, M_a-1], \;\; \forall j,i \in [0, J-1]. \quad (4)$$

The minimum degree polynomial $F_{m,j}(z)$ that satisfies (4) can be uniquely computed by Lagrange interpolation through the M_aJ points $\rho_{\mu,i}$ as follows [4]:

$$F_{m,j}(z) = \prod_{\substack{\mu=0 \\ (\mu,i)\neq(m,j)}}^{M_a-1} \prod_{i=0}^{J-1} \frac{1-\rho_{\mu,i}z^{-1}}{1-\rho_{\mu,i}\rho_{m,j}^{-1}}. \quad (5)$$

Because manipulation of circulant matrices can be performed with FFT, low-complexity transceivers result if $\boldsymbol{F}_m$ is formed by FFT exponentials, which corresponds to choosing $\{\rho_{\mu,i}\}_{i=0}^{J-1}$ in (5) equispaced on the unit circle as: $\rho_{\mu,i} = \exp(j2\pi(\mu+iM_a)/M_aJ)\ \forall \mu, i$. Accounting for the L trailing zeros as per d1), such a choice leads to:

$$\boldsymbol{F}_\mu = \frac{\mathrm{e}^{j2\pi\mu/M_aJ}}{M_aJ}\begin{bmatrix} \boldsymbol{v}^*_{M_aJ}(1) & \boldsymbol{v}^*_{M_aJ}(\mathrm{e}^{j2\pi/J}) & \ldots & \boldsymbol{v}^*_{M_aJ}(\mathrm{e}^{j2\pi(J-1)/J}) \\ & \boldsymbol{0}_{L\times J} & & \end{bmatrix}, \quad (6)$$

where * denotes conjugation. The degree of the mth user's jth code polynomial in (5) is M_aJ-1. Adding the L guard chips to $F_{m,j}(z)$'s inverse Z-transform, sets the number of rows for precoder $\boldsymbol{F}_m$ to $P = M_aJ + L$, which explains our choice in d1).

Next, we substitute (2) into (3) and take account of (4) to obtain the MUI-free:

$$\boldsymbol{y}_\mu(n) = \boldsymbol{D}_{H_\mu}\boldsymbol{\Theta}_\mu\boldsymbol{s}_\mu(n) + \boldsymbol{\eta}_\mu(n) \quad \Rightarrow \quad \hat{s}_\mu(n) = \boldsymbol{\Gamma}_\mu^{zf}\boldsymbol{y}_\mu(n) := \boldsymbol{\Theta}_\mu^\dagger\boldsymbol{D}_{H_\mu}^\dagger\boldsymbol{y}_\mu(n), \quad (7)$$

where $\boldsymbol{D}_{H_\mu} := \mathrm{diag}[H_\mu(\rho_{\mu,0}) \ldots H_\mu(\rho_{\mu,J-1})]$ is a $J\times J$ diagonal matrix with entries $H_\mu(\rho_{\mu,j})$, $\boldsymbol{\eta}_\mu(n) := \boldsymbol{G}_\mu\boldsymbol{w}(n)$, and † denotes pseudoinverse. With $\mathcal{C}^K$ denoting the vector space of complex K-tuples, suppose we design $\boldsymbol{\Theta}_\mu$ in (7) to satisfy:

d2) $J \geq K + L$ and *any* $J-L$ rows of $\boldsymbol{\Theta}_\mu$ span the $\mathcal{C}^K$ row vector space;

Notice that d2) can always be checked and enforced at the transmitter. Under d2), $\boldsymbol{D}_{H_\mu}\boldsymbol{\Theta}_\mu$ in (7) will always be full rank, because the added redundancy ($\geq L$) can afford even L diagonal entries of $\boldsymbol{D}_{H_\mu}$ to be zero (recall that $H_\mu(z)$ has maximum order L and thus at most L nulls). Therefore, identifiability of $\boldsymbol{s}_\mu(n)$ can be guaranteed irrespective of the multipath channel $H_\mu(z)$. Possible choices for $\boldsymbol{\Theta}_\mu$ that are flexible enough for our design include: (a)the $J\times K$ Vandermonde matrix $\boldsymbol{\Theta}_\mu := [\boldsymbol{v}(\rho_{\mu,0}, K) \ldots \boldsymbol{v}(\rho_{\mu,J-1}, K)]^{\mathrm{T}}$ used in the AMOUR system [4], which for $\rho_{\mu,i} = \exp(j2\pi(\mu+iM_a)/(M_aJ))$ becomes $\exp(j2\pi\mu/(M_aJ))$ times a truncated $J\times K$ FFT matrix; (b)a truncated $J\times K$ Walsh-Hadamard (WH) matrix; (c)a $J\times K$ matrix with equiprobable ±1 random entries.

Channel estimation, blind or pilot-based, is needed to build a ZF-equalizer $\boldsymbol{\Gamma}_\mu^{zf}$ in (7), which will guarantee ISI-free detection (MMSE equalizers are also possible). For $\boldsymbol{\Theta}_\mu$'s selected as in (a), a blind channel estimation method was developed in [4]. In the sequel, we will establish the identifiability conditions and derive a more general blind channel estimation algorithm allowing spread-spectrum precoders $\boldsymbol{\Theta}_\mu$ to be chosen as in (b) or (c).

4. BLIND CHANNEL ESTIMATION

We will suppose here that instead of d2), we design $\boldsymbol{\Theta}_\mu$ such that:
d2′) $J \geq K + L$ and *any* K rows of $\boldsymbol{\Theta}_\mu$ span the $\mathcal{C}^K$ row vector space.
Note that when $J = K + L$, d2′) is equivalent to d2). To estimate $H_\mu(z)$ under d2′) in the noiseless case, user μ collects N blocks of $\boldsymbol{y}_\mu(n)$ in a $J \times N$ matrix $\boldsymbol{Y}_\mu := [\boldsymbol{y}_\mu(0) \cdots \boldsymbol{y}_\mu\ (N-1)]$ and forms $\boldsymbol{Y}_\mu \boldsymbol{Y}_\mu^{\mathcal{H}} = \boldsymbol{D}_{H_\mu} \boldsymbol{\Theta}_\mu \boldsymbol{S}_\mu \boldsymbol{S}_\mu^{\mathcal{H}} \boldsymbol{\Theta}_\mu^{\mathcal{H}} \boldsymbol{D}_{H_\mu}^{\mathcal{H}}$, where $\boldsymbol{S}_\mu := [\boldsymbol{s}_\mu(0) \cdots \boldsymbol{s}_\mu(N-1)]_{K \times N}$. User μ also chooses:
d3) N large enough so that $\boldsymbol{S}_\mu \boldsymbol{S}_\mu^{\mathcal{H}}$ is of full rank K.
Under d2′) and d3), we have $\text{rank}(\boldsymbol{Y}_\mu \boldsymbol{Y}_\mu^{\mathcal{H}}) = K$ and range space $\mathcal{R}(\boldsymbol{Y}_\mu \boldsymbol{Y}_\mu^{\mathcal{H}}) = \mathcal{R}(\boldsymbol{D}_{H_\mu} \boldsymbol{\Theta}_\mu)$. Thus, the nullity of $\boldsymbol{Y}_\mu \boldsymbol{Y}_\mu^{\mathcal{H}}$ is $\nu(\boldsymbol{Y}_\mu \boldsymbol{Y}_\mu^{\mathcal{H}}) = J - K$. Further, the eigen-decomposition

$$\boldsymbol{Y}_\mu \boldsymbol{Y}_\mu^{\mathcal{H}} = [\boldsymbol{U}\ \tilde{\boldsymbol{U}}] \begin{bmatrix} \boldsymbol{\Sigma}_{K \times K} & \mathbf{0}_{K \times (J-K)} \\ \mathbf{0}_{(J-K) \times K} & \mathbf{0}_{(J-K) \times (J-K)} \end{bmatrix} \begin{bmatrix} \boldsymbol{U}^{\mathcal{H}} \\ \tilde{\boldsymbol{U}}^{\mathcal{H}} \end{bmatrix} \tag{8}$$

yields the $J \times (J-K)$ matrix $\tilde{\boldsymbol{U}}$ whose columns span the null space $\mathcal{N}(\boldsymbol{Y}_\mu \boldsymbol{Y}_\mu^{\mathcal{H}})$. Because the latter is orthogonal to $\mathcal{R}(\boldsymbol{Y}_\mu \boldsymbol{Y}_\mu^{\mathcal{H}}) = \mathcal{R}(\boldsymbol{D}_{H_\mu} \boldsymbol{\Theta}_\mu)$, it follows that $\tilde{\boldsymbol{u}}_l^{\mathcal{H}} \boldsymbol{D}_{H_\mu} \boldsymbol{\Theta}_\mu = \mathbf{0}_{1 \times K}^{\mathcal{H}}$, $l \in [1, J-K]$, where $\tilde{\boldsymbol{u}}_l$ denotes the lth column of $\tilde{\boldsymbol{U}}$. With $\boldsymbol{D}_{u_l}$ denoting the diagonal matrix $\boldsymbol{D}_{u_l} := \text{diag}[\tilde{\boldsymbol{u}}_l^{\mathcal{H}}]$ and $\boldsymbol{d}_{H_\mu}^T := [H(\rho_{\mu,0}), \ldots, H(\rho_{\mu,J-1})]$, we can write $\tilde{\boldsymbol{u}}_l^{\mathcal{H}} \boldsymbol{D}_{H_\mu} = \boldsymbol{d}_{H_\mu}^T \boldsymbol{D}_{u_l}$. It can be easily verified that with $\boldsymbol{h}_\mu^T := [h_\mu(0), \ldots, h_\mu(L)]$ and with $\boldsymbol{V}_\mu$ being a $(L+1) \times J$ matrix whose $(l+1, j+1)$st entry is $\rho_{\mu,j}^{-l}$, one can write $\boldsymbol{d}_{H_\mu}^T = \boldsymbol{h}_\mu^T \boldsymbol{V}_\mu$. This yields

$$\boldsymbol{h}_\mu^T \boldsymbol{V}_\mu [\boldsymbol{D}_{u_0} \boldsymbol{\Theta}_\mu \vdots \cdots \vdots \boldsymbol{D}_{u_{J-K}} \boldsymbol{\Theta}_\mu] = \mathbf{0}_{1 \times K(J-K)}^{\mathrm{T}}, \tag{9}$$

from which one can solve for $\boldsymbol{h}_\mu$. We have established uniqueness (within a scale) in solving for $\boldsymbol{h}_\mu$, but we omit the proof due to lack of space. In the noisy case, if the covariance matrix of $\boldsymbol{\eta}_\mu$ is known, we can prewhiten $\boldsymbol{Y}_\mu$ before SVD, and a similar blind channel algorithm can be devised. We summarize our results in the following:
Theorem: *i) Design a GMC-CDMA system according to d1) and d2), and suppose that CSI is available at the receiver using pilot symbols. User symbols $\boldsymbol{s}_\mu(n)$ can then be always recovered with linear processing as in (7), irrespective of frequency selective multipath channels up to order L. ii) A GMC-CDMA system designed according to d1), d2′), and d3) guarantees blind identifiability (within a scale) of channels $h_\mu(l)$ with maximum order L, and the channel estimate is found by solving (9) for the null eigenvector.*

Note that unlike [7, 8], even blind channel-irrespective symbol recovery is assured by the Theorem, without bandwidth consuming channel coding/interleaving, and with linear receiver processing (as opposed to the exponentially complex MLSE used in [7, 8]).

5. UNIFYING FRAMEWORK

A number of multiuser multicarrier schemes fall under the model of Fig. 1(a). We outline some of them in this section by describing their baseband discrete-time equivalent models in order to illustrate the generality of GMC-CDMA.

Multicarrier CDMA (MC-CDMA), [3, 11]: For this multicarrier scheme, no blocking of data symbols occurs at the receiver and the first precoding matrix $\mathbf{\Theta}_m$ is a $Q \times 1$ vector $\boldsymbol{\theta}_m$ (the spreading filter). The $\boldsymbol{F}_m$ matrix is no longer user dependent, as it is selected to be a $Q \times Q$ IFFT matrix augmented either by an $L \times Q$ all-zero matrix as in d1), or, by an $L \times Q$ cyclic prefix matrix to yield a $(Q + L) \times Q$ matrix corresponding to an OFDM precoder (modulator). At the OFDM receiver, the cyclic prefix is discarded. For MC-CDMA to have the same bandwidth as GMC-CDMA, Q must be chosen as: $Q = \lfloor P/K \rfloor - L$. The frequency selective channel matrix $\boldsymbol{H}_m$ is $(Q + L) \times (Q + L)$ and for the zero-padded transmissions we have instead of (1): $\boldsymbol{x}(n) = \sum_{m=0}^{M_a-1} \boldsymbol{H}_m \boldsymbol{F} \boldsymbol{\theta}_m s_m(n) + \boldsymbol{w}(n)$. At the receiver end, the matrix $\boldsymbol{G}$ is selected to be a $Q \times (Q+L)$ extended FFT matrix with entries given by powers of $\exp(j2\pi/Q)$, or, a $Q \times Q$ FFT matrix increased by the $L \times Q$ leading zeros to discard the cyclic prefix used at the transmitter. Matrix $\mathbf{\Gamma}_m$ becomes now an $1 \times Q$ vector to be chosen according to the selected multiuser detection technique. Although simpler than GMC-CDMA, because matrix $[\boldsymbol{H}_0 \boldsymbol{F}_0 \boldsymbol{\theta}_0 \ \ldots \ \boldsymbol{H}_{M_a-1} \boldsymbol{F}_{M_a-1} \boldsymbol{\theta}_{M_a-1}]$ is not guaranteed to be invertible, symbol recovery is not assured in MC-CDMA even when CSI is available.

Multicarrier Direct-Sequence CDMA (MC-DS-CDMA), [1, 2]: Both GMC-CDMA and MS-DS-CDMA are multicarrier techniques using a block precoder $\mathbf{\Theta}_m$, except that for MC-DS-CDMA the precoder is particularized to be a $QK \times K$ block diagonal matrix with blocks of size $Q \times 1$. The dimensionality of the OFDM transmitter-receiver pair $\{\boldsymbol{F}_m, \boldsymbol{G}_m\}$ is $(QK + L) \times QK$ and $QK \times (QK + L)$, respectively. Keeping in mind the bandwidth constraint we choose $Q = \lfloor (P - L)/K \rfloor$. Each user transmits K symbols in parallel, spreads them with codes of length Q, and modulates each spread symbol with a specific set of Q subcarriers. Same as the $\mathbf{\Theta}_m$ precoder, $\mathbf{\Gamma}_m$ has a block diagonal structure with K blocks of size $1 \times Q$. If $K = 1$, then $Q = P - L$ and MC-DS-CDMA reduces to MC-CDMA.

Multitone CDMA (MT-CDMA), [9]: This multicarrier technique applies the same data mapping as MC-DS-CDMA. However, in contrast with MC-DS-CDMA, MT-CDMA uses a block diagonal precoder $\mathbf{\Theta}_m$ with K blocks of size $QQ' \times 1$. The $KQ \times KQQ'$ precoder is given by: $\boldsymbol{F}_m = [\boldsymbol{v}_{KQ}^*(1) \ \boldsymbol{v}_{KQ}^*(j2\pi/(KQQ')) \ldots \boldsymbol{v}_{KQ}^*(j2\pi(KQQ' - 1)/(KQQ'))]$, while the receiver is: $\boldsymbol{G}_m = [\mathbf{0}_{KQ \times L} \ \boldsymbol{v}_{KQ}(1) \ \boldsymbol{v}_{KQ}(j2\pi/(KQQ')) \ldots \boldsymbol{v}_{KQ}(j2\pi(KQQ' - 1)/(KQQ'))]^{\mathrm{T}}$. The system has the advantage of allowing a bigger size $\mathbf{\Theta}_m$ matrix (longer spreading codes), which reduces self-interference and MUI at the expense of introducing subcarrier interference.

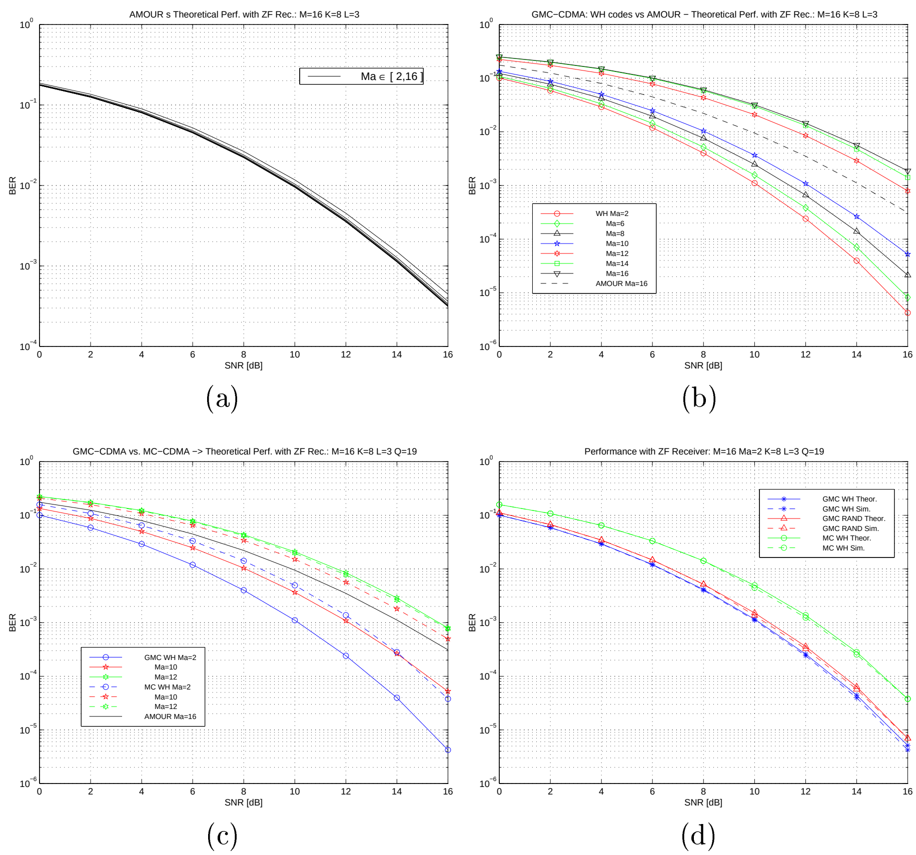

Figure 2 (a) AMOUR: different number of active users. (b) GMC-CDMA: Theoretical WH vs. AMOUR. (c) Theoretical: GMC-CDMA vs. MC-CDMA; (d) Simulation $M_a = 2$: GMC-CDMA vs. MC-CDMA.

6. PERFORMANCE ANALYSIS

The theoretical BER evaluation for an AMOUR system [4] that uses a ZF equalizer can be extended to a GMC-CDMA system with $\mathbf{\Gamma}_{\mu}^{zf}$ as in (7). Perfect knowledge of the channel is assumed. Similar to [4], we choose for simplicity a BPSK constellation to obtain in terms of the Q-function an average bit error rate (BER) $\bar{P}e$:

$$\bar{P}e = \frac{1}{MK} \sum_{m=0}^{M_a-1} \sum_{k=0}^{K-1} \mathrm{Q}\left(\sqrt{\frac{1}{\bar{\boldsymbol{g}}_{m,k}^{\mathrm{H}} \bar{\boldsymbol{g}}_{m,k} E_{m,k}}} \sqrt{\frac{2E_b}{N_0}}\right), \qquad (10)$$

where $\bar{\boldsymbol{g}}_{m,k}^{\mathcal{H}}$ is the kth row of matrix $\mathbf{\Gamma}_m \boldsymbol{G}_m$, $E_{m,k} := \sum_{i=0}^{P-1} |c_{m,k}(i)|^2$ is the energy of the mth user's kth code, and E_b/N_0 is the bit SNR.

First, we plot (10) for an AMOUR system designed for $M = 16$ users with data symbols drawn from a BPSK constellation, each one experiencing a Rayleigh fading channel of order $L = 3$. The length of the $\boldsymbol{s}_m(n)$ blocks is $K = 8$. To avoid channel dependent performance, we averaged (10) over 100 Monte Carlo channel realizations. We decrease gradually the number of active users in the system from 16 down to 2 and each time we redesign AMOUR to incorporate the available bandwidth. The theoretical BER curves from Fig. 2(a) show that under different load conditions there is practically no difference in performance. Next, keeping the same set up we compare GMC-CDMA using WH codes versus AMOUR under different load conditions. It is clear from Fig. 2(b) that under 65% load the WH codes outperform the AMOUR codes. In Fig. 2(c) the theoretical performance of GMC-CDMA with WH codes and same parameters as before is compared with an equivalent MC-CDMA system using OFDM transceivers with trailing zeros. GMC-CDMA has a lower BER than MC-CDMA independent of the number of active users if the WH codes are replaced by AMOUR codes when the system load increases over 65%. Code redesign/switching has the drawback of requiring knowledge of the system load at the mobile transmitter. To verify our theoretical claims, we simulated GMC (both with WH and random codes) and MC with maximum number of users $M = 16$, of which $M_a = 2$ were active and used block size $K = 8$. CSI was assumed to be perfect ($L = 3$). The results plotted in Fig. 2(d) show that at 12% load (value in our region of interest) both WH and random codes exhibit improved performance over MC-CDMA.

Acknowledgments: The work in this paper was supported by NSF CCR grant no. 98-05350 and the NSF Wireless Initiative grant no. 99-79443. The authors also wish to thank prof. S. Barbarossa for discussions on this and related subjects.

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