Scientia Iranica, Vol. 11, No. 3, pp 248–254
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SCIENTIA
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Iterative Interference Cancellation for a Coded Multicarrier Frequency-Hopping CDMA (MC-FH-CDMA) System

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In this paper, a low complexity iterative multi-user receiver for the coded MC-FH-CDMA system, is considered, which has been introduced in [1]. The structure of the proposed receiver consists of a Multi-User Likelihood Calculator (MULC) and a bank of SISO channel decoders, one for each active user. Based on the received signal and a priori information about code bits, provided by SISO decoders in the previous iteration, MULC computes the extrinsic information of the coded symbols of active users in the form of a Log-Likelihood Ratio (LLR). This information is used by channel decoders as a priori information to compute an update of LLRs. The processing proceeds in an iterative manner, similar to the decoding of turbo codes. The performance of the proposed receiver is evaluated in AWGN channels. The numerical results show that the new receiver significantly improves system performance, compared to the conventional single user receiver.

INTRODUCTION

A novel diversity scheme for a phase-coherent frequency-hopping spread spectrum system has been proposed in [2] and the performance of the system has been evaluated in a single-user jamming environment. In this system, which is named the multicarrier frequency-hopping CDMA (MC-FH-CDMA) system, diversity is obtained via both multicarrier transmission and frequency-hopping. The conventional Fast Frequency-Hopping (FFH) systems, which transmit one carrier at a time and change it in fractions of bit duration, make coherent demodulation relatively difficult. However, the system proposed in [2], not only makes carrier-frequency hops slower but, by employing two kinds of diversity simultaneously, imposes each carrier to hop solely at a fraction of the total given bandwidth. Thus, with slow frequency-hopping, with a period at least equal to the bit duration, a coherent reception will be feasible in a slow fading channel [1]. However, in a fast fading channel, a noncoherent reception is inevitable.

In this system, at each bit interval, i, for each user, k, N_s carriers are chosen from N_s distinct subbands. The whole frequency band is partitioned into N_s subbands, each of which has exactly N_h subcarriers spaced apart by f_d. During each bit interval, i, the carrier frequency selected from each subband is determined by the user dedicated signature sequence. Figure 1 shows an example of the subcarriers selected in a MC-FH system with two users, $N_s = 4$ and $N_h = 5$. These N_s carriers are modulated with the ith data bit of user k, using BPSK modulation. These modulated carriers are, then, added together and transmitted through the channel. For the next bit transmission, each of these N_s carriers hops in its subband and another

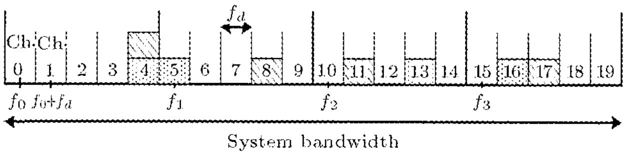

Figure 1. An example of dedicated subcarriers for a MC-FH system with $N_s = 4$ and $N_h = 5$.

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frequency in the subband is chosen. The frequency-hopping pattern, which is determined by a dedicated signature sequence of the user, is assumed to be known to the receiver. After dehopping at the receiver side, N_s subband correlators separate the signal transmitted in different subbands. These N_s detected signals are used to make a decision on the input data bit. To this end, the correlators' weighted outputs are simply combined and the result is compared to threshold zero, to make a decision on the input data bit. Note that the modulating and demodulating of the N_s carriers simultaneously can easily be implemented using IFFT and FFT, respectively.

To exploit the given bandwidth more efficiently, in [1], the authors have proposed to employ a practical, low-rate convolutional code to the uncoded MC-FH-CDMA scheme described above. The idea is as follows: Instead of sending the N_s carriers at each bit interval with an identical phase (0° or 180°) that is determined by the corresponding input data bit, these carriers can be sent with different phases, whose values are determined with output symbols of an encoder. In fact, with the above insight, the uncoded scheme can be considered as a coded scheme with a repetition block code of rate $1/N_s$. Since the repetition code is not a good code, it is expected that, by applying a more powerful code with the same rate $1/N_s$, the system performance will substantially improve, without any bandwidth expansion further than that needed by the uncoded scheme. The low-rate code that has been chosen for demonstrating the performance of the coded scheme is the convolutional superorthogonal code (SOC) [3] of rate $1/N_s$. The performance analysis of the uncoded and coded schemes in [1] has indicated that, at a given bit error rate, the coded scheme increases the number of users by a factor which is logarithmic in N_s.

The previous analyses of both uncoded and coded MC-FH-CDMA systems have been for a single-user correlator receiver, which simply treats multiple-access interference as a noise. In this paper, a low-complexity, iterative receiver is considered for decoding multi-user information data for the coded MC-FH-CDMA scheme, as described above. The receiver structure consists of a Multi-User Likelihood Calculator (MULC), followed by a bank of Soft-Input Soft-Output (SISO) channel decoders. In each iteration, the MULC, first, utilizes the information provided by the SISO decoders in the previous iteration for suppressing and minimizing the effect of the multi-user interference at the correlator output of each user. Then, for each active user, the MULC provides soft information about coded symbols of the user in the form of a Log-Likelihood Ratio (LLR), which is used by the corresponding SISO decoder as a priori information about coded symbols. The processing proceeds in an iterative manner, similar to the decoding of turbo codes. After a few iterations, the user information data is decoded using the hard decision on the SISO channel decoder output. The simulation results indicate that the low complexity iterative receiver proposed significantly improves the coded system performance, compared to the conventional non-iterative receiver considered in [1].

The outline of this paper is as follows. In the next section, a brief description of the system is proposed, for both uncoded and coded schemes. Then, the iterative multi-user receiver structure is presented and after that, an error performance analysis of the system is developed. Finally, the numerical results are proposed and this paper is concluded.

SYSTEM DESCRIPTION

In a MC-FH-CDMA system, every transmitter sends N_s carriers for each data bit using BPSK modulation. The carriers are spaced apart in sequential subbands. In fact, the total given frequency bandwidth is partitioned into N_s subbands of equal bandwidth, where each subband contains N_h different frequency carriers spaced apart by f_d. f_d is chosen such that every pair of carriers is orthogonal, i.e., $f_d = 1/T_s$, where T_s denotes a bit time interval.

The equivalent base band transmitted signal of user k is as follows:

$$s^{(k)}(t) = \sqrt{2P}\sum_n d_k(n) \times e^{j2\pi(f_{n-N_s\lfloor n/N_s\rfloor}+c_k(n)f_d)(t-\lfloor n/N_s\rfloor T_s)} \times P_{T_s}\left(t-\left\lfloor\frac{n}{N_s}\right\rfloor T_s\right), \tag{1}$$

where $\{d_k(n)\}$ is the transmitted binary sequence of user k. This sequence modulates the dedicated carriers. $\{c_k(n)\}$ is the pseudorandom sequence of user k, which determines the carrier frequency selected from each subband $l = n - N_s\lfloor n/N_s\rfloor$ during the $i = \lfloor n/N_s\rfloor$th bit interval. The elements of this sequence are i.i.d random variables, which take on integer values in the interval $[0, N_h-1]$. $P_\lambda(t)$ is a rectangular pulse over the interval $[0,\lambda]$ with amplitude equal to 1. f_l is the first carrier frequency in subband l and is equal to $lN_hf_d, l=0,1,2,\ldots,N_h-1$.

For the uncoded scheme, the stream $\{d_k(n)\}$ is N_s repetitions of the transmitted data sequence, i.e., if the transmitted binary data sequence is $\{D_k(i)\}$, then, one has $d_k(n) = D_k(i)$ for $\lfloor n/N_s\rfloor = i$. Thus, the uncoded scheme can be considered as a coded system, which employs a repetition code of rate $1/N_s$. With this insight, in [1], a coded scheme has been proposed, which employs a near optimal, low-rate, convolutional code, instead of the simple repetition code. This code,

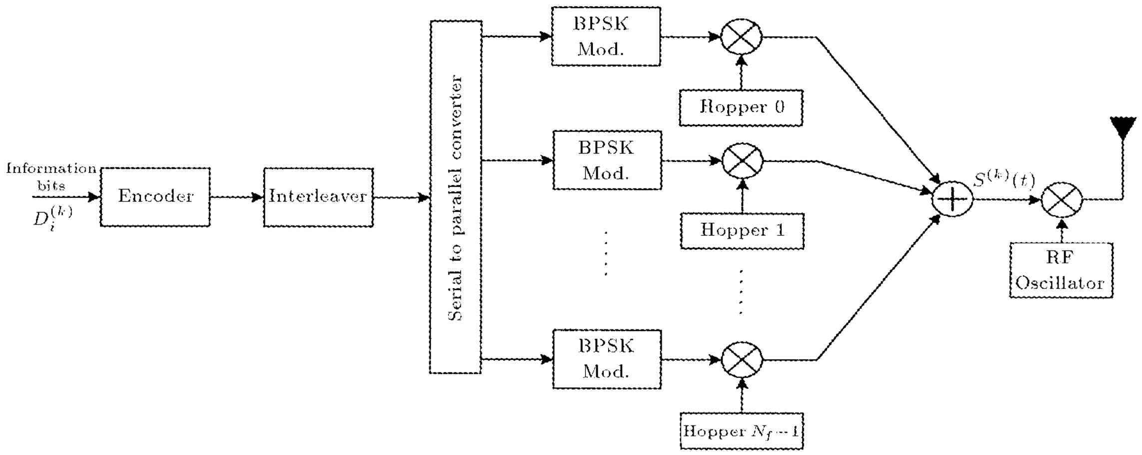

Figure 2. The block diagram of the coded MC-FH systems.

called Super Orthogonal Code (SOC), has a rate of $1/2^{K-2}$, where K is the constraint length of the code. Since a rate of $1/N_s$ is required, $2^{K-2}=N_s$ or $K=\log_2 N_s+2$ must be set. In the proposed coded scheme, $\{d_k(n)\}_{n=iN_s}^{(i+1)N_s-1}$ are the N_s coded symbols of user k at information bit interval i. Figure 2 shows the block-diagram of the coded system.

For simplicity of presentation, a synchronous system is assumed. An AWGN channel is, also, considered. Thus, the total received signal can be written as:

$$r(t)=\sum_{k=1}^{N_u} A_k s^{(k)}(t)+\nu(t), \tag{2}$$

where A_k is the received amplitude of user k, N_u is the number of active users and $\nu(t)$ is AWGN with a two sided power spectral density of $N_0/2$. The conventional correlator receiver, which is an optimum receiver in a single-user environment, ignores the presence of multiple access signals and treats them as a noise. Let the desired user be k. Then, the receiver of the uncoded scheme can be described as:

$$D_k(i)=1 \Leftrightarrow Z_i(k) \triangleq$$

$$\mathrm{Re}\left\{\sum_{m=0}^{N_s-1}\overbrace{\frac{1}{T_s}\int_{iT_s}^{(i+1)T_s} r(t)e^{-j2\pi(f_m+c_k(m+jN_s))(t-iT_s)}dt}^{Y_K(m+iN_s)}\right\}>0. \tag{3}$$

From this equation, it is clear that the correlator receiver consists of N_s branches, each with a subband correlator. The role of the subband correlator of branch m is to extract the symbol transmitted in subband m. In Equation 3, $Y_K(m+iN_s)$ is the subband correlator output of branch m during bit interval i. Here, N_s subband correlator outputs $(\{Y_K(m+iN_s)\}_{m=0}^{N_s-1})$ are added to make the decision variable $Z_i(k)$, which is then compared to zero.

In a coded scheme, the real part of the subband correlator outputs are the inputs of the decoder. Decoding is performed using a soft input Viterbi algorithm. The state diagram of a super orthogonal code, with constraint length K, consists of 2^{K-1} states. In this diagram, from each state, two branches, corresponding to bit zero and bit one, exit. To update the metrics, it is, first, necessary to calculate the trellis diagram branch metrics, using the received signal $r(t)$. For this purpose, in each subband m and during each bit interval I, the real part of the subband correlator output is computed as $y_k(m+iN_s) \triangleq \mathrm{Re}\{Y_k(m+iN_s)\}$. Then, the branch metrics can be simply evaluated based on these quantities.

MULTIUSER RECEIVER STRUCTURE

Figure 3 shows the block diagram of the proposed iterative receiver. The receiver consists of two essential parts, namely, Multiuser Likelihood Calculator (MULC) and SISO channel decoders. These two parts are separated by interleavers and deinterleavers. MULC consists of three separate parts: Correlator receivers, Parallel Soft Interference Canceller (PSIC) and single-user likelihood calculators. In each iteration, the PSIC for each user k utilizes soft information about the coded symbols of interfering users provided by their SISO channel decoders, for interference cancellation at the kth user's correlator receiver output. Then, the result is used by the single-user likelihood calculators to deliver soft information (extrinsic information) about the coded symbols of user k in the form of Log-Likelihood Ratio (LLR). The kth SISO channel decoder uses this extrinsic information, i.e., $L_{12}^e[d_k(n)]$, as a priori information and delivers an update of LLRs for code bits, i.e., $L_{21}^e[d_k(n)]$, based on the code constraint.

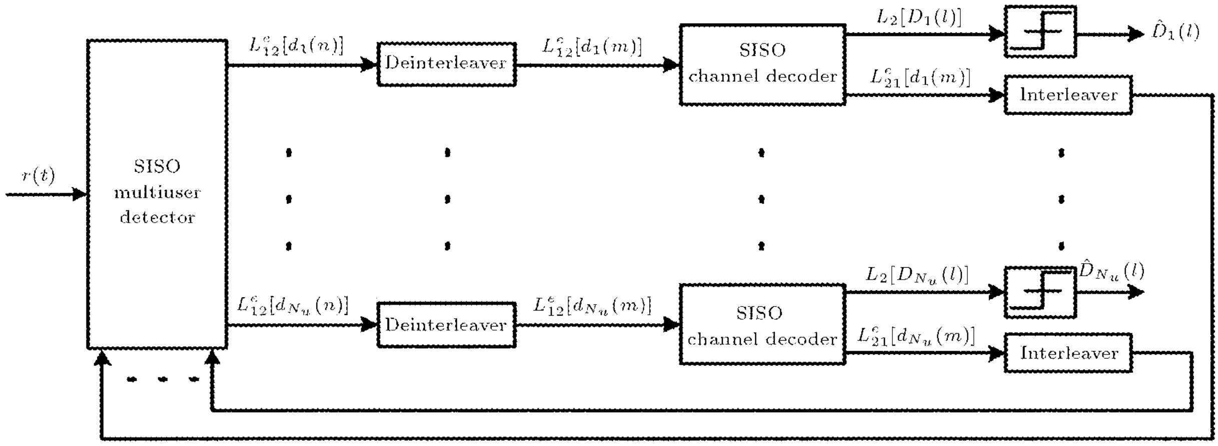

Figure 3. The block diagram of the iterative receiver.

The sequence $\{L_{21}^{e}[d_k(n)]\}$ is, then, used by the PSIC in the next iteration. The SISO channel decoders also compute the LLR for each information bit, i.e., $L_2[D_k(n)]$, which is used to make a decision on the user information bits at the last iteration (see Figure 3). Channel decoders are implemented using the SISO algorithm presented in [4]. In the following, each part of the MULC is described in more detail.

Correlator Receivers

There are N_u correlator receivers, one for each active user. Each of these correlator receivers consists of N_s subband correlators in N_s branches, one for each coded symbol, as described in the previous section. After parallel to serial conversion, the outputs of these branches are fed to PSIC. From Equations 1 and 2, the received signal in bit interval i is equal to:

$$r(t)=\sqrt{2P}\sum_{k=1}^{N_u} A_k \sum_{n=iN_s}^{iN_s+N_s-1} d_k(n)e^{j2\pi(f_{n-iN_s}+c_k(n)f_d)t}+\nu(t), \quad (4)$$

where $i \triangleq \lfloor n/N_s \rfloor$. The output of the subband correlator of user k, corresponding to the nth coded symbol of the user at branch $(n-iN_s)$, is equal to:

$$Y_k(n)=\frac{1}{T_s}\int_{iT_s}^{(i+1)T_s} r(t)e^{-j2\pi(f_{n-iN_s}+c_k(n)f_d)t}dt. \quad (5)$$

By substituting $r(t)$ from Equation 4 in Equation 5 and taking the real part of Equation 5 one obtains:

$$y_k(n)=A_k d_k(n)\sqrt{2P}$$
$$+\sum_{\substack{k'=1\\k'\neq k}}^{N_s} A_{k'} d_{k'}(n)\sqrt{2P}\frac{1}{T_s}\int_{iT_s}^{(i+1)T_s} e^{j2\pi(c_{k'}(n)-c_k(n))f_d t}dt$$
$$+\eta_k(n), \quad (6)$$

where $\eta_k(n) \triangleq \mathrm{Re}(\frac{1}{T_s}\int_{iT_s}^{(i+1)T_s}\nu(t)e^{-j2\pi}(f_{n-iN_s} +c_k(n)f_d)tdt)$ is the white Gaussian noise component, with zero mean and variance equal to $\sigma_\eta^2 = N_0/T_s$. By defining:

$$\delta_{k,k'}(n)=\begin{cases}0 & c_k(n)\neq c_{k'}(n)\\ 1 & c_k(n)=c_{k'}(n)\end{cases}, \quad (7)$$

one can rewrite Equation 6 as:

$$y_k(n)=A_k d_k(n)\sqrt{2P}$$
$$+\sum_{\substack{k'=1\\k'\neq k}}^{N_s} A_{k'} d_{k'}(n)\sqrt{2P}\delta_{k,k'}(n)+\eta_k(n). \quad (8)$$

The first term in this equation is the signal of the desired user, the second term is the Multiple Access Interference (MAI) component and the third term is the noise component, as described previously.

Parallel Soft Interference Canceller (PSIC)

This block makes a soft estimation of the code bits by using a posteriori information about code bits provided by the SISO channel decoders. Then, these estimates are used to cancel the MAI component from the subband correlator output. Suppose that $\tilde{d}_k(n)$ is the estimation of code bit $d_k(n)$. Then, at the output of PSIC for user k, i.e., $\tilde{y}_k(n)$, one has:

$$\tilde{y}_k(n)=y_k(n)-\sum_{\substack{k'=1\\k'\neq k}}^{N_u} A_{k'}\tilde{d}_{k'}(n)\sqrt{2P}\delta_{k,k'}=A_k d_k(n)\sqrt{2P}$$
$$+\sum_{\substack{k'=1\\k'\neq k}}^{N_u} A_{k'}\left(d_{k'}(n)-\tilde{d}_{k'}(n)\right)\sqrt{2P}\delta_{k,k'}(n)+\eta_k(n). \quad (9)$$

Each code bit, $d_k(n)$, is estimated by its expectation, i.e.:

$$\tilde{d}_k(n) = 1 \times P[d_k(n) = +1] - 1 \times P[d_k(n) = -1]. \tag{10}$$

It can be easily shown that [5]:

$$\tilde{d}_k(n) = \tanh\left(\frac{1}{2} L_{21}^e[d_k(n)]\right). \tag{11}$$

Such a soft interference cancellation was first proposed in [5]. Note that incorrect decisions by the channel decoders about code bits usually have small a posteriori LLRs and, as a result, the soft estimates of these incorrect decoded bits are small and do not have much effect on cancellation in Equation 9. Thus, with soft interference cancellation, error propagation is substantially avoided.

Since in the first iteration, there is no a priori information about the code bits, it is well assumed that $\Pr\{d_k(n) = +1\} = \Pr\{d_k(n) = -1\} = 1/2$ and, as a result, $\tilde{d}_k(n)$ is equal to zero.

Since soft estimates of code bits are not exactly equal to code bits, the MAI is not completely suppressed. The second term in Equation 9 is the residual interference in the nth code bit of user k. If $\xi_k(n)$ denotes the residual interference, one has:

$$\xi_k(n) = \sqrt{2P} \sum_{\substack{k'=1 \\ k' \neq k}}^{N_u} A_{k'} \left(d_{k'}(n) - \tilde{d}_{k'}(n)\right) \delta_{k,k'}(n). \tag{12}$$

Single-User Likelihood Calculator

For each user k, the single user likelihood calculator, at code symbol interval n, delivers the extrinsic information about code bit $d_k(n)$ in the form of LLR, as follows:

$$L_{12}^e[d_k(n)] = \log \frac{p\left(\tilde{y}_k(n) | d_k(n) = +1\right)}{p\left(\tilde{y}_k(n) | d_k(n) = -1\right)}. \tag{13}$$

To compute this quantity, the probability density function of $\tilde{y}_k(n)$, conditioned on $d_k(n)$, must be derived. From Equation 9, it can be realized that the conditional density function of $\tilde{y}_k(n)$ depends on distribution of the second and third terms of Equation 9, which are the residual interference and noise at code symbol interval n, respectively.

Since the distributions of code bits $d_k(n)$ of all users are known from the extrinsic information delivered by SISO decoders in the previous iteration, by the assumption that the code bits of different users are independent, the distribution of $\xi_k(n)$ can be evaluated. For simplicity, however, using Central Limit Theorem (C.L.T), it will be well assumed that $\xi_k(n)$ has Gaussian distribution. So, one only needs to compute the mean and variance of $\xi_k(n)$, which are given as:

$$E[\xi_k(n)] = \sum_{\substack{k'=1 \\ k' \neq k}}^{N_u} A_{k'} \left(E[d_{k'}(n) - \tilde{d}_{k'}(n)\right) \sqrt{2P} \delta_{k,k'}(n) = 0, \tag{14}$$

and

$$\begin{aligned} \sigma_k^2(n) &= E[\xi_k^2(n)] - E^2[\xi_k(n)] \\ &= \sum_{\substack{k'=1 \\ k' \neq k}}^{N_u} 2A_{k'}^2 P \delta_{k,k'}(n) E\left[\left(d_{k'}(n) - \tilde{d}_{k'}(n)\right)^2\right] \\ &= \sum_{\substack{k'=1 \\ k' \neq k}}^{N_u} 2A_{k'}^2 P \delta_{k,k'}(n) \left(1 - \tilde{d}_{k'}^2(n)\right). \end{aligned} \tag{15}$$

Since the residual and noise terms are independent, by the above assumption, their sum also has a Gaussian distribution with zero mean and variance equal to $\sigma_k^2(n) + \sigma_\eta^2$, where $\sigma_k^2(n)$ is given in Equation 15 and σ_η^2 is the variance of noise term, which is equal to N_0/T_s. Now, from Equation 13, the single-user likelihood calculator computes extrinsic information about code bit $d_k(n)$ as:

$$L_{12}^e[d_k(n)] = \frac{2A_k\sqrt{2P}}{\sigma_k^2(n) + N_0/T_s} \tilde{y}_k(n). \tag{16}$$

This value is delivered to the SISO channel decoder of user k as a priori information about code bit $d_k(n)$.

PERFORMANCE ANALYSIS

In this section, an upper bound is provided on the bit error rate of the iterative multi-user receiver in a coded MC-FH-CDMA system. Since the decoding rule used in the convolutional decoder is Maximum A Posteriori (MAP), whose bit error rate is less than that of the Maximum Likelihood (ML) decoding rule, one can well use the upper bound on the bit error rate of the ML decoder as an upper bound of the bit error rate of the MAP decoder. By assumption that residual interference, plus noise, at the input of the channel decoder, has Gaussian distribution, the upper bound on the bit error rate of the ML decoder depends on its input Signal to Interference plus Noise Ratio (SINR). So, SINR is first computed at the input of the channel decoder or, equivalently, at the output of the block PSIC.

Using Equations 9 and 15, one can obtain the input SINR per code bit to the channel decoder of user k as:

$$\mathrm{SINR}(k,n) \triangleq \frac{1}{2}\frac{E^2\{\widetilde{y}_k(n)\}}{\mathrm{var}\{\widetilde{y}_k(n)\}} = \frac{A_k^2 P}{\sigma_k^2(n)+N_0/T_s}. \quad (17)$$

As realized, the SINR is not constant for different code bits. So, one can define:

$$\mathrm{SINR}_{\min}(k) \triangleq \min_{n\in\{0,1,2,\cdots,N_b-1\}} \mathrm{SINR}(k,n), \quad (18)$$

where N_b is the total number of code bits per block at the input of the SISO channel decoder. In each iteration, the $\mathrm{SINR}_{\min}(k)$ can be computed as above. Then, an upper bound on the bit error rate of user k in each iteration is derived using the convolutional code weight generating function, i.e., $T(X,Y)$ as:

$$P_b(k) < \frac{1}{m}\exp\{d_{\mathrm{free}}\mathrm{SINR}_{\min}(k)\} \times Q\left(\sqrt{2d_{\mathrm{free}}\mathrm{SINR}_{\min}(k)}\right) \frac{\partial T(X,Y)}{\partial Y}\Big|_{Y=1,X=\exp\{-\mathrm{SINR}_{\min}(k)\}}, \quad (19)$$

where:

d_{free} the free distance of the convolutional code,

$T(X,Y)$ the weight generating function of the convolutional code,

m the number of information bits per each trellis branch = 1.

At high SNRs, after a few iterations, the interference term will be substantially eliminated. That is, $\sigma_k^2(n)$ in Equation 17 will be, approximately, equal to zero and, thus, Equation 17 reduces to:

$$\mathrm{SINR}_{\min}(k) = SNR(k) = \frac{A_k^2 P T_s}{N_0}. \quad (20)$$

As a result, the bit error rate will be limited as:

$$P_b(k) < \frac{1}{m}\exp\{d_{\mathrm{free}}\mathrm{SNR}(k)\} \times Q\left(\sqrt{2d_{\mathrm{free}}\mathrm{SNR}(k)}\right) \frac{\partial T(X,Y)}{\partial Y}\Big|_{Y=1,X=\exp\{-\mathrm{SNR}(k)\}}. \quad (21)$$

NUMERICAL RESULTS

In order to numerically evaluate the performance of the proposed iterative receiver for a coded MC-FH-CDMA system, in this section, some simulation results are presented. For simplicity, a system with full power control has been assumed. Then, without any loss of generality, $A_k = 1$ can be set. Furthermore, the convolutional code used is a SOC with rate of 1/4, i.e. $N_s = 4$, wich has been set. Also, the number of carriers in a subband, i.e. N_h, is equal to 3.

Figures 4 to 6 present the plots of Bit Error Rate (BER) versus the number of users for Signal to Noise Ratio equal to 2, 4, and 6 dB, respectively. In these figures, the plots of the BER of the system, without an iterative receiver, are also given.

As realized, the iterative receiver significantly improves system performance compared to the conventional non-iterative correlator (single-use) receiver. For

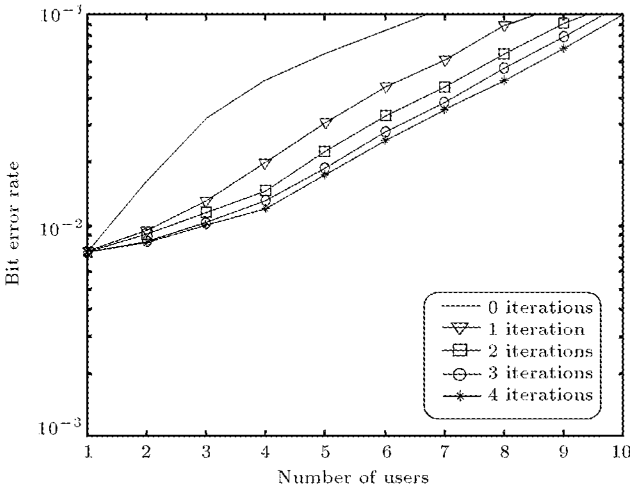

Figure 4. Bit error rate versus the number of users for SNR = 2 dB.

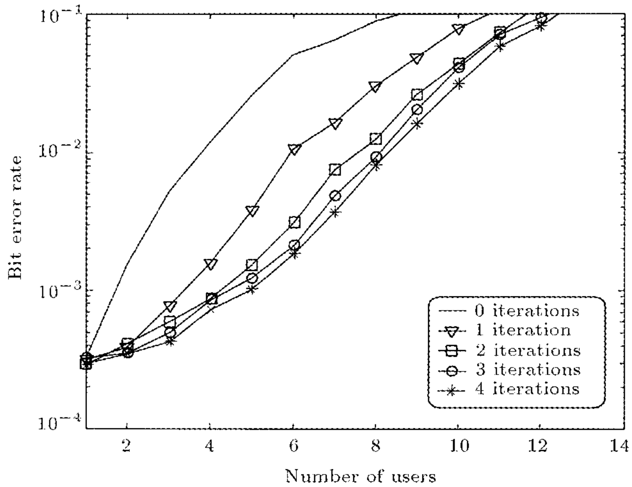

Figure 5. Bit error rate versus the number of users for SNR = 4 dB.

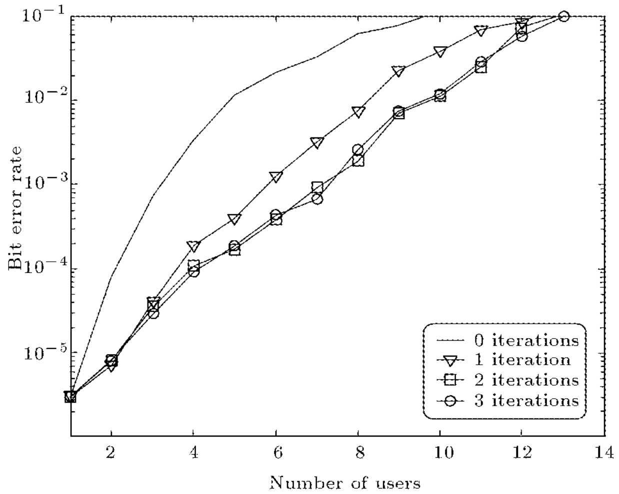

Figure 6. Bit error rate versus the number of users for SNR=6 dB.

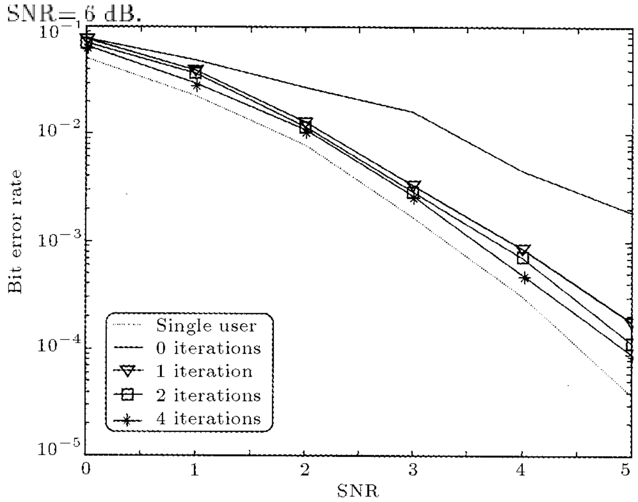

Figure 7. Bit error rate versus the signal to noise ratio for $N_u = 3$.

instance, from Figure 5, at a BER of 10^{-2}, the number of users supported by the non-iterative receiver is up to 4, whereas, with the proposed receiver, it is about 8.

Figure 7 shows the plots of BER versus SNR, when the number of active users is equal to 3. The plot for the single-user system is also given for comparison. It can be observed that the iterative receiver significantly improves system performance compared to the conventional non-iterative receiver. For instance, at a BER of 10^{-3}, the amount of improvement is about 2 db. Performance improvement is higher at higher SNRs. In fact, by increasing the number of iterations, at high SNRs, system performance reaches single-user performance.

CONCLUSION

In this paper, an iterative receiver has been proposed for a convolutionally coded MC-FH-CDMA system in an AWGN channel. The proposed receiver structure consists of a multi-user likelihood calculator and a bank of SISO channel decoders. Based on soft decisions on the coded bits of interfering users, provided by SISO channel decoders from the previous iteration, the MULC, first, reduces the MAI at the output of users' correlators and, then, provides the SISO channel decoders with a priori information about coded bits. The numerical results show that the new receiver significantly improves system performance, compared to the conventional single user receiver. At high SNR, after a few iterations, it is expected that the receiver will reach its ultimate performance, which is very close to the single-user performance.

ACKNOWLEDGMENT

The authors would like to thank Mr. A. Bayesteh for providing some parts of the simulation program. This work has been partially supported by Iran Telecom Research Center under contract T/500/7674.

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