

Time- and frequency-domain spreading assisted MC DS-CDMA using interference rejection spreading codes for quasi-synchronous communications

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Abstract

In this contribution we considered a Time-Frequency (TF)-domain spreading aided MC DS-CDMA system, which employed so-called generalized orthogonal spreading codes exhibiting an interference free window (IFW). Provided that the propagation delay differences of the different users are within this IFW, no multiuser interference is inflicted, and hence a near-single-user performance may be attained without invoking a multiuser detector. As an additional benefit, the proposed MC DS-CDMA system is capable of extending the width of IFW in comparison to a conventional DS-CDMA system. Furthermore, it is capable of achieving a significantly higher frequency diversity gain in comparison to a single-carrier DS-CDMA system while reducing the MUD's complexity.

1. INTRODUCTION

In the context of code-division multiple-access (CDMA) communication, there are two fundamental types of spread-spectrum schemes. The first scheme spreads the original data stream in the time (T)-domain [1, 2], while the second in the frequency (F)-domain, resulting in the scheme known as MC-CDMA [3–5]. In [6], an amalgam of the above spreading schemes has been developed. Explicitly, the original data stream is spread not only in the T-Domain, but also in the F-Domain. Hence each user is assigned two spreading sequences for this operation, namely a T-domain and a F-domain sequence. This system exhibits a high flexibility as well as a reduced Multiuser Detection (MUD) complexity. In [6], several TF-domain spreading assisted MC DS-CDMA MUD schemes have been considered. However, the complexity imposed may become excessive, when the number of users supported is high. In this contribution, we will consider the employment of two specific families of spreading codes as the T-domain spreading code, which are known as generalized orthogonal codes [7]. These codes exhibit a so-called Interference Free Window (IFW). Over the duration of the IFW both the cross-correlation and the auto-correlation of the spreading codes is zero. The benefit of employing these specific codes as the T-domain code is that we are capable of reducing the complexity of the MUD, while achieving a frequency diversity gain. Specifically, the MUD's complexity is reduced because only a small fraction of the total number of users has to be separated and detected by the MUD, which belong to a given MUD group. By contrast, the set of users which are differentiated with the aid of unique user-specific spreading codes having an IFW do not interfere with each other, as a benefit of the IFW provided by the T-Domain codes used. Another advantage of the proposed scheme is that we can significantly extend the width of the IFW in comparison to a single-carrier DS-CDMA system, because as a benefit of

distributing the bits to several subcarriers, MC DS-CDMA has the potential of significantly reducing the chip rate, thereby extending the duration T_c of the chips. This also allows us to extend the width of the IFW, which renders the system more insensitive to timing imperfections, since larger timing errors can be accumulated without imposing interference.

This report is organized as follows. Sections 2 will briefly describe the family of generalized orthogonal codes. Section 3 characterises the philosophy of TF-domain spreading in the context of MC DS-CDMA signals. Section 4 considers two different correlation based detection schemes, while in Section 5 we discuss the beneficial features of this specific system. Finally, in Section 6 we provide simulation results for characterising different generalized orthogonal codes, while in Section 7 we present our conclusions.

2. GENERALIZED ORTHOGONAL CODES

Since the basic properties of generalized orthogonal codes have been characterized in [7], we will concentrate our attention on procedures used for creating generalized orthogonal codes. Firstly, we define a sequence set $\mathbf{c}_1, \dots, \mathbf{c}_N$, where $\mathbf{c}_n = [c_{n,0}, \dots, c_{n,L-1}]$ is a spreading sequence having a length of L . The spreading codes result in an IFW width of τ_{IFW} , if the crosscorrelation of the spreading codes satisfies:

$$R_{jk}(\tau) = \sum_{l=0}^{L-1} c_{j,l} c_{k,(l+\tau) \bmod L} = \begin{cases} L, & \text{if } \tau = 0, j = k \\ 0, & \text{if } \tau = 0, j \neq k \\ 0, & \text{if } 0 < |\tau| \leq \tau_{\text{IFW}} \end{cases} \quad (1)$$

The family of generalized orthogonal binary codes is generated from a pair of so-called complementary sequences also referred to as mates [8, 9], which can be recursively generated as follows:

$$[X_0, Y_0] = [1, 1]. \quad (2)$$

$$[X_m, Y_m] = [X_{m-1} Y_{m-1}, (-X_{m-1}) Y_{m-1}]. \quad (3)$$

Hence, the length of X_m, Y_m is $L_m = 2^m$, and for a given complementary sequence pair $[X_m, Y_m]$, we can construct the 0th order generalized orthogonal code's mother matrix $F^{(0)}$, which can be expressed as [7]:

$$F^{(0)} = \begin{bmatrix} -X_m & Y_m \\ -\tilde{Y}_m & -\tilde{X}_m \end{bmatrix}, \quad (4)$$

where \tilde{Y}_m is generated by reversing the order of the sequence Y_m , while $-\tilde{X}_m$ is the negated version of X_m . Each row of the mother matrix $F^{(0)}$ constitutes a spreading sequence, hence two spreading sequences are hosted by the matrix $F^{(0)}$. Once we obtained the mother matrix $F^{(0)}$, the so-called $(n) + 1$ th-order generalized orthogonal code's matrix can be recursively generated according to:

$$F^{(n+1)} = \begin{bmatrix} F^{(n)} \otimes F^{(n)} & (-F^{(n)}) \otimes F^{(n)} \\ (-F^{(n)}) \otimes F^{(n)} & F^{(n)} \otimes F^{(n)} \end{bmatrix}, \quad (5)$$

¹The financial support of the Mobile VCE, UK; the EPSRC UK and that of the European Union is gratefully acknowledged.

where \otimes denotes an operation referred to as interleaving, and the interleaving interval is 2^{m+n-1} . For the example of two vectors, namely for $A = [a_1, a_2, a_3, a_4, \dots]$, $B = [b_1, b_2, b_3, b_4, \dots]$ and for the interleaving interval of 2 we have:

$$A \otimes B = [a_1, a_2, b_1, b_2, a_3, a_4, b_3, b_4, \dots]. \quad (6)$$

For simplicity, we denote the n th-order generalized orthogonal codes as $F(L, M, Z)$, where L is the length of generalized orthogonal code, M is the number of the codes generated and Z is the width of the IFW. Explicitly, we have $L = 2^{2n+m+1}$, $M = 2^{n+1}$, $Z = 2^{n+m-1}$. As an example, we consider $m = 2$ complementary pairs, which can be obtained with the aid of Equation 2 and 3 as follows:

$$X_m = \{-, -, -, -, -, +\}$$

$$Y_m = \{-, +, +, +, -, -\}.$$

Once we obtained the complementary pair X_m, Y_m , we can construct the 0th-order mother matrix $F^{(0)}$ according to Equation 4, and the matrix $F(32, 4, 4)$ of the codes can be constructed according to Equation 5, while the interleaving interval is $2^{m+n-1} = 2^{2+1-1} = 4$. Therefore, following the interleaving operation, the spreading code $F(32, 4, 4)$ can be constructed as:

$$\begin{aligned} & \{-, -\} \\ & \{-, -\} \\ & \{-, -\} \\ & \{-, -\} \\ & \{-, -\} \\ & \{-, -\} \\ & \{-, -\} \\ & \{-, -\} \end{aligned}$$

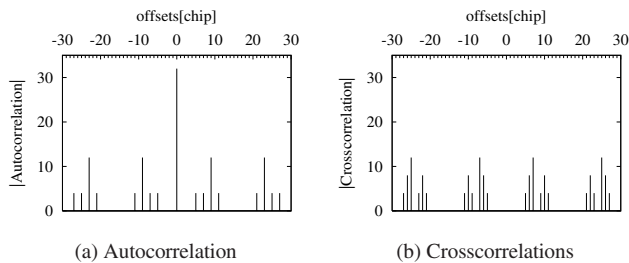


Figure 1: The auto- and cross-correlation magnitudes of the $F(32, 4, 4)$ codes. (a) All four codes of the family exhibit the same autocorrelation magnitude. (b) The crosscorrelation magnitudes of the four codes are also identical.

All four different codes of the $F(32, 4, 4)$ family exhibited the same autocorrelation magnitudes, namely that seen in Figure 1(a). It can be observed in Figure 1(a) that the off-peak autocorrelation $R_p[\tau]$ $p = 1, \dots, 4$ becomes zero for $|\tau| \leq 4$. The crosscorrelation magnitudes $|R_{j,k}(\tau)|$ depicted in Figure 1(b) are also zero for $|\tau| \leq 4$. From the observations made as regards to the aperiodic correlations we may conclude that the $F(32, 4, 4)$ codes exhibit an IFW within an offset duration of ± 4 chip intervals.

Furthermore, we can shorten the generalized orthogonal code set $F(L, M, Z)$ for creating a new generalized orthogonal code set $F'(L', M', Z')$, where we have $L' = 2^{2n+m+1-t}$, $M' = 2^{n+1}$, $Z' = 2^{n+m-1-t}$, provided that we have $t < \min\{m_i\}$. This shortening operation will reduce the code length L , however, it also reduces the width τ_{IFW} of the interference free window. Our general objective is to maximize the relative duration of the IFW in comparison to the code length, while generating the highest possible number of codes. Broadly speaking this allows us to support the highest possible number of users without imposing multiuser interference.

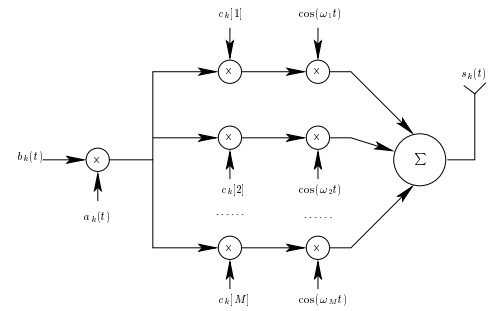


Figure 2: Transmitter model of MC DS-CDMA using both time-domain and frequency-domain spreading. T-domain spreading is achieved using the code $a_k(t)$ and each chip of $a_k(t)$ is spread in the frequency-domain by mapping it to M subcarriers, each carrying one of the M F-domain chips c_k .

3. SYSTEM MODEL

The transmitter schematic of the MC DS-CDMA scheme using both T-domain and F-domain, i.e. TF-domain spreading is shown in Fig.2 in the context of the k th user. At the transmitter side, the binary data stream $b_k(t)$ is first direct-sequence (DS) spread using the T-domain signature sequence $a_k(t)$. Following T-domain spreading, the T-domain DS spread signal is divided into M parallel branches, where each branch-signal is multiplied by a corresponding chip value of the F-domain spreading sequence $\mathbf{c}_k = [c_k[1], c_k[2], \dots, c_k[M]]^T$ of length M . Following F-domain spreading, each of the M branch signals modulates a subcarrier frequency using binary phase shift keying (BPSK). Then, the M number of subcarrier-modulated substreams are added in order to form the transmitted signal. Hence, the transmitted signal of user k can be expressed as

$$s_k(t) = \sqrt{\frac{2P_k}{M}} \sum_{m=1}^M b_k(t) a_k(t) c_k[m] \cos(\omega_m t), \quad k = 1, 2, \dots, K \quad (7)$$

where P represents the transmitted power of each user and $\{\omega_m\}$, $m = 1, \dots, M$ represents the subcarrier frequency set. The binary data stream's waveform $b_k(t) = \sum_{i=0}^{\infty} b_k P_{T_b} (t - iT_b)$, $k = 1, \dots, K$

consists of a sequence of mutually independent rectangular pulses P_{T_b} of duration T_b and of amplitude +1 or -1 with equal probability. The spreading sequence $a_k(t) = \sum_{j=0}^{\infty} a_{kj} P_{T_c} (t - jT_c)$, $k =$

$1, \dots, K$ denotes the T-domain spreading sequence waveform of the k th user, where $P_{T_c}(t)$ is the rectangular chip waveform, which is defined over the interval $[0, T_c)$. We assume that the T-domain spreading factor is $N = T_b/T_c$, which represents the number of chips per bit-duration. Furthermore, we assume that the subcarrier signals are orthogonal and the spectral main-lobes of the subcarrier signals are not overlapping with each other.

We assume that K quasi-synchronous¹ TF-domain spread MC DS-CDMA signals obeying the form of (7) are transmitted over the uplink frequency selective channel, but each subcarrier of each user experiences statistically independent single-path flat Rayleigh fading. Hence, the m th subcarrier's Channel Impulse Response (CIR) can be written as $\gamma_{m,k} e^{j\phi_{m,k}}$, $m = 1, \dots, M$, $k = 1, \dots, K$, where the amplitude $\gamma_{m,k}$ is a Rayleigh distributed random variable and the

¹Quasi-synchronous in this context implies that the delay-differences of the individual users are within the IFW.

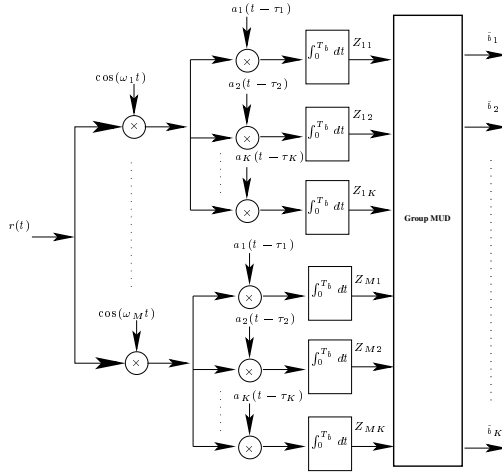


Figure 3: Receiver model of MC DS-CDMA using both time-domain and frequency-domain spreading.

phase $\phi_{m,k}$ is uniformly distributed between $[0, 2\pi)$. We also assume furthermore that each user experienced a different delay of τ_k , which obeys $\tau_k \leq \tau_{\text{IFW}}$. This can be achieved by invoking a Global Positioning System (GPS) assisted synchronization protocol. Then the received signal may be expressed as:

$$r(t) = \sum_{k=1}^K \sqrt{\frac{2P_k}{M}} \sum_{m=1}^M b_k(t) a_k(t - \tau_k) c_k[m] \gamma_{m,k} \times \cos(\omega_m t + \phi_{m,k}) + n(t), \quad (8)$$

where $n(t)$ represents the AWGN having zero mean and double-sided power spectral density of $N_0/2$.

Moreover, let $\{a_1(t), a_2(t), \dots, a_N(t)\}$ and $\{c_1, c_2, \dots, c_M\}$ be the N number of T-domain spreading sequences and M number of F-domain spreading sequences, respectively, where $\mathbf{c}_u = [c_u[1] \ c_u[2] \ \dots \ c_u[M]]^T$, $u = 1, \dots, M$ represents a F-domain spreading code. Furthermore, we assume that the number of active users is K . We also introduce a new variable of $\mathcal{K} = \lfloor K/N \rfloor$, where $\lfloor x \rfloor$ represents the smallest integer not less than x , for denoting the number users associated with a specific T-domain code, which are differentiated by a unique F-domain code. Then, we have $1 \leq \mathcal{K} \leq M$, since the total number of users is less than the product of the T-domain and F-domain spreading factor, *i.e.* we have $K \leq NM$ and $\mathcal{K} \leq M$. Based on the above assumptions, the K number of users supported can be grouped into N user groups, with each group supporting at most \mathcal{K} users. Consequently, it can be readily shown that each of the N user groups can be distinguished by assigning one of the N number of T-domain spreading sequences.

As shown in Fig.2 and Equation (7), each TF-domain spread MC DS-CDMA signal is identified with the aid of two spreading sequences, one applied in the context of the T-domain and one in the F-domain. In the following sections we will analyze the detection of TF-domain spread MC DS-CDMA signals by invoking two different detection schemes. Specifically, in Section 4 we used both Maximum Ratio Combining (MRC) as well as a low-complexity Maximum Likelihood (ML) decision based MUD.

4. DETECTION SCHEMES

In this section we will discuss the receiver model of the MC DS-CDMA schemes employed, which is shown in Fig 3. We consider

a correlation-based receiver, which essentially carries out the inverse operation seen in Fig 2. In Fig 3 the output variable $Z_{m,k}$ corresponding to the m th subcarrier of the k th user can be expressed as:

$$Z_{m,k} = \int_0^{T_b} r(t) a_k(t - \tau_k) \cos(\omega_m t) dt, \quad k = 1, \dots, K, m = 1, \dots, M \quad (9)$$

According to [10], the received signal vector \mathbf{Z}_m at the output of the bank of matched filters related to the m th subcarrier can be expressed as:

$$\begin{aligned} \mathbf{Z}_m^{(i)} &= [z_{m,1}, \dots, z_{m,K}] \\ &= \mathbf{R}[1] \mathbf{W}_m \mathbf{C}_m \mathbf{b}^{(i-1)} + \mathbf{R}[0] \mathbf{W}_m \mathbf{C}_m \mathbf{b}^{(i)} + \mathbf{R}^T[1] \mathbf{W}_m \mathbf{C}_m \mathbf{b}^{(i+1)} + \mathbf{n}_m, \end{aligned} \quad (10)$$

where we have:

$$\begin{aligned} \mathbf{C}_m &= \text{diag}[c_1[m], \dots, c_K[m]] \\ \mathbf{W}_m &= \text{diag}[\gamma_{m,1} e^{j\phi_{m,1}}, \dots, \gamma_{m,K} e^{j\phi_{m,K}}] \\ \mathbf{b} &= [b_1, \dots, b_K]^T \\ \mathbf{n} &= [n_1, \dots, n_K]^T \\ i &\text{ is the time index,} \end{aligned} \quad (11)$$

and the zero-mean Gaussian noise vector \mathbf{n}_m has the crosscorrelation matrix of:

$$E[\mathbf{n}[i] \mathbf{n}^T[j]] = \begin{cases} \sigma^2 \mathbf{R}^T[1], & \text{if } j = i+1 \\ \sigma^2 \mathbf{R}[0], & \text{if } j = i; \\ \sigma^2 \mathbf{R}[1], & \text{if } j = i-1; \\ 0, & \text{otherwise.} \end{cases} \quad (12)$$

The partial crosscorrelation matrix $\mathbf{R}[1]$ and the crosscorrelation matrix $\mathbf{R}[0]$ of the T-domain spreading codes, when communicating over an asynchronous channel, are defined as [10]:

$$\mathbf{R}_{jk}[0] = \begin{cases} 1, & \text{if } j = k; \\ \rho_{jk}, & \text{if } j < k; \\ \rho_{kj}, & \text{if } j > k \end{cases} \quad (13)$$

$$\mathbf{R}_{jk}[1] = \begin{cases} 0, & \text{if } j \geq k; \\ \rho_{kj}, & \text{if } j < k \end{cases} \quad (14)$$

where the coefficients ρ_{kj} and ρ_{jk} are the pair of crosscorrelations of the spreading codes recorded in the asynchronous CDMA system considered, which can be written as [10]:

$$\rho_{jk} = \int_{\tau}^{T_b} a_j(t) a_k(t - \tau) dt \quad (15)$$

$$\rho_{kj} = \int_0^{\tau} a_j(t) a_k(t - \tau + T_b) dt. \quad (16)$$

However, in our system the time-domain spreading codes are generalized orthogonal codes, which exhibit an IFW. Therefore, the partial cross-correlation matrix $\mathbf{R}[1]$ is an all-zero matrix, provided that the delay differences obey $\tau_k \leq \tau_{\text{IFW}}$, $k = 1, \dots, K$. Hence, Equation 10 can be simplified to:

$$\mathbf{Z}_m^{(i)} = \mathbf{R}[0] \mathbf{W}_m \mathbf{C}_m \mathbf{b}^{(i)} + \mathbf{n}_m. \quad (17)$$

In Equation 17 we can ignore the time index (i) in the superscripts. Hence, Equation 17 can be written as:

$$\mathbf{Z}_m = \mathbf{R}[0] \mathbf{W}_m \mathbf{C}_m \mathbf{b} + \mathbf{n}_m. \quad (18)$$

4.1. Low-Complexity Multiuser Detection

Let us now interpret Equation 18 in more detail. In Section 3 we have divided the K number of users supported into \mathcal{K} user groups, each group having N users, which are distinguished by their time-domain spreading code $a_k(t)$. In this scenario, the system activates N time-domain spreading codes and \mathcal{K} frequency-domain spreading codes. Therefore, provided that both the i th and j th user's delay is within the range of the IFW and the i th user and j th user are in the same group, the element ρ_{ij} of the correlation-matrix $\mathbf{R}[0]$ of Equation 18 will satisfy $\rho_{ij} = 0$. More specifically, in this system each user will encounter Multiple Access Interference (MAI) imposed by a reduced number of $\mathcal{K} - 1$ users, rather than K users, since all these \mathcal{K} users of each of the N user groups employ the same T-domain spreading code but a different F-domain spreading code. Therefore, we have to detect a reduced number of \mathcal{K} , rather than K users, which facilitates the employment of low-complexity Multiuser Detection (MUD). For example, let us consider the first user in the context of supporting a total of $K = \mathcal{K}N$ users. Then the first user encounters interference inflicted by the N th, ..., $\mathcal{K}N$ th user, because they share the same T-domain spreading code, but they are identified by the different F-domain spreading codes. Hence, Equation 18 can be simplified as:

$$\tilde{\mathbf{Z}}_m = \tilde{\mathbf{R}}[0] \tilde{\mathbf{W}}_m \tilde{\mathbf{C}}_m \tilde{\mathbf{b}} + \tilde{\mathbf{n}}_m, \quad (19)$$

where we have:

$$\begin{aligned} \tilde{\mathbf{Z}}_m &= [Z_1, Z_N, \dots, Z_{\mathcal{K}N}]^T, \\ \tilde{\mathbf{C}}_m &= \text{diag}[c_1[m], c_N[m], \dots, c_{\mathcal{K}N}[m]], \\ \tilde{\mathbf{W}}_m &= \text{diag}[\gamma_{1,m} e^{j\phi_{1,m}}, \gamma_{N,m} e^{j\phi_{N,m}}, \dots, \gamma_{\mathcal{K}N,m} e^{j\phi_{\mathcal{K}N,m}}], \\ \tilde{\mathbf{b}} &= [b_1, b_N, \dots, b_{\mathcal{K}N}]^T, \\ \tilde{\mathbf{n}}_m &= [n_1, n_N, \dots, n_{\mathcal{K}N}]^T, \end{aligned}$$

$$\tilde{\mathbf{R}}[0] = \begin{bmatrix} \rho_{1,1} & \rho_{1,N} & \cdots & \rho_{1,\mathcal{K}N} \\ \rho_{N,1} & \rho_{N,N} & \cdots & \rho_{N,\mathcal{K}N} \\ \vdots & \vdots & \ddots & \vdots \\ \rho_{\mathcal{K}N,1} & \rho_{\mathcal{K}N,N} & \cdots & \rho_{\mathcal{K}N,\mathcal{K}N} \end{bmatrix}. \quad (20)$$

The Maximum Likelihood (ML) decision based MUD of the m th sub-carrier has to evaluate:

$$\hat{\mathbf{b}} = \arg\{\min_{\mathbf{b}} \{ \|\tilde{\mathbf{Z}}_m - \tilde{\mathbf{R}}[0] \tilde{\mathbf{W}}_m \tilde{\mathbf{C}}_m \tilde{\mathbf{b}}\|^2 \}\}, \quad (21)$$

where $\|\cdot\|^2$ denotes the Euclidean norm. Upon combining all the M subcarriers, the ML based MUD's decision function can be written as:

$$\hat{\mathbf{b}} = \arg\{\min_{\mathbf{b}} \{ \sum_{m=1}^M \|\tilde{\mathbf{Z}}_m - \tilde{\mathbf{R}}[0] \tilde{\mathbf{W}}_m \tilde{\mathbf{C}}_m \tilde{\mathbf{b}}\|^2 \}\}. \quad (22)$$

According to Equation 22, the complexity of the ML decision based MUD invoked in the context of the TF-domain spreading assisted MC DS-CDMA system is on the order of $2^{\mathcal{K}}$, rather than on $2^{\mathcal{K}N}$. Let us now briefly summarize the basic features of the system considered, before we characterize the achievable system performance.

5. CHARACTERISTICS OF TF-DOMAIN SPREADING ASSISTED MC DS-CDMA EMPLOYING GENERALIZED ORTHOGONAL CODES AND LOOSELY SYNCHRONIZED CODES

The system considered in Figure 2 of the previous section can be extended by using $U \cdot M$ number of subcarriers. Specifically, as it is shown in Fig 4, the original serial data stream is first subjected

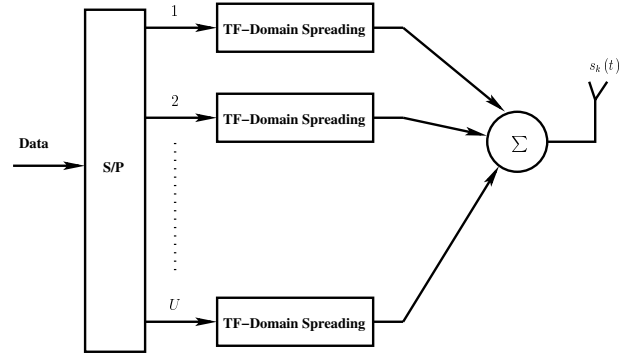


Figure 4: Transmitter model of MC DS-CDMA using serial to parallel conversion.

to serial to parallel conversion, resulting in U independent parallel data streams. Moreover, each parallel data stream is further spread to M subcarriers using the TF-domain spreading philosophy of Fig 2. The M subcarriers are arranged for maintaining the maximum possible frequency spacing, so that they experience independent fading and hence achieve the maximum possible frequency diversity gain. Specifically, let $\{f_1, f_2, \dots, f_{US}\}$ be the $U \cdot M$ number of subcarrier frequencies, where each of the U parallel bits is spread according to the following $U \times M$ -dimensional matrix:

$$\mathbf{F} = \begin{pmatrix} f_1 & f_{U+1} & \cdots & f_{(M-1)U+1} \\ f_2 & f_{U+2} & \cdots & f_{(M-1)U+2} \\ \vdots & \vdots & \ddots & \vdots \\ f_U & f_{2U} & \cdots & f_{MU} \end{pmatrix}. \quad (23)$$

Therefore, the chip rate of the MC DS-CDMA signal can be reduced by a factor of $U \cdot M$ and hence the width of the IFW can be extended by a factor of $U \cdot M$ in comparison to a DS-CDMA system, which can be expressed as:

$$\tau_{IFW,MCDS-CDMA} = U \cdot M \cdot \tau_{IFW,DS-CDMA}. \quad (24)$$

Based on Equation 24, we argue that the width of the interference free window of MC DS-CDMA systems may be significantly higher than that of DS-CDMA. This beneficial feature allows us to have significantly larger cells, which result in higher propagation delay differences and/or allows the system to reliably operate even in case of higher absolute code synchronization errors, as long as they do not exceed the IFW width.

Another advantage of this system is that it is capable of achieving a high frequency diversity gain, because the chips of each bit are transmitted on independently fading subcarriers. Hence, when the transmitted signal experiences frequency selective fading, the chances are that only some of the chips of a F-domain spreading code are corrupted and therefore the corresponding bit conveyed by the specific spreading code concerned may still be recovered. Consequently this system may be capable of achieving M th-order F-domain diversity and benefit from a factor of $2^{\mathcal{K}N}/2^{\mathcal{K}} = 2^N$ lower multiuser detection complexity.

6. SIMULATION RESULTS

Let us consider a system having $M = 4$ subcarriers and using different T-domain spreading codes. The F-domain spreading codes are generated by the 4×4 -dimensional orthogonal Walsh code generator

matrix given by:

$$\mathbf{F} = \begin{pmatrix} + & + & + & + \\ + & - & + & - \\ + & + & - & - \\ + & - & - & + \end{pmatrix}. \quad (25)$$

Furthermore, we considered four different T-domain spreading codes, which are the $F(16, 8, 1)$, $F(16, 4, 2)$, where again, generalized orthogonal codes are defined with the aid of the parameters $F(M, Z, L)$, with L being the length of the code, M being the number of codes generated and Z being the IFW width. Moreover, we assume that each subcarrier of each user experienced independent flat Rayleigh fading.

From Figures 5 and 6, we can observe that the low-complexity MUD is capable of approaching the single-user bound.

7. CONCLUSION

In this report, we employed a specific family of spreading codes, which exhibit an interference free window in the context of the TF-domain spreading assisted MC DS-CDMA system considered. In this system we reduced the complexity of the MUD by a factor of $2^{K \cdot N} / 2^K = 2^N$, while achieving M th-order frequency diversity. The system is capable of significantly extending the width of the interference free window as a benefit of the serial to parallel conversion invoked in Figure 4, which additionally renders the system insensitive to timing imperfections.

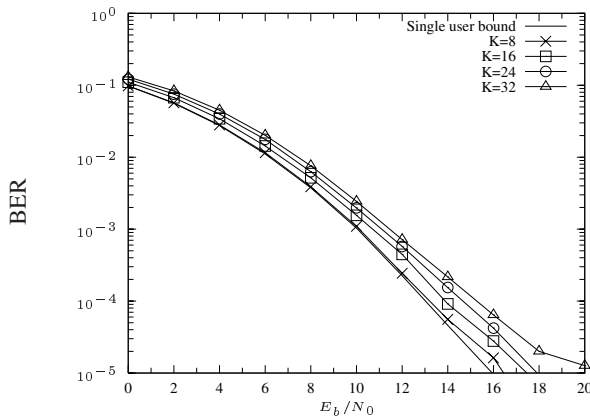


Figure 5: BER performance of the TF-domain spreading aided MC DS-CDMA system in conjunction with low-complexity MUD, while employing the $F(16, 8, 1)$ generalized orthogonal code as the T-domain spreading code. The F-domain spreading code was a 4×4 -dimensional Walsh code, and each of the $M = 4$ subcarriers experienced independent narrowband Rayleigh fading. The MUD complexity reduction factor was $2^N = 2^{82} = 5$.

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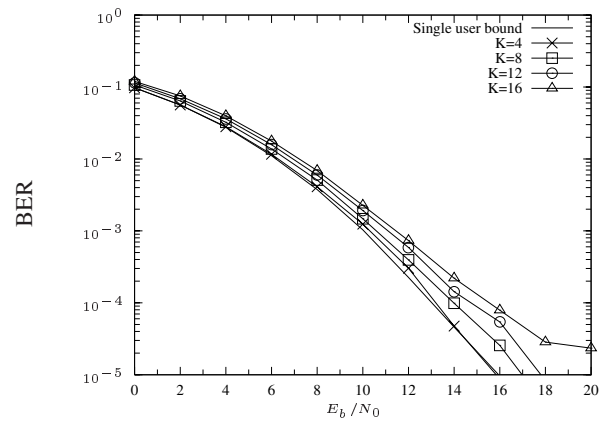


Figure 6: BER performance of the TF-domain spreading aided MC DS-CDMA system in conjunction with low-complexity MUD, while employing the $F(16, 4, 2)$ generalized orthogonal code as the T-domain spreading code. The F-domain spreading code was a 4×4 -dimensional Walsh code, and each of the $M = 4$ subcarriers experienced independent narrowband Rayleigh fading. The MUD complexity reduction factor was $2^N = 2^4 = 1$.

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