Performance Analysis of Time-Hopping Spread-Spectrum Multiple-Access Systems: Uncoded and Coded Schemes

Amir R. Forouzan, *Student Member, IEEE*, Masoumeh Nasiri-Kenari, *Member, IEEE*, and Jawad A. Salehi, *Member, IEEE*

***Abstract*—In [Scholtz (1993)], an ultra-wide bandwidth time-hopping spread-spectrum code division multiple-access system employing a binary PPM signaling has been introduced, and its performance was obtained based on a Gaussian distribution assumption for the multiple-access interference. In this paper, we begin first by proposing to use a practical low-rate error correcting code in the system without any further required bandwidth expansion. We then present a more precise performance analysis of the system for both coded and uncoded schemes. Our analysis shows that the Gaussian assumption is not accurate for predicting bit error rates at high data transmission rates for the uncoded scheme. Furthermore, it indicates that the proposed coded scheme outperforms the uncoded scheme significantly, or more importantly, at a given bit error rate, the coding scheme increases the number of users by a factor which is logarithmic in the number of pulses used in time-hopping spread-spectrum systems.**

***Index Terms*—CDMA, low-rate convolutional codes, spread-spectrum techniques, super-orthogonal codes, time-hopping, ultra-wide bandwidth radio.**

I. Introduction

IN [1], an ultra-wide bandwidth time-hopping spread-spectrum code-division multiple-access system (UWB-TH-CDMA) employing a binary pulse-position modulation (PPM) signaling have been introduced. In this system, data is transmitted using extremely short pulses with duration less than 1 ns. This technique is called impulse radio (IR) and since the transmitted pulses are extremely short, the bandwidth of this system is a few hundred times larger than the bandwidth of other systems for the same applications. This communication system does not use sinusoidal carriers to raise the signal to higher frequencies, and in fact its frequency band is from about dc to several gigahertz. The advantages of this spread-spectrum multiple-access system are briefly power consumption, cost and complexity reductions.

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A. R. Forouzan was with the Department of Electrical Engineering, Sharif University of Technology, Tehran, Iran. He is now with the Institute of Electronics, Faculty of Engineering, University of Tehran, Tehran, Iran (e-mail: a_r_forouzan@yahoo.com).

M. Nasiri-Kenari and J. A. Salehi are with the Department of Electrical Engineering, Sharif University of Technology, Tehran, Iran (e-mail: mnasiri@sharif.edu; jasalehi@sharif.edu).

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The capability of the system to highly resolve the multipaths with differential delays on the order of 1 ns or less, and its ability to penetrate materials makes the UWB system viable for high-quality, fully mobile short-range indoor radio communications [2]. The receiver processing and performance prediction for both analog and digital modulator under an ideal multiple-access channel (without multipath fading) have been investigated in [2] and [3]. Real indoor channel measurements and the system robustness in dense multipath environments have been reported in [4]. For more details on UWB-TH-CDMA systems, see [2]–[5]. In this paper, however, we focus on an ideal multiple-access channel, i.e., an additive white Gaussian noise (AWGN) channel without multipath fading effects.

Pervious studies on the performance of the UWB-TH-CDMA considered Gaussian distribution for the multiuser interference in the system. Under this assumption, the system has a capability to support a relatively very high total transmission rate (or equivalently a very large number of users at a given fixed bit rate for each user) using the well-known single-user correlator receiver in an AWGN channel [1]–[3]. In order to verify and justify the results, an exact analysis without the Gaussian assumption is required. On the other hand, the exact performance analysis of the system, in general, could prove to become a cumbersome and an unwieldy task. In this paper, we attempt to present a more accurate analysis for the system performance and compare the results with those using Gaussian assumption for various cases.

To obtain a more accurate analysis for asynchronous UWB-TH-CDMA, we first begin our analysis for a more simplified system configuration, namely, synchronous UWB-TH-CDMA system. For this simplified case we calculate the exact bit-error rate (BER) and compare the result with the case where the multiuser interference for this system is modeled as a Gaussian random variable. We will in fact show that the Gaussian assumption predicts accurately only when the number of pulse per bit gets large. Once we have developed insight into the mechanism of calculating the exact error rate for the synchronous case, we then relax the condition of synchronous configuration and develop a more precise error rate for an asynchronous UWB-TH-CDMA system.

Our more precise performance analysis indicates that at high bit rates, the Gaussian assumption substantially overestimates the number of users supported by the system. However, at low rates, the Gaussian assumption predicts error rates that are extremely close to the more precise analysis.

In this paper, we also show that in the UWB-TH-CDMA system as described in [1], not all of the system potentials are used. The system can achieve much higher capabilities using some parts of its spread-spectrum bandwidth expansion for channel coding. We propose to employ practical low-rate superorthogonal codes and study their performance in the context of the described system. These codes are near optimal codes with a rate $1/n$ and are suitable for spread-spectrum systems. Our performance analysis indicates that despite of its relatively low complexity in a time-hopping spread-spectrum system (and also in a UWB-TH-CDMA), the novel proposed TH-CDMA system combined with low-rate error control coding presents significant improvement compared with an ordinary uncoded TH-CDMA system and increases the number of supported users at a given BER by a factor equal to $(4+\log_2 N_s)/2$, where N_s is the number of pulses per bit used in TH-CDMA systems.

It should be noted that the coding scheme presented in this paper, in which no extra bandwidth is required further than needed by spread-spectrum modulation, has been previously introduced for DS-CDMA communication system [6], [7]. But in the best knowledge of the authors, this paper is the first one that proposes this coding scheme for TH-SSMA system and presents its performance analysis for an UWB multiple-access communication application. It must be also noted that since in the current application, the code rate is inversely decreased by the number of pulses transmitted per each input information bit, the scheme has very low complexity, and it is completely practical.

The rest of this paper is organized as follows. We present a brief description of the system for uncoded and coded schemes in Section II, then we develop an error performance analysis for these schemes, in Sections III and IV, respectively. We present the numerical results in Section V and, finally, we conclude this paper in Section VI.

II. System Description

A. Uncoded Scheme

Every transmitter sends N_s pulses for each data bit. These pulses are located apart in sequential frames, each with duration T_f. The modulation is binary pulse-position modulation (BPPM), in which the pulses corresponding to bit 1 are sent δ seconds later than the pulses corresponding to bit 0. The locations of the pulses in each frame are determined by a user dedicated pseudorandom sequence. The transmitted signal of user k is

$$s_{tr}^{(k)}(t)=\sum_j w_{tr}\left(t-jT_f-c_j^{(k)}T_c-\delta d_j^{(k)}\right) \qquad (1)$$

where the index j indicates the frame number, $w_{tr}(t)$ represents the transmitted pulse, and $\{c_j^{(k)}\}$ is the dedicated pseudorandom sequence for the user k with integer components. The integer number can take on any values between zero to N_h-1. In the equation, T_c indicates chip duration and satisfies $N_hT_c \le T_f$, and $\{d_j^{(k)}\}$ is the binary sequence of the transmitted symbols corresponding to user k. For the uncoded systems (the scheme presented in [1]), this sequence is N_s repetitions of the transmitted data sequence, i.e., if the transmitted binary data sequence is $\{D_i^{(k)}\}$, then we have $d_j^{(k)}=D_i^{(k)}$ for $iN_s \le j < (i+1)N_s$. We can consider the above-uncoded scheme as a coded scheme with the simple repetition block code of rate $1/N_s$. Since N_s frames are sent by the transmitter during each data bit, transmission rate R_s will be $R_s=(N_sT_f)^{-1}$.

We assume a free-space propagation channel with AWGN. However, the antenna system modifies the shape of the transmitted pulse $w_{tr}(t)$ at the output of the receiver's antenna [2]. In this case, the received signal of the kth user at the receiver antenna output is

$$s_{rec}^{(k)}(t)=\sum_j w_{rec}\left(t-jT_f-c_j^{(k)}T_c-\delta d_j^{(k)}\right) \qquad (2)$$

where $w_{rec}(t)$ is the received pulse with duration T_w, i.e., $w_{rec}(t)$ is zero out of the time duration [0, T_w]. The total received signal is

$$r(t)=\sum_{k=1}^{N_u} A_k s_{rec}^{(k)}(t-\tau_k)+n(t) \qquad (3)$$

where N_u is the number of active users, and A_k and τ_k are the channel attenuation and delay, respectively, corresponding to user k and $n(t)$ is the received noise.

In the following, we briefly present the receiver structure for an uncoded system. We assume that synchronization between the desired transmitter and receiver is established prior to data transmission. Without any loss of generality, we consider the desired user to be user 1. Then, for the uncoded scheme, the correlating receiver 1 decides based on the following rule [2], [3] as shown in (4) at the bottom of the page, where $v(t)$ is called the receiver's template signal and is defined by $v(t)\triangleq w_{\rm rec}(t)-w_{\rm rec}(t-\delta)$. Since $w_{\rm rec}(t)$ has duration T_w, this is evident that the receiver's template signal $v(t)$ has duration $T_w+\delta$. Here, N_s pulse correlator outputs (α_j) are added to make the test statistic α. Then, α is compared with zero. As we will discuss later, these pulse correlator outputs (α_j) are the basic elements of the decision process in our proposed coded scheme, as well.

$$\text{"decide that } D_0^{(1)}=0\text{"} \Leftrightarrow \underbrace{\sum_{j=0}^{N_s-1}\overbrace{\int_{\tau_1+jT_f}^{\tau_1+(j+1)T_f} r(t)v(t-\tau_1-jT_f-c_j^{(1)}T_c)dt}^{\text{pulse correlator output}\triangleq\alpha_j}}_{\text{test statistic}\triangleq\alpha}>0 \qquad (4)$$

B. Coded Scheme

As mentioned in the previous section, we can consider the above simple time-hopping spread-spectrum as a coded system in which a simple repetition block code with rate $1/N_s$ is used. As it is well-known, the repetition code is not a good code. Thus, we expect by applying a near optimal code instead of the above simple repetition code, the system performance will improve significantly. In [8], a class of low-rate superorthogonal convolutional codes that have near optimal performance is introduced.

In a superorthogonal code with constraint length K, the rate is equal to $1/2^{K-2}$. Since in the TH-CDMA (UWB) system N_s pulses are sent for each data bit, we must set $2^{K-2} = N_s$ or $K = 2 + \log_2 N_s$. The location of each pulse in each frame is determined by the user dedicated pseudorandom sequence along with the code symbol corresponding to that frame.

Decoding is performed using Viterbi algorithm. The state diagram of this decoder consists of 2^{K-1} states. Two branches, corresponding to bit zero and bit one, exit from each state in the trellis diagram. To update the state metrics, it is first necessary to calculate the branch metrics, using the received signal $r(t)$. For this purpose, in each frame j the quantity $\alpha_j \triangleq \int_{\tau_1+jT_f}^{\tau_1+(j+1)T_f} r(t)v(t-\tau_1-jT_f-c_j^{(1)}T_c)dt$, which is called pulse correlator output [see (4)], is obtained. Because of special form of the Hadamard–Walsh sequence that is used in the structure of superorthogonal codes, the branch metrics can be simply evaluated based on the outputs of pulse correlators $\alpha_j s$ [8]. The processing complexity of this decoder grows only linearly with K (or logarithmic with N_s); the required memory, however, grows exponentially with K (or equally linearly with N_s) [8]. Since in time-hopping spread-spectrum application, the value of K is relatively low (the typical value is in the range 3 – 12), the system can be considered to be completely practical.

III. Performance Evaluation (Uncoded Scheme)

In this section, we evaluate the system performance for both synchronous and asynchronous uncoded schemes. In this study, the BER is obtained as a function of the number of users with a given transmission bit rate. In this and also in the following sections, we assume no near–far problem (i.e., for all k, $A_k = A_1 = 1$), and we neglect the effect of AWGN $n(t)$. The first assumption enables us to derive the BER as a function of the number of users N_u. However, under the second assumption, we can determine the maximum achievable multiple-access capability of the system. Furthermore, we assume that the elements $\{c_j^{(k)}\}$, for $k = 1, 2, \ldots, N_u$ and for all j, are independent, identically distributed (iid) random variables with a uniform distribution on [0, $N_h - 1$].

For all the cases considered in our studies here, we compute the probability density function (pdf) for the total interference signal due to the undesired users. Since the users transmit their data independently, the pdf for the total interference is the convolution of the interference pdf of each user. Furthermore, under the assumption of no near-far problem, the interference pdf of different users are identical and equal. Therefore, it is sufficient to determine the pdf of the interference caused by only one user.

A. Synchronized Users

The term synchronized users means that for all k, $\tau_k = \tau_1$, see (3). In the following, we first derive the exact BER; we then provide the BER based on a Gaussian distribution assumption for multiuser interference, and then compare the results.

Exact Analysis: The approach we have elected to use in order to obtain the total interference pdf is first by computing the total probability characteristic function and then inverse transform to evaluate the desired pdf. Since the effect of different users can be modeled as iid random variables, then it suffices to obtain the probability characteristic function associated with one interfering user and then raise the resulted characteristic function to the power of the number of interfering users to obtain the total probability characteristic function. The probability characteristic function of the kth interfering user at the output of the desired user 1 receiver can be expressed as

$$\begin{aligned} T(z) =& \text{Pr}(\text{user } k \text{ sends } 0)P(z) + \text{Pr}(\text{user } k \text{ sends } 1)Q(z) \\ =& \frac{1}{2}P(z) + \frac{1}{2}Q(z) \end{aligned} \tag{5}$$

where $P(z)$ and $Q(z)$ are the probability characteristic functions of the interference conditioned on the transmitted bit of user k, being zero and one, respectively. We first compute $P(z)$. In this case, the received signal of user k at frame j is equal to $w_{\text{rec}}(t-\tau_k-jT_f-c_j^{(k)}T_c)$; see (2) and (3). From (4), the effect of this signal at the output of the jth pulse correlator is shown in (6) at the bottom of the page. For the synchronous case where $\tau_k = \tau_1$, we obtain $I_j = \int_0^{T_f} w_{\text{rec}}(t-c_j^{(k)}T_c)v(t-c_j^{(1)}T_c)dt$. Since outside the time interval $[c_j^{(1)}T_c, c_j^{(1)}T_c+T_w+\delta]$, $v(t-c_j^{(1)}T_c)$ is zero, we have $I_j = \int_0^{T_w+\delta} w_{\text{rec}}(t-c_j^{(k)}T_c+c_j^{(1)}T_c)v(t)dt$. We assume $T_c \geq T_w+\delta$, so when $c_j^{(k)} \neq c_j^{(1)}$, $w_{\text{rec}}(t-c_j^{(k)}T_c+c_j^{(1)}T_c)$ will be zero in the time interval $[0, T_w+\delta]$. Thus, we have

$$I_j = \begin{cases} 0, & c_j^{(k)} \neq c_j^{(1)} \\ \int_0^{T_w+\delta} w_{\text{rec}}(t)v(t)dt, & c_j^{(k)} = c_j^{(1)}. \end{cases} \tag{7}$$

We define $m_p \triangleq \int_0^{T_w+\delta} w_{\text{rec}}(t)v(t)dt = \int_{-\infty}^{+\infty} w_{\text{rec}}(t)v(t)dt$. m_p is an important parameter for the system in consideration, and in fact, it indicates the contribution of the desired user signal at each pulse correlator output when the desired user sends bit 0. Since the elements $c_j^{(k)}$ and $c_j^{(1)}$ are assumed iid on [0, $N_h - 1$], the probability of $c_j^{(1)} = c_j^{(k)}$ is $\beta = 1/N_h$ and the probability of $c_j^{(1)} \neq c_j^{(k)}$ is $\alpha = 1 - \beta = 1 - 1/N_h$. Thus, the pdf of

$$I_j = \int_{\tau_1+jT_f}^{\tau_1+(j+1)T_f} w_{rec}(t-\tau_k-jT_f-c_j^{(k)}T_c)v(t-\tau_1-jT_f-c_j^{(1)}T_c)dt. \tag{6}$$

the interference on frame j, conditioned on the transmitted bit being zero, is

$$f_{I_j|0}(x) = \alpha\delta_D(x) + \beta\delta(x - m_p) \tag{8}$$

where $\delta_D(\cdot)$ is Dirac delta function, and from (8), the probability characteristic function of an interference on frame j is $\Phi_{I_j|0}(z) = \alpha + \beta z^{m_p}$, where $z = e^s$. Since the elements of $\{c_j^{(k)}\}$ are iid random variables, the probability characteristic function due to kth user's interfering signal over one bit time duration conditioned on the transmitted bit being zero is equal to $P(z) = (\Phi_{I_j|0}(z))^{N_s} = (\alpha + \beta z^{m_p})^{N_s}$, where N_s is the number of frames in each bit time interval. The conditional probability characteristic function $Q(z)$ can be evaluated in much the same way as $Q(z) = (\alpha + \beta z^{-m_p})^{N_s}$. Substituting $P(z)$ and $Q(z)$ into (5), we obtain

$$T(z) = \frac{1}{2}(\alpha + \beta z^{m_p})^{N_s} + \frac{1}{2}(\alpha + \beta z^{-m_p})^{N_s}. \tag{9}$$

Then, the total probability characteristic function at the output of the correlator receiver due to all users is

$$\begin{aligned} R(z) =& (T(z))^{N_u-1} \\ =& 2^{1-N_u}\left[(\alpha + \beta z^{m_p})^{N_s} + (\alpha + \beta z^{-m_p})^{N_s}\right]^{N_u-1}. \end{aligned} \tag{10}$$

Using the binomial expansion, we have (11) at the bottom of the page. By letting $k = m - n$ in the above equation and subsequently taking the inverse Z transform of (11), the pdf of the total interference due to all remaining users is computed as

$$\begin{aligned} p(x) = \left(\frac{\alpha^{N_s}}{2}\right)^{N_u-1} \sum_k \delta(x - km_p) \sum_{l=0}^{N_u-1} \binom{N_u - 1}{l} \\ \times \sum_{m=0}^{N_s l} \binom{N_s l}{m} \binom{N_s(N_u - 1 - l)}{m - k} \left(\frac{\beta}{\alpha}\right)^{2m-k}. \end{aligned} \tag{12}$$

In the above expression $\binom{s}{r}$ is equal to zero for $s < r$ or $r < 0$. These cases occur when $n = m - k$ is not in the summation range on n in (11).

Without loss of generality, we assume that the desired user transmits bit 0. As we have mentioned previously, the desired user's signal effect on each pulse correlator output is m_p. Since there are N_s frames during a bit time interval, the output of the correlator is equal to $m_p N_s$ plus the interference and noise terms. From (4), the bit error probability P_b is equal to the probability that the correlator output is less than zero. By neglecting the AWGN term, a lower bound on the probability of error P_b is obtained, and it is equal to the probability that the interference is less than $-m_p N_s$. Thus

$$P_b = \int_{-\infty}^{-m_p N_s} p(x)dx = \frac{1}{2}p_{dis}(-N_s) + \sum_{k=-N_s(N_u-1)}^{-N_s-1} p_{dis}(k) \tag{13}$$

where

$$\begin{aligned} p_{dis}(k) = \left(\frac{\alpha^{N_s}}{2}\right)^{N_u-1} \sum_{l=0}^{N_u-1} \binom{N_u - 1}{l} \\ \times \sum_{m=0}^{N_s l} \binom{N_s l}{m} \binom{N_s(N_u - 1 - l)}{m - k} \left(\frac{\beta}{\alpha}\right)^{2m-k}. \end{aligned} \tag{14}$$

Performance Analysis Under Gaussian Assumption: If we assume that the distribution of the interference at the output of the correlator is Gaussian, the BER can be easily evaluated. In this case, the BER can be written as

$$P_b = Q\left(\frac{\eta}{\sqrt{\sigma_n^2}}\right) \tag{15}$$

where $\eta = m_p N_s$ is the mean, and

$$\sigma_n^2 = m_p^2(N_u - 1)N_s\beta(N_s\beta + \alpha) \tag{16}$$

is the variance of the sum of N_s pulse correlator outputs distribution, assuming N_u active users and neglecting the AWGN term (see Appendix A).

B. Unsynchronized Users

In most applications, for instance in a mobile wireless uplink, the system is considered to be asynchronous. This implies that, the time delays τ_k for $k = 1, 2, \ldots, N_u$ are mutually random and iid with $\tau_k \bmod T_f$ being uniformly distributed on the interval [0, T_f]. In such cases, the exact BER calculation is usually cumbersome and unwieldy. In this section, however, we derive the BER of the above asynchronous system with some minor deviation from the exact analysis which will result in an excellent approximation. To compute the pdf of the multiuser interference term at the receiver output, for the BER calculation, it is first required to derive the pdf of the interference term due to an interfering user k ($k \neq 1$). To this end, it is necessary to calculate the interference of the user at the pulse correlator output. Since the frame duration T_f is much greater than the receiver's template signal ($v(t)$) duration, i.e., $T_w + \delta$, we assume that only one pulse from each interfering user in each frame is contributed to the interference term at the output of the desired user's receiver, i.e., user 1. If this interfering pulse has a delay τ with respect to the desired user's receiver template signal $v(t)$, the interference made by this pulse can be written as $X = \int_0^{T_w+\delta} v(t)w_{rec}(t - \tau)dt$. According to the earlier assumptions, τ is a random variable with a uniform distribution on an interval with duration T_f containing [$-T_w$, $T_w + \delta$]. Based on the definition of the pdf $f_X(x) = \lim_{\Delta x \to 0}(\Pr(x \le X < x + \Delta x))/(\Delta x)$, we have computed the pdf of the variable X numerically for a received waveform shape as depicted as in Fig. 1, and it is plotted in Fig. 2. As it can be seen, there is an strong impulse at the origin, which means that the probability of the interference being zero is very high and the absolute value of the interference due to

$$R(z) = 2^{1-N_u} \sum_{l=0}^{N_u-1} \binom{N_u - 1}{l} \sum_{m=0}^{N_s l} \sum_{n=0}^{N_s(N_u-1-l)} \binom{N_s l}{m} \binom{N_s(N_u - 1 - l)}{n} \alpha^{N_s(N_u-1)-m-n}\beta^{m+n}z^{m_p(m-n)}. \tag{11}$$

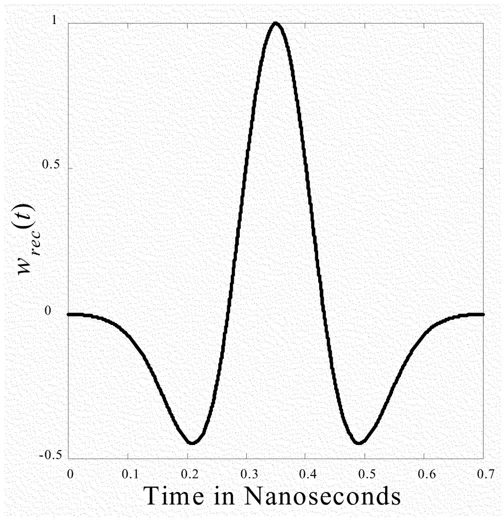


Fig. 1. Received pulse.

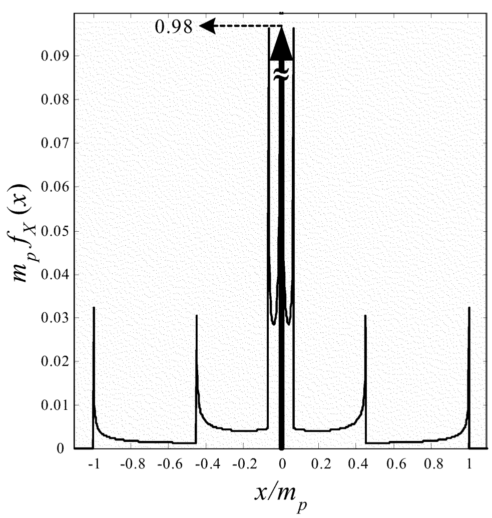


Fig. 2. PDF of interference corresponding to one interfering user.

each interfering user at each pulse correlator output is at most m_p.

The overall interference due to an interfering user at the correlator output is the sum of its interference at each pulse correlator output at each frame [see, (4)]. We assume that the interferences due to each pulse occurring at different frames are independent. Then, the pdf of the overall interference from each interfering user at the receiver output is

$$f_X^{N_s}(x) = \underbrace{f_X(x) * f_X(x) * \cdots * f_X(x)}_{N_s \text{ times}}.$$

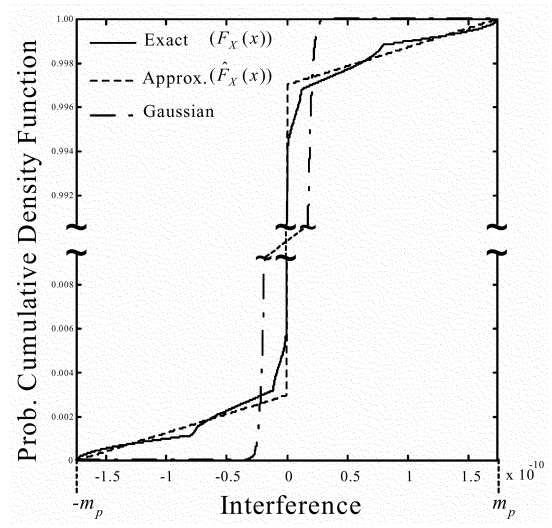


Fig. 3. Cumulative density function of interference in one frame corresponding to one interfering user.

The pdf of the total interference due to all users in the system is

$$f_X^{N_s(N_u-1)}(x) = \underbrace{f_X(x) * f_X(x) * \cdots * f_X(x)}_{N_s(N_u-1) \text{ times}} \tag{17}$$

where N_u is the number of active users. Due to the complicated shape of $f_X(x)$ (Fig. 1), the exact calculation of the expression is cumbersome and unwieldy. To circumvent this difficulty, we use the following relatively good approximation:

$$\hat{f}_X(x) = \alpha \delta_D(x) + \beta\left(u(x+m_p) - u(x-m_p)\right) \tag{18}$$

where $u(\cdot)$ is the unit step function. The parameters α and β are selected such that the variance and the mean of the interference do not change. We denote by σ_a^2 to be the variance of the interference contributed by only one interfering pulse on a frame, and it is easily computed as shown in (19) at the bottom of the page. On the other hand, $\sigma_a^2 = \int_{-\infty}^{+\infty} x^2 \hat{f}_X(x)\,dx = \int_{-m_p}^{m_p} \beta x^2 dx = 2\beta m_p^3/3$, thus, $\beta = 3\sigma_a^2/(2m_p^3)$. Furthermore, $\int_{-\infty}^{+\infty} \hat{f}_X(x)\,dx = \alpha + 2\beta m_p = 1$; therefore, $\alpha = 1 - 3\sigma_a^2/m_p^2$.

For all practical purposes, the proposed approximation has a high precision, especially if we plot the exact and approximated cdf's of interference in a frame and compare it with the cdf resulted by the Gaussian assumption. In Fig. 3, the interference cdf $F_X^n(x)$ and its approximation $\hat{F}_X^n(x)$ and the cdf of a Gaussian function with the same variance is plotted for $n = 1$. It can be observed that the proposed approximation leads to a suitable result and estimates considerably better than the Gaussian assumption the effect of an interfering signal. This is because the

$$\begin{aligned}\sigma_a^2 &= \int_{\text{an interval containinig } [-T_w,\, T_w+\delta]} \frac{1}{T_f}\left[\int_0^{T_w+\delta} v(t) w_{rec}(t-\tau)dt\right]^2 d\tau \\ &= T_f^{-1} \int_{-\infty}^{+\infty} \left[\int_{-\infty}^{+\infty} v(t) w_{rec}(t-\tau)dt\right]^2 d\tau. \end{aligned} \tag{19}$$

probability of interference signal being zero is very high, and it is well captured in our proposed model where it is not well captured or considered in the Gaussian model. Furthermore, according to Fig. 1, excluding a few points, the interference pdf is near a special value in the interval $[-m_p, m_p]$ and is zero outside this interval which are both considered in our model but not in a Gaussian model.

In Appendix B, the nth-order convolution of $\hat{f}_X(x)$ has been derived. The result is as follows :

$$\hat{f}_X^n(x) = \alpha^n \delta_D(x) + \sum_{i=1}^{n} \binom{n}{i} \alpha^{n-i} \beta^i \frac{m_p^{i-1}}{(i-1)!} \times \sum_{j=0}^{i} (-1)^j \binom{i}{j} \left(\frac{x}{m_p} + i - 2j\right)^{i-1} u\left(\frac{x}{m_p} + i - 2j\right). \quad (20)$$

Furthermore, using a similar method as described in [9], we may use a Gaussian approximation for the internal summation of the above equation for large values of i, with much less computation complexity, as follows:

$$\sum_{j=0}^{i} (-1)^j \binom{i}{j} \left(\frac{x}{m_p} + i - 2j\right)^{i-1} u\left(\frac{x}{m_p} + i - 2j\right) \approx 2^i (i-1)! m_p \frac{g\left(\frac{x}{s_i}\right)}{s_i} \quad (21)$$

where $g(x) = (2\pi)^{-0.5} \exp(-x^2/2)$ and $s_i = m_p (i/3)^{0.5}$.

By integrating $\hat{f}_X^n(x)$ in (20), we compute the probability cumulative density function (cdf) of the total interference as

$$\hat{F}_X^n(x) = \alpha^n u(x) + \sum_{i=1}^{n} \binom{n}{i} \alpha^{n-i} \beta^i \frac{m_p^i}{i!} \times \sum_{j=0}^{i} (-1)^j \binom{i}{j} \left(\frac{x}{m_p} + i - 2j\right)^{i} u\left(\frac{x}{m_p} + i - 2j\right) \quad (22)$$

where from (17), $n = N_s(N_u - 1)$. Similarly, the internal summation of this equation can be approximated for large values of i as

$$\sum_{j=0}^{i} (-1)^j \binom{i}{j} \left(\frac{x}{m_p} + i - 2j\right)^{i} u\left(\frac{x}{m_p} + i - 2j\right) \approx 2^i i! G\left(\frac{x}{s_i}\right) \quad (23)$$

where $G(x) = \int_{-\infty}^{x} g(\zeta) d\zeta$. Then, by neglecting the thermal noise term, the probability of bit error is equal to the probability that the receiver output is less than zero, conditioned on the desired user input bit being zero. Since the contribution of the desired user signal at the correlator output is $m_p N_s$, the BER is equal to

$$P_b \cong \hat{F}_X^{N_s(N_u-1)}(-m_p N_s). \quad (24)$$

In [1], the BER of the asynchronous case is derived based on a Gaussian distribution assumption for the total interference at the correlator output. The result is as follows:

$$P_b = Q\left(\frac{N_s m_p}{\sqrt{\sigma^2}}\right) = Q\left(\left(\frac{m_p^2}{\sigma_a^2} \frac{N_s}{N_u - 1}\right)^{1/2}\right) \quad (25)$$

where σ_a^2 is defined as in (19). The computation complexity of (24) is much higher than (25) which is based on Gaussian distribution assumption. However, in a practical system the product βm_p is in the order of 10^{-2} or less, and as a result, the external summation in (22) can be truncated up to moderate value of i. Moreover, (23) can be used for further computational complexity reduction.

IV. Performance Evaluation (Coded Scheme)

Since for a convolutional code only the upper and lower bounds on the BER using ML decoder are available, then for a convolutionally encoded TH-CDMA system, only the upper and lower bounds on the BER can be computed. To this end, the path generating function of the code is required. This function for a superorthogonal code is computed as [8]

$$T_{SO}(\gamma, \beta) = \frac{\beta W^{K+2}(1 - W)}{1 - W[1 + \beta(1 + W^{K-3} - 2W^{K-2})]} \quad (26)$$

in which $W = \gamma^{2^{K-3}}$, and K is the constraint length of the code. Expanding the above expression, we get a polynomial in γ and β. The coefficient and the powers of γ and β in each term of the polynomial indicate the number of paths and output–input path weights, respectively. Free distance of this code is obtained from the first term of the expansion as $d_f = 2^{K-3}(K+2) = N_s(\log_2 N_s + 4)/2$. If we consider the uncoded system presented in [1] as a coded scheme with a repetition code, its free distance will be N_s. Comparing the free distances of these two schemes, it is clear that our proposed coded scheme outperforms the scheme in [1] significantly.

An upper bound on the probability of error per bit for a memoryless channel is obtained using union bound as follows:

$$P_b < \left. \frac{dT_{SO}(W, \beta)}{d\beta} \right|_{\beta=1} = \frac{W^{K+2}}{(1 - 2W)^2} \left(\frac{1 - W}{1 - W^{K-2}}\right)^2 \quad (27)$$

where $W = Z^{2^{K-3}}$. The parameter Z is calculated from the Bhattacharyya bound as

$$Z = \int_{-\infty}^{+\infty} \sqrt{p_0(y) p_1(y)} dy \quad (28)$$

where $p_0(y)$ and $p_1(y)$ are the pdfs of the pulse correlator output conditioned on the input symbol being zero and one, respectively.

A lower bound on the probability of error per bit is obtained by considering only the first term of the path generating function (26). The result is as follows:

$$P_b \geq P_{d_f} \quad (29)$$

where P_{d_f} is the probability of pair wise error in favor of an incorrect path that differs in d_f symbols from the correct path over the unmerged span in the trellis diagram. Without any loss of generality, we assume that the input is all zero sequence, so P_{d_f} is the probability that the summation of d_f pulse correlator outputs is less than zero, when the corresponding input symbols are zero. (We assume that the interferences in these different frames are independent.)

In a binary symmetric Gaussian channels this lower bound states that

$$P_b \geq Q\left(\left(\frac{\eta^2}{\sigma^2}d_f\right)^{1/2}\right). \quad (30)$$

In the following, we compute the upper and lower bounds on the BER for both synchronous and asynchronous case.

A. Synchronized Users

At first, we must determine $p_0(y)$ and $p_1(y)$. If we set N_s equal to one in (10), we obtain the probability characteristic function of the pulse correlator output interference as

$$R^1(z) = 2^{1-N_u}\left[(\alpha+\beta z^{m_p})^1 + (\alpha+\beta z^{-m_p})^1\right]^{N_u-1}. \quad (31)$$

Similarly, if we set N_s equal to one in (12), the pdf of the interference at the pulse correlator output will be determined as

$$p^1(x) = \left(\frac{\alpha}{2}\right)^{N_u-1}\sum_k \delta_D(x-km_p)\sum_{l=0}^{N_u-1}\binom{N_u-1}{l} \times \sum_{m=0}^{l}\binom{l}{m}\binom{N_u-1-l}{n}\left(\frac{\beta}{\alpha}\right)^{2m-k}. \quad (32)$$

Since the effect of the desired user's signal at the pulse correlator output, conditioned on the input bit being zero and one, is m_p and $-m_p$, respectively, we simply have $p_0(y) = p^1(y-m_p)$ and $p_1(y) = p^1(y+m_p)$. Substituting $p_0(y)$ and $p_1(y)$ into (28) and using (27), we obtain an upper bound on the BER.

To determine the lower bound on the BER, the probability P_{d_f}, as defined in (29), must be first computed. The probability characteristic function of the interference on d_f frames, $d_f = N_s(\log_2 N_s + 4)/2$ is simply as follows:

$$\begin{aligned}R_{\text{Coded}}(z) &= \left(R^1(z)\right)^{N_s(\log_2 N_s+4)/2}\\ &= 2^{(1-N_u)N_s(\log_2 N_s+4)/2}\left[(\alpha+\beta z^{-m_p})\right.\\ &\quad \left.+(\alpha+\beta z^{m_p})\right]^{(N_u-1)N_s(\log_2 N_s+4)/2} \quad (33)\end{aligned}$$

where $R^1(z)$ is defined as in (31). Then, it can be shown that the pdf of the interference is

$$p(x) = \left(\frac{\alpha}{2}\right)^{(N_u-1)N_s(\log_2 N_s+4)/2}\sum_k \delta(x-km_p) \times \sum_{l=0}^{(N_u-1)N_s(\log_2 N_s+4)/2}\binom{\frac{(N_u-1)N_s(\log_2 N_s+4)}{2}}{l} \times \sum_{m=0}^{l}\binom{l}{m}\binom{\frac{(N_u-1)N_s(\log_2 N_s+4)}{2}-l}{m-k}\left(\frac{\beta}{\alpha}\right)^{2m-k}. \quad (34)$$

Since the summation of the signal effect at the output of $N_s(\log_2 N_s + 4)/2$ pulse correlators (conditioned on the input symbol being 0) is equal to $m_pN_s(\log_2 N_s + 4)/2$, a lower bound on P_b is obtained

$$P_b \geq P_{d_f} = \int_{-\infty}^{-m_pN_s(\log_2 N_s+4)/2} p(x)dx. \quad (35)$$

An improved upper bound for the probability of error per bit can be used, when the interference is assumed Gaussian. In binary symmetric Gaussian channels, the value of Z as defined in (28) is easily computed as $Z = \exp\left(-\eta^2/(2\sigma^2)\right)$, where η and σ^2 are the mean and variance of the pulse correlator output conditioned on the input symbol being zero.

Using the inequality $Q(\sqrt{x+y}) < Q(x)e^{-y/2}$ and some modifications, the improved upper bound for the binary symmetric Gaussian channels is obtained as [8]

$$\begin{aligned}P_b &< \exp\left(\frac{\eta^2}{2\sigma^2}d_f\right)Q\left(\left(\frac{\eta^2}{\sigma^2}d_f\right)^{1/2}\right) \times \frac{W^{K+2}}{(1-2W)^2}\left(\frac{1-W}{1-W^{K-2}}\right)^2\\ &= Q\left(\left(\frac{\eta^2}{\sigma^2}d_f\right)^{1/2}\right)\left(\frac{1-W}{(1-2W)(1-W^{K-2})}\right)^2. \quad (36)\end{aligned}$$

To use the upper bound (36), we only need to calculate the mean and variance of the pulse correlator output. The mean of the output distribution conditioned on the input bit being zero is m_p. The variance is obtained by substituting N_s by 1 in (16), i.e., $\sigma_n^2 = m_p^2(N_u-1)\beta(\beta+\alpha) = m_p^2(N_u-1)\beta$. Thus, under Gaussian assumption, we have

$$P_b < Q\left(\left(\frac{N_s(\log_2 N_s+4)}{2(N_u-1)\beta}\right)^{1/2}\right) \times \left(\frac{1-W}{(1-2W)(1-W^{K-2})}\right)^2 \quad (37)$$

where $W = \exp\left(N_s/(4(N_u-1)\beta)\right)$.

Similarly, using (29), the lower bound on the BER for a binary symmetric Gaussian channel is computed as

$$P_b \geq Q\left(\left(\frac{\eta^2}{\sigma^2}d_f\right)^{1/2}\right) = Q\left(\left(\frac{N_s(\log_2 N_s+4)}{2(N_u-1)\beta}\right)^{1/2}\right). \quad (38)$$

B. Unsynchronized Users

As in the previous section to obtain upper and lower bounds on the BER, we must first determine $p_0(y)$ and $p_1(y)$. Since, the interference of different users on a frame is independent, the pdf of the total interference on a frame caused by all $N_u - 1$ interfering users is $f_X^{N_u-1}(x)$. By using the approximated pdf $\hat{f}_X(x)$, as proposed in (18), the interference pdf on a frame is $\hat{f}_X^{N_u-1}(x)$, which can be calculated from (20) by setting $n = N_u - 1$. Then, it can be shown that $p_0(y) = \hat{f}_X^{N_u-1}(y-m_p)$ and $p_1(y) = \hat{f}_X^{N_u-1}(y+m_p)$. By substituting $p_0(y)$ and $p_1(y)$ in (27) and (28), we obtain the upper bound on the probability of error per bit.

The lower bound on the probability of error per bit is obtained by considering only the first term of the path generating function expansion of the code. Similar to previous section, it can be shown that the lower bound on the BER is equal to the probability that the interference be less than $-m_pN_s(\log_2 N_s+4)/2$. Thus, the lower bound on the BER is

$$P_b \geq \hat{F}_X^n\left(-m_pN_s\frac{(\log_2 N_s+4)}{2}\right) \quad (39)$$

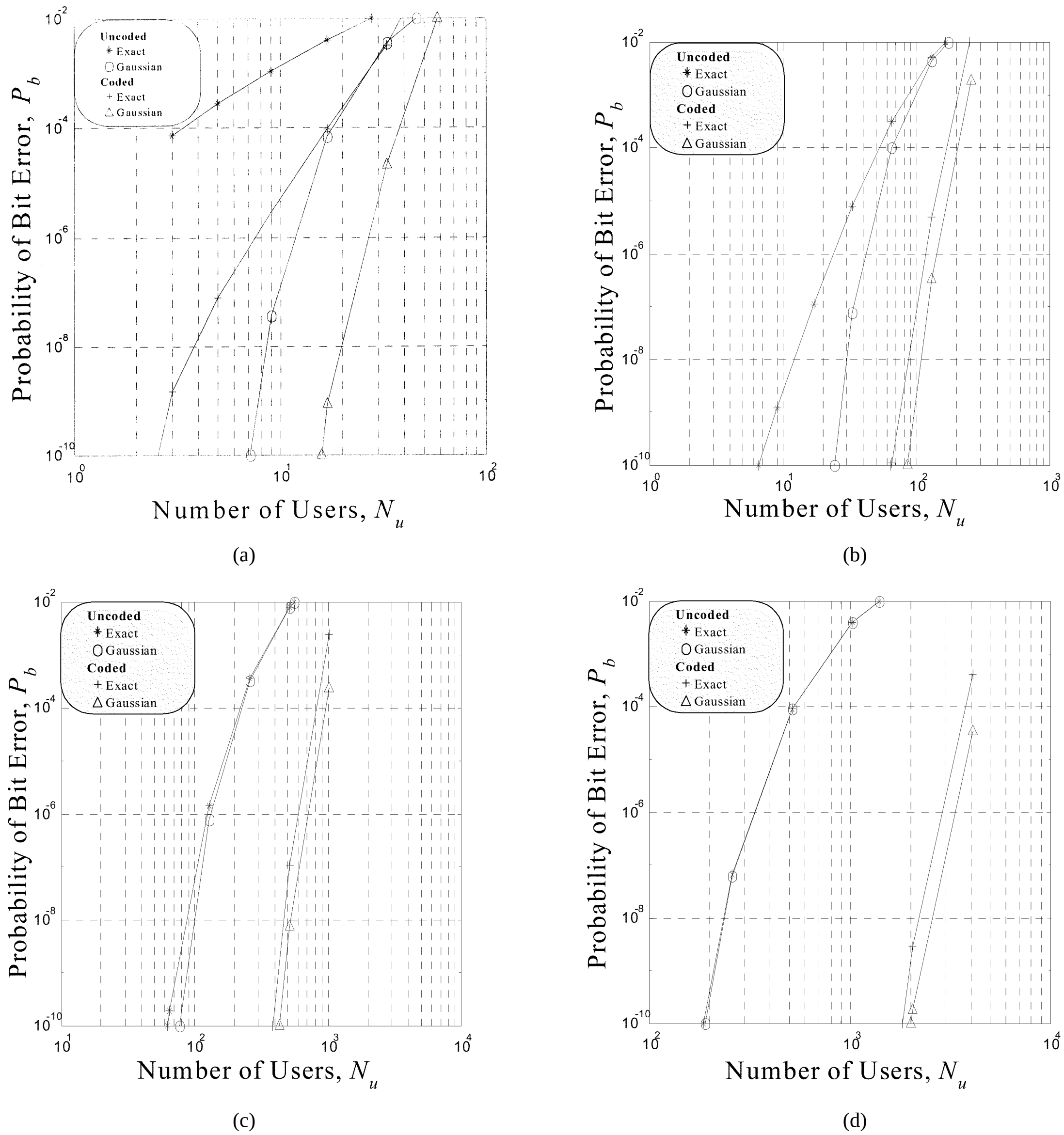


Fig. 4. (a) Probability of bit error as a function of number of users for synchronous uncoded and coded (upper bound) schemes in exact and Gaussian cases at $R_s = 5$ Mb/s ($N_s = 2$). (b) Probability of bit error as a function of number of users for synchronous uncoded and coded (upper bound) schemes in exact and Gaussian cases at $R_s = 1.25$ Mb/s ($N_s = 8$). (c) Probability of bit error as a function of number of users for synchronous uncoded and coded (upper bound) schemes in exact and Gaussian cases at $R_s = 312.5$ kb/s ($N_s = 32$). (d) Probability of bit error as a function of number of users for synchronous uncoded and coded (upper bound) schemes in exact and Gaussian cases at $R_s = 78.1$ kb/s ($N_s = 128$).

where $n = (N_u - 1)N_s(\log_2 N_s + 4)/2$, and $\hat{F}_X^n(x)$ is the probability cumulative distribution function which is computed in (22).

If we assume Gaussian distribution for the interference, we can use the improved upper bound of (36). In this bound the parameter η is m_p and the parameter σ^2, assuming N_u users, is equal to $\sigma^2 = (N_u - 1)\sigma_a^2$, where σ_a^2 is defined as in (19). We denote by d_f to be the free distance of the code and is equal to $N_s(\log_2 N_s + 4)/2$, $W = Z^{N_s/2}$, $K = \log_2 N_s + 2$, and $Z = \exp(-(m_p^2)/(2\sigma_a^2(N_u - 1)))$. Then, we obtain

$$P_b < Q\left(\left(\frac{m_p^2}{\sigma_a^2}\frac{\frac{N_s(\log_2 N_s+4)}{2}}{N_u - 1}\right)^{1/2}\right) \times \left(\frac{1-W}{(1-2W)(1-W^{K-2})}\right)^2 \quad (40)$$

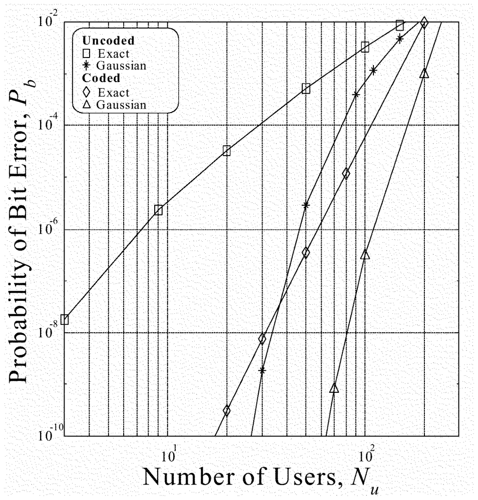
and the lower bound is easily computed as (30)

$$P_b \geq Q\left(\left(\frac{m_p^2}{\sigma_a^2}\frac{\frac{N_s(\log_2 N_s+4)}{2}}{N_u - 1}\right)^{1/2}\right). \quad (41)$$

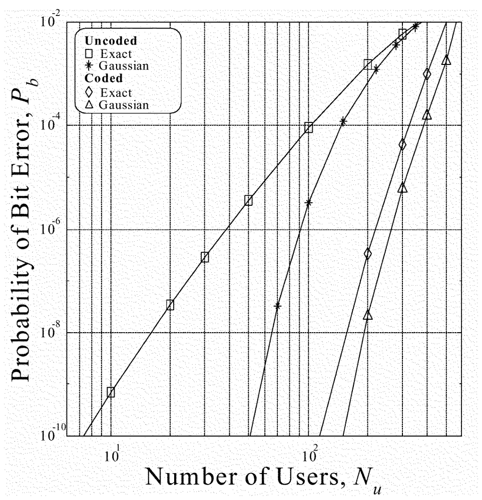
V. Numerical Results

In this section, we present some numerical results. Following the same examples as in [1], the received pulse is modeled as $w_{\text{rec}}(t + T_w/2) = [1 - 4\pi(t/\tau_m)^2]\exp(-2\pi(t/\tau_m)^2)$, where $\tau_m = 0.2877$ ns (the received pulse waveform is plotted in Fig. 2). We also set δ and T_f to 0.156 and 100 ns, respectively. With these selections, the parameter m_p^2/σ_a^2 will be approximately equal to 504.

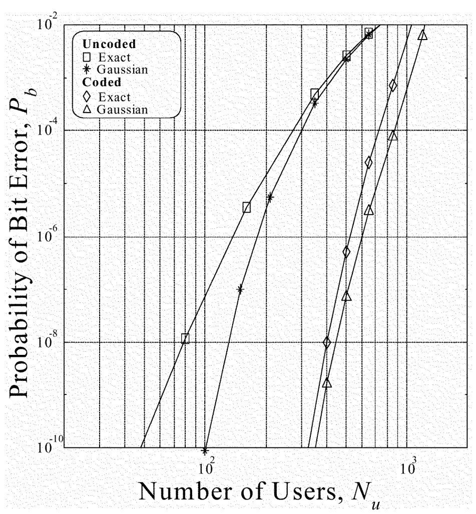
Fig. 4(a)–(d) present the plots of BER versus the number of users for uncoded and coded schemes in synchronous cases,



(a)



(b)



(c)

Fig. 5. (a) Probability of bit error as a function of number of users for asynchronous uncoded and coded (upper bound) schemes in exact and Gaussian cases at $R_s = 5$ Mb/s ($N_s = 2$). (b) Probability of bit error as a function of number of users for asynchronous uncoded and coded (upper bound) schemes in exact and Gaussian cases at $R_s = 2.5$ Mb/s ($N_s = 4$). (c) Probability of bit error as a function of number of users for asynchronous uncoded and coded (upper bound) schemes in exact and Gaussian cases at $R_s = 1.25$ Mb/s ($N_s = 8$).

based on both exact and Gaussian assumption analysis, at input rates of 5 Mb/s, 1.25 Mb/s, 312.5 kb/s, and 78.1 kb/s, respectively. Fig. 5(a)–(c) present the similar plots in asynchronous cases, at input rates of 5, 2.5, and 1.25 Mb/s.

As it can be realized from these figures, at high data transmission rate, the Gaussian assumption overestimates the number of users that can be supported by the system. However, as the transmission rate decreases the Gaussian assumption performs relatively well. We can justify these results as follows. By decreasing the transmission rate, the number of frames (or pulses) per bit increases, and the conditions for applying the central limit theorem to the distribution of the total interference at the receiver output holds more accurately.

From Figs. 4(a)–(d) and 5(a)–(c), it can be observed that the coded scheme performs significantly better than the uncoded scheme. For example [from Fig. 5(a)], at rate 5 Mb/s, for 30 users, the bit error probability for uncoded scheme is about 10^{-4}, but for the coded scheme, it is about 10^{-8}. As expected, the performance of coded system improves by increasing N_s, but even for $N_s = 2$ (rate 5 Mb/s), the number of users supported by our proposed coded scheme increases by a factor of 2.5–15 in BERs of 10^{-3} to 10^{-8} compared with the uncoded scheme. By comparing (15) and (38) and (25) and (41), and noting that the lower and upper bounds of the bit error probability of coded systems are relatively close to each other, it is evident that the superorthogonal codes can increase the number of users by a factor of $(4 + \log_2 N_s)/2$ at a given moderate to low BER.

VI. Conclusion

In this paper, we first proposed to use relatively low-complexity superorthogonal convolutional codes in time-hopping CDMA systems (specifically in the UWB radio system). The receiver for our proposed scheme employs only one pulse correlator and a superorthogonal decoder. The processing complexity of this decoder is proportional to $\log_2 N_s$, and the required memory is proportional to N_s, where N_s is the number of pulses transmitted by TH-spread-spectrum system for each input bit.

We have, then, provided a more precise performance analysis of the system for both uncoded and coded schemes. Our performance analysis first indicates that at high bit rate, the Gaussian distribution assumption for the total interference at a single user receiver output, as obtained in [1]–[3], overestimates the number of users supported by the system. It then shows that our proposed coded scheme significantly outperforms an uncoded scheme and increases the number of supported users at a given BER by a factor equal to $(4 + \log_2 N_s)/2$, without any increase in the required bandwidth expansion factor.

Appendix A
Computation of the Interference Variance in a Synchronous System

Substituting z by e^x in the interference probability characteristic function of each interfering user (9), we get

$$\Phi(s) = T(e^s) = \frac{1}{2}(\alpha+\beta e^{sm_p})^{N_s} + \frac{1}{2}(\alpha+\beta e^{-sm_p})^{N_s}. \quad \text{(A.1)}$$

Then, the variance will be

$$\begin{aligned}\sigma^2 =& E\{X^2\} - E^2\{X\} = \left.\frac{d^2\Phi(s)}{ds^2}\right|_{s=0} - \left(\left.\frac{d\Phi(s)}{ds}\right|_{s=0}\right)^2 \\ =& m_p^2 \beta N_s(\beta N_s + \alpha). \end{aligned} \quad \text{(A.2)}$$

With an N_u independent active users, the variance of the total interference is

$$\sigma_n^2 = (N_u - 1)\sigma^2 = m_p^2 (N_u - 1) N_s \beta (N_s \beta + \alpha). \quad \text{(A.3)}$$

Appendix B
Calculation of the nth-Order Convolution of $\hat{f}_X(x)$

We can rewrite

$$\hat{f}_X(x) = \alpha\delta_D(x) + \beta\left[u(x+m_p) - u(x-m_p)\right]$$

as

$$\hat{f}_X(x) = \alpha\delta_D(x) + \beta h(x) = \alpha\delta_D(x) + \beta h_S\left(\frac{x}{m_p}\right) \quad \text{(B.1)}$$

where $h_S(x) = h(xm_p) = u(x+1) - u(x-1)$.

First, we compute the nth-order convolution of $h_S(x)$, i.e., $h_S^n(x)$. By using mathematical induction, we show that

$$h_S^n(x) = \frac{1}{(n-1)!}\sum_{i=0}^{n}(-1)^i\binom{n}{i}(x+n-2i)^{n-1}u(x+n-2i). \quad \text{(B.2)}$$

Proof

We start with $h_S^1(x)$. We have

$$\begin{aligned} h_S^1(x) =& h_S(x) = u(x+1) - u(x-1) \\ =& \frac{1}{0!}\sum_{i=0}^{1}(-1)^i\binom{1}{i}(x+1-2i)^0 u(x+1-2i). \end{aligned} \quad \text{(B.3)}$$

Thus, $h_S^1(x)$ satisfies the formula. Assuming the formula is correct for value n, we must show that it is also correct for value $n+1$. We have the equation at the bottom of the page.

Since $h(x) = h_S(x/m_p)$, the nth-order convolution of $h(x)$, i.e., $h^n(x)$, is related to $h_S^n(x)$ by $h^n(x) = m_p^{n-1} h_S^n(x/m_p)$. Thus

$$h^n(x) = \frac{m_p^{n-1}}{(n-1)!} \cdot \sum_{i=0}^{n}(-1)^i\binom{n}{i}\left(\frac{x}{m_p} + n - 2i\right)^{n-1} u\left(\frac{x}{m_p} + n - 2i\right). \quad \text{(B.4)}$$

Now, we compute $\hat{f}_X^n(x)$ using $h^n(x)$. It can easily be shown that

$$\hat{f}_X^n(x) = \alpha^n\delta_D(x) + \sum_{i=1}^{n}\binom{n}{i}\alpha^{n-i}\beta^i h^i(x). \quad \text{(B.5)}$$

Substituting $h^n(x)$ from (B.4) in the above expression, we get

$$\begin{aligned}\hat{f}_X^n(x) = & \alpha^n\delta_D(x) + \sum_{i=1}^{n}\binom{n}{i}\alpha^{n-i}\beta^i\frac{m_p^{i-1}}{(i-1)!} \\ & \times \sum_{j=0}^{i}(-1)^j\binom{i}{j}\left(\frac{x}{m_p} + i - 2j\right)^{i-1} u\left(\frac{x}{m_p} + i - 2j\right). \end{aligned} \quad \text{(B.6)}$$

$$\begin{aligned} h_S^{n+1}(x) =& h_S * h_S^n(x) \\ =& \int_{-\infty}^{+\infty}\left[u(a+1) - u(a-1)\right]\cdot\frac{1}{(n-1)!}\sum_{i=0}^{n}(-1)^i\binom{n}{i}(x-a+n-2i)^{n-1}u(x-a+n-2i)da \\ =& \int_{-1}^{1}\frac{1}{(n-1)!}\sum_{i=0}^{n}(-1)^i\binom{n}{i}(x-a+n-2i)^{n-1}u(x-a+n-2i)da \\ =& \frac{1}{n!}\sum_{i=0}^{n}(-1)^{i+1}\binom{n}{i}\left[(x-1+n-2i)^n u(x-1+n-2i) - (x+1+n-2i)^n u(x+1+n-2i)\right] \\ =& \frac{1}{n!}\sum_{i=1}^{n}(-1)^i\left[\binom{n}{i-1} + \binom{n}{i}\right](x+n+1-2i)^n u(x+n+1-2i) \\ & + \frac{1}{n!}(-1)^{n+1}\binom{n}{n}(x-n-1)^n u(x-n-1) + \frac{1}{n!}(-1)^0\binom{n}{0}(x+n+1)^n u(x+n+1) \\ =& \frac{1}{n!}\sum_{i=1}^{n}(-1)^i\binom{n+1}{i}(x+n+1-2i)^n u(x+n+1-2i) \\ & + \frac{1}{n!}(-1)^{n+1}\binom{n+1}{n+1}(x-n-1)^n u(x-n-1) + \frac{1}{n!}(-1)^0\binom{n+1}{0}(x+n+1)^n u(x+n+1) \\ =& \frac{1}{((n+1)-1)!}\sum_{i=0}^{n+1}(-1)^i\binom{n+1}{i}(x+(n+1)-2i)^{(n+1)-1}u(x+(n+1)-2i). \end{aligned}$$

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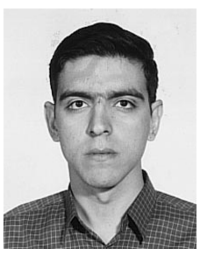


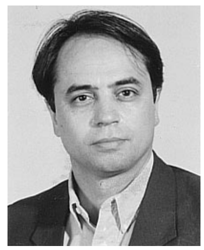
Amir R. Forouzan (S'99) was born on June 17, 1976 in Naein, Iran. He received the B.S. and M.S. degrees from Sharif University of Technology (SUT), Tehran, Iran, in 1998 and 2000, respectively.

He is currently with the Institute of Electronics, University of Tehran, Tehran, Iran. Since August 1999, he has been with the technical staff of Advanced CDMA Research Lab, Iran Telecom Research Center (ITRC), Tehran. His research interests include ultrawideband radio, radio and optical CDMA, and error-control coding.

Mr. Forouzan is a recipient of the gold medal of the Iranian Olympiad in Informatics.

Masoumeh Nasiri-Kenari (S'92–M'93) received the B.S. and M.S. degrees in electrical engineering from Isfahan University of Technology, Isfahan, Iran, in 1986 and 1987, respectively, and the Ph.D. degree in electrical engineering from the University of Utah, Salt Lake City, in 1993.

From 1987 to 1988, she was a Technical Instructor and Research Assistant at Isfahan University of Technology. Since 1994, she has been with the Department of Electrical Engineering, Sharif University of Technology, Tehran, Iran, where she is now an Associate Professor. From 1999 to 2001, she has also been a Co-Director of the Advanced CDMA Lab at the Iran Telecom Research Center (ITRC), Tehran. Her research interests are in radio and optical communications, and error-control coding.



Jawad A. Salehi (S'80–M'84) received the B.S. degree in electrical engineering from the University of California, Irvine, in 1979 and the M.S. and Ph.D. degrees from the University of Southern California (USC), Los Angeles, in 1980 and 1984, respectively.

From 1981 to 1984, he was a full-time Research Assistant at the Communication Science Institute at USC, engaged in research in the area of spread-spectrum systems. From 1984 to 1993, he was a Member of the Technical Staff with the Applied Research Area at Bell Communications Research (Bellcore), Morristown, NJ. From February to May 1990, he was with the Laboratory for Information and Decision Systems at the Massachusetts Institute of Technology, Cambridge, as a Visiting Research Scientist conducting research on optical multiple-access networks. Currently, he is an Associate Professor with the Department of Electrical Engineering, Sharif University of Technology, Tehran, Iran. From fall 1999 until fall 2001, he was the head of the Mobile Communications Division and a Co-Director of the Advanced and Wideband CDMA Lab at Iran Telecom Research Center (ITRC), Tehran, conducting research in the area of advanced CDMA techniques for optical and radio communications systems. His current research interests are in the areas of optical multiaccess networks, in particular, fiber-optic CDMA, femtosecond or ultra-short light pulse CDMA, spread time CDMA, holographic CDMA, wireless indoor infrared CDMA systems, applications of EDFA in optical systems, and optical orthogonal codes (OOC). His work on optical CDMA systems resulted in ten U.S. patents.

Dr. Salehi is a recipient of Bellcore's Award of Excellence. He assisted, as a member of the organizing committee, to organize the first and the second IEEE Conferences on Neural Information. In May 2001, he was appointed to serve as Editor for Optical CDMA of the IEEE TRANSACTIONS ON COMMUNICATIONS.