A New Multicarrier Multicode DS-CDMA Scheme for Time and Frequency Selective Fading Channels

Yewen Cao, Tjeng Thiang Tjhung, and Chi Chung Ko

Abstract: **In this paper, a new multicarrier, direct sequence code division multiple access (MC-DS-CDMA) system is proposed. Our new signal construction is based on convolutional encoding of the transmitted data, serial-to-parallel (S/P) conversion of the encoded data, Walsh-Hadamard-transformation (WHT), a second S/P conversion of the WHT outputs, spread spectrum (SS) modulation with a common pseudo-noise (PN) sequence, and then multicarrier transmission. The system bit error rate (BER) performance in frequency selective fading channel in the presence of additive white Gaussian noise (AWGN) and a jamming tone is analyzed and simulated. The numerical results are compared with those from an orthogonal MC-DS-CDMA system of Sourour and Nakagawa [7]. It is shown that the two systems have almost the same BER performance, but the proposed scheme has better anti-jamming ability.**

Index Terms: **DS-CDMA, multicarrier CDMA system, multicode transmission.**

I. INTRODUCTION

Recently, much attention has been directed to the use of multicarrier modulation techniques to obtain frequency diversity in digital communication systems. This technique was proposed for code division multiple access (CDMA) systems in 1993 [1], [2], leading to the basic multicarrier CDMA system, for which in depth investigations have subsequently been made in [3]–[10].

Another novel technique in CDMA systems that has attracted recent interests is the use of multicodes. In the next generation of DS-CDMA communication systems, variable and high data rate services with user-specified quality of services requirements need to be provided. Multicode transmission can provide these services without decreasing the processing gain or increasing the spreading bandwidth [11].

In this paper, we propose a new multicarrier DS-CDMA scheme in which the idea of multicode transmission and channel coding are considered together. In our scheme, a convolutional encoder with a low code rate $(1/R)$ is used. After encoding, the R coded bits are serial-to-parallel (S/P) converted, interleaved and then multiplexed with R Walsh-Hadamard (WH) codes to form M super-bits. The super-bits are again S/P converted to form M parallel super-bits sub-streams. Then, all the M super-bits sub-streams are direct sequence spread spectrum (DS-SS) modulated and transmitted using M orthogonal carriers. In the receiver, a correlator is provided for each carrier, and after inverse WH transforming, de-interleaving, and self interference canceling (SIC), the Viterbi algorithm is adopted for convolutional decoding.

Rowitch and Milstein [12] first conceived the idea of using convolutional code in multicarrier DS-CDMA. In [12], the convolutionally coded symbols are duplicated and transmitted through multicarriers. In our proposed scheme, however, we multiplex the convolutionally coded bits with multicodes, in a so-called code-division-multiplexing (CDM) scheme. We then S/P convert the CDM coded symbols for transmission using all the subcarriers. Several years ago, Matsutani and Nakagawa [16] have proposed the use of CDM in multi-carrier DS-CDMA system. However, unlike in [16] where the CDM processing is applied on the parallel data bits, we apply the CDM on the convolutionally encoded bit stream in our new scheme.

The motivation in proposing our above multi-code, multicarrier DS-CDMA (MM-DS-CDMA) scheme is to achieve both time and frequency diversity for flexible high data rate transmission over multipath fading channels. We use WH transformation, a well-known approach recently reported in [22] and [23], to obtain frequency diversity. We also employ convolutional encoding and interleaving, another well-known technique formalized in [20], to attain time diversity. Finally, we address the need for flexible high data rate services by including the use of multi-code and multi-carrier transmission.

This paper is organized as follows. In Section II, we introduce the system model for the proposed muticarrier multicode DS-CDMA scheme. In Section III, we present the channel model. In Section IV, we analyze the performance of the proposed system. In Section V, numerical results are presented. Finally, in Section VI, we present our conclusions.

II. SYSTEM MODEL

A. Transmitter Signal Construction

The transmitter block diagram of the proposed system is shown in Fig. 1. The input data $d(t)$ with bit duration T is convolutionally encoded at rate $1/R$ and then serial-to-parallel converted to R parallel coded bit sub-streams with bit duration T. These R sub-streams, after undergoing interleaving, are then multiplexed by R Walsh-Hadamard (WH) codes of length M. This results in M multi-level symbols in a time period of T.

For simplicity, we refer to these multilevel symbols as code division multiplexing (CDM) super-bits. The M CDM super-bits are serial-to-parallel converted, then modulated by a common PN sequence and transmitted using M subcarriers so that both time and frequency diversity are achieved.

Manuscript received May 28, 2003; approved for publication by Gi-Hong Im, Division I Editor, June 23, 2004.

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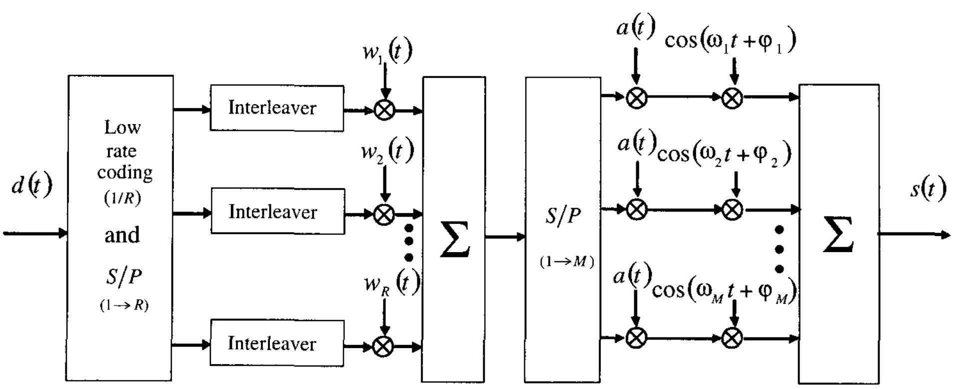

Fig. 1. Transmitter of proposed multicarrier multicode DS-CDMA system.

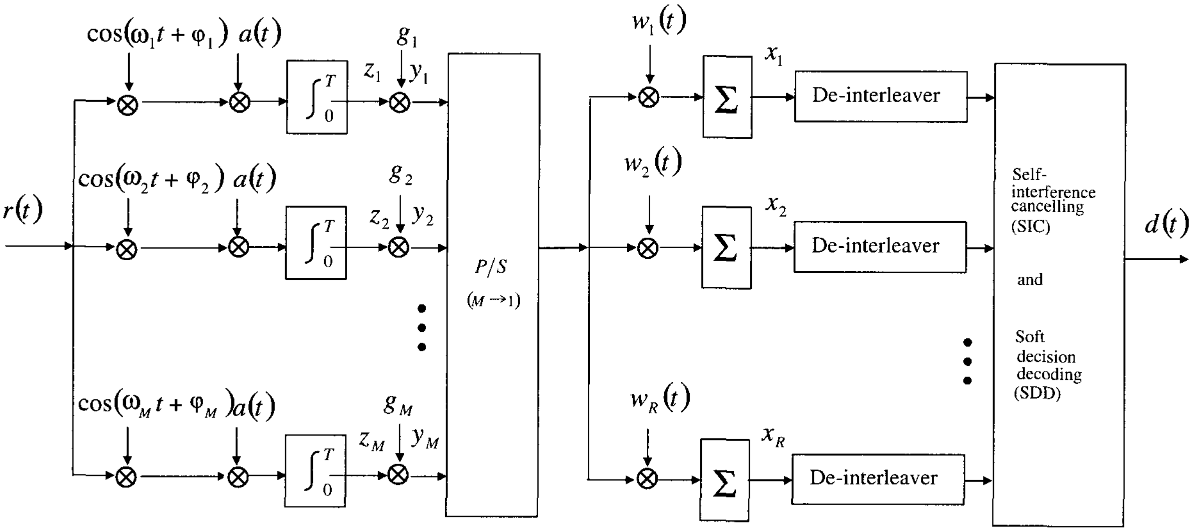

Fig. 2. Receiver of proposed multicarrier multicode DS-CDMA system.

To describe this modulation process, we note that the super-bit signal $b_{k,m}(t)$ of the k-th user is given by

$$b_{k,m} = \sum_{j=-\infty}^{\infty} p_T(t - jT), \quad m = 1, \cdots, T, \qquad (1)$$

where $b_{k,m} = \sum_{q=l}^{R} c^j_{k,q} w_{q,m}$ is the m-th CDM super-bits, $c^j_{k,q}$ is the q-th convolutionally encoded bit for the k-th user during the j-th bit duration, $w_{q,m}$ is the m-th code bit in the q-th Walsh-Hadamard sequence of length M, and $p_T(t)$ is a unit rectangular pulse of duration T. The transmitted signal for the k-th user is therefore

$$s_k(t) = \sum_{m=1}^{M} \sqrt{2P_k} b_{k,m}(t) a_k(t) \cos(\omega_m t + \varphi_{k,m}), \qquad (2)$$

where RP_k is the transmitted signal power. $a_k(t) = \sum_{n=0}^{N-1} a_{k,n} p_{T_c}(t - nT_c)$ is a pseudo-noise (PN) spreading signal with chip duration of $T_c = T/N$, $a_{k,n} \in \{+1, -1\}$ is a PN code sequence with length of N, and $p_{T_c}(t)$ is a unit rectangular pulse of duration T_c. $\omega_m = \omega_0 + 2\pi m F/T_c$ and $\varphi_{k,m}$ are, respectively, the frequency and phase of the m-th subcarrier for the k-th user, where ω_0 is the lowest carrier frequency, and the parameter F is an integer. Note that the parameter F should be such that the carrier separation is larger than the coherence bandwidth $(\Delta f)_c$ of the multipath fading channel so as to ensure independent fading among different subcarriers [1].

B. Receiver Structure

In Fig. 2, we show the block diagram of the receiver for our proposed system. A conventional correlator is used to demodulate the received signal for each carrier. After correlation demodulation, the outputs are weighted by combining gains and then parallel-to-serial converted for inverse Walsh-Hadamard (IWH) transformation, de-interleaving, self-interference canceling (SIC), and soft decision Viterbi decoding. The SIC is used to cancel the interference due to the use of multicodes and will be explained in Section V.

C. Receiver Structure

In this paper, we shall compare the bit error rate performance of our proposed muticarrier multicode DS-CDMA system with that of the orthogonal multicarrier DS-CDMA scheme of [7]. We assume that the same convolutional encoders are used for both of these systems. The overall bandwidth is $BW = N(M+$

$1)F/T$ for our proposed system, and that for the system of [7] is given by $BW_1 = N_1(M+1)F/T$, where $T = RT_1$, and N and N_1 are the lengths of the PN sequences in the two systems. Therefore, to have the same overall bandwidth, we require $N = RN_1$.

III. CHANNEL MODEL

We model the channel as a flat fading process for each of the subcarriers [3]. The low-pass equivalent channel impulse response for the m-th carrier of the k-th user is

$$h_{k,m}(t) = \beta_{k,m} \exp\{j\gamma_{k,m}\}\delta(t-\tau_k), \quad m = 1, \cdots, M, \quad (3)$$

where τ_k is the delay of the k-th user. Assuming that different subcarriers undergo the same fading process, each subcarrier will experience the same τ_k. The channel gain, $\beta_{k,m}\exp(j\gamma_{k,m})$ is a zero mean complex Gaussian random variable (r. v.). For different user k, $\beta_{k,m}$ is an independent identically distributed (i.i.d.) Rayleigh random variable with unity second moment, and $\gamma_{k,m}$ is an i.i.d. uniform random variable over $[0, 2\pi)$.

We also assume that the fading is slow so that the bit period in each subcarrier is much less than the coherence time of the channel. To meet all these conditions, we have included an interleaver and a de-interleaver in the system. However, since the addition of an interleaver and a de-interleaver will not change the basic characteristic of the proposed scheme, we will ignore them in the following discussion for convenience as in [7].

In this paper, we shall also consider the presence of a strong jamming signal in the channel [17]–[19]. The jamming will be taken to be an unmodulated single tone residing somewhere in the given frequency band. In the presence of fading, the received jamming signal is

$$J(t) = \beta_J\sqrt{2P_J}\cos(\omega_J t + \varphi_J), \quad (4)$$

where P_J is the jamming power, ω_J is the jamming frequency, β_J is a Rayleigh distributed random variable with unity second moment, and φ_J is a uniformly distributed random variable in $[0, 2\pi)$. Taking the first user as the desired one, the jamming to the desired user signal power ratio is given by α, where $P_j = \alpha P_1$.

IV. PERFORMANCE ANALYSIS

A. Decision Statistics

Assume that there are K asynchronous users with AWGN in our system. The signal received in the above-mentioned fading channel and in the presence of a single tone jamming can be expressed as

$$r(t) = \sum_{k=1}^{K}\sqrt{2P_k}\sum_{m=1}^{M}\beta_{k,m}b_{k,m}(t-\tau_k)a_k(t-\tau_k) \times \cos(\omega_m t + \phi_{k,m}) + J(t) + n(t), \quad (5)$$

where $\phi_{k,m} = \varphi_{k,m} + \gamma_{k,m} - \omega_m\tau_k$, and $n(t)$ is a Gaussian white noise with zero mean and double-side power density $N_0/2$.

We assume that the channel can be estimated perfectly. This assumption is reasonable for ensuring mathematical tractability [15]. We consider the first user to be the desired user. For a bit transmitted in, say, $[0, 2T)$, the output of the first user's correlator for the q-th carrier is

$$z_q = \int_{\tau_1}^{T+\tau_1} r(t)a_1(t-\tau_1)\cos(\omega_q t + \phi_{1,q})dt = z_0 + \eta_q + J_q + I_q, \quad (6)$$

where η_q is a Gaussian variable with zero mean and variance $N_0T/4$,

$$z_0 = \frac{\sqrt{2P_1}}{2}T\beta_{1,q}b_{1,q}^0, \quad (7)$$

$$J_q = \int_{\tau_1}^{T+\tau_1} J(t)a_1(t-\tau_1)\cos(\omega_q t + \phi_{1,q})dt, \quad (8)$$

and I_q is the sum of all interferences and can be divided into two components, $I_q = I_q^{(1)} + I_q^{(2)}$. $I_q^{(1)}$ is the multiple access interference (MAI) from the same carrier and $I_q^{(2)}$ is the MAI due to all other carriers and all other users. Note that self-interference from the same subcarrier and other subcarriers of the desired user are equal to zero due to the use of orthogonal subcarriers. $I_q^{(1)}$ and $I_q^{(2)}$ can be further expressed as

$$I_q^{(1)} = \sum_{k=2}^{K}\frac{\sqrt{2P_k}}{2}\beta_{k,q}\cos(\phi_{k,q} - \phi_{1,q}) \times \left[b_{k,q}^{(-1)}R_{1,k}(\tau) + b_{k,q}^{(0)}\widehat{R}_{1,k}(\tau)\right], \quad (9)$$

$$I_q^{(2)} = \sum_{k=2}^{K}\frac{\sqrt{2P_k}}{2}\sum_{m=1, m\neq q}^{M}\beta_{k,m} \times \left[b_{k,q}^{(-1)}Q_{1,k,m,q}(\tau) + b_{k,q}^{(0)}\widehat{Q}_{1,k,m,q}(\tau)\right], \quad (10)$$

where $\tau = \tau_k - \tau_1$, and $R_{1,k}$ and $\widehat{R}_{1,k}(t)$ are the continuous time partial cross-correlation functions as defined in [13]. $Q_{1,k,m,q}$ and $\widehat{Q}_{1,k,m,q}$ are given by

$$Q_{1,k,m,q}(\tau) = \int_0^{\tau_1} a_k(t-\tau)a_1(t) \times \cos\left[(\omega_m - \omega_q)t + \Theta_{k,1,m,q}\right]dt, \quad (11)$$

$$\widehat{Q}_{1,k,m,q}(\tau) = \int_{\tau_1}^{T} a_k(t-\tau)a_1(t) \times \cos\left[(\omega_m - \omega_q)t + \Theta_{k,1,m,q}\right]dt, \quad (12)$$

where $\Theta_{k,1,m,q} = \phi_{k,m} - \phi_{1,q}$. Setting $\tau_1 = 0$ for the first user, τ will be a random variable uniformly distributed in $[0, T)$. Now, referring to [13], we have

$$R_{1,k}(\tau) = C_{k,1}(\Gamma_k - N)(1-\Delta_k)T_c + C_{k,1}(\Gamma_k - N + 1)\Delta_k T_c, \quad (13)$$

$$\widehat{R}_{1,k}(\tau) = C_{k,1}\Gamma_k(1-\Delta_k)T_c + C_{k,1}(\Gamma_k+1)\Delta_k T_c, \quad (14)$$

where $\Gamma_k = \lfloor \tau_k/T_c \rfloor$, $\Delta_k = \tau_k/T_c - \Gamma_k$, Γ_k is an integer uniformly distributed in $[0, N-1]$, and Δ_k is uniformly distributed in $[0,1]$. As defined in [13], $C_{k,1}$, for $k=1$, is the discrete aperiodic partial auto-correlation function for the pseudo random signature $a_1(t)$. For $k \neq 1$, it is equal to the cross-correlation between $a_1(t)$ and $a_k(t)$.

Similar to that in [7], after carrying out the integration given by (11) and (12), and performing some further mathematical manipulation, we obtain

$$Q_{1,k,m,q}(\tau) = T_c\left[C_{k,1}(\Gamma_k - N + 1) - C_{k,1}(\Gamma_k - N)\right] \times D_{k,1,m,q}, \quad (15)$$

$$\widehat{Q}_{1,k,m,q}(\tau) = T_c[C_{k,1}(\Gamma_k+1) - C_{k,1}\Gamma_k]D_{k,1,m,q}, \quad (16)$$

where $D_{k,1,m,q} = \cos[\pi(m-q)\Delta_k + \Theta_{k,1,m,q}]\Delta_k \mathrm{sinc}[\pi(m-q)\Delta_k]$.

To implement frequency diversity, combining methods such as equal gain combining (EGC) and maximum ratio combining (MRC) may be considered. After combining, we get

$$y_q = g_{1,q} z_q, \quad q = 1, \cdots, M, \quad (17)$$

where $g_{1,q}$ is the combining gain for the q-th subcarrier of the first user. As shown in the receiver of Fig. 2, $\{y_q, q = 1, \cdots, M\}$ are parallel-to-serial converted and multiplied by R Walsh-Hadamard code sequences to form the decision signals $\{x_r\}$ before decoding. Therefore, the decision statistic for the r-th transmitted coded bit in bit duration $[0, T)$ is

$$x_r = \sum_{q=1}^{M} w_{r,q} y_q, \quad r = 1, \cdots, R, \quad (18)$$

where $w_{r,q}$ is the q-th code bit in the r-th inverse Walsh-Hadamard sequence of length M.

Using (1), (6), (17), and (18), we get

$$x_r = \frac{\sqrt{2P_1}T}{2}\left\{\left(\sum_{q=1}^{M} g_{1,q}\beta_{1,q}\right)c_{1,r}^0 + \sum_{i=1,i\neq r}^{R}\left(\sum_{q=1}^{M} g_{1,q}\beta_{1,q}w_{i,q}w_{r,q}\right)c_{1,i}^0\right\} + \widetilde{N}_r, \quad (19)$$

where

$$\widetilde{N}_r = \widetilde{I}_r + \widetilde{J}_r + \widetilde{\eta}_r, \quad (20)$$

is the sum of all of interferences, $\widetilde{I}_r = \sum_{q=1}^{M} g_{1,q} w_{r,q}\left(I_q^{(1)} + I_q^{(2)}\right)$, $\widetilde{J}_r = \sum_{q=1}^{M} g_{1,q} w_{r,q} J_q$, and $\widetilde{\eta}_r = \sum_{q=1}^{M} g_{1,q} w_{r,q}\eta_q$. Furthermore, $\widetilde{N}_r$ can be expressed as follows:

$$\widetilde{N}_r = \sum_{q=1}^{M} g_{1,q} w_{r,q} N_q, \quad r = 1, \cdots, R, \quad (21)$$

where $N_q = I_q + J_q + \eta_q, q = 1, \cdots, M$ are mutually independent. This is because $\{I_q, q = 1, \cdots, M\}$, $\{J_q, q = 1, \cdots, M\}$, and $\{\eta_q, q = 1, \cdots, M\}$ are mutually independent.

Now by defining $u_{ij} = 1/S\left(\sum_{q=1}^{M} g_{1,q}\beta_{1,q}w_{i,q}w_{j,q}\right)$, $S = \sum_{q=1}^{M} g_{1,q}\beta_{1,q}$, and $\sqrt{\varepsilon_1} = \sqrt{E_b T/2S}$, where $E_b = P_1 T$ is the energy per bit, we can re-write (19) in matrix form as follows:

$$\boldsymbol{X} = \boldsymbol{UC} + \widetilde{\boldsymbol{N}}, \quad (22)$$

where $\boldsymbol{X} = [x_1 \cdots x_R]^{\mathrm{T}}$, $\widetilde{\boldsymbol{N}} = [\widetilde{N}_1 \cdots \widetilde{N}_R]^{\mathrm{T}}$, $\boldsymbol{C} = [\sqrt{\varepsilon_1}c_{1,1}^0 \cdots \sqrt{\varepsilon_1}c_{1,R}^0]^{\mathrm{T}}$, and $\boldsymbol{U}$ is a R-by-R matrix with elements u_{ij}.

In the decoding procedure depicted in Fig. 2, the transmitted binary data $d(t)$ is recovered by using a soft decision Viterbi decoder based on the decision statistics $\boldsymbol{X}$. Note, however, that there exists self-interference due to loss of orthogonality between multicodes in the desired user signal. This self-interference corresponds to the second term in the brackets of (19), and may be removed by using the following procedure. From $\boldsymbol{U}$ and $\boldsymbol{X}$, we can calculate

$$\boldsymbol{X}' = \boldsymbol{U}^{-1}\boldsymbol{X} = \boldsymbol{U}^{-1}\boldsymbol{UC} + \boldsymbol{U}^{-1}\widetilde{\boldsymbol{N}} = \boldsymbol{C} + \widetilde{\boldsymbol{N}}', \quad (23)$$

Expressing (23) in component form, we have

$$x_r' = \sqrt{\varepsilon_1}c_{1,r}^0 + \widetilde{N}_r', \quad (24)$$

where $\widetilde{N}_r'$ is the r-th element of $\boldsymbol{U}^{-1}\widetilde{\boldsymbol{N}}$. Equation (24) shows that the self-interference due to multicodes has been cancelled completely. This processing is referred to as self-interference canceling (SIC) in our proposed scheme. The SIC is used before the Viterbi decoding as depicted in Fig. 2.

However, the interference $\widetilde{\boldsymbol{N}}$ has now been transformed to become $\widetilde{N}'$ (i.e., $\boldsymbol{U}^{-1}\widetilde{\boldsymbol{N}}$). This means that the processing of SIC results in noise coloring, which is inevitable due to the use of WH transformation, and will degrade the performance of our proposed system. To prevent this performance degradation, the use of maximum likelihood (ML) detection is one solution. In this paper, we do not intend to further investigate this problem due to limited space.

B. Mean and Variance

In this section, we will derive the statistics of the signal vector $\boldsymbol{X} = [x_1 \cdots x_R]^{\mathrm{T}}$ before decoding. Firstly, from (19), the mean of each component of $\boldsymbol{X}$, conditioned on the fading channel parameter $\beta_1 = \{\beta_{1,q}, q = 1, \cdots, M\}$ for the first user, is given by

$$\mathrm{E}[x_r|\beta_1] = \sqrt{\varepsilon_1}c_{1,r}^0, \quad (25)$$

where $c_{1,r}^0$ is the r-th coded bit for the first user in duration $[0, T)$.

Now, we will determine the conditional variance of $\boldsymbol{X}$. Assuming that the EGC detection strategy given by (17) is used, it is easy to see from (20) that the contribution due to AWGN leads to a Gaussian random variable $\widetilde{\eta}_r$ with $\mathrm{E}[\widetilde{\eta}_r] = 0$ and $\mathrm{var}[\widetilde{\eta}_r] = N_0 TM/4$.

Let us consider the component $\widetilde{I}_r$ in (20). It can be readily shown that $\widetilde{I}_r$ has a mean of zero. Using the analysis similar to that in [7] and [13], the variance of $\widetilde{I}_r$ is given as follows

$$\text{var}[\widetilde{I}_r] = \frac{RT^2M}{4N^3}\left\{\frac{1}{3}\sum_{k=2}^{K} P_k r_{k,1} + \frac{1}{2\pi^2 M}\sum_{k=2}^{K} P_k \times[\mu_{k,1}(0) - \mu_{k,1}(1)]\sum_{q=1}^{M}\sum_{m=1,\neq q}^{M}\frac{1}{(m-q)^2}\right\}, \quad (26)$$

where $r_{k,1}$ and $\mu_{k,1}(n)$ are defined in [13]. Note that the fading process is constrained to have unity energy, and that fading over all subcarriers is independent and identically distributed. From (26), it is easy to see that var$[\widetilde{I}_r]$ depends on the PN code sequence used.

We now turn to the component $\widetilde{J}_r$ in (20). Substituting (4) into (8), we obtain

$$J_q = \beta_J\sqrt{2P_J}\int_{\tau_1}^{T+\tau_1} a_1(t-\tau_1)\cos(\omega_J t + \varphi_J) \times \cos(\omega_q t + \phi_{1,q})dt, \approx \frac{\beta_J\sqrt{2P_J}}{2}\int_0^T a_1(t)\cos[(\omega_q - \omega_J)t + \theta]dt, \quad (27)$$

where $\theta = \varphi_J - \phi_{1,q} + (\omega_q - \omega_J)\tau_1$. Recalling that β_J is Rayleigh distributed with unity second moment, and that φ_J is uniformly distributed in the interval $[0, 2\pi)$, it is easy to obtain $\text{E}[J_q] = 0$. In order to obtain the variance of J_q, we evaluate the integral in (27) over each chip interval as follows [19]:

$$J_q \approx \frac{\beta_J\sqrt{2P_J}}{2}\sum_{i=0}^{N-1} a_{1,i}\int_{iT_c}^{(i+1)T_c}\cos[(\omega_q - \omega_J)t + \theta]dt = \frac{\beta_J\sqrt{2P_J}T_c}{2}\text{sinc}\left(\frac{(\omega_q - \omega_J)T_c}{2}\right) \times \sum_{i=0}^{N-1} a_{1,i}\cos[(\omega_q - \omega_J)iT_c + \theta'], \quad (28)$$

where $\theta' = (\omega_q - \omega_J)T_c/2 - \theta$. We note that θ' is uniformly distributed in $[0, 2\pi)$. Based on (28), since $a_{1,i}$ takes on the values of ± 1 with equal probability, it is easy to obtain

$$\text{var}[J_q] \approx \frac{\alpha E_b T}{4N}\text{sinc}^2\left(\frac{(\omega_q - \omega_J)T}{2N}\right). \quad (29)$$

Using (20) and (29), the variance of $\widetilde{J}_r$ can be obtained as

$$\text{var}[\widetilde{J}_r] = \sum_{q=1}^{M}\text{var}[J_q] \approx \frac{\alpha E_b T}{4N}\sum_{q=1}^{M}\text{sinc}^2\left(\frac{(\omega_q - \omega_J)T}{2N}\right). \quad (30)$$

As $\widetilde{\eta}_r$, $\widetilde{I}_r$, and $\widetilde{J}_r$ are independent mutually, we finally have

$$\sigma_{x_r}^2 = \text{var}[\widetilde{N}_r] = \text{var}[\widetilde{\eta}_r] + \text{var}[\widetilde{I}_r] + \text{var}[\widetilde{J}_r]. \quad (31)$$

Note that the variances of $\widetilde{I}_r$ and $\widetilde{J}_r$, as given by (26) and (30), as well as that of $\widetilde{\eta}_r$, var$[\widetilde{\eta}_r] = N_0TM/4$, do not depend on the set of fading amplitudes. It is also interesting to see that var$[\widetilde{N}_1]$ = var$[\widetilde{N}_2] = \cdots =$ var$[\widetilde{N}_R]$. This result is because the interference due to all the factors has been averaged by the use of multicodes in the proposed scheme.

As mentioned above, in order to detect the coded bits $\{c_{1,r}^0, r = 1, \cdots, R\}$ from the decision statistics $\{x_r, r = 1, \cdots, R\}$, we need to find $\boldsymbol{U}^{-1}$ to remove the self-interference. We will now discuss the implication of this procedure. From (23), the r-th element of $\boldsymbol{U}^{-1}\widetilde{\boldsymbol{N}}$ is

$$\widetilde{N}_r' = \boldsymbol{V}_r\widetilde{\boldsymbol{N}} = \sum_{i=1}^{R}\nu_{ri}\widetilde{N}_r, \quad (32)$$

where $\boldsymbol{V}_r = [\nu_{r1}\cdots\nu_{rR}]$ is the r-th row of $\boldsymbol{U}^{-1}$ and is dependent only on the channel parameter $\beta_1 = \{\beta_{1,q}, q = 1, \cdots, M\}$ associated with the first user. Therefore, given β_1, we have

$$\text{var}[x_r'] = \sigma_{x_r'}^2 = \text{var}\left[\widetilde{N}_r'|\beta_1\right] = \lambda_r\text{var}\left[\widetilde{N}_r\right], \quad (33)$$

where the coefficient λ_r can be considered as a degradation factor due to the inverse matrix operation. If we use random instead of Walsh-Hadamard code sequences, we will get λ_r simply as

$$\lambda_r = \sum_{i=1}^{R}(\nu_{ri})^2. \quad (34)$$

From the simulation results to be presented in Section V, λ_r is a r. v., has a mean of approximately 1, and a small variance if R is chosen to be less than 8 and the number of subcarriers M is larger enough.

Finally, defining γ_r to be the conditioned SNR for the r-th coded bit before decoding and using the above results, we get

$$\gamma_r = \frac{\left(\sqrt{\varepsilon_1}c_{1,r}^0\right)^2}{\text{var}\left[x_r'\right]} = \frac{E_bTS^2}{2\lambda_r\sigma_{x_r}^2}. \quad (35)$$

C. Viterbi Decoder and BER Upper Bound

We employ Gaussian approximation in the following analysis to derive the bit error rate (BER) of the proposed system. As discussed above, conditioned on β_1, the signal x_r in Fig. 2 is given by a sum of uncorrelated and independent Gaussian random variables. Hence, it is also Gaussian, and so is x_r'. Specifically, conditioned on β_1, x_r' is $N(\pm\sqrt{\varepsilon_1}, \lambda_r\sigma_{x_r}^2)$ as the number of users $K \to \infty$. The r-th convolutionally coded bit determines the $\pm$ sign.

Assume that $T(D_1, D_2, \cdots, D_R, B)$ is the transfer function of the convolutional encoder as defined in [12], in which the label D_i is used to denote the accumulated Hamming distance of the i-th code symbol and B is used to indicate whether the signal branch corresponds to a decoder bit error. Following the same method given in [12], after Viterbi decoding, the BER conditioned on β_1, can be upper bounded as

$$P(e|\beta_1) < \left.\frac{\partial T(D_1, D_2, \cdots, D_R, B)}{\partial B}\right|_{B=1, D_r=\xi_r, r=1,\cdots,R}, \quad (36)$$

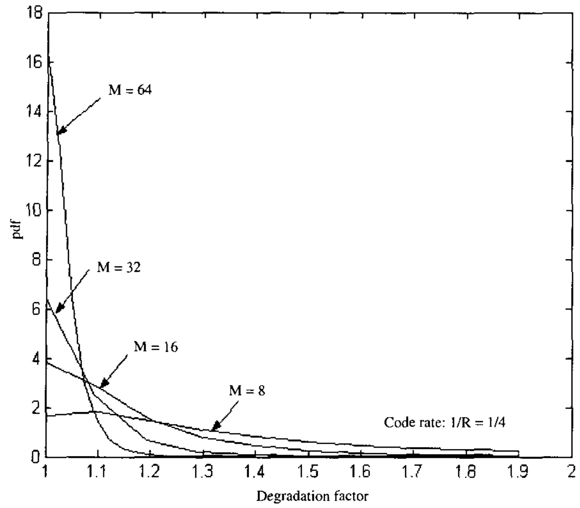

Fig. 3. Pdf of degradation factor (obtained by simulation).

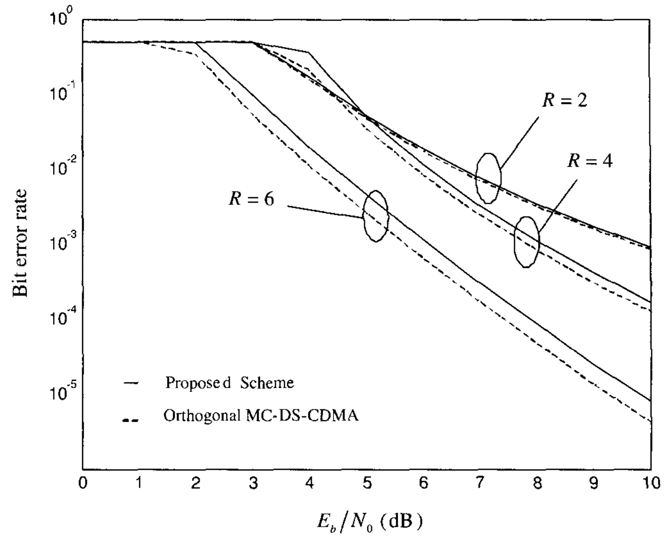

Fig. 4. BER upper bound ($M = 64$, $N = 32$, and $K = 100$).

where $\xi_r = (1 + \gamma_r)^{-1}$. Finally, the BER upper bound is given by

$$P_e = \int_0^{\infty} P(e|\beta_1)p(S)dS, \tag{37}$$

where $p(S)$ is the pdf of r. v. S.

V. NUMERICAL RESULTS

A. System Parameters

We assume perfect power control so that the received signal powers for all the users are equal. We make use of maximal-length PN spreading sequences with periods exceeding N or N_1, where N and N_1 are the processing gains per subcarrier in the proposed and the orthogonal MC-DS-CDMA scheme, respectively. In our simulation, the PN generator is chosen with a length of 255 and a primitive polynomial of $f(x) = 1 + x^2 + x^3 + x^4 + x^8$.

As discussed in [7], the system performance for the orthogonal MC-DS-CDMA system will depend on the spreading code sequences chosen for the users. This will also be the case for our proposed system. But since our aim is not to search for the best code, we have simply used some random sequences to be the spreading codes in this paper.

Only four simple convolutional encoders are considered. The generator polynomials (in octal) for these convolutional coders are $G_1 = (5,7)$, $G_2 = (5,7,7,7)$, $G_3 = (7,7,7,7,5,5)$, and $G_4 = (7,7,5,5,5,7,7,7)$ for code rate $1/R = 1/2, 1/4, 1/6$, and $1/8$, respectively, where the same constraint length of 3 is used [14].

B. Degradation Factor

It is obvious that λ_r, the degradation factor due to the use of (23), is a r. v. that depends on the matrix $\boldsymbol{U}$. Since the pdf of λ_r is difficult to derive mathematically, Monte Carlo simulation is used to obtain this pdf. In the simulation, we generated $1,000,000$ realizations of $\boldsymbol{U}$ using independent sets of Rayleigh fading channel parameters $\beta_1 = \{\beta_{1,q}, q = 1, \cdots, M\}$. Then, the inverse matrix $\boldsymbol{U}^{-1}$ and λ_r are calculated. The measured pdf of λ_r is shown in Fig. 3. It is easy to see that λ_r fall in a narrow range near 1 with a high probability as M increases. As an example, for $M = 32$ and $R = 4$, we have the probability of $P(1 < \lambda < 1.2) > 98\%$. This means that λ_r is practically equal to 1, and the performance degradation due to SIC processing in (23) and the resulting noise coloring is negligible.

C. BER Performance

As mentioned above, we assume that the EGC detection strategy given by (17) is used. The BER results are obtained from (37) with the pdf $p(S)$ obtained numerically from $1,000,000$ realizations of S. In Fig. 4, we plot the BER upper bound from (37) against E_b/N_0 for R equal to 2, 4, and 6. The number of users (K) is 100, with $M = 54, N = 32$, and an encoder with rate $1/R$ is used. We can see that increasing the number of multicodes R gives a better performance for both our proposed scheme and the orthogonal MC-DS-CDMA system of [7]. Generally, the orthogonal MC-DS-CDMA system slightly outperforms our proposed scheme. This is because the use of inverse matrix transform $\boldsymbol{U}^{-1}$ in our proposed scheme results in a slight increase in the total interference. The difference of performance between the two systems gradually vanishes as R decreases.

Next, we consider the effect of the number of subcarriers M. In Fig. 5, for the case of $R = 2$ and $K = 100$, we plot the BER upper bound versus E_b/N_0, with M taking a value of 8, 16, or 32, respectively. We can see that the BER performance of our proposed scheme is slightly improved as M increases. This is because the degradation due to the inverse matrix transformation $\boldsymbol{U}^{-1}$ decreases as more subcarriers are used. However, the performance of the orthogonal MC-DS-CDMA system of [7] is degraded slightly as M increases. A reasonable explanation is that the larger the number of subcarriers, the larger the interference between subcariers for that system due to the loss of orthogonality from fading. Note that the BER performance of

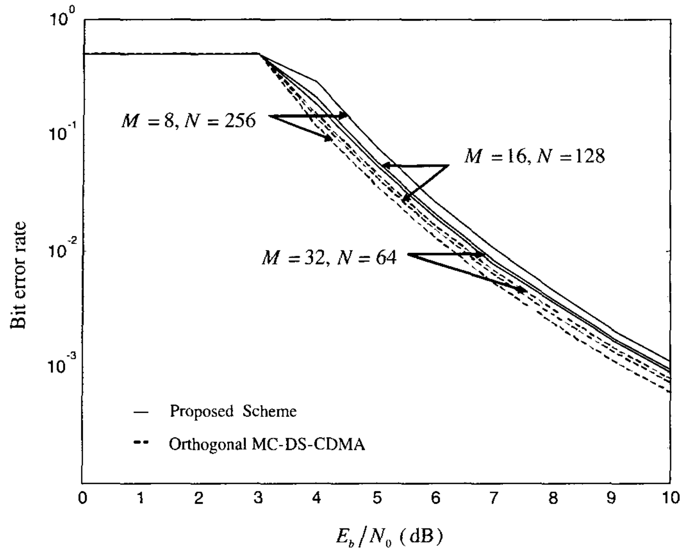

Fig. 5. BER upper bound ($R = 2$ and $K = 100$).

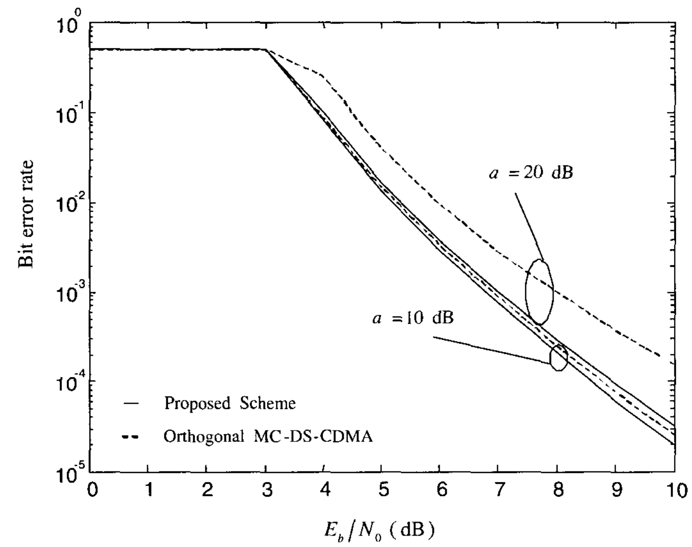

Fig. 6. BER upper bound ($M = 64$, $N = 32$, $R = 4$, and $K = 100$).

both systems is almost the same if the number of subcarriers is large, say, 32.

In Fig. 6, we plot the BER performance of the two systems in the presence of fading and a strong single tone jamming. The latter has the same frequency as the 32nd subcarrier, and a strength α of 10 or 20 dB. The number of users is 100, with $M = 64$ and $R = 4$. We can see that our proposed system has a better BER performance than the orthogonal MC-DS-CDMA system of [7] in the presence of strong jamming. This is because the length of the PN code in the orthogonal MC-DS-CDMA system will be decreased by a factor of R if a convolutional encoder with a rate $1/R$ is used. On the other hand, the length of the PN code in our proposed system is independent of the code rate.

Finally, we compare the BER performance obtained from both the upper bound analysis and simulation in Fig. 7. An independent Rayleigh fading channel without jamming is assumed, with $K = 10$, $M = 8$, and $R = 2$. The length of the pseudo random sequences for our proposed system is 16, while that for the orthogonal MC-DS-CDMA system is 8. From this figure, we can see that the simulation result is consistent with the upper bound, especially at high E_b/N_0.

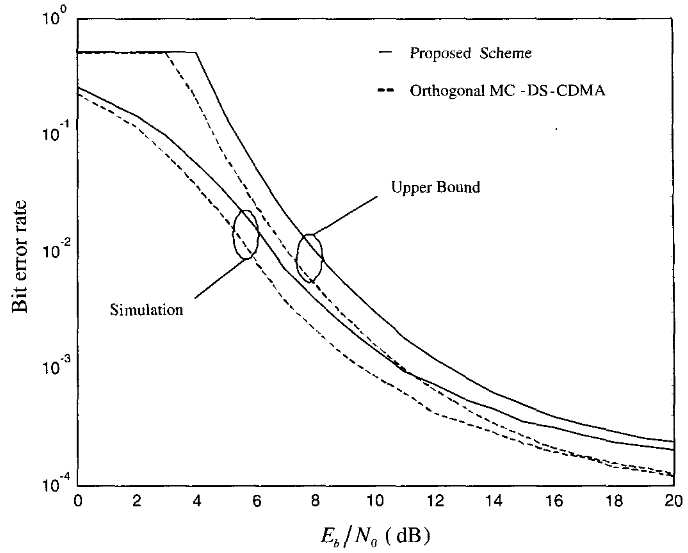

Fig. 7. BER upper bound and simulation results ($M = 8$, $N = 16$, $R = 2$, and $K = 10$)

D. Discussions

As discussed in Section II-C, we can use a PN sequence that is R times longer than the PN sequence used in the system in [7] because of the use of multicodes in our proposed scheme. In the meantime, we keep the same overall system bandwidth whatever the data rate is. This is why our proposed MM-DS-CDMA system outperforms the conventional multicarrier CDMA system in [7]. On the other hand, the length of PN sequence can also be easily adjusted by combining two conventional techniques: The conventional multicode and the conventional variable spreading gain codes [21]. However, if this approach is taken, the longer the PN sequence is, the smaller will the real data transmission rate that can be supported be. Alternatively, the longer the PN sequence is, the wider will the needed transmission bandwidth be, if the data rate is kept unchanged. It is clear that it is worthwhile to make a comparison between our scheme and the conventional variable spreading coded systems. Unfortunately, it is difficult to make this comparison, because the data transmission rate in the conventional variable spreading coded systems must be changed as the spreading gain is adjusted, while in our proposed scheme, the spreading gain (that is, the length of the PN sequence) has nothing to do with the data rate.

In addition, in our proposed scheme, the length of the PN code will increase as the code rate decreases. In contrast, the length of the PN sequence will not change with the code rate in an orthogonal MC-DS-CDMA system. Or, at the same code rate, the length of the PN code in our proposed scheme will be larger than that in an orthogonal MC-DS-CDMA system. This will lead to a better bit synchronization performance because a higher correlation peak can be realized. Finally, requirements such as high- and variable-rate transmission in the next generation of CDMA system can be more easily satisfied with our proposed system. This is because the symbol duration T in our proposed scheme is R times longer than the symbol duration T_1 in an orthogonal MC-DS-CDMA system and can be changed easily by choosing the code rate.

VI. CONCLUDING REMARKS

In this paper, we have proposed and analyzed a new multicarrier, direct sequence code division multiple access (MC-DS-CDMA) system, which may be suitable for realizing future wide-band CDMA system. It is shown that the proposed scheme has almost the same BER performance as that of an orthogonal MC-DS-CDMA system, but has better anti-jamming ability. In addition, due to the use of longer PN code, the proposed scheme may have better performance in bit synchronization and may be able to support high- and variable-rate transmission more easily.

ACKNOWLEDGMENTS

The authors thank the anonymous reviewers for their excellent feedback, which has been used to improve our works.

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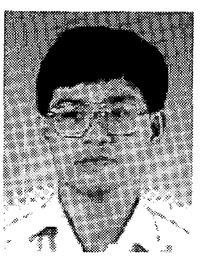

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