VOL. 4, NO. 7, SEPTEMBER 2009 ISSN 1819-6608
ARPN Journal of Engineering and Applied Sciences

www.arpnjournals.com

PRE-RAKE DIVERSITY WITH GENERALIZED ORTHOGONAL CODES AND IMPERFECT CHANNEL CONDITIONS FOR FDD/DS-CDMA SYSTEMS

C. Subhas[1], K. Rama Naidu[2] and Y. Venkatarami Reddy[3]
[1]Department of Electronics and Communication Engineering, Sree Vidyanikethan Engineering College, Tirupati, A. P., India
[2]Department of Electronics and Communication Engineering, JNTU College of Engineering, Anantapur, A. P., India
[3]Andhra Pradesh Public Service Commission, Hyderabad, A. P., India
E-Mail: schennapalli@gmail.com

ABSTRACT
The Pre-Rake diversity combining technique decreases the complexity, size and cost of the mobile unit while achieving the same inter symbol interference (ISI) mitigation effects of conventional Rake receiver for direct sequence code division multiple access (DS-CDMA) systems. The technique is based on preprocessing of signal at the transmitter relying on knowledge of the channel state information. This *a priori* information is available in time division duplex (TDD) mode due to channel reciprocity. In frequency division duplex (FDD) mode the channel has to be estimated at mobile unit (MU) for the downlink and fed to base station to predict the channel for downlink time slot. In this paper, we compare the performance of Pre-Rake system with different spreading codes. We will also propose a method for channel prediction in FDD mode and evaluate the system under ideal and predicted channel conditions using generalized orthogonal (GO) and Walsh codes through analytical and computer simulations for DS-CDMA downlink.

Keywords: channel prediction, DS-CDMA, FDD, ISI, pre-rake, rake.

INTRODUCTION
In recent years, direct sequence code division multiple access (DS-CDMA) has become more and more important for modern wireless communications. One interesting feature of DS-CDMA systems is the use of Rake receivers in multipath environment to achieve multipath diversity gain [1]. Rake diversity combining, a receiver-based technique, will effectively combat self interference (SI) or inter symbol interference (ISI) and increases data throughput. However, this method requires channel state information (CSI) and complex signal processing at the mobile unit (MU) in the downlink which make the MU more bulky, expensive and unreliable. Hence, for the downlink, one can transfer the signal processing for interference suppression from the MU receiver to the base station (BS) transmitter by using Pre-Rake diversity combining technique, thereby reducing complexity of the MU (only a matched-filter to the own spreading sequence is required) with equal performance for both the single-user and multiuser systems employing orthogonal codes [1]. The Pre-Rake system with generalized orthogonal (GO) codes outperforms the one with other spreading codes [1]. In time division duplex (TDD) mode because the uplink and downlink transmissions share the same frequency channel with different time slot, it can be assumed that the uplink and downlink undergo the same channel fading. For operation in frequency division duplex (FDD), the downlink channel has to be predicted at the BS after obtaining CSI from MU through feedback.

In very few papers in the literature, the Pre-Rake system's performance using various modulation schemes is evaluated with orthogonal codes under imperfect channel conditions [2-5] that too in TDD mode. GO codes are applied to the Pre-Rake system to suppress system interference in this paper. The generalized orthogonality describes the zero correlation zone (ZCZ) or orthogonal zone property of GO codes, instead of the zero in-phase correlation property owned by traditional orthogonal codes such as Walsh sequences [6, 7]. The length of the orthogonal zone Z_{cz} represents the degree of the generalized orthogonality. Because, in the DS-CDMA system, the multiple access interference (MAI) and the SI are determined by the cross and autocorrelation functions, respectively [1], the concept of the orthogonal zone, in fact, opens a new direction in spreading sequence design. In this paper, we evaluate the performance of multiuser Pre-Rake system [1] using BPSK for various spreading sequences including GO codes. We extend our study of Pre-Rake technique in FDD mode for the DS-CDMA downlink. Since for FDD operation, the downlink channel needs to be predicted to compensate inherent feedback delay involved in receiving its CSI from MU, we will propose a very effective prediction method for obtaining CSI.

MULTIPATH CHANNEL MODELING
We utilize the simplified tapped delay line multipath channel model of [1]. The up-link channels are assumed to be statistically independent for all users. Also, with the utilization of up-link power control we assume that all channels are statistically identical, even if the MU's are at different distances from the BS. The complex low-pass impulse response of the channel of user k is given by

$$h_k(t) = \sum_{l=0}^{L-1} \beta_{k,l} \exp(j\gamma_{k,l})\delta(t - lT_c) \ , \tag{1}$$

Where L is the number of channel paths, the path gains $\beta_{k,l}$ are independent identically distributed (i.i.d.) Rayleigh random variables (r.v.'s) for all k and l, the angles $\gamma_{k,l}$ are i.i.d. uniformly distributed in $[0, 2\pi]$, and T_c is the chip duration. Without any loss of generality we can take the normalization $E[\beta_{k,l}^2] = 1$. In a TDD system under slow fading conditions, which are typical for portable communication systems, we assume that $h_k(t)$ does not change during two successive up and down time slots. In particular, when a slot is received at the BS through $h_k(t)$, BS estimates $h_k(t)$ for use in its own Rake receiver. The $h_k(t)$ might change due to time variability when the BS transmits during the next time slot to the MU of user k.

GENERALIZED ORTHOGONAL CODES

It is well known that in DS-CDMA system, the MAI is mainly caused by the discrete periodic and aperiodic correlation properties between the spreading sequences [1]. Therefore, the discrete correlation functions of GO sequences [1] are presented briefly in this section before further discussion.

For the GO code set $\{a_n^{(k)}$, *n = 0, 1. . . N; k = 0, 1, . . . , M -1*$\}$ of family size *M* and length *N*, the discrete periodic correlation function can be written as

$$\theta_{k,i}(l) = \sum_{n=0}^{N} a_n^{(k)} a_{n+l}^{(i)} = \begin{cases} N for l=0, k=i \\ 0 for l=o, k\neq i \\ 0 for o<|l|\leq Z_{cz}, k\neq i \end{cases} \tag{2}$$

Where Z_{cz} is the orthogonal zone which represents the degree of the generalized orthogonality. It should be noted that the traditional orthogonal codes, such as Walsh-Hadamard codes and orthogonal variable spreading factor (OVSF) codes, can be treated as GO codes with $Z_{cz} = 0$. In other words, the traditional orthogonal code set is only a special case of the GO code set.

In the following sections, the discrete aperiodic correlation function given below is used

$$C_{k,i}(l) = \begin{cases} \sum_{n=0}^{N-l-1} a_n^{(k)} a_{n+1}^{(i)}, 0\leq l\leq N-1 \\ \sum_{n=0}^{N+l-1} a_{n-1}^{(k)} a_n^{(i)}, -N+1\leq l\leq 0 \\ 0, |l|\geq N \end{cases} \tag{3}$$

(Instead of $\theta_{k,i}(l)$) to evaluate the system performance. However, notice that when $l = N$, from (2) and (3), $\theta_{k,i}(l) \approx C_{k,i}(l)$ can be obtained. Thus, for GO codes, we have $C_{k,i}(l) \approx 0$ for $|l| \leq Z_{cz}$.

The GO codes with required length and orthogonal zone can be generated using the following recursive relation [7].

$$\Delta_1^{(n)} = \begin{bmatrix} \Delta_1^{(n-1)}\Delta_1^{(n-1)}\Delta_1^{(n-1)}(-\Delta_1^{(n-1)}) \\ \Delta_1^{(n-1)}(-\Delta_1^{(n-1)})\Delta_1^{(n-1)}\Delta_1^{(n-1)} \end{bmatrix} \tag{4}$$

Where -Δ denotes the matrix the ij^{th} entry of which is the ij^{th} negation of Δ, and $\Delta_1\Delta_2$ denotes the matrix the ij^{th} entry of which is the concatenation of the ij^{th} entry of Δ_1 and the ij^{th} entry of Δ_2. The starter $\Delta_1^{(0)}$ is 4x4 Hadamard matrix.

PERFORMANCE ANALYSIS OF PRE-RAKE SYSTEM WITH GO CODES

In a multipath fading channel, conventionally a Rake receiver is used. It consists of a matched filter (MF) that is matched to the chip waveform of the spreading code, followed by a number of Rake fingers. Each Rake finger is synchronized to one of the channel paths. The Rake receiver fingers provide matching to the channel impulse response. Each Rake finger despreads the spread spectrum (SS) signal on its assigned channel path. Using Rake combining, the despread signals on all fingers are multiplied with the time reversed complex conjugate of the path gains and all outputs are then added [3]. Hence, the receiver in both BS and MUs must be equipped with sufficient Rake fingers and the corresponding channel estimation process.

In TDD mode, we can utilize the fact that for a period of slot time the channel impulse response is the same for the up and down links. Hence, only the BS needs to estimate it. As shown in Figure-1, to Pre-Rake the down-link signal of a user, the BS convolves this signal with the time reversed complex conjugate of the uplink channel impulse response of that user. Estimation of the uplink multipath complex gains can be practically achieved using pilot symbols aided techniques. The BS receiver extracts the pilot symbols and uses them to estimate the uplink channel impulse response. Alternate

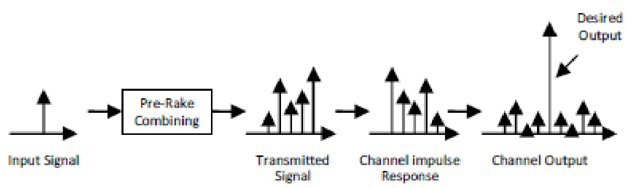

Figure-1. The pre-rake diversity combination process with ideal channel.

VOL. 4, NO. 7, SEPTEMBER 2009 ISSN 1819-6608
ARPN Journal of Engineering and Applied Sciences

method is to use dedicated pilot channels. Channel estimation errors cause performance degradation in both the Pre-Rake and the conventional Rake systems. For ideal channel, the conventional transmitted signal without Pre-Rake with binary phase-shift keying (BPSK) modulation is given by

$$s_k(t) = \sqrt{2P}b_k(t)a_k(t)\exp(j\omega t)\,, \tag{5}$$

Where P is the transmitted power, ω is the carrier frequency, and $b_k(t)$ is the data stream for user k consisting of a train of i.i.d. data bits with duration T. The current bit is denoted by b_k^0 while next or previous bits are denoted by adding or subtracting the superscript by one. $a_k(t)$ is the PN code of user k with chip duration of T_c and code length $N = T/T_c$. Now, in a Pre-Rake system, instead of (5) the downlink transmitted signal will be

$$s_k(t) = \sqrt{\frac{2P}{U_k}}\sum_{l=0}^{L-1} \beta_{k,L-l-1} b_k(t-lT_c) \cdot a_k(t-lT_c)\exp(j\omega(t-lT_c)-j\gamma_{k,L-l-1}) \tag{6}$$

Where U_k is a normalizing factor that keeps the instantaneous transmitted power constant regardless of the number of paths, and is given by

$$U_k = \sum_{n=0}^{L-1}\beta_{k,n}^2 \tag{7}$$

Based on the analysis of the study in [3], assuming a DS-CDMA system with K users, the received signal of user i during the downlink time slot can be written as

$$r_i(t) = n(t) + \mathrm{Re}\sum_{k=1}^{K}\sum_{j=0}^{L-1}\beta_{i,j}s_k(t-jTc)\exp(j\gamma_{i,j}) \tag{8}$$

Where $n(t)$ is the zero-mean additive white Gaussian noise (AWGN) with two-sided power spectral density $N_0/2$. From (6), (8), and Figure-1, it is shown clearly that the desired part of the received signal is the strongest peak among the $2L$-1 paths. Because, in the Pre-Rake system, the receiver no longer processes the path combination, the remaining $2L$-2 paths should be treated as the interference to user i and the other users.

Without loss of generality, we assume that user 1 is the desired user. The output of the correlation receiver of user 1 can be written as

$$Z = \int_{(L-1)T_c}^{(L-1)T_c+T}\eta(t)a_1(t-(L-1)T_c) \cdot \cos[\omega t - \omega T_c(L-1)]dt = D + S + A + \eta \tag{9}$$

Where D is the desired part of received signal, S is the SI, A is the MAI, and η is the zero-mean AWGN with variance $TN_0/4$.

As shown in Figure-1, the desired part of the received signal can be obtained from (9) when $j = L$-1-l

$$D = \sqrt{\frac{P}{2}}b_1^0 T\sqrt{U_1} \tag{10}$$

Where b_1^0 denotes the current bit of user 1. By employing Gaussian approximation, S and A can be treated as independent zero-mean Gaussian r.v's, and to estimate the bit error performance, the variances of S and A conditioned on $\beta_{1,j}$ are discussed in the following parts.

Self Interference (SI)

Because of multipath propagation, the SI exists even in a single-user system. In (6) and (8), when $j \neq$ L-1-l and k = 1, which means the undesired part of user 1, termed as the SI, S can be obtained as

$$S = \sqrt{\frac{P}{2U_1}}\sum_{j=0}^{L-1}\sum_{m=0,\neq j}^{L-1}\beta_{1,j}\beta_{1,m}\cos[\omega T_c(j-m)+\gamma_{1,m}-\gamma_{1,j}] \cdot \int_0^T b_1(t-(j-m)T_c)a_1[t-(j-m)T_c]a_1(t)dt \tag{11}$$

S can be divided into two parts: $S|_{|m-j|\le Z_{CZ}}$ and $S|_{|m-j|>Z_{CZ}}$. Because the aperiodic autocorrelation value of GO codes is almost equal to zero within the orthogonal zone, the SI contributes by paths arriving within the zero zone can be eliminated. After some manipulations, S can be further written as,

$$S \approx T_c\sqrt{\frac{P}{2U_1}} \cdot \sum_{j=0}^{L-2-Z_{cz}}\sum_{m=j+1+Z_{cz}}^{L-1}\{\beta_{1,j}\beta_{1,m}\cos[\omega T_c(j-m)+\gamma_{1,m}-\gamma_{1,j}] \cdot [(b_1^{-1}+b_1^{1})C_1(N-m+j)+2b_1^0 C_1(m-j)]\} \tag{12}$$

Where $C_k(m) = C_{k,k}(m)$ is the discrete aperiodic autocorrelation function of the kth code. For any value of j and m, the mean of S is zero, and all terms in (12) are uncorrelated due to the independent phase shifts. When GO codes are considered, the variances of S conditioned on $\beta_{1,j}$ can be obtained as

$$E[S^2 \mid \{\beta_{1,j}\}] = \frac{PT_c^2}{2U_1} \cdot \sum_{j=0}^{L-2-Zcz}\sum_{m=j+1+Z_{cz}}^{L-1}\beta_{1,j}^2\beta_{1,m}^2\{E[C_1^2(N-m+j)] + 2E[C_1^2(m-j)]\} = \frac{PT_c^2}{2U_1}[2N_\chi - \mu] \tag{13}$$

where $\chi = \sum_{j=0}^{L-2-Z_{cz}}\sum_{m=j+1+Z_{cz}}^{L-1}\beta_{1,j}^2\beta_{1,m}^2$ and

$\mu = \sum_{j=0}^{L-2-Z_{cz}}\sum_{m=j+1+Z_{cz}}^{L-1}\beta_{1,j}^2\beta_{1,m}^2(m-j)$.

Multiple access interference (MAI)

In the Pre-Rake system, because the transmitted signal has been Pre-Rake processed according to the channel impulse response, the received signal has more path components and stronger MAI than the conventional Rake system.

Based on (6) and (8), when $k \neq 1$, the MAI can be obtained as

$$A = \sqrt{\frac{P}{2}} \sum_{k=2}^{K} \sum_{j=0}^{L-1} \sum_{m=0}^{L-1} \frac{\beta_{1,j}\beta_{k,m}}{\sqrt{U_k}} .\cos[\omega T_c (j-m)+\gamma_{k,m}-\gamma_{1,j}]$$
$$.\int_0^T b_k[t-(j-m)T_c] a_k[t-(j-m)T_c] a_1(t)dt \qquad (14)$$

A can be divided into two parts: $A|_{|m-j|\leq Z_{CZ}}$ and $A|_{|m-j|>Z_{CZ}}$. When codes GO are used, because the aperiodic cross-correlation value of GO codes is almost equal to zero within the orthogonal zone, $A|_{|m-j|\leq Z_{CZ}} \approx 0$ can be obtained. Therefore, after some manipulations, A can be written as,

$$A = \sqrt{\frac{P}{2}} \sum_{k=2}^{K} \sum_{j=0}^{L-2-Zcz} \sum_{m=j+1+Zcz}^{L-1} \frac{T_c}{\sqrt{U_k}} \{\beta_{1,j}\beta_{k,m}$$
$$.\cos[\omega T_c (j-m)+\gamma_{k,m}-\gamma_{1,j}]$$
$$.[b_k^0 C_{k,1}(j-m)+b_k^1 C_{k,1}(N+j-m)]$$
$$+\beta_{1,m}\beta_{k,j} \cos[\omega T_c (m-j)+\gamma_{k,j}-\gamma_{1,m}]$$
$$.[b_k^{-1} C_{k,1}(m-j-N)+b_k^0 C_{k,1}(m-j)]\} \qquad (15)$$

Furthermore, the variances of A conditioned on $\beta_{1,j}$ can be obtained as

$$E[A^2 \mid \{\beta_{1,j}\}] \approx$$
$$\frac{T_c^2 P(K-1)}{4L} \sum_{j=0}^{L-2-Z_{cz}} \sum_{m=j+1+Z_{cz}}^{L-1} \{\beta_{1,j}^2 [E[C_{k,1}^2(j-m)]+$$
$$E[C_{k,1}^2(N+j-m)]]+\beta_{1,m}^2 E[C_{k,1}^2(m-j-N)]+$$
$$E[C_{k,1}^2(m-j)]]\} = \frac{U_1 PNT_c^2 (K-1)(L-1)}{4L} \qquad (16)$$

Probability of error [Bit error rate (BER)]

The signal to interference and noise ratio (SINR) can be written as,

$$Y = \frac{D^2}{2Var(Z)} = \frac{D^2}{2[Var(A)+Var(s)+Var(\eta)]} \qquad (17)$$

Where Var (Z) is the variance of r.v. Z.

Based on (13), (16), and (17), it is not hard to get

$$Y = \left[\frac{L}{\bar{\gamma}_b U_1} + \frac{4\chi}{NU_1^2} - \frac{2\mu}{N^2 U_1^2} + \frac{(K-1)(L-1)}{NL}\right]^{-1} \qquad (18)$$

Where $\bar{\gamma}_b$ = PTL/N_0 is the average received signal to AWGN ratio.

All items in (15) and (16) are assumed to be zero-mean Gaussian r.v. The conditioned probability of error can be written as,

$$P(e \mid \{\beta_{1,j}\}) = 0.5 erfc(\sqrt{Y}) \qquad (19)$$

For GO codes, however, it is hard to obtain the analysis results of E $[C^2_{k,j}(m)]$. To evaluate the error performance, the value of E $[C^2_{k,j}(m)]$ would be obtained by numerical computation for certain GO codes. Therefore, by randomly generating the Rayleigh r.v. $\beta_{1,j}$, j = 0, 1, . . . , L-1, Y can be evaluated according to (18), and the conditioned error probability can be obtained based on (19). Averaged on more than 10 millions of conditioned error probability, the averaged error probability P (e) can be obtained.

CHANNEL PREDICTION

In this section, we discuss the procedure for prediction of fading channel impulse response at MU for the downlink using Burg's algorithm assuming ideal channel conditions as inputs in order to estimate the performance of the system in FDD mode. In our work, we predict the channel information for the next time slot based on ideal channel samples of the previous slots at the MU. We assume that predicted channel information is available at BS through control channel fed by MU. We also assume that the feedback delay corresponds to one time slot duration which would be optimistic.

The objective of the prediction operation is to forecast future values of the fading channel coefficients ahead. To accomplish this task, a linear prediction method based on the autoregressive model (AR) of fading is proposed [8]. Assume that the equivalent complex Rayleigh fading process $h(t)$ is sampled at the rate $f_s = 1/T_s$, where f_s is at least twice the maximum Doppler shift. The linear MMSE prediction of the future channel sample $h(n)$ based on its G previously estimated channel samples $\{h(n-1), h(n-2), ..., h(n-G)\}$ can be determined as

$$\hat{h}(n) = \sum_{j=1}^{G} a_G(j) h(n-j) \qquad (20)$$

Where $a_G(j)'s$ are the coefficients of the prediction filter and G is the order of the predictor [8]. The optimum values of these coefficients are computed using Burg's method. This method can be viewed as an order-recursive least squares lattice method, based on the minimization of the forward and backward errors in linear predictors with the constraint that these coefficients satisfy the Levinson-Durbin recursions [8]. The Burg's algorithm is summarized as follows:

Step 1. Initialize the forward and backward prediction errors with the estimated channel coefficients.

$$f_0(n) = q_0(n) = h(n), \tag{21}$$

Step 2. For m = 1, 2, 3,….., G , compute the following:

$$f_m(n) = f_{m-1}(n) + K_m q_{m-1}(n-1), \tag{22}$$

$$q_m(n) = K_m^* f_{m-1}(n) + q_{m-1}(n-1) \tag{23}$$

Where

$$K_m = \frac{-\sum_{n=m}^{M-1} f_{m-1}(n) q_{m-1}^*(n-1)}{0.5\sum_{n=m}^{M-1}[|f_{m-1}(n)|^2 + |q_{m-1}(n-1)|^2]} \tag{24}$$

Here, K_m is the m^{th} reflection coefficient of the lattice filter. 'M' is the number of channel estimates used as an input for estimating the lattice filter. The denominator of the above equation is the least square error. This is minimized by computing the prediction coefficients such that, they satisfy the recursive equation given by

$$a_m(k) = a_{m-1}(k) + K_m a_{m-1}^*(m-k) \quad for 1 \le k \le m-1 \tag{25}$$

The $a_G(j)'s$ obtained from the above algorithm are substituted in (22) to obtain the predicted channel sample at time 'n'.

SIMULATION RESULTS

To see the performance of the Pre-Rake diversity combining under channel estimation errors, we consider a

Table-1. Parameters used in simulation.

Parameter	Value
Bit rate	65000 bits/sec
Spreading code types	Random, Walsh, and GO codes
Process gain	64
Modulation	BPSK
Doppler frequency	50, 100, 150 Hz
Number of paths	20
Power delay profile	As per IEEE standards
Frame length	10 ms
Number of slots	15
Feedback delay	0.667 ms (1 time slot)

DS-CDMA system with time varying Rayleigh fading channel. The absolute path gains are i.i.d. Rayleigh random variables and angles are i.i.d. uniformly distributed in $[0, 2\pi]$. The channel tap weight vector is normalized such that $\|h\| = 1$. Simulation parameters used in simulation are given in Table-1.

Subsections on results are organized as follows: First, we compare and discuss the performance of the multiuser Pre-Rake system with various spreading sequences under ideal channel conditions. Subsequently, the system is evaluated under imperfect channel conditions in FDD mode.

Multiuser pre-rake system with various spreading sequences

In Figure-2, the BER versus Eb/No performance of the Pre-Rake system is compared with that of Rake system for random, orthogonal and GO codes under ideal channel conditions. Clearly, when GO codes are used, there is substantial improvement in the performance of both the systems which can be explained by the fact that the SI and MAI arriving within the orthogonal zone are eliminated when compared with other two codes, with Rake receiver performing better than the Pre-Rake. However, the difference is marginal, and the use of Pre-Rake is still justified to reduce the cost and size of the MU. Next, the performance of both the systems is moderate with orthogonal codes and both the systems perform similarly at higher values of Eb/No. At lower side, the difference is not much and hence use of Pre-Rake is again justified. At Eb/No of 30 dB, the BER is approximately 6 $\times 10^{-04}$ for GO codes and 10^{-02} for orthogonal codes. When random codes are used, the Pre-Rake receiver greatly over performs the Rake receiver since the latter maximum ratio combines the interference as well along with desired signal.

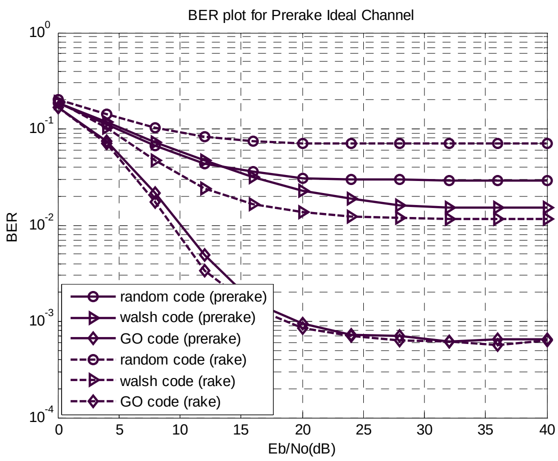

Figure-2. BER performance of ideal multiuser pre-rake system with various spreading sequences. The system parameters are N=64, L=20, K=16 and Z_{cz} = 2.

Multiuser pre-rake system with predicted channel in FDD mode

In FDD mode, the channel is predicted at the MU for the current slot with lattice filter of length, G=8 and fed to the BS. We have considered one time slot feedback delay which is typically used in 3G standards. In Figure-3, BER versus Eb/No performance of the system with predicted channel is compared with that of ideal channel using both orthogonal and GO codes. Figure-4 shows the BER versus number of user's performance of the Pre-Rake system using both types of codes. System performance saturates as the number of users increases. In both the graphs, the system performance with predicted channel has coincided with that of ideal channel showing that our prediction method is more effective. However, in both the cases, the performance with GO codes is much better due

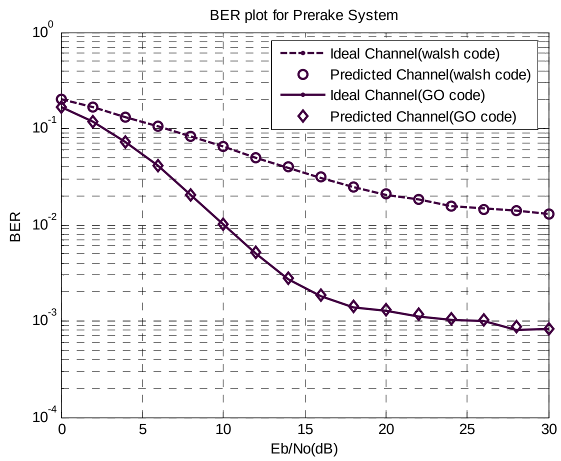

Figure-3. BER performance of pre-rake system versus Eb/No with predicted channel in FDD mode. The system parameters are N=64,L=20, K=16, and Z_{cz} =2.

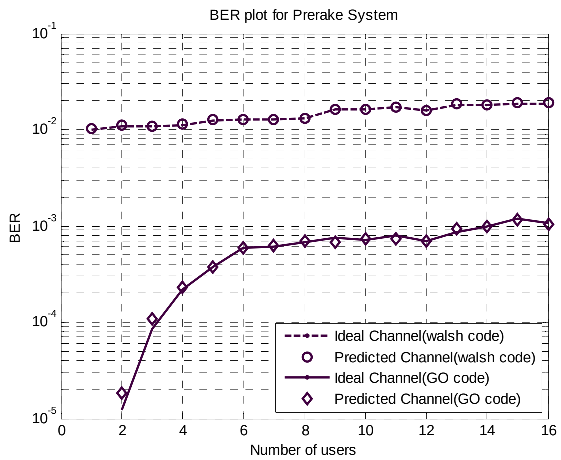

Figure-4. BER performance of pre-rake system versus number of users with predicted channel in FDD mode. The system parameters are N=64, L=20, andZ_{cz}=2.

to the reason given in Section 6.2. At BER of 2 x10^{-02}, Eb/No is better by about 12 dB and at 16 users BER is better by 20 times when GO codes are used.

CONCLUSIONS

To simplify the complexity of MU, Pre-Rake processing at the BS makes the transmitted signal more complex than the conventional Rake system, which introduces more interference at the receiver. The generalized orthogonal codes will help the Pre-Rake system to reduce the self and multiple access interference effectively because of their zero correlation zone property. It is shown that the performance of the system has improved significantly with generalized orthogonal codes as compared to that with the conventional orthogonal codes. An effective method for channel prediction is presented and the performance of the system under predicted channel conditions with generalized orthogonal codes is compared to that with conventional orthogonal codes for FDD/DS-CDMA systems. It is shown that the performance with predicted channel is close to that of ideal channel for both the codes which indicates that the prediction method we have proposed is effective and can be applied in realistic scenario. At BER of 2x10^{-02}, Eb/No is better by about 12 dB and at 16 users BER is better by 20 times when GO codes are used.

REFERENCES

[1] Ke Wan Li Hao and Pingzhi Fan. 2007. Performance Analysis of Pre-Rake Diversity with Generalized Orthogonal Codes for DS-CDMA. IEEE Trans. on Vehicular Technology. 56(4): 1896-1901.

[2] Essam E. Sourour, Tamer A.Kadous and Said E. El-Khamy. 1997. The Performance of TDD-CDMA Systems Using Pre-Rake Combining with Different Diversity Techniques and Imperfect Channel Estimation. 2nd IEEE symposium on computers and communications. pp. 280-284. July.

[3] Said E. El-Khamy, Essam E. Sourour and Tamer A. Kadous. 1997. Wireless Portable Communications using Pre-Rake CDMA/TDD/QPSK Systems with Different Combining Techniques and Imperfect Channel Estimation. 8th IEEE intl. Symposium on PIMRC. 2: 529-533. September.

[4] Yue Song and Yang Xiao. 2002. The Performance of Prerake Diversity Combining Under Imperfect Channel Estimation based on QPSK Modulation in TDD CDMA System. Proceedings of IEEE TENCON. pp. 1126-1129.

[5] Norharyati binti Harum, Yuh Tamura, Sigit P. W. Jarot, Ria Esmailzadeh and Masao Nakagawa. 2005. Analysis of Performance Degradation due to Channel Estimation Error in Pre-Rake TDD/CDMA. IEICE Trans. Comm. Vol. E88-B. No.6. June.

[6] Pingzhi Fan. 2004. Spreading Sequence Design and Theoretical Limits for Quasi-synchronous CDMA Systems. EURASIP Jr. on wireless comm. and networking. pp. 19-31.

[7] Rathinakumar and A.K. Chaturvedi. 2004. Mutually Orthogonal Sets of ZCZ Sequences. Electronics Letters. 40(18): September.

[8] C. Subhas, K. Rama Naidu and Y. Venkatarami Reddy. 2009. Performance Degradation due to Imperfect Channel Conditions in Pre-Rake DSSS Systems. International Journal of Systemics, Cybernetics and Informatics. Pentagram Research Publications, Hyderabad, India. ISSN 0973-4864. pp. 23-29. April.