Transmitter-Based Filters for Interference Mitigation under Imperfect Channel Conditions for the FDD / DS - CDMA Systems

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ABSTRACT

Precoding techniques reduce complexity, size and cost of the mobile stations, while achieving the same interference mitigation effects of receiver based methods for Direct Sequence Code Division Multiple Access (DS-CDMA) downlink. The technique is based on processing of signal before transmission relying on knowledge of the channel state information. In most of the previous works, this a priori information is assumed. In practice, there will be performance degradation due to time varying nature of the radio mobile channels and methods used for channel estimation/ prediction. In this paper, we will present the performance evaluation of transmitter-based filters as precoding elements for mitigating inter symbol interference and multiple access interference under imperfect channel conditions in frequency-division duplex mode for DS-CDMA downlink through analysis and computer simulations.

Keywords : channel prediction, DS-CDMA, frequency-division duplex, inter-symbol interference, multiple access interference, precoding.

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1. INTRODUCTION

Direct Sequence-CDMA (DS-CDMA) is one of the important techniques used in mobile communications. CDMA offers many advantages over other multiple access techniques, such as sharing of same frequency band by multiple users, no absolute limit on the number of users, scope for reduction of multipath fading, high data rates etc. However CDMA suffers from three major intracell problems: (i) inter symbol interference (ISI) due to multipath propagation, (ii) multiple access interference (MAI) due to degradation in the orthogonality of spreading codes caused by multipath effects and (iii) near-far interference when an undesired user has a high detected power as compared to the desired user. The near-far interference can be controlled by proper power control in the uplink [1].

Number of receiver based techniques like multiuser detection (MUD) methods are investigated to combat ISI and MAI and to increase data throughput. However these methods require channel state information (CSI) and complex signal processing at the mobile station (MS) for the downlink which make the MS more expensive and unreliable. Hence, for the downlink, one can transfer the signal processing for interference suppression from the MS receiver to the base station (BS) transmitter by using precoding techniques thereby reducing the complexity of the MS with similar or better performance [2]. Many types of receiver-based MUD methods and transmitter-based precoding techniques are discussed and compared in [3].

The precoding techniques are classified as blockwise or bitwise, based on realization of block diagram. The bitwise techniques demonstrate less computational complexity and better performance [2] and Pre-Rake is one such technique which mitigates ISI but fails to reduce MAI [4]. Block-wise precoding algorithms namely joint transmission (JT) and transmitter precoding (TP) are investigated in [5] and [6] respectively. A bitwise technique called zero forcing (ZF) decorrelating prefilter (DPF) is demonstrated in [7]. Various other types of bitwise and blockwise precoding techniques are discussed in [8] – [11].

In large part of the literature, the perfect channel is assumed to demonstrate the performance of the various precoding techniques proposed. However in practical scenario, the channel needs to be estimated for time-division duplex (TDD) mode and predicted for frequency-division duplex (FDD) mode at the receiver. In this paper we will present the performance evaluation of a new bitwise technique called transmitter-based filters (TF) under imperfect channel conditions for the downlink of DS-CDMA system in FDD mode.

Rest of the paper is organized as follows : Section 2 presents the system modeling used in the paper. In section 3, the system performance for mitigating interference using TF is investigated. Section 4 discusses methods used for channel prediction. Simulation results are compared and discussed in section 5. Finally we conclude our work in section 6.

2. System Modeling

The system considered here is a multiuser DS-CDMA with K active users in a cell communicating through a common BS. The BS has one antenna element and transmits synchronously over frequency selective channels to MSs. The discrete time transmission model that follows is sampled at the chip rate $1/T_c$. For the sake of clarity, we state that for any vector **x**, $\|x\|^2$ denotes the inner product $\mathbf{x}^T\mathbf{x}$ and for any matrix **X**, $X^{i,j}$ denotes the element in the i^{th} row and j^{th} column.

The l^{th} symbol $d_k^l c_k(n)$ for user k, $k \in \{1,2,....K\}$ is BPSK modulated data d_k^l spread with the spreading code $c_k(n)$. The spreading code consists of L_q chips and it is described by the vector $c_k = \left[c_k(0), c_k(1),....c_k(L_q-1)\right]^T$ which is normalized such that $\|c_k\| = 1$. The symbol interval is T_s such that $T_c = T_s / L_q$. The k^{th} channel response $h_k(n)$. It is considered as a finite-impulse response (FIR) filter whose chip spaced tap coefficients are the vector $h_k = \left[h_k(0), h_k(1),....h_k(L-1)\right]^T$, where L is the delay spread. Assuming that the l^{th} symbol of power u_k^l is transmitted for the k^{th} user, the total transmitted signal for the conventional CDMA downlink is described as

$$s(n) = \sum_{k=1}^{K} \sqrt{u_k^l}\, d_k^l c_k(n). \qquad (1)$$

To simplify the analysis, we will assume that all the users are transmitted with equal power and thus u_k^l is normalized to one. Furthermore, data are assumed to be independent from bit to bit and between users

$$E\left\{d_\alpha^{l_1} d_\beta^{l_2}\right\} = \begin{cases} 1, \text{ for } \alpha = \beta, \text{ and } l_1 = l_2 \\ 0, \text{otherwise.} \end{cases} \qquad (2)$$

Let d_k be a vector sequence of $D = 2B+1$ transmitted bits for user k, $d_k = \left[d_k^{-B}....d_k^{0}....d_k^{B}\right]^T$. The vector **d** of length KD contains bits of all the users and is defined as $d = \left[d_1^T...d_k^T.....d_K^T\right]^T$. By defining, now, the $DL_q \times D$ matrix C_k as $C_K = blockdiag(c_1,....c_k,....c_k)$ and the $DL_q \times DK$ matrix C as $C = [C_1.....C_k......C_K]$, the DL_q-length vector **s** for signal s(n) for a conventional CDMA downlink system and its corresponding total energy per

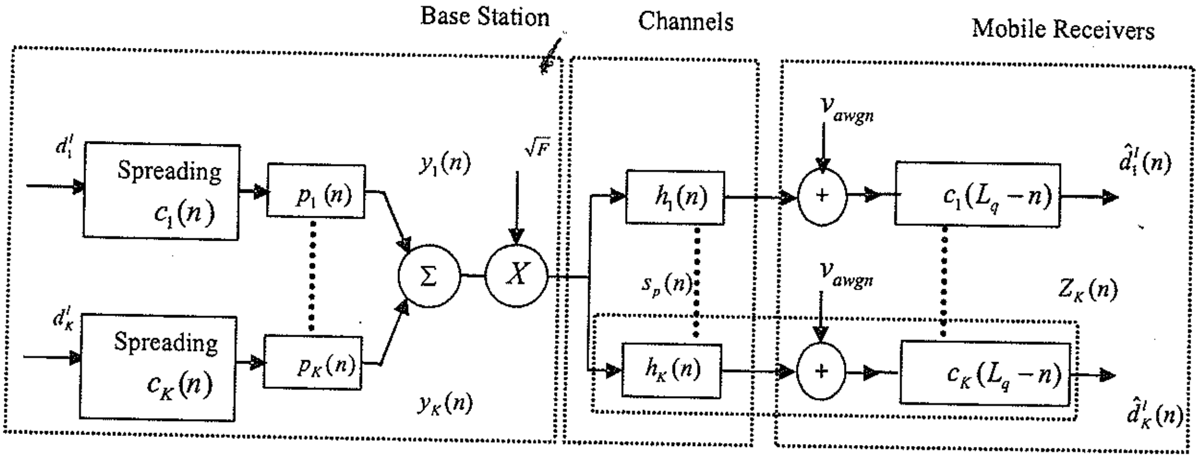


Figure 1 : Overall System with Inverse Filters to Mitigate Interference

transmitted symbol ε_c, are given by $s = \mathbf{C}\boldsymbol{d}$ and $\varepsilon_c = S^T s / D = K$, for $\|c_k\| = 1$, respectively.

The receiver is a filter matched to the spreading code of the desired user with impulse response $c_k(L_q - n)$. Additive white Gaussian noise (AWGN) **v**, with zero mean and correlation matrix $\sigma^2 I$, is added to the received signal. The multipath channel length is assumed to be slightly more than one symbol period and thus ISI is extended to the next two transmitted symbols. Throughout our analysis the users are synchronous (realistic for the downlink) and the BS has exact knowledge about all users spreading codes. The channel variation in wireless communications can of course be a problem, but accurate *a priori* CSI can be obtained even in a mobile environment if prediction techniques are employed. This is naturally possible only as long as the channel does not change too rapidly, i.e., if its coherence time is much larger than the feedback signaling interval in FDD systems [8].

Vector $p_k = [p_k(0), p_k(1), ... p_k(L_p - 1)]^T$ represents the tap coefficients of the k^{th} user specific FIR prefilter. The transmitted symbol is now expanded to $d_k^l c_k(n) * p_k(n) = d_k^l g_k(n)$. Written in vector form, $g_k(n)$ has a length of $L_g = L_q + L_p - 1$ chips and is denoted as $g_k = [g_k(0), g_k(1), ... g_k(L_g - 1)]^T$. Since the technique is bitwise, we define d_k^0 as the desired data bits. The channel imposes ISI on the system and we must take into account B transmitted bits before and after the desired one. The right choice for is $B = \langle (L_g - L_q) / L_q \rangle$, where $\langle x \rangle$ denotes that x is rounded up to the nearest integer. Let $y_k(n)$ be the signal at the output of the k^{th} prefilter as shown in Fig. 1. The L_g-length transmitted vector for user k, y_k is written as $y_k = G_k d_k$, where the $L_g \times D, (D = 2B + 1)$, matrix G_k is $G_k = \{G_k^{i,(j+B+1)}\}; i = 1, 2, ... L_g, \ j = -B, ... B$

$$G_K^{i,(j+B+1)} = \begin{cases} g_k(i - jL_q - 1), for \ 0 \le i - jL_q - 1 \le L_g - 1 \\ 0 \qquad ,otherwise \end{cases} \quad (3)$$

The transmitted signal s_p can be written as the L_g-length vector $s_p = \sum_{k=1}^{K} y_k = Gd$, where the $L_g \times DK$ matrix G is defined as $G = [G_1 ..... G_k ...... G_K]$.

3. Performance Of Transmitter-Based Filters For Mitigation Of Interference

The precoding technique, ignoring noise and assuming a unit gain channel, aims at $\hat{d}_k^l = d_k^l$. A certain transmit power is required for interference elimination which is usually greater than the one required for a conventional spreading. Thus, the transmitted signal s_p must be scaled with the appropriate factor $\sqrt{F}$. The reference power used is the one for the corresponding conventional CDMA system, ε_c. If $F < 1.0$, the scaling results in a decrease in the signal-to-noise ratio (SNR) at the receiver decision point. With this in mind, the optimum precoding technique is one that minimizes the MAI and ISI and maximizes F, which implies the necessity of a power constraint. If the unscaled total transmission power per symbol produced by the algorithm is denoted as E_f, then F is given by $F = \varepsilon_c / E_f$. To incorporate power constraint in our bitwise algorithm, the E_f must be expressed as a function of p, which is a KL_p-length vector of prefilter taps defined as $p = \left[p_1^T....p_k^T......p_K^T\right]^T$. It can be shown that

$$E_f(p) = p^T.Z^T.Zp \quad (4)$$

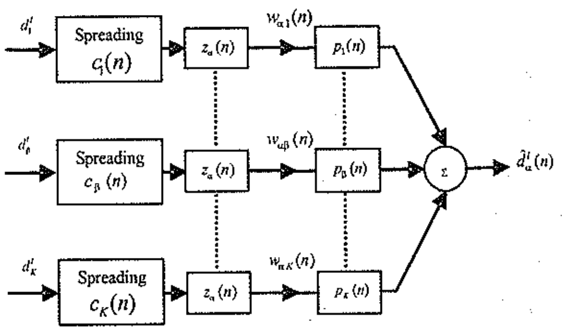
where matrix $Z = blockdiag(Z_1.....Z_k.....Z_K)$ is of dimension $K(L_q + L_p - 1) \times KL_p$ and matrix Z_k is of dimension $(L_q + L_p - 1) \times L_p$ and is defined as

$$Z_k = \{Z_k^{i,j}\};\ i = 1,2,....L_q + L_p - 1,\ j = 1,2,...L_p$$

$$Z_k^{i,j} = \begin{cases} c_k(i-j), & for\ 0 \le i-j \le L_q - 1 \\ o, & otherwise \end{cases} \quad (5)$$

Equation (4) is used in what follows to develop the TF algorithm using MMSE criterion with power constraint [12] in order to determine vector p. The method is essentially different from the ZF-DPF [7], although they follow the same block diagram.

For simplicity, we substitute the cascaded filters $h_k(n)$ and $c_k(L_q - n)$ with an FIR filter of length $Y = L_q + L - 1$ and impulse response $z_k(n) = h_k(n) * c_k(L_q - n)$, as shown in Fig. 1. After rearrangement (only to facilitate the mathematical analysis), the system can equivalently be illustrated as in Fig. 2 for the α^{th} receiver. Signals $w_{\alpha\beta}(n)$ are defined as "filtered reference signals" and are produced by passing the sequence of symbols $d_\beta^l c_\beta(n)$ through the filters with impulse response $z_\alpha(n)$ [13]. This block diagram arrangement is used to determine the optimal, in MMSE terms, FIR



filters, p_k. The fact that noise is not included in the analysis does not affect the solution because the noise is not filtered by the inverse filters. The power constraint term imposed in the solution gives good noise performance.

Vectors $w_{\alpha\beta}(n) = \left[w_{\alpha\beta}(n)......w_{\alpha\beta}(n - L_p + 1)\right]^T$ correspond to the filtered reference signals. Following the block diagram in Fig. 2, the $w_{\alpha\beta}(n)$ are passed through the prefilters p to yield the estimated desired data. The outputs $\hat{d}_0(n)$ can be written in matrix form as $\hat{d}_0(n) = W(n)p$ (6)

$$\hat{d}_0(n)=\begin{bmatrix}\hat{d}_1^0(n)\\ \hat{d}_2^0(n)\\ \vdots \\ \hat{d}_k^0(n)\end{bmatrix},$$

$$W(n)=\begin{bmatrix} w_{11}^T(n) & \dots & w_{1K}^T(n)\\ \vdots & \ddots & \vdots \\ w_{K1}^T(n) & \dots & w_{KK}^T(n)\end{bmatrix}. \quad (7)$$

We now seek the optimal filter vector P to minimize the time-averaged squared error between the actual and desired outputs. The desired outputs are sampled every symbol period, $T_{SS}=nL_q+\chi, n=1/2,\dots,$ where χ is an imposed delay to the system which assists the equalization of the multipath channels. In the final cost function, we must include, using Lagrange multipliers, the constraint that E_f [see (4)] should not exceed the reference transmitted power per symbol ε_c. The cost function to be minimized is

$$J=E\left[\left(d_0-\hat{d}_0(T_{SS})\right)^T\left(d_0-\hat{d}_0(T_{SS})\right)\right]+\lambda\left(E_f(p)-\varepsilon_c\right) \quad (8)$$

where E denotes the expectation and d_0 is the vector of desired data defined as $d_0=\left[d_1^0\dots d_k^0\dots d_K^0\right]^T$ By using (6) and after expansion, (8) takes the quadratic form

$$J=E\left[d_0^Td_0\right]-2E\left[d_0^TW(T_{SS})\right]p+ \\ p^TE\left[W^T(T_{SS})W(T_{SS})\right]p+\lambda\left(E_f(p)-\varepsilon_c\right) \quad (9)$$

From vector gradient theory we find that $\delta E_f(p)/\delta p=Z^TZp$. Therefore, the Wiener solution for the taps of the prefilters p_0 is given by

$$p_0=\left[E\left[W^T(T_{ss})W(T_{SS})\right]\right]+\lambda Z^TZ]^{-1}E\left[W^T(T_{SS})d_0\right] \quad (10)$$

The assumption that users are transmitted with powers $u_k^l=1$ does not affect the final Wiener solution. For different powers, the Wiener solution is still p_0 and the received powers will be proportionally different.

In order to complete the analysis, the expectation terms $E\left[W^T(T_{SS})W(T_{SS})\right]$ and $E\left[W^T(T_{SS})d_0\right]$ in (10) must be calculated. As stated before, $w_{\alpha\beta}=d_\beta^lc_\beta(n)*h_\alpha(n)*c_\alpha(L_q-n)$. Let $\psi_{\alpha\beta}(n)=c_\beta(n)*h_\alpha(n)*c_\alpha(L_q-n)$. The corresponding symbol in vector form is denoted as $\psi_{\alpha\beta}(n)=\left[\psi_{\alpha\beta}(0),\psi_{\alpha\beta}(1)\dots,\psi_{\alpha\beta}(L_\psi-1)\right]^T$ of length $L_\psi=2L_q+L-2$ due to the convolution effect. Vector $w_{\alpha\beta}(n)$, of length L_p, is given next as a function of the $D=2B+1$ vector d_β accounting for the ISI effect. To assist the illustration of the Wiener solution it is assumed that $\chi+L_q\le L_\psi, \chi+L_q\ge L_p$ (in fact they can be arbitrarily chosen). In (11), $\psi_{\alpha\beta}$ is a $L_p\times D$ matrix and $w_{\alpha\beta}$ is written as

$$w_{\alpha\beta}=\psi_{\alpha\beta}d_\beta \quad (11)$$

where $\psi_{\alpha,\beta}=\left\{\psi_{\alpha\beta}^{i,(j+B+1)}\right\}; i=1,\dots L_p, j=-B,\dots B$

$$\psi_{\alpha\beta}^{i,(j+B+1)}= \begin{cases}\psi_{\alpha\beta}\left(\chi+L_q-i-jL_q\right), & for\ 0\le\chi+L_q-i-jL_q\le L_\psi-1\\ 0, & otherwise\end{cases} \quad (12)$$

By combining (7) and (11) we obtain

$$W(T_{SS})=\begin{bmatrix} d_1^T\psi_{11}^T & \dots & d_K^T\psi_{1K}^T\\ \vdots & \ddots & \vdots\\ d_1^T\psi_{K1}^T & \dots & d_K^T\psi_{KK}^T\end{bmatrix} \quad (13)$$

Taking into account the assumption in (2), it is proved that

$$E\left[W^T(T_{SS})W(T_{SS})\right]=\begin{bmatrix}\sum_{\alpha=1}^K\psi_{\alpha1}\psi_{\alpha1}^T & 0 & 0\\ 0 & \ddots & 0\\ 0 & 0 & \sum_{\alpha=1}^K\psi_{\alpha K}\psi_{\alpha K}^T\end{bmatrix} \quad (14)$$

$$E\left[W^T(T_{SS})d_0\right]=\left[\bar{\psi}_{11}^T,\bar{\psi}_{22}^T,\dots,\bar{\psi}_{KK}^T\right]^T \quad (15)$$

where $\bar{\psi}_{\alpha\beta}$ is a L_p-length vector defined as $\bar{\psi}_{\alpha\beta}=\left[\psi_{\alpha\beta}^{1,(B+1)}\dots\psi_{\alpha\beta}^{L_p,(B+1)}\right]^T.\psi_{\alpha\beta}^{1,(B+1)}$ is given by (12). By replacing the expectation terms in (10) with the right-hand side of (14) and (15), the Wiener solution is fully determined in a closed form.

4. Channel Prediction

In this section, we discuss the procedure for prediction of fading channel impulse response using Burg's algorithm with ideal channel coefficients as inputs in order to asses the performance of the system in FDD mode. 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 MS 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 our work, we consider ideal CSI at the MS for each slot and assume that this information is available at the BS through an uplink signaling channel. We also assume that the feedback delay corresponds to one time slot, which would be optimistic, in getting the feedback from MS and predict CSI at BS accordingly for the current slot.

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 $f_{sa} = 1/T_{sa}$ where f_{sa} is at least twice the maximum Doppler shift. The linear MMSE prediction of the future channel sample $h(n)$ based on its L_d previously estimated channel samples $\{h(n-1), h(n-2).....h(n-L_d)\}$ can be determined as

$$\hat{h}(n) = \sum_{j=1}^{L_d} a_{L_d}(j) h(n-j) \qquad (17)$$

where $a_{L_d}(j)$'s are the coefficients of the prediction filter and L_d 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 minimization 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 summarized as follows:

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

$$f_0(n) = q_0(n) = h(n) \qquad (18)$$

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

$$f_m(n) = f_{m-1}(n) + Q_m q_{m-1}(n-1) \qquad (19)$$

$$q_m(n) = Q_m^* f_{m-1}(n) + q_{m-1}(n-1) \qquad (20)$$

where

$$Q_m = \frac{-\sum_{n=m}^{N-1} f_{m-1}(n) q_{m-1}^*(n-1)}{0.5\sum_{n=m}^{N-1}\left[\left|f_{m-1}(n)\right|^2 + \left|q_{m-1}(n-1)\right|^2\right]} \qquad (21)$$

Here, Q_m is the m^{th} reflection coefficient of the lattice filter. 'N' 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

$$a_m(k) = a_{m-1}(k) + Q_m a_{m-1}^*(m-k) \;\; for 1 \le k \le m-1 \qquad (22)$$

The $a_{L_d}(j)'s$ obtained from the above algorithm are substituted in (17) to obtain the predicted channel sample at time 'n'.

5. Simulation Results

To see the performance of the system with TF under channel prediction errors, we consider a DS-CDMA system with Rayleigh fading channel which is static during a slot. Other parameters considered are uncoded BPSK with carrier frequency f_c = 2GHz, Walsh spreading sequence with process gain L_q = 16, Slot rate $1/T_{slot}$ = 1500 Hz, Number of symbols in a slot = 160 (symbol rate = 240 kbps), Feedback delay in FDD mode = 0.66667 ms (1 slot time), and linear predictor length L_d = 8.

BPSK data symbols d_k^l are spread by Walsh binary spreading sequences c_k and transmitted synchronously over the downlink. The mobile radio channels are different for each user and the tap weights are normalized such that $\|h_k\| = 1$. Also the spreading sequence of each user is normalized such that $\|c_k\| = 1$. The power and delay profiles adopted are as per IEEE standards and has a delay spread of L = 19 chips. Simulation studies are carried out at 100Hz Doppler frequency. The filter length L_p is set at 32. The performance of the filter can be improved by increasing the value of L_p. The delay χ in sampling the output at the receiver is set to half of the filter length L_p. When the filter performance with respect to number of users is considered, Eb/No = 10 dB is taken. The Wiener solution as stated in (10) contains the power control term m $Z^T Z$. The value for λ is set at 0.05.

Subsections on simulation results are organized as follows: First, we discuss and compare the performance of the system in time ideal wireless channels at different Doppler frequencies. Subsequently, channel prediction errors are considered in FDD mode for system evaluation. Here, the term 'error' means, it is the deviation from the ideal value.

5.1 Ideal Channel at various Doppler frequencies

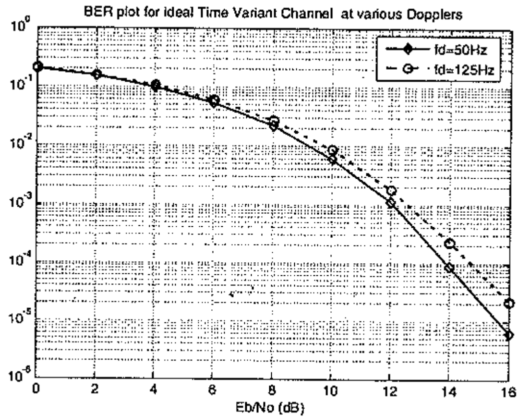


Figure 3. BER performance of ideal Time Variant Channel at various Dopplers for K = 15 users. The system parameters are $L_q = 16$, L = 19, $L_p = 32$, $\lambda = 0.05$.

In Fig. 3, it is observed that as the Doppler frequency increases from 50 Hz to 125 Hz, obviously the BER performance of system gets degraded. However, at BER of 1×10^{-4}, the difference in Eb/No is only about 0.5 dB which indicates the effectiveness of the filter.

5.2. Predicted Time Variant Channel in FDD mode

In FDD mode, since the uplink and downlink channels are different, the ideal downlink channel is considered at the MS and the information is fed back to the BS. In this work, it is assumed that the channel information is available at BS for simplicity. Here, one time slot delay, which is typically used in 3G standards, is considered. The channel is predicted at the BS for the current slot with lattice filter of

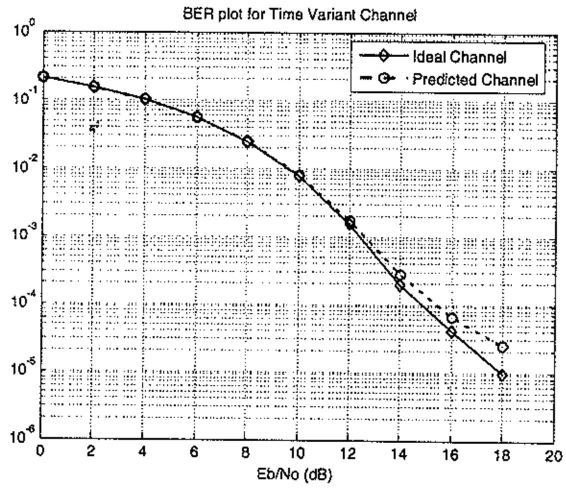


Figure 4 : BER performance of predicted channel versus Eb/No in FDD mode for K = 15 users. The system parameters are L_d =8, L_q = 16, L = 19, L_p = 32, λ = 0.05.

Length, $L_d = 8$, In Fig. 4, BER performance of the system with predicted channel is compared with that of perfect channel. The difference in Eb/No at BER of 1×10^{-4} is about 0.5 dB only. BER performance versus number of users is compared in Fig. 5 and found that errors saturate

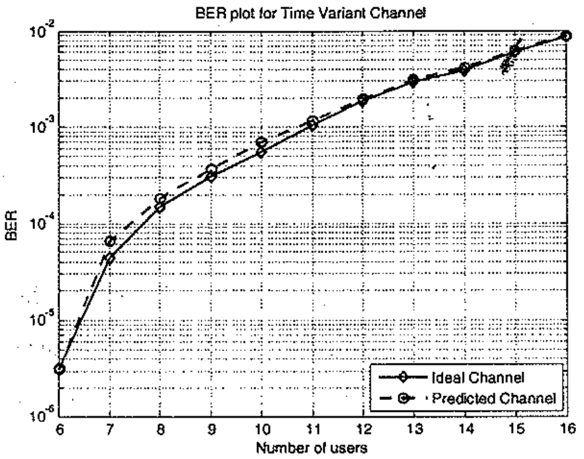


Figure 5 : BER performance of predicted channel versus number of users in FDD mode. The system parameters are L_d =8, L_q = 16, L = 19, L_p = 32, λ = 0.05, Eb/No = 10 dB.

slowly as the number of users increases. At 16 users, the difference in BER between predicted and perfect channels is almost zero. These two results indicate that the prediction method applied in this paper is very effective.

6. Conclusion

Most of the previous works on precoding have not taken into consideration the effects of imperfect channel conditions. In this paper, performance of the system using transmitter-based filters as precoding elements has been evaluated under imperfect channel conditions for DS-CDMA downlink in FDD mode. It is shown that due to channel prediction, performance degradation of the system is marginal as compared to that of the case where perfect channel is assumed. The difference in Eb/No at BER = $1x10^{-4}$ is about 0.5 dB only. At 16 users, the difference in BER between predicted and perfect channel is almost zero. Hence the method applied in this paper for performance evaluation of the system under imperfect channel conditions can be applied in a realistic scenario.

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