Chip-Interleaved Block-Spread Code Division Multiple Access

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Abstract — **A novel MUI-free CDMA transceiver for frequency-selective multipath channels is developed in this paper. Relying on chip-interleaving and zero padded transmissions, mutual orthogonality between different users' spreading codes is maintained at the receiver even after frequency selective propagation, which leads to deterministic multiuser separation with low-complexity code-matched filtering without loss of maximum likelihood optimality. In addition to MUI-free reception, the proposed system guarantees channel-irrespective symbol recovery and achieves high bandwidth efficiency by increasing the symbol block size. Non-redundant precoded alternatives and (semi-) blind channel estimation algorithms are also discussed. Simulation results demonstrate improved performance of the proposed system relative to competing alternatives.**

I. Introduction

Relying on orthogonal spreading codes, Code Division Multiple Access (CDMA) systems allow simultaneous transmissions from multiple users over the same bandwidth. However, when the chip rate increases, the underlying multipath channel becomes frequency selective; it introduces inter chip interference (ICI) and thus destroys code orthogonality at the receiver. The latter gives rise to Multiuser Interference (MUI). To suppress MUI, various multiuser detectors are available [9], e.g., the linear decorrelator or Zero Forcing (ZF), the Minimum Mean Square Error (MMSE), as well as the nonlinear Decision Feedback (DF) and Maximum Likelihood (ML) receivers. However, these schemes require knowledge of the multipath channels for all users and/or suppress MUI statistically (except for the ZF option) even with exact Channel State Information (CSI). In addition to increased complexity that comes with multichannel estimation and multiuser detection, there even exist frequency-selective channels preventing symbol detection no matter what receiver is used [10].

To remove MUI deterministically *regardless* of the underlying multipath channels, several alternatives have been proposed recently. Those include the Orthogonal Frequency Division Multiple Access (OFDMA) [7], where complex exponentials are utilized as information-bearing subcarriers and thus retain their shape and orthogonality when passing through multipath channels. However, when the channels have nulls (or deep fades) on some subcarriers, the information symbols on those subcarriers will be lost. Therefore, OFDMA-like transceivers require extra diversity (such as frequency hopping or channel coding) to ameliorate fading effects. To guarantee channel-irrespective MUI-free reception and symbol recovery, A Mutually-Orthogonal Usercode-Receiver (AMOUR) system was proposed in [3]. However, AMOUR transmissions are not Constant-Modulus (C-M) in general, and its codes are generally complex valued. To maintain C-M at the transmitter and facilitate low-complexity receivers for MUI-free reception, the so-called shift orthogonal codes (which are not only orthogonal to each other but also to their shifted versions) were proposed in [4, 5]. However, to maintain shift orthogonality, a 50% bandwidth efficiency penalty is paid for both the real and the complex codes in [4, 5].

In this paper, we develop novel MUI-free CDMA transceivers based on chip-interleaving. Thanks to chip-interleaving and zero padding at the transmitter, mutual orthogonality between different users' codes is preserved even after multipath propagation, which allows for deterministic multiuser separation through low-complexity code-matched filtering without loss of maximum likelihood optimality; thus the multiuser detection has been successfully converted to a set of equivalent single user detection problems (Section III). In addition to MUI-free reception, channel-irrespective symbol recovery is also guaranteed. By increasing the symbol block size, the proposed system achieves high bandwidth efficiency. Thanks to MUI-free reception, non-redundant precoding alternatives can be developed for each single user. Transforming the Multiple-Input Multiple-Output (MIMO) channels to parallel Single-Input Single-Output (SISO) channels, it is also shown that the receiver can employ various training-based or (semi-) blind channel estimators developed for single user systems. Simulations are then performed in Section IV to collaborate the improved performance of the proposed system.

II. System Modeling

The block diagram in Fig. 1 describes a CDMA system model in the downlink or uplink scenario, where only one user (the mth user out of a maximum M users) is shown. Unlike traditional spreading which is performed over a *single symbol*, we here use block spreading that operates on a *block of symbols*; in other words, the information stream of the mth user $s_m(n)$ is first parsed into K-long blocks $\mathbf{s}_m(i) := [s(iK), s(iK+1), \ldots, s(iK+K-1)]^T$ and then block spread by a $P \times K$ spreading matrix $\mathbf{C}_m$ to obtain the $P \times 1$ output vector $\mathbf{u}_m(i) := \mathbf{C}_m \mathbf{s}_m(i)$. After Parallel to Serial (P/S) conversion, the mth user's coded chip sequence $u_m(n)$ is pulse shaped to the corresponding continuous time signal $u_m(t)$. The latter propagates through a (possibly *unknown*) channel $h_m(t)$ and is filtered by the receive filter. Let $h_m(l)$ (with order L_m) denote the equivalent discrete time channel impulse response that includes the mth user's asynchronism in the form of delay factors as well as transmit-receive filters and multipath effects, and $\eta(n)$ be the sampled noise. The received

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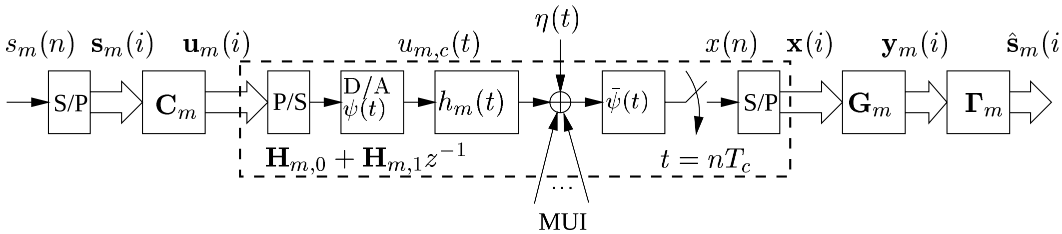

Figure 1: Continuous and discrete-time equivalent baseband system model (only mth user shown)

signal from all users at the chip rate can then be written as

$$x(n) = \sum_{m=0}^{M-1} \sum_{l=0}^{L_m} h_m(l) u_m(n-l) + \eta(n). \qquad (1)$$

Similar to [3, 10, 4], we here focus on a quasi synchronous (QS) system, where mobile users attempt to synchronize with the base-station's pilot waveform. Thus, the maximum channel order $L := \max_m L_m$ for all users can be considered small (relative to the block size K) provided that we select K and the transmitted block length $P \gg L$.

The received samples $x(n)$ are serial to parallel converted to form the $P \times 1$ vector: $\mathbf{x}(i) := [x(iP), x(iP+1), \ldots, x(iP+P-1)]^T$ (similar to the noise vector $\boldsymbol{\eta}(i)$). Let $\mathbf{H}_{m,0}$ be the $P \times P$ lower triangular Toeplitz matrix with first column $[h_m(0), \ldots, h_m(L), 0, \ldots, 0]^T$, and $\mathbf{H}_{m,1}$ be the $P \times P$ upper triangular Toeplitz matrix with first row $[0, \ldots, 0, h_m(L), \ldots, h_m(1)]$. The Input-Output block relationship of the channels can be described by [10]:

$$\mathbf{x}(i) = \sum_{m=0}^{M-1} [\mathbf{H}_{m,0}\mathbf{C}_m \mathbf{s}_m(i) + \mathbf{H}_{m,1}\mathbf{C}_m \mathbf{s}_m(i-1)] + \boldsymbol{\eta}(i), \qquad (2)$$

where the second term in the sum accounts for the Inter Block Interference (IBI).

To extract the μ-th user from $\mathbf{x}(i)$, a linear multiuser separating front-end for user μ, denoted by the matrix $\mathbf{G}_\mu$, is applied to $\mathbf{x}(i)$ to suppress MUI; the MUI-free output is then equalized by a linear single user equalizer Γ_μ as follows:

$$\mathbf{y}_\mu(i) = \mathbf{G}_\mu \mathbf{x}(i), \qquad \hat{\mathbf{s}}_\mu(i) = \Gamma_\mu \mathbf{y}_\mu(i). \qquad (3)$$

In the next section, we will design the transceiver pairs $\{\mathbf{C}_m, \mathbf{G}_m\}_{m=0}^{M-1}$, to separate the superposition of multiuser signals deterministically regardless of multipath propagation of (quasi-) synchronous transmissions through FIR channels of maximum order L.

III. Transceiver design

For each of the M users, we assign a distinct $M \times 1$ vector $\mathbf{c}_m := [c_m(0), \ldots, c_m(M-1)]^T$. These vectors are selected to be mutually orthogonal which can be formally stated as a design constraint ($\delta(\cdot)$ stands for Kronecker's delta):

d1) select code vectors: $\mathbf{c}_\mu^{\mathcal{H}} \mathbf{c}_m = \delta(m-\mu)$, $\forall m, \mu \in [0, M-1]$.

Let $\otimes$ denote the Kronecker's product, $\mathbf{I}_{K+L}$ stand for the $(K+L) \times (K+L)$ identity matrix and define the $(K+L) \times K$ zero-padding (ZP) matrix $\mathbf{T}_{zp} := [\mathbf{I}_K, \mathbf{0}_{K \times L}]^T$. With these notations, we design our $P \times K$ user code matrices $\mathbf{C}_m$ and correspondingly the $(K+L) \times P$ separating matrices $\mathbf{G}_m$ as:

$$\mathbf{C}_m = \mathbf{c}_m \otimes \mathbf{T}_{zp}, \qquad \mathbf{G}_m = \mathbf{c}_m^{\mathcal{H}} \otimes \mathbf{I}_{K+L}, \qquad (4)$$

where $(\cdot)^{\mathcal{H}}$ stands for conjugate transpose. Using the Kronecker product definition each block-spreading code (column of $\mathbf{C}_m$) turns out to have length $P = M(K+L)$.

Our block-spread transmission $\mathbf{u}_m(i) = \mathbf{C}_m \mathbf{s}_m(i)$ with the $\mathbf{C}_m$ defined in (4) can be viewed also as a symbol-spread transmission followed by a chip interleaver with guards [1] as shown in Fig. 2. DS-CDMA corresponds to transmitting the interleaver entries row-wise, while our proposed transmitter outputs the interleaver entries column-wise. This not only explains the acronym Chip-Interleaved Block-Spread CDMA (CIBS-CDMA), but also highlights how readily implementable (and backward-compatible) the proposed system is by simply cascading to an existing DS-CDMA system a chip interleaver with guards (see also Fig. 3). Note that if $\mathbf{C}_m$ in (4) is replaced by $\mathbf{C}'_m = \mathbf{c}_m \otimes \mathbf{I}_K$, the block spreading operation corresponds to the chip-interleaving transmission of [2]. However, the choice in [2] does not lead to the low-complexity multiuser separation and symbol recovery guarantees that this paper will turn out to possess regardless of multipath. Instrumental to these important multipath-transparent properties are the guard times of length L (see Fig. 2). They introduce redundancy that can be made arbitrarily negligible if one increases the block size K.

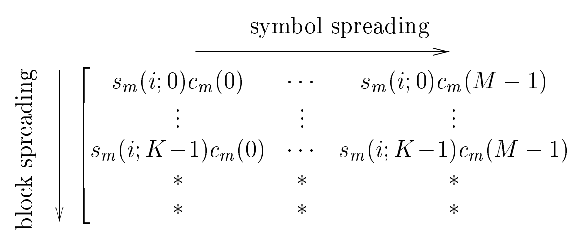

Figure 2: Redundant Chip Interleaver

Note that the receive matrix $\mathbf{G}_m$ in (4) depends only on the user's code and constitutes the code-matching user-separating front-end for user m. Multiplying $\mathbf{x}(i)$ by the matrix $\mathbf{G}_m$ amounts to passing $\mathbf{x}(i)$ through a de-interleaver (that is common to all users) and processing the output with a single-user de-spreading filter matched to $\mathbf{c}_m$ (see also Fig. 3). Next, we will show how this simple user-specific matrix $\mathbf{G}_m$ eliminates MUI deterministically (by design) regardless of *unknown* FIR frequency-selective channels.

[1] The $P \times 1$ transmitted block $\mathbf{u}_m(i) = \mathbf{C}_m \mathbf{s}_m(i)$ has M zero sub-blocks of length L evenly distributed across each block, thus has constant modulus except at the guard zeros. However, we can make up for perfectly constant modulus by filling in the zero gaps using entries drawn from the same constellation as $\mathbf{s}_m(i)$ as detailed in [12].

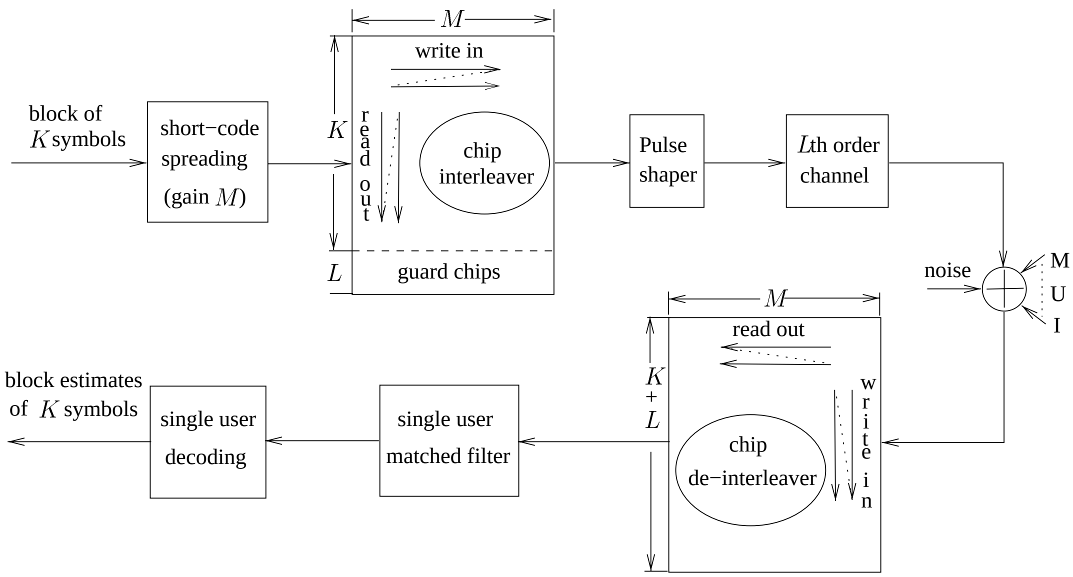

Figure 3: CIBS-CDMA transceiver for a single user

Let $\bar{\mathbf{H}}_{m,0}$ be the $(K+L)\times(K+L)$ lower triangular Toeplitz matrix with first column $[h_m(0),\ldots,h_m(L),0,\ldots,0]^T$, $\bar{\mathbf{H}}_{m,1}$ the $(K+L)\times(K+L)$ upper triangular Toeplitz matrix with first row $[0,\ldots,0,h_m(L),\ldots,h_m(1)]$, and $\mathbf{J}_M$ an $M\times M$ shift matrix which is defined as the lower triangular Toeplitz matrix with first column $[0,1,0,\ldots,0]^T$. We can then split the $P\times P$ matrix $\mathbf{H}_{m,0}$ into smaller blocks and rewrite it as:

$$\mathbf{H}_{m,0} = \mathbf{I}_M \otimes \bar{\mathbf{H}}_{m,0} + \mathbf{J}_M \otimes \bar{\mathbf{H}}_{m,1}. \tag{5}$$

Now, let us recall the identity of Kronecker products applied to matrices (with matching dimensions): $(\mathbf{A}_1\otimes\mathbf{A}_2)(\mathbf{A}_3\otimes\mathbf{A}_4) = (\mathbf{A}_1\mathbf{A}_3)\otimes(\mathbf{A}_2\mathbf{A}_4)$. Applying the latter twice, we obtain:

$$\begin{aligned}\mathbf{G}_\mu\mathbf{H}_m\mathbf{C}_m &= (\mathbf{c}_\mu^{\mathcal{H}}\otimes\mathbf{I}_{K+L})(\mathbf{I}_M\otimes\bar{\mathbf{H}}_{m,0}+\mathbf{J}_M\otimes\bar{\mathbf{H}}_{m,1})(\mathbf{c}_m\otimes\mathbf{T}_{zp})\\ &= (\mathbf{c}_\mu^{\mathcal{H}}\mathbf{c}_m)\otimes(\bar{\mathbf{H}}_{m,0}\mathbf{T}_{zp}) + (\mathbf{c}_\mu^{\mathcal{H}}\mathbf{J}\mathbf{c}_m)\otimes(\bar{\mathbf{H}}_{m,1}\mathbf{T}_{zp})\\ &= \delta(m-\mu)\bar{\mathbf{H}}_{m,0}\mathbf{T}_{zp},\end{aligned} \tag{6}$$

where we also used the fact that $\bar{\mathbf{H}}_{m,1}\mathbf{T}_{zp} = \mathbf{0}_{(K+L)\times(K+L)}$, which can be readily verified by direct substitution.

Because the last L rows of the matrix $\mathbf{C}_m$ are zero, the IBI is eliminated since $\mathbf{H}_{m,1}\mathbf{C}_m = \mathbf{0}_{P\times K}$. Therefore, using (2) and (6), we can express (3) as:

$$\mathbf{y}_\mu(i) = \mathbf{G}_\mu\mathbf{x}(i) = \bar{\mathbf{H}}_{\mu,0}\mathbf{T}_{zp}\mathbf{s}_\mu(i) + \mathbf{G}_\mu\boldsymbol{\eta}(i). \tag{7}$$

Equation (7) shows how the superposition of received signals from multiple users can be separated deterministically regardless of the unknown FIR multipath channels. Under d1), the matrix $\mathbf{G}_\mu$ in (4) depends only on the user's code, and is unitary; thus, if $\boldsymbol{\eta}(i)$ is white, then $\mathbf{G}_\mu\boldsymbol{\eta}(i)$ remains white and *multiuser separation with the MF* $\mathbf{G}_\mu$ *in* (7) *preserves maximum likelihood optimality*; i.e.,

$$P\left(\mathbf{x}(i)|\,\{\mathbf{s}_\mu(i)\}_{\mu=0}^{M-1}\right) = \prod_{\mu=0}^{M-1} P\left(\mathbf{y}_\mu(i)|\mathbf{s}_\mu(i)\right). \tag{8}$$

Relying on (de-)interleaving and matched filtering (MF) operations only, we have successfully converted multiuser detection to a set of equivalent single user detection problems without loss of optimality.

After MUI elimination, any single user equalizer can be applied to $\mathbf{y}_\mu(i)$ of (7) to remove the Inter Symbol Interference (ISI). Since the tall Toeplitz matrix $\bar{\mathbf{H}}_{m,0}\mathbf{T}_{zp}$ of size $(K+L)\times K$ has *always full rank*, the symbol block $\mathbf{s}_\mu(i)$ is guaranteed to be recoverable. For example, we can adopt the linear ZF equalizer given by († denotes matrix pseudo-inverse):

$$\Gamma_\mu^{zf} = (\bar{\mathbf{H}}_{\mu,0}\mathbf{T}_{zp})^{\dagger}, \tag{9}$$

or, the linear MMSE equalizers. Nonlinear approaches including DF and ML equalization can also be applied. Trading off performance with complexity, simple frequency domain equalization (similar to that used in OFDM [1]) is also possible, as we detail next.

Let $\mathbf{I}_{zp}$ denote the first L columns of $\mathbf{I}_K$, and define the $K\times(K+L)$ matrix $\mathbf{R}_{zp} := [\mathbf{I}_K, \mathbf{I}_{zp}]$. We then can easily verify that $\mathbf{R}_{zp}$ converts the Toeplitz matrix $\bar{\mathbf{H}}_{\mu,0}\mathbf{T}_{zp}$ into a circulant one: $\tilde{\mathbf{H}}_\mu := \mathbf{R}_{zp}\bar{\mathbf{H}}_{\mu,0}\mathbf{T}_{zp}$. Because (I)FFTs diagonalize circulant matrices, the circulant matrix $\tilde{\mathbf{H}}_\mu$ can be decomposed (see [10] for details) as $\tilde{\mathbf{H}}_\mu = \mathbf{F}_K^{\mathcal{H}}\mathbf{D}(\mathbf{h}_\mu)\mathbf{F}_K$, where: $\mathbf{F}_K$ is the $K\times K$ FFT matrix with $(k+1,l+1)$th entry $e^{-j2\pi kl/K}$; $\mathbf{h}_\mu := [h_\mu(0),\ldots,h_\mu(L)]^T$; and $\mathbf{D}(\mathbf{h}_\mu)$ is a diagonal matrix with kth diagonal entry: $H_\mu(e^{-j2\pi k/K}) := \sum_{l=0}^{L} h_\mu(l)e^{-j2\pi kl/K}$. Therefore, we can design our low complexity (LC) equalizer as:

$$\Gamma_\mu^{lc} = \mathbf{F}_K^{\mathcal{H}}\mathbf{D}^{\dagger}(\mathbf{h}_\mu)\mathbf{F}_K\mathbf{R}_{zp}, \quad \mu\in[0,M-1]. \tag{10}$$

Matrix Γ_μ^{lc} in (10) includes two FFT operations and yields at its output $\Gamma_\mu^{lc}\mathbf{y}_\mu(i) = \mathbf{s}_\mu(i) + \Gamma_\mu^{lc}\mathbf{G}_\mu\boldsymbol{\eta}(i)$. However, symbol recovery is not guaranteed for this low-complexity equalizer if the channel $H_\mu(z)$ has nulls located at $z = e^{-j2\pi k/K}$, because $\mathbf{D}(\mathbf{h}_\mu)$ (and hence $\tilde{\mathbf{H}}_\mu$) looses rank and becomes non-invertible.

We next compare the proposed design with those MUI-free transceivers in [3, 10] and [4, 5], and discuss possible extensions and channel estimation issues.

A. Bandwidth Efficiency

One way to look at bandwidth efficiency is to calculate the maximum number of users (with guaranteed MUI elimination and symbol recovery) that can be accommodated by the available system bandwidth. Suppose the system is allocated bandwidth W and the information rate is R_b. The spreading gain is thus $N = W/R_b$. For Additive White Gaussian Noise (AWGN) channels, the maximum number of users that do not interfere with each other is N when employing either TDMA, FDMA, or CDMA. In our system (and [3]), the spreading gain is $P/K = M(K+L)/K$ and thus the maximum number of MUI-free users is

$$M_1 = KN/(K+L). \tag{11}$$

The maximum number of users in the AMOUR system is $M_1' = (NK–L)/(K+L)$. When $L \ll NK$, we obtain $M_1 \approx M_1'$. On the other hand, choosing $K \gg L$ enables M_1 and M_1' to approach N, e.g., $M_1 = N-1$ when $K = L(N-1)$. In contrast, the codes in [4, 5] are designed to be shift orthogonal, which constrains their length to equal twice the number of users plus one. Thus, the maximum number is $M_2 = (N-1)/2$ and $N-1$ should also be a power of 2 [4, 5]. Selecting the information block size $K \geq L$ as in [4, 5], we find that

$$M_1 \approx M_1' \geq N/2 > M_2. \tag{12}$$

Certainly, with increasing K decoding delays increase, but our system can accommodate many more users than [4, 5].

B. Complexity and Design Flexibility

Since MUI-free transceivers decouple MIMO transmissions into multiple parallel SISO transmissions, they have in general lower complexity than joint multiuser detectors (see, e.g., [5] for a comparison). Here we will compare the computational complexity among the three different MUI-resilient transceivers: this paper's and those in [3, 10] and [4, 5]. Because all three transceivers arrive at the same single user block model (7) and hence the complexity of blind channel estimation and equalization is identical, we only concentrate on the complexity differences among the three front-ends ($\mathbf{G}_m$).

For each user in our proposed transceiver, the front-end consists of $K+L$ inner products between $M \times 1$ vectors. Since each M-long correlator requires M multiplies and $M-1$ adds, its complexity is $O(M)$. So the complexity per user is $O(M(K+L))$ and the overall complexity for all users (e.g., at the Base Station (BS) where we need to demodulate all users' information) is $O(M^2(K+L))$. With M users in [4, 5], the shift-orthogonal codes have length $2M+1$ and subsequently each receiver needs $K+L$ inner products between $2M \times 1$ vectors after discarding 1 cyclic prefix. Therefore, each user in [4, 5] has the complexity of $O(2M(K+L))$, and the overall complexity for M users will be $O(2M^2(K+L))$.

Note that the code design herein is very flexible because the only requirement on our spreading codes is their mutual orthogonality. Since the design of orthogonal codes has been well developed in the literature (at least for multipath-free propagation), there are many fast algorithms available. Using such algorithms, the complexity of our transceivers can be lowered even further. For example, we can adopt IFFT or Walsh-Hadamard (W-H) codes for the code-generating vector $\mathbf{c}_m$ in (4). Then at the BS, we can apply FFT or Fast Walsh Transform (FWT) to separate users. If the FFT is employed, the complexity for all users is $O(M(\log_2 M)(K+L))$, while the FWT is even faster than the FFT.

For each user in [3], the front-end consists of $K+L$ inner products between the $P \times 1$ received block and different Vandermonde vectors. So the complexity per user is $O(P(K+L))$ and the overall complexity for M users will be $O(MP(K+L))$. Note that AMOUR codes can also be constructed based on FFTs to reduce computational complexity. The simplest AMOUR code is proposed in [10, eq. (25)] as: $\mathbf{C}_m = \mathbf{f}_m \otimes \mathbf{T}_{zp}$ with $\mathbf{f}_m$ the mth column of $M \times M$ FFT matrix. Since $\{\mathbf{f}_m\}_{m=0}^{M-1}$ satisfy d1), it is subsumed in our code designs herein and the resulting receiver has complexity $O(M(\log_2 M)(K+L))$.

Therefore, with the same number of users, our system has less complexity than [4, 5] and [3]. The difference becomes more pronounced if special codes (e.g., W-H or FFT) are employed. Fast algorithms for the codes in [4, 5] have not been reported.

C. Non-redundant precoded alternatives

The equalizer (10) entails two FFTs at the receiver. As with single user OFDM systems [1], one FFT operation can be moved to the transmitter, which amounts to linear precoding with the IFFT matrix as follows:

$$\mathbf{s}'_m(i) = \mathbf{F}_K^{\mathcal{H}} \mathbf{s}_m(i), \tag{13}$$

where $\mathbf{s}'_m(i)$ now denotes the $K \times 1$ transmitted block. Then, easy frequency-domain equalization can be accomplished with one FFT and scalar division at the receiver. By using non-redundant IFFT precoding at the transmitter, the frequency-selective channel *for each user* is further converted to parallel flat-faded sub-channels as in single user OFDM [1]. Therefore, power and bit loading can be applied across sub-channels for each user, exactly as with Discrete Multiple Tone (DMT) modulation (see e.g. [6]). That is, we can replace (13) by

$$\mathbf{s}'_m(i) = \mathbf{F}_K^{\mathcal{H}} \mathbf{\Lambda}_m \mathbf{s}_m(i), \tag{14}$$

where $\mathbf{\Lambda}_m$ is a diagonal matrix with diagonal entries allocating power across the single user sub-channels.

More generally, we can precode the information block $\mathbf{s}_m(i)$ by any full rank $K \times K$ matrix (let us denote it by $\bar{\mathbf{F}}_K$ here) instead of the FFT matrix $\mathbf{F}_K^{\mathcal{H}}$: $\mathbf{s}'_m(i) = \bar{\mathbf{F}}_K \mathbf{s}_m(i)$. Under several criteria, the optimal loading matrices $\bar{\mathbf{F}}_K$ were developed in [8]. Therefore, with the low-complexity multiuser separating front-end, we recognize that schemes developed for single user systems can be applied directly to multiuser scenarios.

D. Channel Estimation Issues

To perform channel equalization, we need to acquire CSI at the receiver. Since multiuser transmissions are converted to parallel single user transmissions, blind or non-blind channel estimation methods developed for single user transmissions can be applied directly. Certainly, training-based CSI acquisition is one candidate. However, training sequences consume bandwidth especially when the underlying channel is rapidly varying and frequent re-training is required. (Semi-) blind channel estimators attract growing interest for such cases.

Based on the data model (7), the subspace based method of [8] is directly applicable and guarantees channel identifiability regardless of the FIR channel zero locations. Because

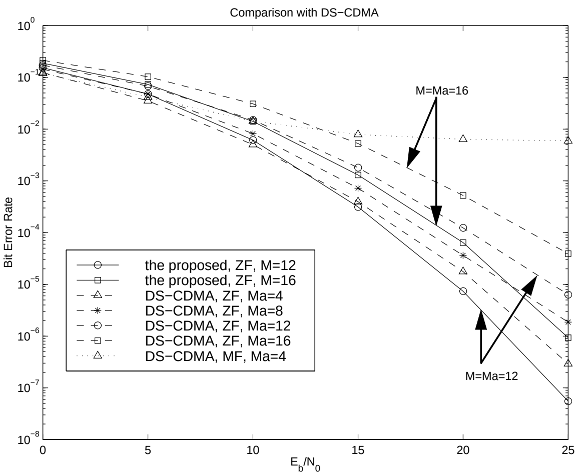

Figure 4: Comparison with DS-CDMA

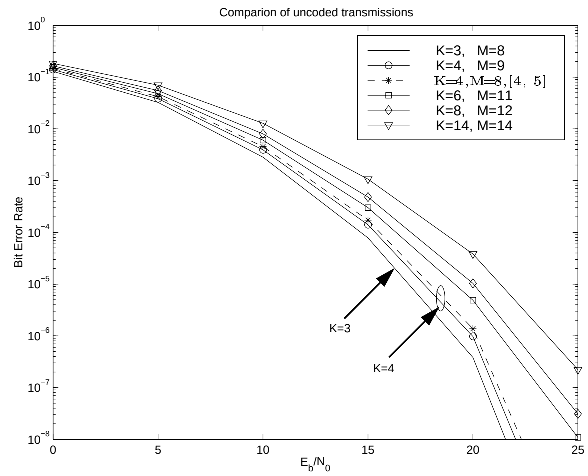

Figure 5: BER comparison with [4, 5]

subspace based methods capitalize only on data structure information, they are appropriate for any signal constellation. They can still be applied to the systems with non-redundant precoding that we described in Section C. The drawbacks of subspace based approaches are: i) they require many data blocks, which assumes that the channel is varying sufficiently slow; and ii) when K is large, their complexity increases since they involve matrix Singular Value Decomposition (SVD).

When the appropriate precoding matrix is applied, low complexity channel estimators become possible. For example, if the FFT precoder of (13) (or (14)) is applied, we arrive at a single user OFDM transmission model for each user. Based on the Finite Alphabet (FA) property of the source symbols, a new low-complexity channel estimation method was proposed recently in [11], which can estimate the channel with only one block for PSK constellations. If used together with training sequences at the beginning of the data transmission, the semi-blind adaptive implementation of [11] can track slow channel variations with high accuracy and surprisingly low complexity. Therefore, with FA-based channel estimation, the overall multiuser receiver has very low complexity.

IV. Simulations

To illustrate the merits of the proposed system, we here present several simulation results.

Test Case 1 (comparison with multiuser detectors): To compare the block-spread MUI-free transceivers with symbol-spread multiuser detectors, we simulate a Direct Sequence (DS) CDMA system with spreading gain $N = 19$ under FIR channels with maximum order $L = 3$. The DS-CDMA exploit W-H codes with length 16 and insert 3 zeros to avoid IBI. When only the channel of the desired user is available, the single-user matched filter (MF) receiver can be employed. The corresponding BER curve levels off and a high error floor appears due to the MUI, as shown in Fig. 4. To remove MUI, we also simulated ZF multiuser detectors which require knowledge of the codes and the channels of all users at the receiver-end. For the proposed transceiver, we simulated two scenarios. The first uses $K = 6$ to accommodate $M = 12$ users [c.f. (11)]. The second adopts $K = 16$, which enables MUI-free reception of $M = 16$ users. Thanks to the MUI-free reception, the BER remains unchanged when the number of active users M_a ($M_a \leq M$) varies. However, when the number of active users M_a increases in DS-CDMA, the performance of each user degrades, as shown in Fig. 4. When the system is heavily loaded, e.g., $M_a \approx M$, the proposed MUI-free transceiver outperforms the multiuser-detectors as confirmed by Fig. 4. When the system is lightly loaded, i.e., $M_a \ll M$, the multiuser detectors could exhibit better performance since MUI is less severe in this case. A final remark here is that multiuser detectors in general require higher computational complexity for both channel estimation and equalization, as illustrated in [5].

Test Case 2 (comparison with the MUI-free transceiver of [4, 5]): To compare with [4, 5], we here adopt the design parameters in [4]: $L = 3$, $K = 4$ and $M_2 = 8$ users. The code length (spreading gain) for each user in [4, 5] is then $N = 2M+1 = 17$ and the block length $P_2 = NK = 68$.

To make a fair comparison, we let both transceivers have the same spreading gain $N = W/R_b = 17$. Various system designs can be afforded by the proposed transceiver as follows.

We first fix the number of users to $M_1 = M_2 = 8$. Then the block length K should be chosen to satisfy: $M_1(K+L) < KN$, which requires $K > LM_1/(N - M_1)$. Therefore, we can set $K = L = 3$ here and $P_1 = 8 \times 6 = 48$, which results in smaller decoding delays and leads to better BER performance, which is shown in Fig. 5 with $K = 3$.

Second, we fix our system to have the same decoding delay by adopting $K = 4$. Therefore, we can afford $M_1 = 9 > M_2$ users [c.f. (11)]. However, since both systems reach the same block model of (7), each user has the same performance as confirmed by Fig. 5 with $K = 4$. The only difference here is that the system of [4, 5] incurs a small power loss $10\log_{10}(17/16) = 0.26$dB due to the cyclic prefix of length 1 (out of $N = 17$) discarded at the receiver.

Finally, we note from (11) that the number of users increases when K increases, which correspondingly increases the decoding delay. For example, with $K = 6, 8, 14$ we allow for $11, 12, 14$ MUI-free users within a block of length $P_1 = 11 \times 9 = 99, 12 \times 11 = 132, 14 \times 17 = 238$, respectively. In addition to longer decoding delays, the BER performance degrades as shown in Fig. 5 as the maximum number of users increases. Clearly, we see the trade-offs among bandwidth efficiency, decoding delays and BER performance. These trade-

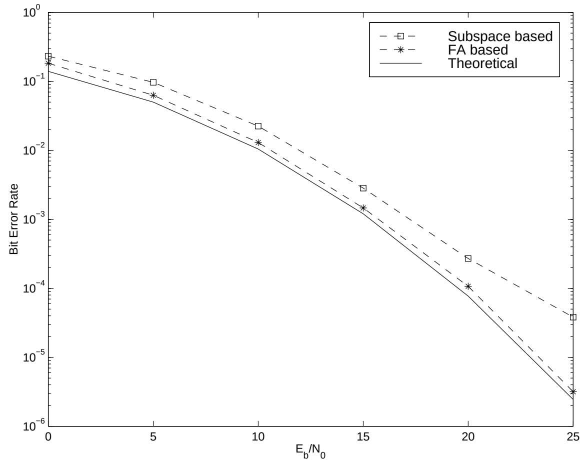

Figure 6: Channel estimation with 15 blocks

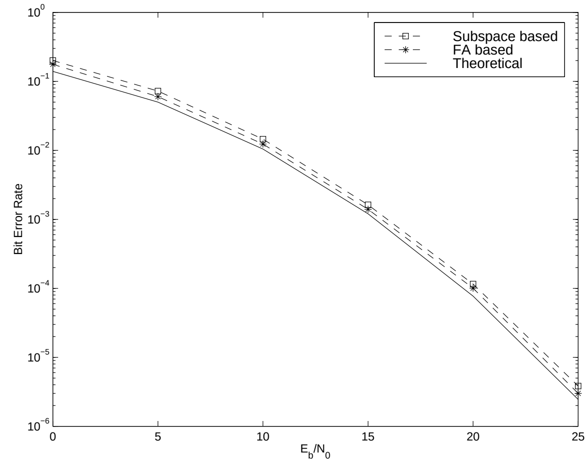

Figure 7: Channel estimation with 30 blocks

offs are not available in [4, 5] where the number of users is limited to half of the spreading gain.

Test Case 3 (comparison of (semi-) blind channel estimators): To test the blind channel estimators outlined in Section D, we here simulate a system with parameters: $K = 8$, $L = 3$. To provide a very low-complexity channel estimator based on the Finite-Alphabet property of the source symbols, we also precode our transmission with a K-point IFFT as in (13) and choose to initialize the FA-based estimator with one training block. When only 15 blocks are collected, the FA-based channel estimator outperforms the subspace based alternative considerably as verified by Fig. 6, which also corroborates the fast convergence of the FA-based method. When more blocks become available, Fig. 7 illustrates that subspace based methods can also approach the benchmark theoretical performance with known channels. As mentioned in Section D and confirmed by Figs. 6 and 7, the drawback of subspace based methods is their slow convergence and high computational complexity. On the positive side, the subspace based method is constellation-independent and thus it can be applied to any of the non-redundant precoded and loaded transmissions that we discussed in Section C. In contrast, the FA-based channel estimator is tailored to the FFT precoded transmission in (13), to offer fast convergence and low complexity.

V. Conclusions

A novel MUI-Free CDMA transceiver was developed in this paper for frequency-selective multipath channels. Relying on chip-interleaving and zero-padded transmissions, multiuser signals were extracted via low-complexity code-matched filtering, which preserves maximum likelihood optimality and converts multiuser detection to a set of equivalent single user detection problems. In addition to MUI-free reception, symbols were also guaranteed to be recoverable. By increasing the symbol block size, the proposed system achieves high bandwidth efficiency and by filling the zero gaps with known symbols it enables transmissions with perfectly constant modulus [12]. Separating the superposition of multiuser transmissions through frequency-selective multipath enabled application of single-user (semi-) blind channel estimation and equalization algorithms for intersymbol interference mitigation. Simulation results corroborated the improved performance relative to competing alternatives.

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