FREQUENCY-HOPPED GENERALIZED MC-CDMA FOR MULTIPATH AND INTERFERENCE SUPPRESSION†

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ABSTRACT

Generalized multi-carrier (GMC) CDMA transceivers equipped with frequency hopping are developed in this paper. Through judicious code design, multiuser interference (MUI) is eliminated deterministically and symbol recovery is guaranteed, relying on a minimum number of redundant subcarriers. This, together with subcarrier frequency hopping (FH), improves BER performance in the presence of frequency-selective multipath fading channels and enhances resistance to narrow-band interference. Blind channel estimation methods are developed with guaranteed channel identifiability, regardless of the locations of the zeros of the FIR channel. Performance analysis and simulation results illustrate the merits of the proposed FH-GMC-CDMA transceivers relative to competing OFDMA and multicarrier CDMA alternatives.

I. INTRODUCTION

A variant of spread-spectrum (SS) signaling, called frequency-diversity spread spectrum (FD-SS) was recently proposed and shown to be more resistant than direct-sequence SS (DS-SS) to partial-band interference (PBI) [3]. FD-SS, with disjoint frequency support for each subcarrier, is in fact the analog counterpart of the OFDM spread spectrum (OFDM-SS) proposed in [6] and the underlying multicarrier spread spectrum (MC-SS) for MC-CDMA [8] systems with overlapping subcarriers. Assuming digital implementations via DFTs, both OFDM-SS and MC-SS systems can be seen as special cases of a unifying framework [2].

In MC-CDMA, users transmit simultaneously using the entire system bandwidth; in the down-link, user separation is achieved by use of orthogonal spreading codes. However, in the up-link orthogonality is lost due to multipath, resulting in MUI and performance degradation as the system load increases. In [5], FH-OFDMA was proposed for uplink CATV transmission, and was shown to achieve MUI elimination. However, it suffers from frequency-selective fading and requires extra diversity to ameliorate the effects of channel nulls.

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In this paper, we generalize MC-CDMA and FH-OFDMA, and propose a generalized MC-CDMA (GMC-CDMA) system. By judicious transceiver design, MUI is eliminated deterministically. Use of redundant subcarriers guarantees recovery of the symbols, regardless of the locations of the zeros of the multipath channel. Further, GMC-CDMA permits blind estimation of the channel, without imposing any restrictions on the zeros. Resistance to both narrow-band interference (NBI) and PBI is guaranteed.

Section II provides a general model, with Section III specializing it to existing MC-CDMA and FH-OFDMA. Frequency-hopped GMC-CDMA transceivers are developed in Section IV and simulation results are provided in Section V.

II. DISCRETE TIME SYSTEM MODEL

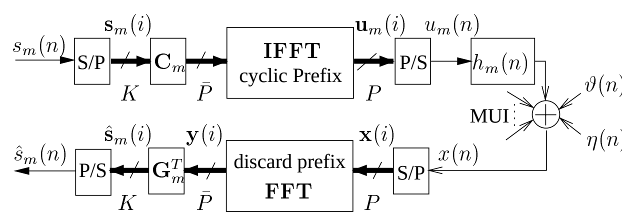

Fig. 1. Discrete time system model

The baseband discrete-time equivalent transmitter and receiver model for the mth user is depicted in Fig. 1, where $m \in [0, M-1]$ and M is the number of users. The information stream $s_m(n)$, at symbol rate $1/T_s$, is first parsed into K-long blocks, $\mathbf{s}_m(i) := [s(iK), s(iK+1), \ldots, s(iK+K-1)]^T$, spread by the $\bar{P} \times K$ code matrix $\mathbf{C}_m$, and then IFFT processed to obtain the $\bar{P} \times 1$ vector $\mathbf{F}^{\mathcal{H}}\mathbf{C}_m\mathbf{s}_m(i)$. Here, $\bar{P}$ is the block spreading length, and $\mathbf{F}$ is the $\bar{P} \times \bar{P}$ FFT matrix with (m, n) entry $(1/\sqrt{\bar{P}})\exp(-j2\pi mn/\bar{P})$.

To avoid channel-induced inter symbol (block) interference, we use a cyclic prefix (CP) (as in conventional OFDM systems, e.g., [1]). The received signal, after conversion to baseband and receive filtering, is sampled at the chip rate, to yield

$$x(n) = \sum_{m=0}^{M-1} \sum_{l=0}^{L} h_m(l) u_m(n-l) + \vartheta(n) + \eta(n), \quad (1)$$

where $h_m(\ell)$ is the overall channel (transmit and receive filters, and propagation channel) encountered by the mth user's signal, $\vartheta(n)$ is the filtered PBI, $\eta(n)$ is the filtered additive Gaussian noise (AGN), and L is the maximum order of FIR channels for all users, i.e., $h_m(l) = 0, \forall l \notin [0, L], \forall m$. Additive noise $\eta(t)$ is assumed to have flat power spectral density $N_0/2$ over the system bandwidth B. PBI $\vartheta(t)$ is assumed to occupy εB bandwidth ($0 < \varepsilon < 1$) with large power density $\mathcal{I}_0/2$.

In order to avoid inter-block interference (IBI), we assume that the prefix length is longer than the maximum channel order for all users:

(a0) $P - \bar{P} \geq L$.

To cast (1) into a convenient matrix-vector form, we define the $P \times 1$ vector $\mathbf{x}_m(i) := [x_m(iP), x_m(iP+1), \ldots, x_m(iP+P-1)]^T$, and similarly $\boldsymbol{\vartheta}(i)$ and $\boldsymbol{\eta}(i)$; define the $P \times P$ Toeplitz channel matrices $\mathbf{H}_m, \mathbf{H}_m^{(1)}$ with (k,l)th entries $h_m(k-l)$ and $h_m(k-l+P)$, respectively. Because of (a0), we can write (1) as

$$\mathbf{x}(i) = \textstyle\sum_{m=0}^{M-1} \mathbf{H}_m \mathbf{u}_m(i) + \mathbf{H}_m^{(1)} \mathbf{u}_m(i-1) + \boldsymbol{\vartheta}(i) + \boldsymbol{\eta}(i),$$

where the second term represents IBI. For convenience, we define $\mathbf{w}(i) := \boldsymbol{\vartheta}(i) + \boldsymbol{\eta}(i)$ as the equivalent (colored) additive noise vector.

At the receiver, the CP is removed by dropping the first $P - \bar{P}$ elements of $\mathbf{x}(i)$, thus eliminating IBI. After FFT processing, we have

$$\mathbf{y}(i) = \sum_{m=0}^{M-1} \mathbf{F}\tilde{\mathbf{H}}_m \mathbf{F}^{\mathcal{H}} \mathbf{C}_m \mathbf{s}_m(i) + \tilde{\mathbf{w}}(i), \qquad (2)$$

where $\tilde{\mathbf{w}}(i)$ is the filtered interference and noise vector, and $\tilde{\mathbf{H}}_m$ is the resulting channel matrix. $\tilde{\mathbf{H}}_m$ is a $\bar{P} \times \bar{P}$ circulant matrix with its (k,l)th entry given by $h_m((k-l) \bmod \bar{P})$ (see also [7]). Since $\tilde{\mathbf{H}}_m$ is circulant, $\tilde{\mathbf{D}}_m(H) := \mathbf{F}\tilde{\mathbf{H}}_m\mathbf{F}$ is a diagonal matrix; the diagonal elements, $H_m(\rho_p)$, are values of the channel frequency response $H_m(z) = \sum_{l=0}^{L} h_m(l) z^{-l}$ evaluated at $z = \rho_p = \exp(j2\pi p/\bar{P})$, $p = 0, 1, \ldots, \bar{P}-1$. Therefore, we can rewrite (2) as

$$\mathbf{y}(i) = \sum_{m=0}^{M-1} \tilde{\mathbf{D}}_m(H) \mathbf{C}_m \mathbf{s}_m(i) + \tilde{\mathbf{w}}(i). \qquad (3)$$

Subsequently, for the mth user, multiplication by the equalization matrix $\mathbf{G}_m$ yields symbol block estimates $\hat{\mathbf{s}}_m(i)$ as

$$\hat{\mathbf{s}}_m(i) = \mathbf{G}_m \mathbf{y}(i). \qquad (4)$$

Instead of adding a CP to avoid IBI, we can add trailing zeros as in [7].

Equation (3) generalizes both conventional MC-CDMA and coherent FH-OFDMA.

III. SPECIAL CASES

In this section, we discuss two special cases of our general model obtained by setting $K = 1$ (no symbol blocking): MC-CDMA and FH-OFDMA.

A. MC-CDMA

The block spreading matrix $\mathbf{C}_m$ reduces to a spreading vector $\mathbf{c}_m$, and the receiver matrix $\mathbf{G}_m$ reduces to a vector $\mathbf{g}_m^T$. Therefore, (3) reduces to

$$\mathbf{y}(i) = \tilde{\mathbf{D}}_m(H)\mathbf{c}_m s(i) + \sum_{\mu=0,\mu\neq m}^{M-1} \tilde{\mathbf{D}}_\mu(H)\mathbf{c}_\mu s_\mu(i) + \mathbf{w}(i).$$

To gain insight into how MC-CDMA copes with frequency-selective channels, we focus on the single user case, i.e., $M = 1$, and assume that $\mathbf{w}(i)$ is white (no MUI and no PBI). The Maximum Ratio Combiner is known to achieve the maximum output SNR:

$$\mathbf{G}_m = \mathbf{g}_m^T = \mathbf{c}_m^{\mathcal{H}} \mathbf{D}_m^{\mathcal{H}}(H) = \tilde{\mathbf{h}}^{\mathcal{H}} \mathbf{D}_m^{\mathcal{H}}(\mathbf{c}) \ , \qquad (5)$$

where $\mathbf{D}_m(\mathbf{c}) = \text{diag}(c_m(0), c_m(1), \ldots, c_m(\bar{P}-1))$, $\tilde{\mathbf{h}}_m = [H_m(\rho_0), \ldots, H_m(\rho_{\bar{P}-1})]^T$. Define the Vandermonde matrix $\mathbf{V}$ with $(l+1, k+1)$th entry $[\mathbf{V}]_{l+1,k+1} = \exp(-j2\pi lk/\bar{P})$; hence, $\tilde{\mathbf{h}}_m = \mathbf{V}\mathbf{h}_m$, and the output SNR is given by

$$SNR = \mathbf{h}_m^{\mathcal{H}} \mathbf{V}^{\mathcal{H}} \mathbf{D}_m^{\mathcal{H}}(\mathbf{c}) \mathbf{D}_m(\mathbf{c}) \mathbf{V} \mathbf{h}_m \sigma_s^2/\sigma_v^2 = \mathbf{h}_m^{\mathcal{H}} \mathbf{h}_m \sigma_s^2/\sigma_v^2,$$

which reveals that frequency diversity due to transmitting replicas through $\bar{P}$ subcarriers comes only from the $L+1$ channel taps. With $\bar{P} > L+1$, the vector $\tilde{\mathbf{h}}_m$ is correlated and it does not provide more diversity than the original $(L+1)$ channel taps. More explicitly, for single user MC-SS, we do not need to exploit all the $\bar{P}$ subcarriers. It suffices to choose $\bar{P}_1 \geq L+1$; if $\bar{P}/\bar{P}_1$ is an integer, we select $\bar{P}_1$ equispaced subcarriers out of the $\bar{P}$ subcarriers; vector $\mathbf{c}_m$ will have $\bar{P}_1$ non-zero entries with amplitude $1/\sqrt{\bar{P}_1}$. We verify that $\mathbf{h}_m^{\mathcal{H}} \mathbf{V}^{\mathcal{H}} \mathbf{D}_m(\mathbf{c}) \mathbf{D}_m(\mathbf{c}) \mathbf{V} \mathbf{h}_m = \mathbf{h}_m^{\mathcal{H}} \mathbf{h}_m$, which indicates that choosing $L < \bar{P}_1 < \bar{P}$ subcarriers results in the same performance as choosing $\bar{P}$ subcarriers. Since an $L+1$ tap FIR channel can have a maximum of L nulls, we note that the symbol recovery is not guaranteed if we choose fewer than $L+1$ subcarriers. We summarize this discussion below:

Result 1: *If the FIR channel has order L, and the additive noise is white, full diversity gain is achieved by MC-SS by using more than L equispaced subcarriers. Using fewer subcarriers leads to loss of diversity gain; using more subcarriers does not improve performance.*

Thus, MC-SS exploits the full frequency diversity of the multipath channel, as long as $\bar{P} \geq L+1$. However, for multiple access, each user utilizes all available subcarriers and MUI occurs as soon as $M \geq 2$. The separation between users relies on having a distinct $\bar{P} \times 1$ spreading code $\mathbf{c}_m$ (e.g., Walsh-Hadamard code) for each user. Then choosing $\bar{P} \gg L+1$ serves to suppress the MUI by decreasing the MUI level or possible NBI (colored noise).

Because of MUI, MC-CDMA suffers from the near-far problem and thus degraded performance in comparison with the single-user case. This motivates our MUI-resilient GMC-CDMA system in Section IV.

B. FH-OFDMA

For OFDMA, we set $K = 1$; $\mathbf{c}_m = \mathbf{e}_m$ is the mth Euclidean basis vector and has only one non-zero element. Thus, $\mathbf{g}_m = \mathbf{e}_m$ and the equivalent model is

$$y_m(i) := \mathbf{g}_m^T \mathbf{y}(i) = H(\rho_m) s_m(i) + w_m(i) \;, \qquad (6)$$

which reveals that MUI is eliminated.

However, FH-OFDMA suffers from frequency-selective fading. Specifically, when the channel has zeros close to ρ_m, the symbol will suffer from severe fading. To avoid consistent fading, frequency hopping is proposed (coding or fast hopping is beyond the scope of this paper). Specifically, $\mathbf{c}_m$ is changed frequently (and thus depends upon a time index i) according to some prescribed hopping pattern. Multiple users are allowed to transmit information simultaneously using different hopping sequences, and MUI can be avoided if no frequencies are employed by two users at the same time. For example, we can set $\mathbf{c}_m(i) = \mathbf{e}_{(m+i) \bmod \bar{P}}$.

However, symbol recovery is not guaranteed and frequency diversity is not exploited in FH-OFDMA because it uses only a single subcarrier. We saw in Result 1 that $L+1$ subcarriers are needed to fully exploit the frequency diversity of the order L FIR channel.

GMC-CDMA system considered next overcomes the limitations due to MUI, NBI, PBI, and multipath effects.

IV. GMC-CDMA TRANSCEIVER DESIGN

We will transmit K symbols per block, using $\bar{P} \le P - L$ subcarriers such that $J := \bar{P}/M$ is an integer. Thus, each of the M users can be assigned a distinct set of J subcarriers. We denote by $\rho_{m,q}$, $0 \le q \le J-1$, the J distinct subcarriers assigned to the m^{th} user.

To unravel the attractive features of MUI resilience progressively, we factor our spreading and despreading matrices $\{\mathbf{C}_m, \mathbf{G}_m\}_{m=0}^{M-1}$ in the following forms:

$$\mathbf{C}_m = \mathbf{\Phi}_m \mathbf{\Theta}_m, \quad \mathbf{G}_m = \mathbf{\Gamma}_m \mathbf{\Phi}_m^T \;, \qquad (7)$$

with each matrix factor playing a different role: $\mathbf{\Theta}_m$ is a $J \times K$ matrix that linearly maps the K information symbols of the ith block $\mathbf{s}_m(i)$ to J ($J > K$) symbols $\mathbf{\Theta}_m \mathbf{s}_m(i)$; these in turn are mapped to the user's J signature subcarriers via the $\bar{P} \times J$ selector matrix $\mathbf{\Phi}_m$, to yield $\mathbf{\Phi}_m \mathbf{\Theta}_m \mathbf{s}_m(i)$. We have $[\mathbf{\Phi}_m]_{p+1,q+1} = 1$ if $\rho_{m,q} = \exp(j2\pi p/\bar{P})$, and 0 otherwise; note that we will have J non-zero entries. Per Result 1, we would like to choose equispaced subcarriers for each user such that $[\mathbf{\Phi}_m]_{p+1,q+1} = 1$ if $p = m + qM, q = 0, 1, \ldots, \bar{P}-1$. Thanks to the non-overlapping frequency allocation, the corresponding subcarrier selector matrices $\mathbf{\Phi}_m$ are mutually orthogonal by construction. Since $\mathbf{\Phi}_m$ has a single non-zero (unity) entry per column, it can be readily verified that $\tilde{\mathbf{D}}_m(H)\mathbf{\Phi}_m = \mathbf{\Phi}_m \mathbf{D}_m(H)$, where $\mathbf{D}_m(H) := \text{diag}[H_m(\rho_{m,0}, \ldots, H_m(\rho_{m,J-1})]$. This fact, together with the orthogonality of $\mathbf{\Phi}$, allows us to simplify (4) for user m to [7]:

$$\hat{\mathbf{s}}_m(i) = \mathbf{\Gamma}_m \mathbf{\Phi}_m^T \sum_{\mu=0}^{M-1} \mathbf{\Phi}_\mu \mathbf{D}_\mu(H) \mathbf{\Theta}_\mu \mathbf{s}_\mu(i) + \mathbf{\Gamma}_m \mathbf{\Phi}_m^T \tilde{\mathbf{w}}(i)$$

$$= \mathbf{\Gamma}_m [\mathbf{D}_m(H) \mathbf{\Theta}_m \mathbf{s}_m(i) + \mathbf{\Phi}_m^T \tilde{\mathbf{w}}(i)] := \mathbf{\Gamma}_m \tilde{\mathbf{y}}_m(i) \qquad (8)$$

where $\tilde{\mathbf{y}}_m(i)$ is the $J \times 1$ MUI-free vector corresponding to user m. Equation (8) reveals that MUI is eliminated deterministically regardless of the multipath channels.

To guarantee recovery of the K symbols in $\mathbf{s}_m(i)$ regardless of the signal constellation, the matrix $\mathbf{D}_m(H)\mathbf{\Theta}_m$ in (8) must have full rank, regardless of the mth channel. Hence, we require $\text{rank}(\mathbf{D}_m(H)\mathbf{\Theta}_m) = K$, $\forall m \in [0, M-1]$, so that zero forcing (ZF) equalization based on $\mathbf{\Gamma}_m = (\mathbf{D}_m(H)\mathbf{\Theta}_m)^\dagger$ will recover $\hat{\mathbf{s}}_m(i)$. If the noise covariance matrix $\mathbf{R}_{\tilde{w}\tilde{w}}$ is known, we can apply the MMSE receiver $\mathbf{\Gamma}_m = \mathbf{\Theta}_m^{\mathcal{H}} \mathbf{D}_m^{\mathcal{H}}(H)[\mathbf{R}_{\tilde{w}\tilde{w}} + \mathbf{D}_m(H)\mathbf{\Theta}_m \mathbf{\Theta}_m^{\mathcal{H}} \mathbf{D}_m^{\mathcal{H}}(H)]^{-1}$. For small K, the ML solution $\hat{\mathbf{s}}_m(i) = \text{argmin}_{\mathbf{s}_m(i)} \|\mathbf{y}_m(i) - \mathbf{D}_m(H)\mathbf{\Theta}_m \mathbf{s}_m(i)\|^2$ is affordable.

Since each user's channel can have at most L zeros, $\mathbf{D}_m(H)$ can have at most L zero diagonal entries. The above rank condition will be satisfied if any $J - L = K$ rows of $\mathbf{\Theta}_m$ are linearly independent. To meet this rank requirement, each column of $\mathbf{\Theta}_m$ should have at least $L+1$ non-zero entries, i.e., each symbol is transmitted over $L+1$ or more subcarriers, so that the frequency diversity of the multipath channel is fully exploited.

Notice that this rank condition is not a condition on the channels; instead, it is a guideline for designing $\mathbf{\Theta}_m$. For example, we can choose $\mathbf{\Theta}_m$ with entries:

$$[\mathbf{\Theta}_m]_{l+1,k+1} = (1/\sqrt{J})\rho_{m,l}^{-k}. \qquad (9)$$

Since the signature frequencies are distinct, any K rows of the matrix will yield a full rank matrix.

How do we choose the symbol block size K? If $N := T_s/T_c$ denotes the spreading gain, then we need

$$J = \frac{KN - L}{M} \ge K + L \Longrightarrow K \ge \frac{L(M+1)}{N - M} \qquad (10)$$

to achieve MUI elimination and guarantee symbol recovery in the presence of unknown multipath. Recall that M is the number of users, N the spreading gain, L the FIR channel length, and J the number of subcarriers ($P = KN$ here).

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Increasing K increases the value of $J - K = [K(N - M) - L]/M$, which indicates more freedom in choosing the redundant mapping matrix $\mathbf{\Theta}_m$.

We illustrate the advantages of GMC-CDMA over MC-CDMA, via a simple example.

Example 1: Suppose we design an under-loaded system with spreading gain $N = 67$ for $M = 16$ users, and channel order $L = 3$. The system load is approximately $16/67 \approx 25\%$. Even in AWGN channel, MC-CDMA will exhibit BER larger than $Q(\sqrt{1/.25})$ due to MUI [4]. For GMC-CDMA, we have $K \geq 1$ from (10). If we choose the minimum $K = 1$, so that $J = 4 \geq L + 1$, then MUI is eliminated; further, each user fully exploits the frequency diversity of the channel (see Result 1), and obtains the best performance that can be achieved by MC-SS with a single user. Therefore, GMC-CDMA considerably outperforms MC-CDMA in this setting.

As the system load increases, $N - M$ decreases and $K > 1$ is needed [c.f. (10)]. As K increases, the redundancy $J - K$ increases, and we have more flexibility in our design. For example, with $J = K + 2L$, instead of using the linear precoder $\mathbf{\Theta}_m$, we can use a nonlinear precoder (e.g., a block channel encoder) to correct L faded symbols. We summarize our observations as follows:

Result 2: *Since the block size K can be varied, GMC-CDMA can improve its performance depending on system load. When the system load is $M \leq \bar{P}/(L+1)$, by setting $K = 1$ and $J = \lfloor \bar{P}/M \rfloor$, each GMC-CDMA user achieves performance equivalent to a single-user MC-SS system. When M increases, K and J must increase according to* (10) *in order to maintain MUI elimination and symbol recovery guarantees. Increasing K increases $J - K$, which in turn allows channel encoding (over the GF) to be used instead of (or joint with) the $\mathbf{\Theta}_m$-spreading (over the complex field).*

A. Frequency-hopping capability

To randomize channel effects and avoid consistent fading on subcarriers, each user can be assigned different subcarriers at different time intervals. Thus, our GMC-CDMA system has frequency-hopping capabilities. In such a system, the frequency selector matrix, $\mathbf{\Phi}_m(i)$, varies with the block (time) index i. We consider a simple example for illustrative purposes.

Example 2: The $(p+1, q+1)$ entries of $\mathbf{\Phi}_m(i)$ are nonzero, and equal to unity, if

$$p = (m + i + qM) \bmod \bar{P}, \ \forall q \in [0, J-1]. \qquad (11)$$

Hopping the frequency selector matrix $\mathbf{\Phi}_m$ prevents a user from being consistently hit by channel nulls; by varying the precoding matrix $\mathbf{\Theta}_m$ from block symbol to symbol, we increase system security by providing additional coding.

B. NBI suppression

Assume that the PBI has bandwidth $\epsilon B = \epsilon N/T_s$. If we design our system as in (11), each subcarrier has bandwidth $1/(\bar{P}T_c)$, and the distance between two consecutive subcarriers is $\bar{P}/J/(\bar{P}T_c) = N/(JT_s)$. So the number of subcarriers that will be hit by PBI is $N_I = [\epsilon N/T_s]/[N/(JT_s)] = \epsilon J$. We can treat the subcarriers hit by the strong PBI as having encountered a deep fade; thus symbol recovery is guaranteed if $J \geq K + L + N_I$, regardless of the power of the PBI. GMC-CDMA gains resistance to PBI jamming, by narrowing the bandwidth of each subcarrier (from $1/T_s$ to $N/(\bar{P}T_s)$) and increasing the spacing between the subcarriers; for a jammer to be effective, it must know the user's subcarrier set. In the case of NBI (e.g., single tone jammer), at most one of the J ($J \gg 1$) subcarriers will be hit, and the symbols can be recovered from the other $J - 1$ interference-free subcarriers.

C. Blind channel estimation

GMC-CDMA system design facilitates blind channel estimation at the receiver, obviating the need for bandwidth-consuming training sequences. Especially if we select the mapping matrix $\mathbf{\Theta}_m$ to satisfy $J \geq K + L$, and any K rows of $\mathbf{\Theta}_m$ span the $\mathcal{C}^K$ row vector space, e.g., as in (9), then we can use the subspace-based blind channel estimation algorithm of [2]. If we choose $\mathbf{\Theta}_m := [\mathbf{I}_K, \mathbf{V}^T_{K\times(J-K)}]^T$, where $\mathbf{V}_{(J-K)\times K}$ is the Vandermonde matrix constructed through $J - K$ distinct points other than $\{\exp(j2\pi k/K)\}_{k=1}^K$, then any K rows of $\mathbf{\Theta}_m$ are linearly independent, and we can estimate the channel blindly along the lines of the finite alphabet (FA) based approach of [9].

V. SIMULATIONS

Because of space limitations, we do not present the results of our theoretical performance analysis; instead, we show several numerical results.

Test Case 1 (multipath effects): Here, $N = 18$ and $L = 2$ (3-ray channels), so that the maximum number of users in FH-OFDMA is $N - L = 16$. We compare MC-CDMA, FH-OFDMA, and GMC-CDMA under four different loads: $M = 2, 4, 8, 16$ (12.5% to 75%). Each MC-CDMA user picks up one Walsh-Hadamard sequence of length 16. For GMC-CDMA, we use the minimum K according to (10), so that the (M, J, K) triplets are $(2, 8, 1)$, $(4, 4, 1)$, $(8, 5, 2)$, $(12, 7, 5)$. Fig. 2 shows that GMC-CDMA outperforms MC-CDMA and FH-OFDMA considerably. As long as the number of users $M \leq 5$ and $K = 1$, each user of GMC-CDMA achieves the best performance that MC-CDMA can achieve in a single-user setting. Because of MUI, MC-CDMA suffers from near-far effects, and the BER curve levels off at high SNR (the floor effect). With MUI free reception, FH-OFDMA has constant performance regardless of system load. However, it suffers from deep channel fading due to the lack of frequency diversity.

Test Case 2 (PBI suppression): Assume that MC-CDMA is on the down-link, where all users experience the same channel so that MUI can be eliminated by the orthogonality of spreading codes after channel equalization. Here, we set $N = 18, L = 2, M = 8(50\% \text{ load})$. For GMC-CDMA, we set $K = 4, J = 8$, so that $J - K - L = 2$, indicating that we can afford to have at most two subcarriers hit by the PBI. In Fig. 3, we show the performance of the two systems as the relative bandwidth of the strong PBI ($\mathcal{I}_0 \gg N_0$) is varied: $\varepsilon = 0/16$ (no interference), $2/16,\ 4/16,\ 6/16,\ 8/16$. For GMC-CDMA, $N_I = 0,\ 1,\ 2,\ 3,\ 4$ subcarriers are hit by the strong PBI. For both MC-CDMA and GMC-CDMA, we assume that the receiver can detect the presence of strong interference and remove the contaminated subcarriers. As shown in Fig. 3, GMC-CDMA outperforms MC-CDMA when $\epsilon \leq 6/16$. When the PBI occupies half the system bandwidth, GMC-CDMA becomes worse because now $J - K < L + N_I$, so that symbol recovery is no longer guaranteed. However, even when $\epsilon = 1/2$ ($J/2$ subcarriers will be hit by the PBI), we can increase K, J so that symbol recovery can be guaranteed, e.g., with the pair $(K, J) = (17, 38)$.

Test Case 3 (Blind channel estimation): The test system has spreading gain $N = 16$, $M = 12$ users, and channel order $L = 2$. We design GMC-CDMA with $J = 10$, $K = 8$, $\mathbf{\Theta}_m = [\mathbf{I}_K, \mathbf{V}^T_{K \times J - K}]^T$, where the last two rows are Vandermonde vectors constructed from the roots $\exp(j2\pi/2K)$ and $\exp(j2\pi(K+1)/2K)$. BPSK signals are used, and the channel is estimated from 20 symbol blocks, using the subspace-based and FA-based blind channel estimators. Fig. 4 shows the BER curves corresponding to these estimators; we note that the performance is close to the ideal BER, corresponding to known channel.

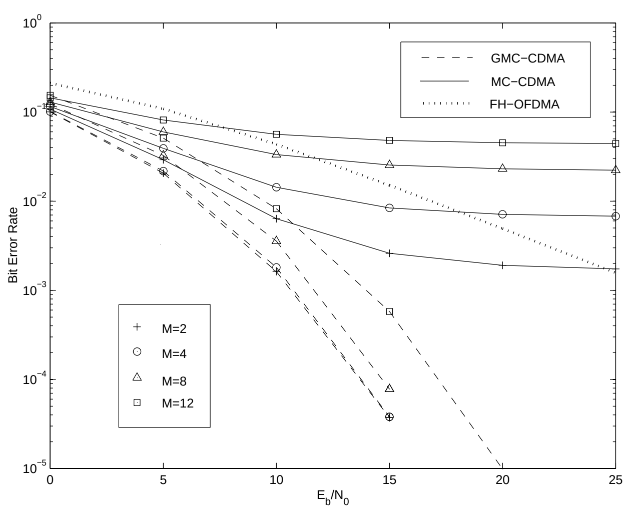

Fig. 2. Comparisons under multipath channels

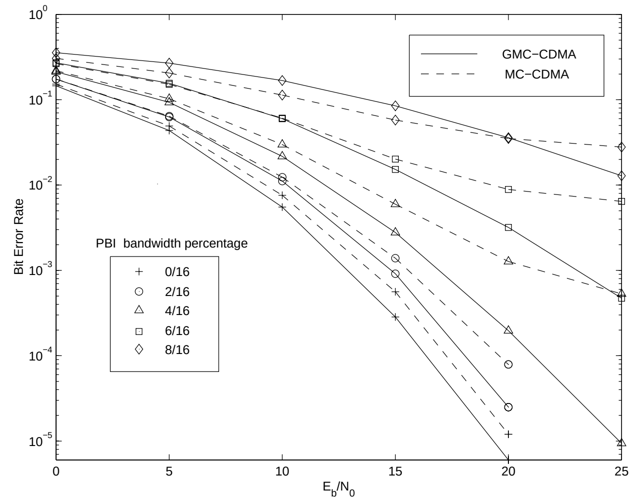

Fig. 3. Comparisons under partial band interference

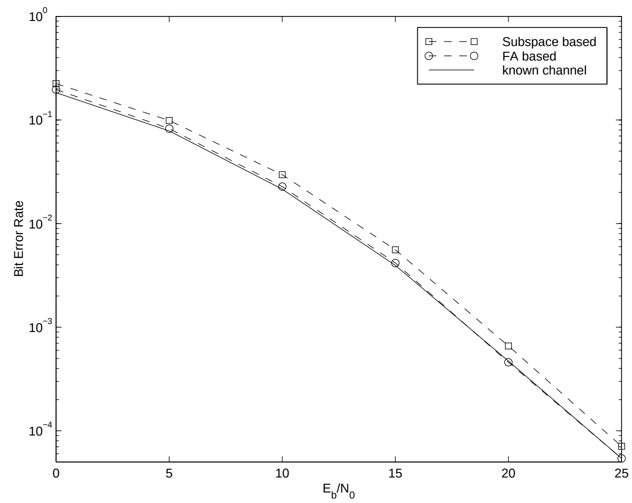

Fig. 4. Blind channel estimation

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