WIDEBAND GENERALIZED MULTI-CARRIER CDMA OVER FREQUENCY-SELECTIVE WIRELESS CHANNELS*

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ABSTRACT

Relying on symbol blocking and judicious design of user codes, this paper builds on a Generalized Multicarrier (GMC) quasi-synchronous CDMA system capable of multiuser interference (MUI) elimination and intersymbol interference (ISI) suppression with guaranteed symbol recovery, regardless of the wireless frequency selective channels. GMC-CDMA provides a unifying framework for multicarrier (MC) CDMA systems and offers flexibility in full load (maximum number of users allowed by the available bandwidth) as well in reduced load settings. 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 [1, 6, 8] have gained popularity because they mitigate both MUI and ISI caused by frequency selective channels. A spread-spectrum multicarrier multiple access scheme was developed in [7] for MUI elimination and mitigation of frequency selective uplink channels. But no existing MC or DS CDMA scheme guarantees (blind or not) symbol recovery in the uplink without imposing constraints on the unknown multipath channel nulls.

A generalized multicarrier (GMC) CDMA system is developed in this paper with the following distinct features:

i) it shows that existing DS-CDMA and various MC-CDMA schemes are special cases of a general all-digital baseband model;

ii) without channel coding and symbol interleaving, it establishes conditions that guarantee FIR-channel-irrespective symbol recovery with FIR equalizers;

iii) it offers capability for blind channel estimation and equalization by exploiting the redundant GMC-precoded transmission [4];

iv) it has low (linear) complexity and does not trade bandwidth efficiency for lowering the exponential complexity of MLSE receivers as in [7].

Features ii)-iv) are present also in the so called AMOUR system of [5], which however, is designed for fully loaded systems. GMC-CDMA designed here retains AMOUR's low complexity and bandwidth efficiency while at the same time it combines spread-spectrum with multicarrier features to improve performance when the system is not fully loaded.

2. BLOCK SYSTEM MODELING

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 symbols $s_m(k)$ with rate $1/T_s$ are first serial-to-parallel (S/P) converted to blocks $\boldsymbol{s}_m(n)$ 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 $P \times 1$ 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, 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.

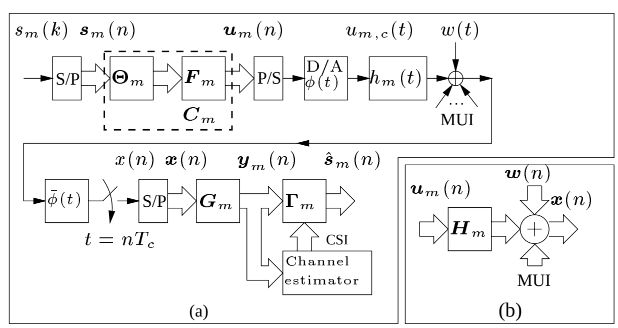

Figure 1: Baseband transceiver model

From the multichannel input-output viewpoint depicted in Fig. 1(b), the transmitted block $\boldsymbol{u}_m(n)$ propagates through an equivalent channel described by the $P \times P$ lower triangular Toeplitz (convolution) matrix $\boldsymbol{H}_m$ with (i,j)th

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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 $L = L_a + L_d$, where L_a captures asynchronism ($\ll P$ chips) relative to a reference user, and L_d 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) the $L \times J$ lower submatrix of $\boldsymbol{F}_m$ is set to zero.
Under d1), the received $P \times 1$ vector $\boldsymbol{x}(n)$ in AGN $\boldsymbol{w}(n)$ is:

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

Processing $\boldsymbol{x}(n)$ by the mth user's receiver amounts to multiplying it with the $J \times P$ matrix $\boldsymbol{G}_m$ that yields $\boldsymbol{y}_m(n) = \boldsymbol{G}_m \boldsymbol{x}(n)$. Similar to [5], the precoder/decoder matrices $\{\boldsymbol{F}_m, \boldsymbol{G}_m\}$ will be designed such that MUI is eliminated from $\boldsymbol{x}(n)$ regardless 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.

Before we design such precoders/decoders for a reduced-load system (as opposed to the full load design in [5]), let us first show the generality of such a precoding structure.

3. AN ALL-DIGITAL UNIFICATION

A number of CDMA schemes fall under the model of Fig. 1(a). Single-carrier DS-CDMA with processing gain P and spreading code $\mathbf{c}_m := [c_m(0) \ldots c_m(P-1)]^T$ can be implemented with our model by setting $K = 1$, $J = P$ (without considering d1). The $P \times K$ matrix $\boldsymbol{C}_m$ in (1) reduces to the $P \times 1$ symbol-periodic code vector $\mathbf{c}_m$. In the sequel, we will also show how three multicarrier schemes can be implemented digitally as special cases of GMC-CDMA. With their discrete-time equivalent models, it becomes clear that symbol recovery in the uplink is not always possible.
Multicarrier CDMA (MC-CDMA) [3, 9]: MC-CDMA combines DS-CDMA with OFDM modulation. For this scheme, no blocking occurs at the transmitter (i.e., $K = 1$) and $\boldsymbol{\Theta}_m$ is a $J \times 1$ vector $\boldsymbol{\theta}_m$ (the spreading filter). The $\boldsymbol{F}_m$ matrix is no longer user dependent, and with $P = J + L$, it is selected to be a $P \times J$ matrix consisting of an $J \times J$ IFFT matrix augmented either by an $L \times J$ all-zero matrix as in d1), or, by an $L \times J$ cyclic prefix matrix. In the latter, $\boldsymbol{F}_m$ corresponds to an OFDM precoder, and the cyclic prefix is discarded at the receiver by choosing $\boldsymbol{G}_m = \boldsymbol{G}$ $\forall m$ to be a $J \times P$ matrix composed of L all-zero columns followed by a $J \times J$ FFT matrix. For zero-padded transmissions, (1) becomes $\boldsymbol{x}(n) = \sum_{m=0}^{M_a-1} \boldsymbol{H}_m \boldsymbol{F} \boldsymbol{\theta}_m s_m(n) + \boldsymbol{w}(n)$. In the downlink, $\boldsymbol{H}_m$ are common to all the users (i.e., $\boldsymbol{H}_m = \boldsymbol{H}$ $\forall m$) and $\boldsymbol{H}\boldsymbol{F}$ is invertible. Therefore, the user symbols can be recovered if and only if the vectors $\boldsymbol{\theta}_m$, $m = 0, \ldots, M_a - 1$, are linearly independent. In the uplink, because matrix $[\boldsymbol{H}_0 \boldsymbol{F} \boldsymbol{\theta}_0 \ \ldots \ \boldsymbol{H}_{M_a-1} \boldsymbol{F} \boldsymbol{\theta}_{M_a-1}]$ is not guaranteed to be invertible, symbol recovery is not assured in MC-CDMA even when CSI is available. At the receiver end, $\boldsymbol{G}_m = \boldsymbol{G}$ is selected to be a $J \times P$ extended FFT matrix with (k, l) entry $\exp(j2\pi kl/J)$. In either case, we can write the output of receiver matrix $\boldsymbol{G}$ as:

$$\boldsymbol{y}(n) = \sum_{m=0}^{M_a-1} \operatorname{diag}[H_m(e^{j0}) \cdots H_m(e^{j\frac{2\pi}{J}(J-1)})] \; \boldsymbol{\theta}_m s_m(n) + \boldsymbol{\eta}(n). \quad (2)$$

Matrix $\boldsymbol{\Gamma}_m$ becomes now an $1 \times J$ vector to be chosen according to the selected multiuser detector (RAKE, MMSE, or decorrelator). Although simpler than GMC-CDMA, MC-CDMA is prone to MUI in the uplink and power control is required to cope with near-far effects. Compared to multicarrier systems that can guarantee MUI-free reception regardless of the encountered multipath [5, 7], MC-CDMA also requires a more complex channel estimation algorithm especially in the uplink (see [7] and references therein).
Multicarrier DS-CDMA (MC-DS-CDMA) [1, 2]: As GMC-CDMA, MC-DS-CDMA uses a block precoder $\boldsymbol{\Theta}_m$. For MC-DS-CDMA the $QK \times K$ precoder is the concatenation of $QK \times K$ diagonal matrices. The qth diagonal matrix in the stack is formed as $\theta(q)\boldsymbol{I}_K$ where $\theta(q)$ is the qth chip of the Q-long spreading code, $q \in [0, Q-1]$ and $\boldsymbol{I}_K$ denotes a $K \times K$ identity matrix. The precoder $\boldsymbol{F}$ reduces to a $QJ \times QK$ block diagonal matrix with blocks of size $J \times K$, each corresponding to the same OFDM modulator (i.e., a $K \times K$ FFT matrix augmented by either an $L \times K$ cyclic prefix, or, an all zero $L \times K$ matrix). So, each user transmits K symbols in parallel, spreads them with codes of length Q, and modulates the resulting chip sequences on a set of K subcarriers. For MC-DS-CDMA to have same bandwidth as GMC-CDMA, we must choose $Q = \lfloor P/J \rfloor$. Similar to the transmitter, the $QK \times QJ$ receiver matrix $\boldsymbol{G}$ is block diagonal with blocks of size $K \times J$ that are identical OFDM receivers. The output of $\boldsymbol{G}$ is a $QK \times 1$ vector $\boldsymbol{y}(n) = [\boldsymbol{y}_0^T(n) \cdots \boldsymbol{y}_{Q-1}^T(n)]^T$, where

$$\boldsymbol{y}_q(n) = \sum_{m=0}^{M_a-1} \operatorname{diag}[H_m(e^{j0}) \cdots H_m(e^{j\frac{2\pi}{K}(K-1)})] \cdot \theta(q) \boldsymbol{s}_m(n) + \boldsymbol{\eta}(n). \quad (3)$$

Same as the $\boldsymbol{\Theta}_m$ precoder, the $K \times QK$ matrix $\boldsymbol{\Gamma}_m$ is formed by concatenating Q $K \times K$ diagonal matrices. For $K = 1$, MC-DS-CDMA reduces to DS-CDMA for flat fading channels. Proposed as a scheme for reducing MUI in quasi-synchronous uplink transmissions through frequency selective channels, MC-DS-CDMA turns out to be more sensitive than MC-CDMA to channel multipath. If in the worst case $\{H_m(e^{j2\pi k/K})\}_{k=0}^{K-1}$ is zero at L frequencies (channel order is assumed to be L) then L information symbols are lost at the MC-DS-CDMA receiver, whereas in MC-CDMA only L chips out of Q are lost. Therefore, MC-DS-CDMA does not offer frequency diversity gain, and error correcting codes are necessary to recover lost symbols [1].
Multitone CDMA (MT-CDMA) [8]: MT-CDMA is similar to MC-DS-CDMA, but uses a frequency separation between subcarriers equal to the inverse of the symbol's duration prior to the spreading operation, leading to a closer subcarrier spacing than MC-DS-CDMA.

Let us define $\boldsymbol{v}_P(z) := [1 \ z^{-1} \ ... \ z^{-P+1}]^T$ ($^{\mathrm{T}}$ denotes transpose) and let $f_m(n)$ denote the mth user's Q-long spreading code. The mth user's transmission can be described as $\boldsymbol{u}_m(n) = \boldsymbol{F}_m \boldsymbol{\Theta} \boldsymbol{s}_m(n)$, where $\boldsymbol{F}_m = \mathrm{diag}[f_m(0) \cdots f_m(Q-1)]$ and $\boldsymbol{\Theta} = [\boldsymbol{v}_Q(e^{j0}) \ ... \ \boldsymbol{v}_Q(\exp(j2\pi(K-1)/Q)]$. Because $P > L$, we have that IBI due to channel memory entails no more than two blocks, namely $\boldsymbol{s}_m(n)$ and $\boldsymbol{s}_m(n-1)$. Thus, the $Q \times 1$ receiver data block can be written as: $\boldsymbol{x}(n) = \sum_{m=0}^{M_a-1} \left(\boldsymbol{H}_m^{(0)} \boldsymbol{F}_m \boldsymbol{\Theta} \boldsymbol{s}_m(n) + \boldsymbol{H}_m^{(1)} \boldsymbol{F}_m \boldsymbol{\Theta} \boldsymbol{s}_m(n-1) \right) + \boldsymbol{w}(n)$, where $\boldsymbol{H}_m^{(0)}$ and $\boldsymbol{H}_m^{(1)}$ are $Q \times Q$ convolution matrices. If an all-zero $L \times Q$ matrix is appended as a suffix to $\boldsymbol{F}_m$ (direction not pursued in [8]) then IBI is avoided Thus, the received vector becomes:

$$\boldsymbol{x}(n) = \sum_{m=0}^{M_a-1} \boldsymbol{H}_m^{(0)} \boldsymbol{F}_m \boldsymbol{\Theta} \boldsymbol{s}_m(n) + \boldsymbol{w}(n). \tag{4}$$

Since the matrix $[\boldsymbol{H}_0^{(0)} \boldsymbol{F}_0 \boldsymbol{\Theta}, \ldots, \boldsymbol{H}_{M_a-1}^{(0)} \boldsymbol{F}_{M_a-1} \boldsymbol{\Theta}]$ is not always invertible, symbol recovery is not assured.

4. MUI/ISI ELIMINATING GMC CODES

In this section, we address the symbol recovery problem present in multicarrier CDMA schemes and design an MUI-free GMC-CDMA system that guarantees symbol recovery regardless of the multipath channels.

First, we pursue MUI elimination from $\boldsymbol{x}(n)$ in the Z-domain [5]. Let us define the Z-transform of $\boldsymbol{x}(n)$ in (1) as: $X(n;z) := \boldsymbol{v}_P^{\mathrm{T}}(z) \boldsymbol{x}(n)$. Substituting $\boldsymbol{x}(n)$ from (1), we find:

$$\begin{aligned} X(n;z) = \sum_{m=0}^{M_a-1} & H_m(z) [F_{m,0}(z) \ ... \ F_{m,J-1}(z)] \\ & \cdot \boldsymbol{\Theta}_m \boldsymbol{s}_m(n) + \boldsymbol{v}_P^T(z) \boldsymbol{w}(n), \end{aligned} \tag{5}$$

where $H_m(z) := \sum_{l=0}^{L} h_m(l) z^{-l}$ and $F_{m,j}(z)$ is the Z-transform of $\boldsymbol{F}_m$'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 operation $\boldsymbol{v}_P^T(\rho_{\mu,i}) \boldsymbol{x}(n)$. Hence, forming $\boldsymbol{y}_\mu := [X(n;\rho_{\mu,0}) \ X(n;\rho_{\mu,1}) \ \ldots \ X(n;\rho_{\mu,J-1})]^T$ requires a receiver $\boldsymbol{G}_\mu := [\boldsymbol{v}_P(\rho_{\mu,0}) \ ... \ \boldsymbol{v}_P(\rho_{\mu,J-1})]^T$ to obtain: $\boldsymbol{y}_\mu(n) = \boldsymbol{G}_\mu \boldsymbol{x}(n)$.

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 $\forall m, \mu \in [0, M_a-1]$, and $\forall j, i \in [0, J-1]$:

$$F_{m,j}(\rho_{\mu,i}) = \delta(j-i)\delta(m-\mu). \tag{6}$$

The minimum degree polynomial $F_{m,j}(z)$ that satisfies (6) can be uniquely determined if we choose $F_{m,j}(z)$ to have order $M_a J - 1$, which gives the code length $P = M_a J + L$ considering the L trailing zeros as per d1). Codes $F_{m,j}(z)$ can be computed by Lagrange interpolation through the $M_a J$ points $\rho_{\mu,i}$ as follows:

$$F_{m,j}(z) = \prod_{\substack{\mu=0 \\ (\mu,i)\neq(m,j)}}^{M_a-1} \prod_{i=0}^{J-1} \frac{1 - \rho_{\mu,i} z^{-1}}{1 - \rho_{\mu,i} \rho_{m,j}^{-1}}. \tag{7}$$

Because manipulation of circulant matrices can be performed with FFT, low-complexity transceivers result if $\boldsymbol{F}_m$ is an FFT matrix, which corresponds to choosing $\{\rho_{\mu,i}\}_{i=0}^{J-1}$ in (7) equispaced on the unit circle as: $\rho_{\mu,i} = \exp(j2\pi(\mu + iM_a)/M_a J) \ \forall \mu, i$. With such $\rho_{\mu,j}$ and the code design (6) we can obtain the MUI-free:

$$\begin{aligned} & \boldsymbol{y}_\mu(n) = \boldsymbol{D}_{H_\mu} \boldsymbol{\Theta}_\mu \boldsymbol{s}_\mu(n) + \boldsymbol{\eta}_\mu(n) \\ \Rightarrow \quad & \hat{\boldsymbol{s}}_\mu(n) = \boldsymbol{\Gamma}_\mu^{zf} \boldsymbol{y}_\mu(n) := \boldsymbol{\Theta}_\mu^\dagger \boldsymbol{D}_{H_\mu}^\dagger \boldsymbol{y}_\mu(n), \end{aligned} \tag{8}$$

where $\boldsymbol{D}_{H_\mu} := \mathrm{diag}[H_\mu(\rho_{\mu,0}) \ ... \ H_\mu(\rho_{\mu,J-1})]$ 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 complex K-tuples, suppose we design $\boldsymbol{\Theta}_\mu$ in (8) to satisfy:

d2) $J \geq K + L$ and *any* $J - L$ rows of $\boldsymbol{\Theta}_\mu$ span $\mathcal{C}^K$, which can always be checked and enforced at the transmitter. Under d2), $\boldsymbol{D}_{H_\mu} \boldsymbol{\Theta}_\mu$ in (8) will always be full rank, because the 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). Therefore, recovery of $\boldsymbol{s}_\mu(n)$ can be guaranteed regardless of $H_\mu(z)$. Possible choices for $\boldsymbol{\Theta}_\mu$ that are flexible enough for our design include:

(a) the $J \times K$ Vandermonde matrix $\boldsymbol{\Theta}_\mu := [\boldsymbol{v}(\rho_{\mu,0}, K) \cdots \boldsymbol{v}(\rho_{\mu,J-1}, K)]^T$ used in the AMOUR system [5], which for $\rho_{\mu,i} = \exp(j2\pi(\mu + iM_a)/(M_a J))$ becomes $\exp(j2\pi\mu/(M_a J))$ 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 (8), which will guarantee ISI-free detection (RAKE and MMSE equalizers are also possible). For $\boldsymbol{\Theta}_\mu$'s selected as in (a), a blind channel estimation method was developed in [5]. For $\boldsymbol{\Theta}_\mu$'s selected as in (b) and (c), channel identifiability conditions and a more general blind channel estimation algorithm were given in [4].

We remark that our code design is parameterized by M_a, the designed number of active users. Therefore, when the system load changes, it is possible (but not necessary) to redesign the codes and take full advantage of the available bandwidth which is characterized in our system as the users' signature points $\rho_{\mu,j}$.

5. PERFORMANCE ANALYSIS

We extend the theoretical bit error rate (BER) evaluation of [5] to an GMC-CDMA system that has acquired perfect CSI at the receiver and uses a ZF-equalizer $\boldsymbol{\Gamma}_\mu^{zf}$ as in (8) to remove ISI and obtain symbol estimates $\hat{\boldsymbol{s}}_m(n)$. We choose for simplicity a BPSK constellation to obtain in terms of the Q-function an average BER $\bar{P}_e$:

$$\bar{P}e = \frac{1}{MK} \sum_{m=0}^{M_a-1} \sum_{k=0}^{K-1} \mathrm{Q}\Bigg(\sqrt{\frac{1}{\bar{\boldsymbol{g}}_{m,k}^{\mathrm{H}} \bar{\boldsymbol{g}}_{m,k} E_{m,k}}} \sqrt{\frac{2E_b}{N_0}} \Bigg), \tag{9}$$

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

To avoid channel dependent performance, we average (9) over 100 Monte Carlo channel realizations. Fig. 2 depicts (9) for a GMC-CDMA system designed with $K = 8$

and $M = 6$ users (only $M_a = 3$ or 4 are active) each experiencing a Rayleigh fading channel of order $L = 3$, with and without d2) enforced at the transmitter. We observe significant degradation in performance if d2) is not satisfied.

Next, we compare a GMC-CDMA system with $K = 8$, $L = 3$, $M = 16$, and $M_a = 11$ and a DS-CDMA system with spreading $P = 19$ (chosen to make the two systems occupy the same bandwidth) for both RAKE and MMSE receivers. Both systems use WH codes. Fig. 3 shows comparable performance if a RAKE receiver is used whereas GMC-CDMA's performance is one order of magnitude better at 16 dB than DS-CDMA's when a MMSE receiver is used instead of the RAKE.

In Fig. 4, we compare: a GMC-CDMA system using WH with $M = 16$, $K = 8$ and $L = 3$; an AMOUR system at full load; and an MC-CDMA system with $P = 19$ using WH codes and OFDM transmissions. Although AMOUR outperforms competing techniques in full load (not shown in the figure), the GMC-CDMA of Section 4 offers performance improvement over AMOUR when the system exhibits reduced load. Fig. 4 also shows that GMC-CDMA has a lower BER than MC-CDMA for all the simulated cases of active users.

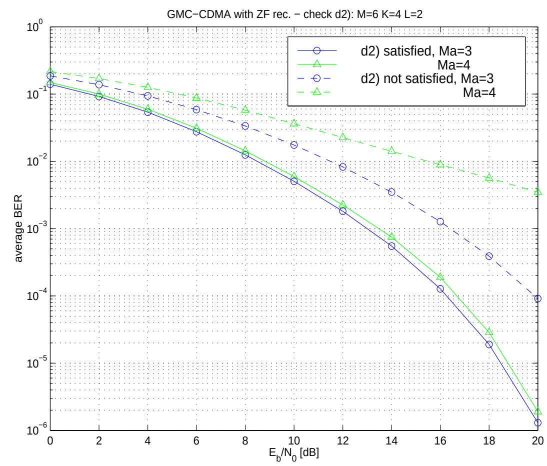

Figure 2: Effect of d2) on GMC-CDMA

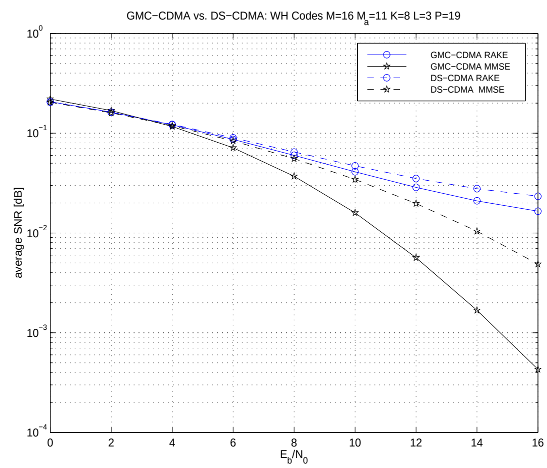

Figure 3: GMC-CDMA vs. DS-CDMA (WH)

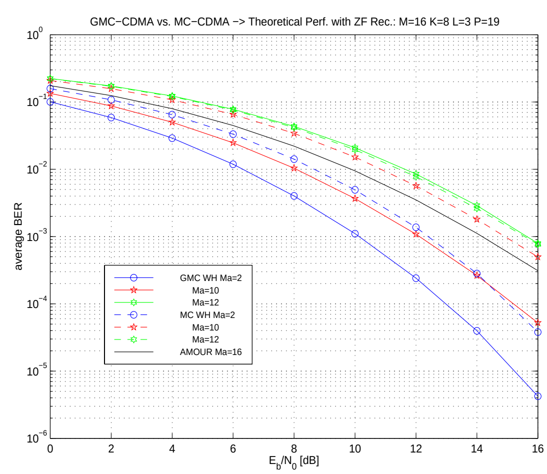

Figure 4: GMC-CDMA vs. MC-CDMA (WH)

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