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Performance Degradation in Pre-rake Frequency-division Duplex/ Direct Sequence-code Division Multiple Access Systems

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ABSTRACT

The transmitter-based pre-rake diversity combining technique reduces the complexity, size and cost of the mobile unit (MU), while achieving the same inter symbol interference (ISI) mitigation effects of rake receiver for direct sequence-code division multiple access (DS-CDMA) systems. The technique is based on preprocessing of transmitted signal relying on knowledge of the channel state information (CSI) before transmission. In most of the previous works, this *a priori* information is either assumed or estimated for the uplink and the same is applied to the downlink in time division duplex (TDD) systems due to channel reciprocity. In this paper, a method for channel prediction to evaluate the pre-rake system using binary phase-shift keying (BPSK) modulation in frequency-division duplex (FDD) through analytical and computer simulations for DS-CDMA downlink has been proposed. The performance of the system was also evaluated under ideal and predicted channel conditions using different spreading codes. The findings will have widespread applications in defence communication equipment.

Keywords: Direct sequence-code division multiple access, DS-CDMA, rake receiver, pre-rake, inter-symbol interference, frequency-division duplex, FDD, channel prediction

1. INTRODUCTION

Direct sequence-code division multiple access (DS-CDMA) offers many advantages, over other modulation techniques, such as anti-jamming, low probability of intercept, and resilience to multipath propagation, etc[1], which have wide spread applications in defence. One interesting feature of DS-CDMA systems is the use of rake receivers in multipath environment to achieve multipath diversity gain[2-3]. However, DS-CDMA suffers from inter-symbol interference (ISI) due to multipath propagation.

Receiver-based rake diversity combining technique will effectively combat 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 using pre-rake diversity combining technique, thereby reducing the complexity of the MU (only a matched-filter to the own spreading sequence is required) with equal performance[4-5].

Pre-rake technique for direct sequence spread spectrum (DSSS) communication systems using differential phase shift keying (DPSK) modulation scheme was discussed by Kadous[6], *et al.* and Esmail-Zadeh[7], *et al.* for single user. The rake function is performed pre-transmission in pre-rake. It is shown that performance of pre-rake technique is on par with rake technique, provided that the good channel estimates are available before transmission. Additional advantages with pre-rake method are reduced cost and complexity, and increased reliability of the MU. Multiuser pre-rake estimates are available before transmission. Esmail-zadeh[8], *et al.* and Song[9], *et al.* have shown that the performance of the pre-rake receiver (pre-rake function in transmitter and single-finger correlator in the receiver) is comparable to the rake receiver when orthogonal spreading codes are used, while it outperforms the rake receiver when non-orthogonal codes are used. Wan[10], *et al.* have shown that the pre-rake with generalised orthogonal (GO) codes outperforms the one with other spreading codes.

A novel diversity combining technique called fore-partial combining (FPC) with GO codes is discussed by Wan[11], *et al.* wherein only few first-arriving paths are considered. It can be seen clearly that this method outperforms all existing multipath diversity techniques such as selection combining (SC), maximum ratio combining (MRC) and equal gain combining (EGC). To tackle ISI problem, an adaptive interference-avoidance technique with pre-rake diversity combining is suggested by Liao[12], *et al.* for high-data-rate ultra wide band (UWB) systems. Torabi[13], *et al.* have proposed two novel pre-equalisation schemes for multiple-input single-output (MISO) direct sequence-UWB systems with pre-

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rake combining. The first scheme employs one pre-equalisation filter (PEF) per transmit antenna, whereas in the second scheme, the simplified PEF (SPEF) scheme, all transmit antennas share the same PEF. In all the above works, the true downlink channel is assumed to demonstrate the performance of the various techniques proposed. Moreover, only time division duplex (TDD) mode of operation is considered.

In this paper, the performance of multiuser pre-rake system is evaluated using BPSK for various spreading sequences including gold codes. The study of pre-rake technique in FDD mode has been extended 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, a very effective prediction method for obtaining CSI has been proposed.

The findings of the study have widespread applications in defence equipment. They can be applied in any practical wireless environment to effectively mitigate interference effects, that develop due to harsh nature of the wireless medium, for reliable transmission between the communicating devices. In general, these methods can be applied in radar, SATCOM, electronic warfare, and communication equipment. More specifically, the findings can be used in mobile communication system (MCS), target update transmitter (TUT) and C[4]I equipment used in defence for reliable communication.

2. MULTIPATH CHANNEL MODELLING

The simplified tapped delay line multipath channel model given by Proakis[1] has been used. The uplink channels are assumed to be statistically independent for all users. Also, with the utilisation of uplink power control, it is assumed that all channels are statistically identical, even if the MUs are at different distances from the BS. The complex low-pass impulse response of the channel of user k is given by

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

where, L is the number of channel paths, the path gains $\beta_{k,l}$ are independent identically distributed Rayleigh random variables for all k and l, the angles $\gamma_{k,l}$ are independent identically distributed uniformly distributed in $[0,2\pi]$, and T_c is the chip duration. Without any loss of generality, one can take normalisation $E[\beta_{k,l}^2]=1$. In a TDD system, under slow-fading conditions, which are typical for portable communication systems, it was assumed 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 $h_k(t)$, it estimates $h_k(t)$ for use in its own rake receiver. It has been assumed that $h_k(t)$ will not change when the BS transmits the following time slot to the MU of user k. For FDD mode, the downlink channel is predicted.

3. PERFORMANCE ANALYSIS OF PRE-RAKE SYSTEM

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 synchronised 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[8]. Hence, the receiver in both BS and MUs must be equipped with sufficient rake fingers and the corresponding channel estimation process.

In TDD mode, the fact that for a period of slot time, the channel impulse response is the same for the up and down links, can be utilised. Hence, only the BS needs to estimate it. As shown in Fig. 1, to pre-rake the down-link signal of a user, the BS multiplies this signal by 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 these to estimate the uplink channel impulse response. Alternate method is to use dedicated pilot channels. In FDD mode, the downlink channel is estimated and the channel information for the next slot is predicted based on estimated channel samples of previous slots at the MU receiver and the information is fed to the BS through control channel. Channel estimation and prediction errors cause performance degradation in both the pre-rake and

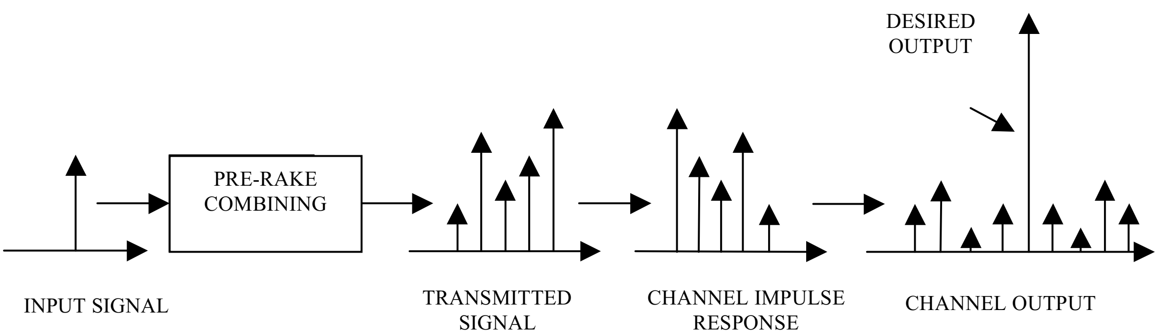

Figure 1. Pre-rake combination process.

the conventional rake systems. For ideal channel, the conventional transmitted signal without pre-rake with binary phase-shift keying (BPSK) modulation is given by

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

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 independent identically distributed 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 1. $a_k(t)$ 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 Eqn (2) the downlink transmitted signal will be

$$s_k(t) = \sqrt{\frac{2P}{U_k}} \sum_{l=0}^{L-1} \beta_{k,L-l-1}\, b_k(t-lT_c) a_k(t-lT_c). \exp\left(j\omega(t-lT_c) - j\gamma_{k,L-l-1}\right) \quad (3)$$

where, U_k is a normalising factor that keeps the instantaneous transmitted power constant regardless of the number of paths. U_k is given by

$$U_k = \sum_{n=0}^{L-1} \beta_{k,n}^2 \quad (4)$$

Note that the uplink channel impulse response is independent for all users. The received signal by user k is Eqn (3) convolved with the channel impulse response of user k. This produces a strong peak at the output of the channel, equivalent to the conventional rake receiver's combining. Therefore, the receiver of the MU does not need to estimate the channel impulse response and can only use one rake finger tuned to this peak. The pre-rake concept is shown graphically in Fig. 1 for a $\delta(t)$ transmitted signal. It is worthwhile to mention that an optimal receiver for this pre-raked signal must provide a match with all components at the output of the channel. However, since the pre-rake system puts most of the power in one component, matching with this component provides sufficient performance.

Assuming a CDMA system with K users, the received signal at i^{th} user during the downlink time slot is given by

$$r_i(t) = n(t) + Re \sum_{k=1}^{K} \sum_{j=0}^{L-1} \beta_{i,j} s_k(t-jT_c) \exp(j\gamma_{i,j}) \quad (5)$$

where, $n(t)$ is the zero-mean additive white Gaussian noise (AWGN) with two-sided power spectral density $N_0/2$. Using Eqn (3) in Eqn (5), one can see that the channel output includes $2L-1$ paths with a strong peak when $j+l = L-1$. Only one rake finger is needed in the MU to synchronize to this path. Without loss of generality, one can assume that $i = 1$ as the desired user.

The output of the finger employed in the receiver of user $i = 1$ is given by

$$Z = \int_{(L-1)T_c}^{(L-1)T_c+T} r_1(t)\; a_1(t-(L-1)T_c). \cos[\omega t - \omega T_c(L-1)]\, dt = D + S + A + \eta \quad (6)$$

where, η is a zero-mean Gaussian random variable with variance $N_0T/4$. D is the desired part for the current bit given by the k=1 part of $r_1(t)$ and $j+l=L-1$ in Eqn (6). After simple manipulations, D is found as

$$D = \sqrt{\frac{P}{2}}\, b_1^0 T \sqrt{U_1} \quad (7)$$

S and A in Eqn (6) are the self- and multiple-access interference, respectively. Using the Gaussian approximation, S and A are treated as Gaussian random variables, and it is readily shown that these are statistically independent with zero mean. Hence, one is only interested in their variances. Similar to the methodology given by Esmail-zadeh[8], *et al.* the variances of S and A and are evaluated conditioned on $\{\beta_{1,l}\}$ and the final results are averaged over $\{\beta_{1,l}\}$.

3.1 Self-interference

This interference exists even in a single-user system and is caused by the multipath. From Eqns (3)-(5), S is found by putting k=1 (self-interference), $m=L-l-1$ and $j+l \neq L-1$ (undesired part). S can be written after some manipulations as:

$$S = \sqrt{\frac{P}{2U_1}} \sum_{j=0}^{L-1} \sum_{m=0,\neq j}^{L-1} \beta_{1,j}\beta_{1,m} \cos[\omega T_c(j-m) + \gamma_{1,m} - \gamma_{1,j}] \int_0^T b_1(t-(j-m)T_c) a_1[t-(j-m)T_c] a_1(t)\, dt \quad (8)$$

The summation of Eqn (8) includes correlated terms when j and m swap their values. The following identity is used:

$$\sum_{j=0}^{L-1} \sum_{m=0,\neq j}^{L-1} f(j,m) = \sum_{j=0}^{L-2} \sum_{m=j+1}^{L-1} f(j,m) + f(m,j) \quad (9)$$

For any function $f(j, m)$. Also one has

$$\int_0^T b_1(t-mT_c) a_1[t-mT_c] a_1(t)\, dt = \begin{cases} T_c\left[b_k^{-1} C_{k,1}(m-N) + b_k^0 C_{k,1}(m)\right] & m \geq 0 \\ T_c\left[b_k^0 C_{k,1}(m) + b_k^1 C_{k,1}(N+m)\right] & m < 0 \end{cases} \quad (10)$$

where, $C_{k,1}(m)$ is the discrete aperiodic cross-correlation function defined by Song and Xiao[9]. Denoting $C_{i,i}(m)$ by $C_i(m)$ and utilising $C_i(m)=C_i(-m)$, one can get

$$S = T_c \sqrt{\frac{P}{2U_1}} \sum_{j=0}^{L-2} \sum_{m=j+1}^{L-1} \beta_{1,j}\beta_{1,m} \cos[\omega T_c(j-m) + \gamma_{1,m} - \gamma_{1,j}] \{b_1^{-1} C_1(N-m+j) + b_1^1 C_1(N-m+j) + 2b_1^0 C_1(m-j)\} \quad (11)$$

For any value of j and m, any term in Eqn (11) is a zero mean and all terms are uncorrelated due to the independent phase angles. Hence, their variances can be added. Taking the second moment of Eqn (11), one gets

$$E\left[S^2\middle|\{\beta_{1,l}\}\right] = \frac{PT_c^2}{2U_1}\sum_{j=0}^{L-2}\sum_{m=j+1}^{L-1}\beta_{1,j}^2\beta_{1,m}^2$$
$$[C_1^2(N-m+j) + 2C_1^2(m-j)] \quad (12)$$

3.2 Multiple-access Interference

The multiple-access interference A due to other users is found by the $k > 1$ part of Eqns (5) in (6) and can be written after some manipulation as

$$A = \sqrt{\frac{P}{2}}\sum_{k=2}^{K}\sum_{j=0}^{L-1}\sum_{m=0}^{L-1}\frac{\beta_{1,j}\beta_{k,m}}{\sqrt{U_k}}$$
$$\cos[\omega T_c(j-m) + \gamma_{k,m} - \gamma_{1,j}]$$
$$\int_0^T b_k[t-(j-m)T_c]a_k[t-(j-m)T_c]a_1(t)\,dt \quad (13)$$

Equation (13) can be divided into two parts–one for $m = j$ and one for $m \neq j$. The first part can be written as

$$A\mid_{m=j} = T_c\sqrt{\frac{P}{2}}\sum_{k=2}^{K}\sum_{j=0}^{L-1}\frac{\beta_{1,j}\,\beta_{k,j}}{\sqrt{U_k}}\cos(\gamma_{k,j}-\gamma_{1,j})b_k^0\;C_{k,1}(0) \quad (14)$$

For $m \neq j$, using Eqns (9) and (10), one obtains

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

It is easy to check that both Eqns (14) and (15) have zero mean and all terms are uncorrelated since all angles inside cos(.) are independent. Also, it is important to note that in Eqn (14) the full-period cross correlation $C_{k,1}$=0 if orthogonal codes such as the Walsh–Hadmard sequences are used. A pointer W is introducedfor this purpose. Taking the second moment of Eqns (14) and (15), one can get

$$E\left[A^2\middle|\{\beta_{1,l}\}\right] = \frac{PT_c^2Q}{4}\sum_{k=2}^{K}\;\{WC_{k,1}^2(0)\sum_{m=0}^{L-1}\beta_{1,m}^2\,.$$
$$\sum_{j=0}^{L-2}\sum_{m=j+1}^{L-1}\beta_{1,j}^2\,.\,[C_{k,1}^2(j-m) +$$
$$C_{k,1}^2(N+j-m)].\sum_{j=0}^{L-2}\sum_{m=j+1}^{L-1}\beta_{1,m}^2$$
$$[C_{k,l}^2(m-j-N) + C_{k,1}^2(m-j)]\} \quad (16)$$

where W is a pointer with W = 0 if orthogonal codes are used and W = 1 otherwise, and Q is given by

$$Q = Q_{k,j} = E\left[\frac{\beta_{k,j}^2}{U_k}\right] = \frac{1}{L}, \quad j = 0,1,2,\dots\dots\dots,L-1$$

Q is easily found by noting that since $\{\beta_{k,j}\}$ are independent identically distributed, $Q = Q_{k,0} = Q_{k,1} = Q_{k,2} = \ldots = Q_{k,l-1}$, and from the definition in Eqn(4), $Q_{k,0} + Q_{k,1} + Q_{k,2} + \ldots + Q_{k,l-1} = 1$ To evaluate Eqns (12) and (16), one can assume random code sequences. Random codes provide a comparison between the best and the worst code selections. In this case, all $C_{k,1}^{\;2}(m)$ are replaced in Eqns (12) and (16) by their expected values given by

$$E[C_i^2(m)] = N-|\,m\,|, \quad m \neq 0$$
$$E\left[C_{k,j}^2(m)\right] = N-|\,m\,|$$
$$E\left[C_{k,i}(m)C_{k,i}(n)\right] = 0 \quad m \neq n\;\&\;k \neq i. \quad (17)$$

The last part in Eqn (17) will be needed later. If orthogonal codes are used, W=0 was set [or equivalently $C(k, i)(0)$=0 was taken. However, for the out-of-phase cross correlations of orthogonal codes [i.e., $m \neq 0$ in Eqn (17)], still Eqn (17) is used. Using Eqn (17) in Eqns (12) and (16), and assuming Z in Eqn (6) as Gaussian random variables, one can find the probability of error conditioned on $\{\beta_{1,n}$, n=0, 1, 2,..., $L-1\}$ is given by

$$P(\,e\mid\{\beta_{1,n}\}\,) = 0.5\,erfc(\sqrt{Y}) \quad (18)$$

where, Y is the signal-to-noise ratio, given by $Y=D^2/2\mathrm{Var}(Z)$, where, Var(Z) is the variance of the random variables Z. Noise means AWGN noise, self-interference, and multiple-access interference. After some manipulation, Y is found to be

$$Y = \left[\frac{L}{\overline{\gamma_b}U_1} + \frac{4\chi}{NU_1^2} - \frac{2\mu}{N^2U_1^2} + \frac{(K-1)(L-1+W)}{NL}\right]^{-1} \quad (19)$$

where, $\overline{\gamma_b} = PTL/N_0$ is the average received signal to AWGN ratio, $\sum_{j=0}^{L-2}\sum_{m=j+1}^{L-1}\beta_{1,j}^2\beta_{1,m}^2$, and $\mu = \sum_{j=0}^{L-2}\sum_{m=j+1}^{L-1}(m-j)\,\beta_{1,j}^2\beta_{1,m}^2$. The probability density functions of x and μ are very difficult to obtain. The final probability of error $P(e)$ is evaluated from Eqns(18) and (19) using the Monte Carlo integration[9]. At each iteration, L Rayleigh r.v.s are computer generated, U_1, x and μ are evaluated, and Y is found and substituted in Eqn(18). For each value, Eqn (18) is averaged over a sufficiently large number of iterations[9].

4. CHANNEL PREDICTION

The procedure for prediction of fading channel impulse response at MU for the downlink using Burg's algorithm assuming ideal channel conditions as inputs to estimate the performance of the system in FDD mode is discussed here. If an increased signaling load is acceptable, the channel can be known by the BS before transmission in FDD mode, provided CSI is regularly sent by the MU through an uplink signaling channel. This is naturally possible only as long as the channel does not change too rapidly, i.e., if coherence time is much higher than the feedback signaling interval. In this paper, the channel information for the next time slot based on ideal channel samples of the previous slots at the MU is predicted. It is assumed that predicted channel information is available at BS through control channel fed by MU. It is also assumed 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[14]. 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 $\{h(n-l), h(n-2), \ldots, h(n-G)\}$ can be determined as

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

where, $a_G(j)$'s are the coefficients of the prediction filter and G is the order of the predictor[15]. 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 minimisation of the forward and backward errors in linear predictors with the constraint that these coefficients satisfy the Levinson-Durbin recursions[16]. The Burg's algorithm is summarised as follows:

Step 1. Initialise 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 and 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 minimised by computing the prediction coefficients such that, they satisfy the recursive equation given by

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

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

The channel prediction in FDD mode is considered to compensate the feedback delay. Indeed, the channel prediction may be necessary even in TDD mode, if the time slot is sufficiently long, due to small delays involved in switching of the same channel between uplink and downlink and also due to delays involved in processing. Prediction techniques can also be used in place of pilot-based estimation techniques which result in saving of channel bandwidth.

5. SIMULATION RESULTS

To see the performance of the pre-rake diversity combining under channel prediction errors, a DS-CDMA system with time-varying Rayleigh fading channel was considered. The absolute path gains were independent identically distributed Rayleigh random variables and angles are independent identically distributed uniformly distributed in [0, 2]. Other parameters considered are BPSK with carrier frequency of 2 GHz, spreading sequences with process gain N of 64, slot rate of 1500 Hz, symbols per slot of 40 (symbol rate = 60 kbps), feedback delay in FDD mode of 0.66667 ms (1 slot duration), number of users $K = 20$, and linear predictor length G of 8. The BPSK data vector b is spread by spreading sequence vector a and transmitted over the downlink. The channel tap weight vector is normalised such that $||h|| = 1$. The power and delay profiles adopted are as per SUI standards and have a delay spread (number of paths) of $L = 20$ chips. Simulation studies were carried out at 100 Hz Doppler frequency. Software tool used for simulation was MATLAB 7.0.

Subsections on results have been organised as follows: First, the analytical and simulation results of single user pre-rake diversity combining with rake and conventional (without any equalisation) receivers were compared and discussed. Next, performance of the multiuser pre-rake system was evaluated for various spreading sequences, viz., random, orthogonal and gold codes, and subsequently, the system was evaluated in FDD mode under channel prediction errors. Here, the term 'error' means, it is the deviation from the ideal value.

5.1 Single User Pre-rake System

In Fig. 2, bit error rate (BER) performance of single user pre-rake scheme has been evaluated for time-varying channel at 100 Hz Doppler frequency and compared with that of a rake receiver having the same number of fingers. The results revealed that the performance of pre-rake system is quite similar to that of the rake system. Hence with pre-rake technique, one can reduce complexity and size of the MU while retaining the same performance of rake receiver. Also it was observed that the pre-rake simulation and analytical results are close to each other at higher values of Eb/No that are typically used in realistic scenario. From the Fig. 2, it was also observed that at lower Eb/No values, small deviation is justified since the computer-generated Rayleigh random variables are used for analysis as justified at the end of Section 3. At BER of $1\text{x}10^{-4}$, the difference between analysis and simulation results is almost zero. To set the performance limit, BER curve is also shown for conventional receiver (unequalised Rayleigh fading channel). It is noted that the performance of the pre-rake scheme

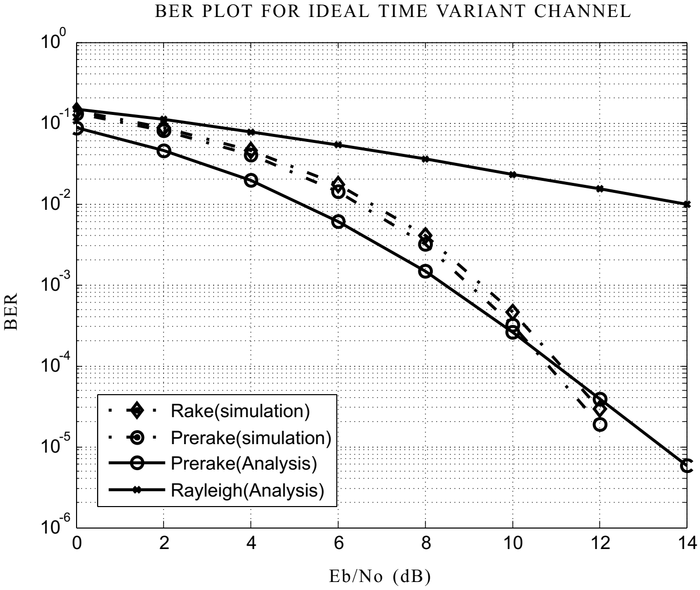

Figure 2. BER versus Eb/No performance of pre-rake, rake, and conventional receivers. The system parameters are N=13, and L = 17.

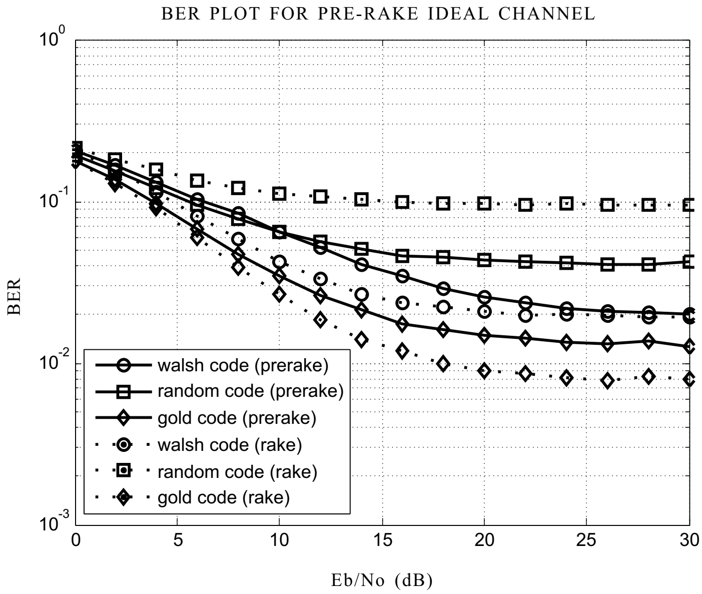

Figure 3. BER performance of ideal pre-rake system with various spreading sequences. The system parameters are N=64, L=20, and K = 20.

is much superior to that of conventional receiver. At Eb/No of 11 dB, BER is 1×10^{-4} for pre-rake and rake, and 2×10^{-2} for conventional receiver. Hence, pre-rake system performs much better than the conventional receiver.

5.2 Multiuser Pre-rake System with Various Spreading Sequences

In Fig. 3, the BER versus Eb/No performance of the pre-rake system has been compared with that of rake system for random, orthogonal and gold codes. Clearly, when gold codes are used, the performance is the best for both the systems, since these codes will have low autocorrelation side-lobes and low cross-correlations[17], 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, 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 justified again. When random codes are used, the pre-rake receiver greatly over performs the rake receiver since the latter rake combines the interference as well along with desired signal.

5.3 Multiuser Pre-rake System with Predicted Channel in FDD Mode

In FDD mode, since the uplink and downlink channels are different, the downlink channel is assumed to be ideal, predicted at the MU for next time slot and the information is fed to the BS. Conversely, channel can be predicted at BS with ideal values fed by the MU. In the first case, predicted channel values will be accurate as the previous ideal channel values required for prediction are stored at MU itself avoiding possible errors in the feedback path. But it is at the cost of increased MU complexity. It is other way around in the second case. Either of the techniques can be applied depending upon the situation. One has considered one-time slot feedback delay which is typically used in 3G standards. The channel is predicted at the MU for the current slot with lattice filter of length, G = 8. In Fig. 4, BER versus Eb/No performance of the system with predicted channel has been compared with that of ideal channel using both orthogonal and gold codes. Figure 5 shows the BER versus number of users performance of the pre-rake system using both types of codes. System performance saturates as expected for users beyond 35

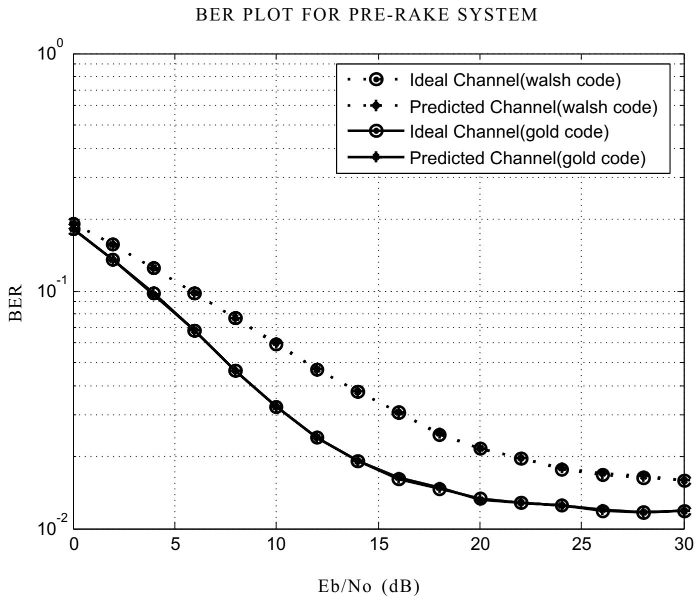

Figure 4. BER performance of pre-rake system versus Eb/No with predicted channel in FDD mode. The system parameters are N=64, L=20, G=8, and K = 20.

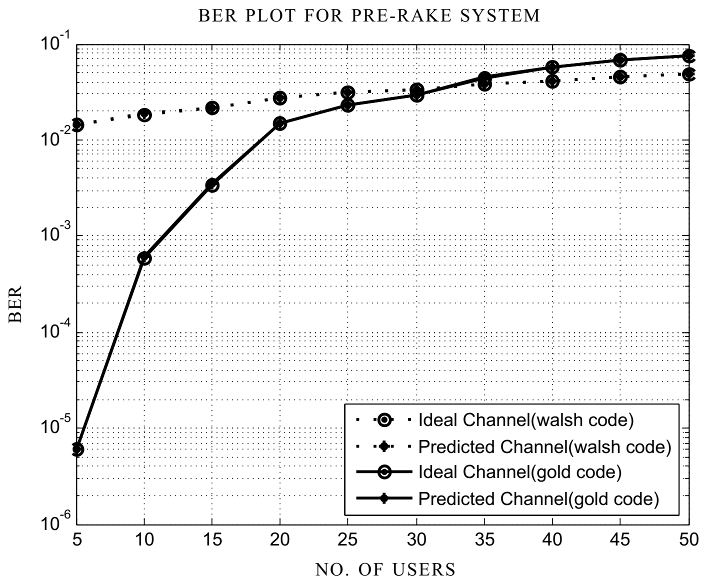

Figure 5. BER performance of pre-rake system versus number of users with predicted channel in FDD mode. The system parameters are *N*=64, *L*=20, and *G*=20.

with marginal difference between the two codes. In both the graphs, the system performance with predicted channel has perfectly coincided with that of ideal channel showing that our prediction method is efficient. However, in both the cases, the performance with gold codes is much better due to the reason given in Section 5.2. At BER of $2x10^{-2}$ Eb/No is better by about 8 dB and 15 more users can be accommodated at BER of $1.5x10^{-2}$, using gold codes.

6. CONCLUSIONS

The pre-rake system with random and orthogonal spreading sequences for DS-CDMA has been analysed, evaluated, and compared with gold codes. To simplify the complexity of MU, pre-rake processing at the BS makes the transmitted signal complex than the conventional rake system, which introduces more interference at the receiver. The gold codes can help the pre-rake system to reduce the self- and multiple- access interference effectively because of their low autocorrelation side-lobes and low cross-correlations. It is shown that the performance of the system has been improved significantly with gold codes as compared to that with the orthogonal codes. In this paper, the system's performance with channel prediction in FDD mode has also been evaluated. It is shown that the performance with predicted channel coincides with that of ideal channel which indicates that the proposed prediction method is efficient and can be applied in realistic scenario. It can be seen clearly that gold codes have performed better than orthogonal codes for predicted channel also. At BER of $2x10^{-2}$, Eb/No is better by about 8 dB and 15 more users can be accommodated at BER of $1.5x10^{-2}$ using gold codes. This work can be extended with some other superior spreading sequences like generalised orthogonal codes for better performance. The findings have wide spread applications in mobile communication systems and target update transmitter developed and used by defence. In fact, spread spectrum techniques were first invented for defence applications which subsequently spread to commercial sector.

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