A Simulation Comparison of Multiuser Receivers for Cellular CDMA

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***Abstract*—In recent years, multiuser detection has gained significant notoriety as a potential advanced enabling technology for the next generation of CDMA systems. Currently, due to the limitations of the conventional correlation receiver, the capacity of a single cell using CDMA is limited by self-interference and is subject to the near–far problem. To overcome these drawbacks, several advanced receiver structures have been proposed. Unlike the conventional receiver which treats multiple access interference (MAI) as if it were AWGN, multiuser receivers treat MAI as additional information to aid in detection. Although each of the multiuser types have been the subject of much recent literature, there is little published work comparing all structures on the basis of common assumptions. We present a comparison of five of the most discussed receiver structures: the decorrelator, the minimum mean square error (MMSE) receiver, the multistage parallel interference cancellation receiver, the successive interference cancellation receiver, and the decorrelating decision feedback receiver. Comparisons are based on both theoretical analysis and simulation results, examining bit error rate (BER) performance in AWGN, Rayleigh fading, and Near/Far channels. Additionally, receiver structures are compared on the basis of computational complexity as well as robustness to code phase misalignment. Finally, we present simulation results for noncoherent architectures of the aforementioned receivers.**

***Index Terms*—CDMA, interference cancellation, multiple access, multiuser detection.**

I. INTRODUCTION

THE tremendous increase in demand for wireless services has caused a search for techniques to improve the capacity of current digital cellular systems. One promising technique for CDMA is multiuser detection [1]–[3]. Conventional receivers treat multiple access interference (MAI) which is inherent in CDMA, as if it were additive random noise. However, this MAI is generally correlated with the desired user's signal due to the likely lack of synchronism between users on the reverse link, from mobile unit to base station, significantly degrading performance. In addition, in a conventional receiver, if an interferer is significantly stronger than the desired user it can potentially dominate performance as even a small amount of correlation will lead to significant interference. This is commonly termed the near-far problem. To ameliorate this problem, stringent power control is required in current CDMA system designs. Precise power control is difficult to maintain and is a significant complexity burden. Multiuser detection is a very promising approach to overcoming the limitations of the conventional DS-CDMA receiver and improving system capacity.

This paper presents a comparison of five multiuser receivers and their usefulness for CDMA, particularly at the base station. The five structures of interest are the decorrelator, the minimum mean square error (MMSE) receiver, the multistage parallel interference cancellation receiver, the successive interference cancellation receiver, and the decorrelating decision feedback receiver. Cellular or Personal Communication System (PCS) design consists of two distinct problems, namely the design of the forward link from the base station to the mobile and the design of the reverse link from the mobile to the base station. The forward link can be designed so that users transmit with orthogonal spreading codes with signals arriving at the mobile receiver with identical energy. To assure marketability the receiver must be inexpensive and have low power requirements. The reverse channel is harsher, but can support a more sophisticated receiver. User signals arrive at the receiver asynchronously and may have differing energies, resulting in the near-far problem. The base station receiver can be larger and more complex, have higher power consumption and use information about the interfering signals. We focus on this latter situation where the receiver jointly detects signals from all users. One may observe that many of the problems observed in CDMA are the result of the approach used by the conventional receiver, rather than being inherent to CDMA.

A brief overview of multiuser receivers is presented in Section II. Section III describes the theoretical performance of the five multiuser receiver techniques in AWGN channels and compares the theoretical performance with simulation results. In Section IV, simulation and theoretical results are presented for near/far channels, flat Rayleigh fading, frequency selective fading, along with a presentation of the effects of code synchronization errors. Computational complexity of the proposed schemes is discussed in Section V. Multi-cell scenarios are considered in Section VI. Section VII discusses noncoherent multiuser detection. Conclusions are presented in Section VIII.

II. THE MULTIUSER RECEIVER: A BRIEF HISTORY

One of the first investigations into multiuser reception of CDMA signals was presented in [4]. Here it was shown that optimal performance in synchronous situations requires estimates of all users for the bit period under consideration. It

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was also asserted that optimal detection in the asynchronous case requires knowledge of the entire transmitted sequence for each user. The optimal detector was developed more rigorously in [5], where Verdu presents the optimal solution for the asynchronous case. It is shown that the complexity per binary decision is $O(2^{K-1})$ where K is the number of simultaneous users. For a large number of users, this level of complexity is impractical. However, the gains provided by the optimal receiver were significant enough to stimulate research on sub-optimal solutions. Some of the first sub-optimal solutions were the linear multiuser detectors of [6] and [7], as well as those presented in [8]. The decorrelating detector of [6], while sub-optimal with respect to probability of error was still optimal with respect to near-far resistance, a measure developed in [7] and [9]. Sub-optimal nonlinear detectors using a multistage approach were proposed [10]–[12], while structures employing successive cancellation were presented in [13] and [14] and similar nonlinear receivers employing decision feedback were proposed in [15].

The previous papers commonly considered AWGN channels with coherent demodulation. Subsequent work adapted these ideas to channels which suffered from time-varying multipath fading and noncoherent demodulation. The optimal receiver of [5] was extended to Rayleigh fading channels in [16] as well as to asynchronous and synchronous Rician channels in [17] and [18]. The decorrelator was also extended to noncoherent modulation, Rayleigh fading [19], [16], and Rician fading [17], [18]. Successive cancellation and multistage cancellation were also extended to fading channels in [14] and [20]. More thorough surveys can be found in [1]–[3].

III. Performance in AWGN Channels

In this section, we derive the decision statistic (or metric) for each of the five receiver structures under consideration. We present expressions for the probability of bit error for each in an AWGN channel with constant (although not necessarily equal) received user energies. These analytical results are compared with simulation results.

A. System Model

For the development of the performance of different detectors, we require a common system model. The received baseband signal from user k is defined as a binary phase modulated waveform

$$s_k(t) = \sqrt{P_k}a_k(t)b_k(t)e^{j\theta_k} \tag{1}$$

where P_k is the kth user's received signal power, $a_k(t)$ and $b_k(t)$ are the spreading and data waveforms, respectively (we assume rectangular pulses for both), and θ_k is the received phase of the kth user relative to some reference phase. Due to the asynchronous nature of the system uplink, the received signal is

$$r(t) = \sum_{k=1}^{K} s_k(t-\tau_k) + n(t) \tag{2}$$

where τ_k is the time delay parameter that accounts for the time of arrival of the kth user's signal, and $n(t)$ represents the additive white Gaussian noise experienced at receiver.

Assuming perfect knowledge of the channel state vector, the set of sufficient statistics for determining the transmitted bits $\mathbf{b}$ can be shown to be the matched filter outputs $\mathbf{y}$ where the filters are matched to each user's spreading code [3]. In terms of the complex envelope, the vector of sufficient statistics is

$$\mathbf{y} = \cos(\mathbf{\Theta})\mathbf{y_I} + \sin(\mathbf{\Theta})\mathbf{y_Q} \tag{3}$$

where the ith matched filter output of the kth user is the $(i-1)K+k$th element of the vector $\mathbf{y}$, $\mathbf{\Theta}$ is a $KN_b \times KN_b$ diagonal matrix where the diagonal elements $\theta_{j,j}$ are the phases of the ith bit of the kth user, and $j=(i-1)K+k$, K is the number of users in the system, $\cos(\mathbf{\Theta})$ and $\sin(\mathbf{\Theta})$ are defined to be diagonal matrices, N_b is the number of bits in the sequence under consideration and the in-phase and quadrature components are defined as

$$y_{I_{(i-1)K+k}} = \int_{(i-1)T+\tau_k}^{iT+\tau_k} r_I(t)a_k(t-\tau_k)\,dt \tag{4}$$

and

$$y_{Q_{(i-1)K+k}} = \int_{(i-1)T+\tau_k}^{iT+\tau_k} r_Q(t)a_k(t-\tau_k)\,dt \tag{5}$$

where $r_I(t) = Re[r(t)]$, $r_Q(t) = Im[r(t)]$ and τ_k is the relative delay of the kth user.

In matrix form we represent the set of matched filter outputs as

$$\mathbf{y_I} = \mathcal{R}\mathcal{W}\cos(\mathbf{\Theta})\mathbf{b} + \mathbf{n_I} \tag{6}$$

and

$$\mathbf{y_Q} = \mathcal{R}\mathcal{W}\sin(\mathbf{\Theta})\mathbf{b} + \mathbf{n_Q} \tag{7}$$

where $\mathcal{R}$ is a $KN_b \times KN_b$ matrix

$$\mathcal{R} = \begin{pmatrix} \mathbf{R}(0) & \mathbf{R}(-1) & 0 & \cdots & 0 \\ \mathbf{R}(1) & \mathbf{R}(0) & \mathbf{R}(-1) & & \vdots \\ 0 & \mathbf{R}(1) & \mathbf{R}(0) & \ddots & 0 \\ \vdots & & \ddots & \ddots & \mathbf{R}(-1) \\ 0 & \cdots & 0 & \mathbf{R}(1) & \mathbf{R}(0) \end{pmatrix} \tag{8}$$

the (k,l)th element of the $K \times K$ matrix $\mathbf{R}(i)$ is defined by

$$\rho_{k,l}(i) = \int_{-\infty}^{-\infty} a_k(t-\tau_k)a_l(t+iT-\tau_l)\,dt. \tag{9}$$

$\mathcal{W}$ is a $KN_b \times KN_b$ diagonal matrix of the square root of user received energies defined similar to $\mathbf{\Theta}$, $\mathbf{b}$ is a KN_b length vector with the $j=(i-1)K+k$th element equal to the ith data symbol of the kth user, and $\mathbf{n}_I$ and $\mathbf{n}_Q$ are vectors of colored noise samples at the matched filter outputs. If the users are numbered such that $\tau_1 < \tau_2 < \cdots < \tau_K$, then $\mathbf{R}(1)$ will be an upper triangular matrix with zeros along the diagonal, $\mathbf{R}(-1) = \mathbf{R}^T(1)$ and $\mathbf{R}(i) = 0\ \forall |i| > 1$. In our notation, parameter estimates are represented with the symbol $\hat{}$. For example, $\hat{\mathbf{\Theta}}$ represents the diagonal matrix of estimated received phases.

B. The Decorrelator

A linear detector is a detector where the decision metric is a linear transformation of the sufficient statistics, i.e.,

$$\hat{\mathbf{b}} = \mathrm{sgn}[\cos(\hat{\Theta})\mathbf{T}\mathbf{y}_I + \sin(\hat{\Theta})\mathbf{T}\mathbf{y}_Q]. \tag{10}$$

The decorrelator is a linear detector where $\mathbf{T} = \mathcal{R}^{-1}$. As the name implies, the decorrelator removes the correlation between the elements of $\mathbf{y}$. More explicitly

$$\hat{\mathbf{b}} = \mathrm{sgn}[\cos(\hat{\Theta})\mathcal{R}^{-1}\mathbf{y}_I + \sin(\hat{\Theta})\mathcal{R}^{-1}\mathbf{y}_Q] \tag{11}$$
$$= \mathrm{sgn}[\mathcal{W}\mathbf{b} + \mathcal{R}^{-1}\mathbf{n}] \tag{12}$$

assuming perfect phase estimates and delay estimates.

This transformation is derived from the maximization of the likelihood function or equivalently the minimization of $(\mathbf{y} - \mathcal{R}\mathbf{b})^T\mathcal{R}^{-1}(\mathbf{y} - \mathcal{R}\mathbf{b})$ [8]. The probability of symbol error (equivalent to bit error in BPSK) of the kth user can be represented as [7]

$$P_{k,i}(E) = Q\left(\sqrt{\frac{E[\mathbf{y}_{k,i}]^2}{\mathrm{var}[\mathbf{y_{k,i}}]}}\right) \tag{13}$$

where $\mathbf{y}$ is the decision metric, $E[\mathbf{y}] = \mathcal{W}\mathbf{b}$, $\mathrm{var}[\mathbf{y}]$ needs to be determined and $Q(\cdot)$ is the standard Q-function. We define $\mathbf{\Sigma}_y = \mathrm{var}[\mathbf{y}]$ and

$$\begin{aligned}\mathbf{\Sigma_y} &= E[\mathbf{y}\mathbf{y}^\mathbf{T}] - E[\mathbf{y}]E[y^T]\\ &= E[(\mathcal{W}\mathbf{b} + \mathcal{R}^{-1}\mathbf{n})(\mathcal{W}\mathbf{b} + \mathcal{R}^{-1}\mathbf{n})^T] - (\mathcal{W}\mathbf{b})(\mathcal{W}\mathbf{b})^T\\ &= \mathcal{R}^{-1}\sigma^2\mathcal{R}(\mathcal{R}^{-1})^T\\ &= \sigma^2(\mathcal{R}^{-1})^T\end{aligned} \tag{14}$$

where we have used $E[\mathbf{n}] = 0$, $E[\mathbf{n}\mathbf{n}^\mathbf{T}] = \sigma_n = \sigma^2\mathbf{R}$ and σ^2 is the power of the AWGN at the receiver. Using this result in (13) results in

$$P_{k,i}(E) = Q\left(\sqrt{\frac{w_{j,j}}{(R^{-1})_{j,j}N_o}}\right) \tag{15}$$

where $j = (i-1)K + k$, N_o is the one-sided noise power spectral density of the additive white Gaussian noise, and we have assumed all data symbols are transmitted with equal probability. Thus, the performance of the decorrelator is identical to the single user case with the exception of the noise enhancement factor $(R^{-1})_{j,j}$. Since all of the elements of $\mathcal{R}$ are less than or equal to one, we find that $(R^{-1})_{j,j} > 1$. Unfortunately, general statistics of $\mathcal{R}^{-1}$ are not easily found, thus predicting error probabilities is best done using the actual correlation matrix of a known set of user codes and relative delays. In this work, we obtain an estimate of the performance by calculating the average of the elements along the diagonal of the inverse during runs. It should be noted that the decorrelator is a generalized version of the zero-forcing equalizer for ISI channels.

C. Minimum Mean Square Error Receiver

A similar receiver structure can be obtained if the linear transformation is sought which minimizes the mean square error between the transmitted bits and the outputs of the transformation, $E[(\mathbf{b} - \mathbf{T}\mathbf{y})^T(\mathbf{b} - \mathbf{T}\mathbf{y})]$. In this case, the linear transformation $\mathbf{T} = \mathcal{R}^{-1}$ used in (11) is replaced by $\mathbf{T} = (\mathcal{R} + \mathbf{N_o}/2\mathcal{W}^2)^{-1}$. The performance of the MMSE detector approaches the decorrelator as $N_o \to 0$. As N_o grows large, $\mathbf{T}$ approaches an identity matrix scaled by $2\mathcal{W}^2/N_o$ and is thus reduced to the conventional receiver. Thus, the MMSE detector seeks to strike a balance between removing the interference and not enhancing the noise. At low E_b/N_o the MMSE receiver outperforms the decorrelator, while at high E_b/N_o the performance of the decorrelator approaches that of the MMSE receiver. However, direct implementation of the MMSE requires knowledge (or estimation) of SNR which can degrade performance. In such cases, the performance of the decorrelator can be superior. This receiver is analogous to an equalizer which attempts to balance the removal of ISI and reduction of the noise enhancement. An attractive feature of the MMSE receiver is that it lends itself to adaptive implementation [3].

D. Multistage Parallel Interference Cancellation

Multistage receivers are receivers which have multiple stages of interference estimation and cancellation. Although any receiver can be used at each stage (particularly the first) this approach is most often used in conjunction with parallel interference cancellation. In one version of this approach a conventional receiver is used in the first stage to estimate the channel gain and data symbol. These estimates along with independent delay estimates are used to remove the interference from each of the desired users' signals. The gain and data estimates can also be independent estimates derived from an outside source. This cancellation approach was first suggested in [21] and further developed in [10] and [11]. In [11], it was suggested that each user's signal could be iteratively estimated in parallel. The estimates for each user can then be used to reduce the interference of the other users by subtracting the estimate of each interferer from the desired user's signal. Ideally, this would allow the elimination of all interfering signals from the desired user. However, due to the inaccuracy of the estimates, this interference will be subtracted imperfectly. Thus to overcome this, the entire process can be repeated for several stages. At each stage, better estimates of each user are produced, allowing more effective interference cancellation. In this paper, we assume the use of matched filters at each stage for estimation. This allows a single estimate (the matched filter output) to be used for both the data symbol and the channel gain and alleviates the need for any outside estimates. While more robust receivers could be used in the first stage to improve performance, this approach is the most straightforward and allows reasonable complexity.

Mathematically we can represent the decision metric for an S-stage parallel cancellation scheme as

$$\hat{\mathbf{b}} = \mathrm{sgn}\left[\cos(\hat{\Theta})\mathbf{y}_\mathbf{I}^{(S)} + \sin(\hat{\Theta})\mathbf{y}_\mathbf{Q}^{(S)}\right] \tag{16}$$

where

$$y_{I_j}^{(s)} = \frac{1}{T}\int_{(i-1)T+\hat{\tau}_k}^{iT+\hat{\tau}_k} \hat{r}_{I_k}^s(t)a_k(t-\hat{\tau}_k)\,dt \tag{17}$$

$$y_{Q_j}^{(s)} = \frac{1}{T}\int_{(i-1)T+\hat{\tau}_k}^{iT+\hat{\tau}_k} \hat{r}_{Q_k}^s(t)a_k(t-\hat{\tau}_k)\,dt \tag{18}$$

and

$$\hat{r}^s_{I_k}(t) = r_I(t) - \sum_{j \neq k} y_j^{s-1} a_j(t - \hat{\tau}_j)\cos(\hat{\theta}_j) \tag{19}$$

$$\hat{r}^s_{Q_k}(t) = r_Q(t) - \sum_{j \neq k} y_j^{s-1} a_j(t - \hat{\tau}_j)\sin(\hat{\theta}_j) \tag{20}$$

with $j = (i-1)K + k$ for the ith bit of the kth user, $\hat{r}^s_{I_k}(t)$ is the kth user's signal after $s-1$ stages of cancellation, and $\hat{\tau}_j$ and $\hat{\theta}_j$ are the estimated delay and phase of user j.

This implementation requires the estimation, regeneration, and cancellation of each interferer from each of the desired users. Since we must regenerate each of the wideband signals we refer to this as the wideband implementation. An alternate implementation which we call a narrowband implementation does not require regeneration. Instead, interference is cancelled from the narrowband outputs (the matched filter outputs) by using the estimates of the data symbol and channel gains as well as the known cross-correlations between users (derived using delay estimates). This narrowband implementation may require more memory to store the matrix $\mathcal{R}$ but requires less computations because it avoids regeneration of the wideband signal. The computational trade-offs will be examined in Section V. The theoretical performance of the two implementations is the same.

The analytical performance of this multistage parallel cancellation approach was derived in [22]. It was shown that in an AWGN channel the bit error rate (BER) of the receiver employing the standard Gaussian approximation for MAI at stage s is

$$P_k^s(E) = Q\left(\left[\frac{1}{2E_b/N_o}\left(\frac{1-\left(\frac{K-1}{3N}\right)^s}{1-\frac{K-1}{3N}}\right) + \frac{1}{(3N)^s}\left(\frac{(K-1)^s-(-1)^s}{K}\frac{\sum_j P_j}{P_k} + (-1)^s\right)\right]^{-1/2}\right) \tag{21}$$

where K is the number of users and N is the processing gain.

The development of this equation assumes that $y_{k,i}^{(s)}$ is an unbiased estimate of the $\sqrt{P_k}b$ at each stage. Unfortunately, it is found that this is not the case. Rather, $y_{k,i}^{(s)}$ is biased particularly after the first stage of cancellation with the bias increasing with system loading [23], [24].

One method of alleviating this problem is to multiply the estimate by a back-off factor with a value in the range $[0, 1)$. This can also be interpreted as modifying the complete cancellation scheme of (19) and (20) to include the partial-cancellation factor $C_K^{(s)}$ as follows:

$$\hat{r}^s_{I_k}(t) = r_I(t) - C_K^{(s)} \sum_{j \neq k} y_j^{s-1} a_j(t - \hat{\tau}_j)\cos(\hat{\theta}_j) \tag{22}$$

and

$$\hat{r}^s_{Q_k}(t) = r_Q(t) - C_K^{(s)} \sum_{j \neq k} y_j^{s-1} a_j(t - \hat{\tau}_j)\sin(\hat{\theta}_j). \tag{23}$$

This factor significantly reduces the bias at each stage and substantially improves performance in heavily loaded systems. This back-off factor varies with both the stage of cancellation and the system loading [24], [25]. It is found that a factor of 0.5 in the first stage of cancellation is a good trade-off and results in performance improvements of an order of magnitude for loadings above $0.6N$ [25].

E. Successive Interference Cancellation

While the previous scheme suggests a cancellation of interference done in parallel and in multiple stages, a somewhat simpler approach is to estimate and cancel interference successively using feedback. In this approach users are first ranked according to their received powers,[1] and then estimated and cancelled in order from strongest to weakest. This cancellation approach has two advantages. First, the strongest users cause the most interference. Thus it is most beneficial to eliminate these interferers first. Second, the strongest users provide the most reliable estimates and thus cause the least error in cancellation. The result is that each user is estimated and only cancelled once as opposed to s times in the parallel cancellation approach. This can provide a savings in computational complexity depending on the implementation. The performance relative to parallel cancellation is dependent on the spread of user powers. That is, for the equal power case the successive cancellation scheme performs significantly worse than the parallel approach.[2] However, as the user powers get more widely distributed, the relative performance of the successive scheme improves.

The decision metric of the kth user during the ith bit interval is found to be

$$y_{k,i} = y_{I_{k,i}}\cos(\hat{\theta}_{k,i}) + y_{Q_{k,i}}\sin(\hat{\theta}_{k,i}) \tag{24}$$

where

$$y_{I_{k,i}} = \int_{(i-1)T+\hat{\tau}_k}^{iT+\hat{\tau}_k} \hat{r}_I^{(k)}(t) a_k(t-\hat{\tau}_k)\,dt \tag{25}$$

$$y_{Q_{k,i}} = \int_{(i-1)T+\hat{\tau}_k}^{iT+\hat{\tau}_k} \hat{r}_Q^{(k)}(t) a_k(t-\hat{\tau}_k)\,dt \tag{26}$$

and

$$\hat{r}_I^{(k)}(t) = r_I(t) - \sum_{i=1}^{N_b}\sum_{j=1}^{k-1} \frac{y_{j,i}}{T} a_j(t-\hat{\tau}_j)\cos(\hat{\theta}_j) \tag{27}$$

$$\hat{r}_Q^{(k)}(t) = r_Q(t) - \sum_{i=1}^{N_b}\sum_{j=1}^{k-1} \frac{y_{j,i}}{T} a_j(t-\hat{\tau}_j)\sin(\hat{\theta}_j) \tag{28}$$

and $\hat{r}^{(k)}(t)$ is the kth user's received signal after users 1 through $k-1$ have been estimated and cancelled, $\hat{\tau}_j$ and $\hat{\theta}_j$ represent the estimated delay and phase for user j, and we have assumed block processing. A theoretically equivalent approach is to perform cancellation without wideband regeneration.

The performance of the successive cancellation detector in an AWGN channel can be predicted using (13) and $E[y_{k,i}] =$

[1]An alternative is to order cancellation by reliability which requires re-ordering after each cancellation but significantly improves performance [26].

[2]If reliability ordering is performed after every cancellation successive cancellation can out-perform parallel cancellation but at a large complexity cost. Due to its increased complexity, we have not considered reliability ordering.

$w_{k,i}b_{k,i}$. The variance of the decision metric can be shown to be [27]:

$$\mathrm{var}[y_{k,i}] = \left[\frac{N_oT}{4} + \frac{T^2}{6N}\sum_{j=2}^{K} P_j\right]\left(1+\frac{1}{3N}\right)^{k-1} - \frac{T^2}{6N}\sum_{j=2}^{k}\left(1+\frac{1}{3N}\right)^{k-l} P_j \tag{29}$$

where N is the processing gain, K is the number of users and N_o is the one-sided noise power spectral density. This variance can be used in the Q-function to produce an analytical performance estimate.

F. Decorrelating Decision Feedback Receiver

Decision feedback multiuser detectors are nonlinear receivers similar to decision feedback equalizers employed in single user channels with ISI [15], [28]. In decision feedback multiuser detection, users are ranked in order of decreasing received power levels. Previous decisions are then used together with current statistics to estimate the current output bits. Decision feedback can be characterized by two matrix transformations: a whitening feedforward filter that operates on the matched filter outputs, and a feedback filter fed by the vector of previously made bit decisions.

The forward filter is obtained by first factoring the matrix of cross-correlations $\mathcal{R}$ from (8) using a Cholesky decomposition.

$$\mathcal{R} = \mathcal{F}^T\mathcal{F} \tag{30}$$

where $\mathcal{F}$ is a lower triangular matrix.

The optimal feedforward filter [29] is the noise whitening filter

$$\mathcal{G} = \left(\mathcal{F}^T\right)^{-1} \tag{31}$$

which eliminates multiuser interference due to future inputs. When this filter is applied to the decision statistics $\mathbf{y}$ the resulting output vector is given by

$$\hat{\mathbf{y}}_I = \mathcal{F}\mathcal{W}\cos(\hat{\boldsymbol{\Theta}})\mathbf{b} + \mathbf{n}_I \tag{32}$$

and

$$\hat{\mathbf{y}}_Q = \mathcal{F}\mathcal{W}\sin(\hat{\boldsymbol{\Theta}})\mathbf{b} + \mathbf{n}_Q \tag{33}$$

where $\mathbf{n}_I$ and $\mathbf{n}_Q$ are white Gaussian noise vectors with autocorrelation $\mathcal{R}(\mathbf{n}) = \sigma^2\mathcal{I}$.

The optimal feedback filter

$$\mathcal{B} = (\mathcal{F} - \mathbf{diag}(\mathcal{F})) \tag{34}$$

operates on previously made bit decisions ($\mathcal{B}$ is a lower triangular matrix with zeros along its diagonal). The feedback terms at the output of this filter eliminate all MAI provided that feedback data is correct. After feedback, the input to the decision devices is given by

$$\mathbf{z} = \cos(\hat{\Theta})\mathbf{z}_I + \sin(\hat{\Theta})\mathbf{z}_Q \tag{35}$$

where

$$\mathbf{z}_I = \hat{\mathbf{y}}_I - \mathcal{B}\mathcal{W}\cos(\hat{\Theta})\hat{\mathbf{b}} \tag{36}$$

$$\mathbf{z}_Q = \hat{\mathbf{y}}_Q - \mathcal{B}\mathcal{W}\sin(\hat{\Theta})\hat{\mathbf{b}} \tag{37}$$

and $\hat{\mathbf{b}} = \mathrm{sgn}[\mathbf{z}]$.

The probability of error for the decision feedback, assuming that the feedback terms are correct was presented in [15]

$$P_{k,i}(E) = Q\left(\sqrt{\frac{2E_bF_{k,k}^2}{N_o}}\right). \tag{38}$$

It can be shown that $F_{k,k}^2 \geq 1/(R_{k,k}^{-1})$ where equality is obtained for user 1. Thus, for the strongest user, the decorrelator and the decision feedback receivers present the same probability of error. For the other users the performance of the decision feedback strategy is superior compared to that of the decorrelator provided feedback estimates are correct. For the weakest user, the ideal performance of decision feedback matches the single user bound since $F_{KN_b,KN_b} = 1$.

G. Simulation Results

This section presents the results of our Monte Carlo simulations for AWGN channels. Monte Carlo simulations are inviting for analyzing complex systems such as those employing multiuser detection because they allow the underlying system architecture to be closely modeled. A distinctive feature of these type of simulations is the large amount of computer time required to generate accurate bit error estimates. In this study, we typically run simulations such that approximately 10–50 errors would occur for the lowest BER desired. A widely used approximation for estimating the accuracy of Monte Carlo simulations is that for a large number of trials ten independent errors will result in a 95% confidence interval. Confidence intervals are discussed in more detail in the Appendix.

The first set of results are capacity curves for with $E_b/N_o =$ 8-dB, $N = 31$, and perfect power control. The simulation results are plotted along with the theoretical curves in Fig. 1. The parallel scheme uses two stages of cancellation $(S = 3)$ and a back-off factor of 0.5 in stage 2. The simulation results show excellent agreement with theoretical performance. We have not plotted the theoretical curves for the MMSE or decorrelating DF receivers to limit the number of plots. The MMSE performance is upper bounded by the decorrelator, while the decorrelating DF receiver will have a theoretical performance between the decorrelator and the single user bound.

For the perfect power control case we find that the decorrelator, MMSE, parallel canceller, and decorrelating DF detectors all provide roughly similar performance, with the nonlinear detectors providing better performance than the linear detectors. The linear detectors provide a capacity improvement of roughly three relative to the matched filter while the two nonlinear detectors (excluding SIC) provide an improvement of four times. The performance of the successive canceller is significantly poorer due to the lack of variance in the received powers. Since the successive canceller orders cancellation based on average powers

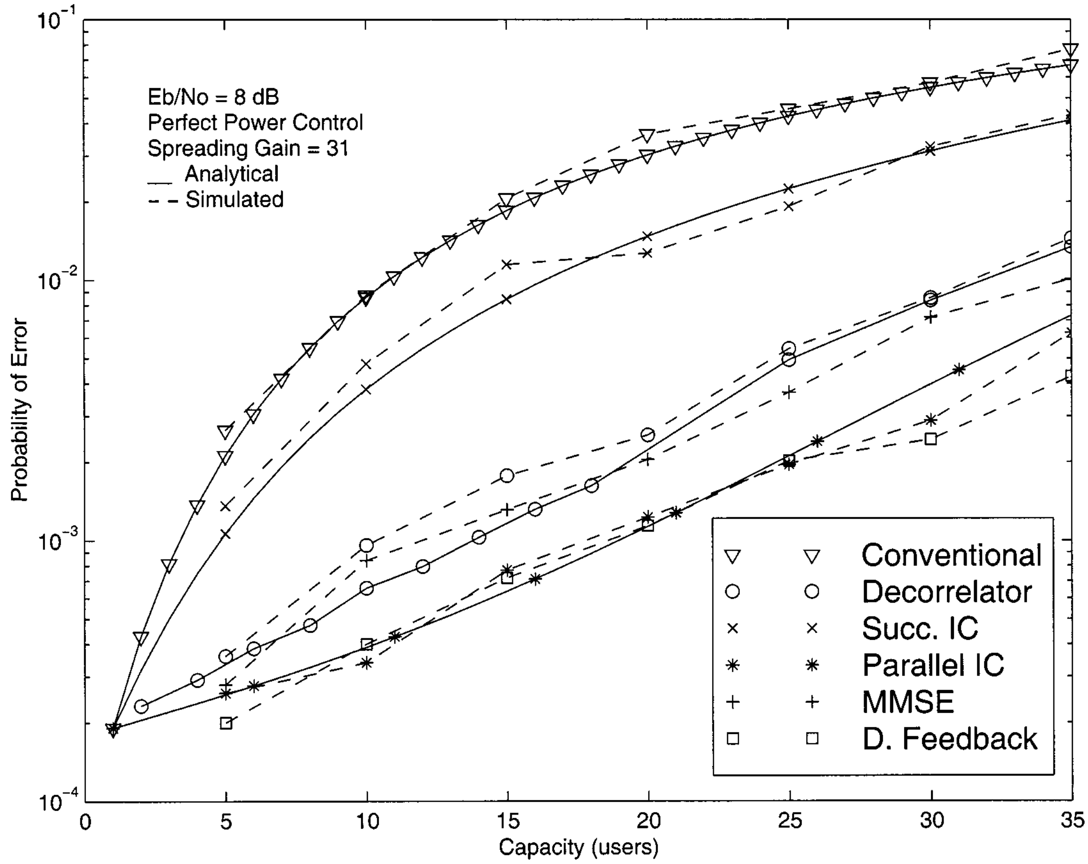

Fig. 1. Capacity Curves for Perfect Power Control ($E_b/N_o = 8$ dB and processing gain $= 31$).

and average powers are the same, the performance of signals detected early in the cancellation process suffers. In fact, the performance of the first user is equivalent to that of the conventional receiver. This can be ameliorated by ordering based on reliability. That is, the matched filter outputs which are furthest from the decision boundary are the most reliable and should be detected and cancelled first. However, this requires ordering to be re-calculated after each cancellation which significantly increases complexity [26].

The BER performance versus E_b/N_o is given in Fig. 2 for $K = 10, N = 31$, and perfect power control. Again we find significant improvement for the decorrelator, the parallel canceller, the MMSE and decorrelating DF receivers with each providing gains of over an order of magnitude at 10-dB, while the successive canceller provides a small improvement. Thus, in AWGN nonlinear detectors provide superior performance although the difference is minor at low loading levels.

IV. Realistic Channel Impairments

A. Near–Far Channels

Previously we examined the improvement in capacity and BER performance possible with multiuser structures in perfect power control. However, as asserted earlier, one of the drawbacks of the conventional receiver is its subjection to the near–far problem. Thus we wish to examine the performance of each of the receiver structures in situations where a single interferer dominates the received power. Fig. 3 presents the performance of the receivers in the presence of two interferers, one with equal power to the desired user, and one with a power which varies from 10-dB below the desired user to 30-dB above the desired user. As expected, the conventional receiver degrades quickly in the presence of strong interference. The successive canceller and the decision feedback receiver which benefit from diverse powers are found to be robust to strong interferers, as is the decorrelator which has a performance which is independent of user energies. The MMSE having a theoretical near–far resistance identical to the decorrelator also displays robustness. The parallel canceller is not as robust and shows slow degradation for high interference power. The parallel canceller suffers due to the fact that cancellation of the weak user is inaccurate in the first stage of cancellation due to the dominating interference. This poor cancellation serves to degrade the channel gain estimate of the strong user in the succeeding stage. Consequently, when the strong user is cancelled from the weak user's signal in the second stage of cancellation it is done inaccurately, causing problems for the weak user. This continues from stage to stage with slight improvement each time. Thus we find that as $s \rightarrow \infty$ the parallel scheme approaches the near–far resistance of the decorrelator and successive canceller. It can be shown that due to the channel estimation, the parallel approach is not near–far resistant in general [30]. However, it shows significant robustness when compared to the conventional receiver. One way of improving the parallel receiver in such situations is to avoid cancelling the weak user since its channel gain is unreliable. This provides

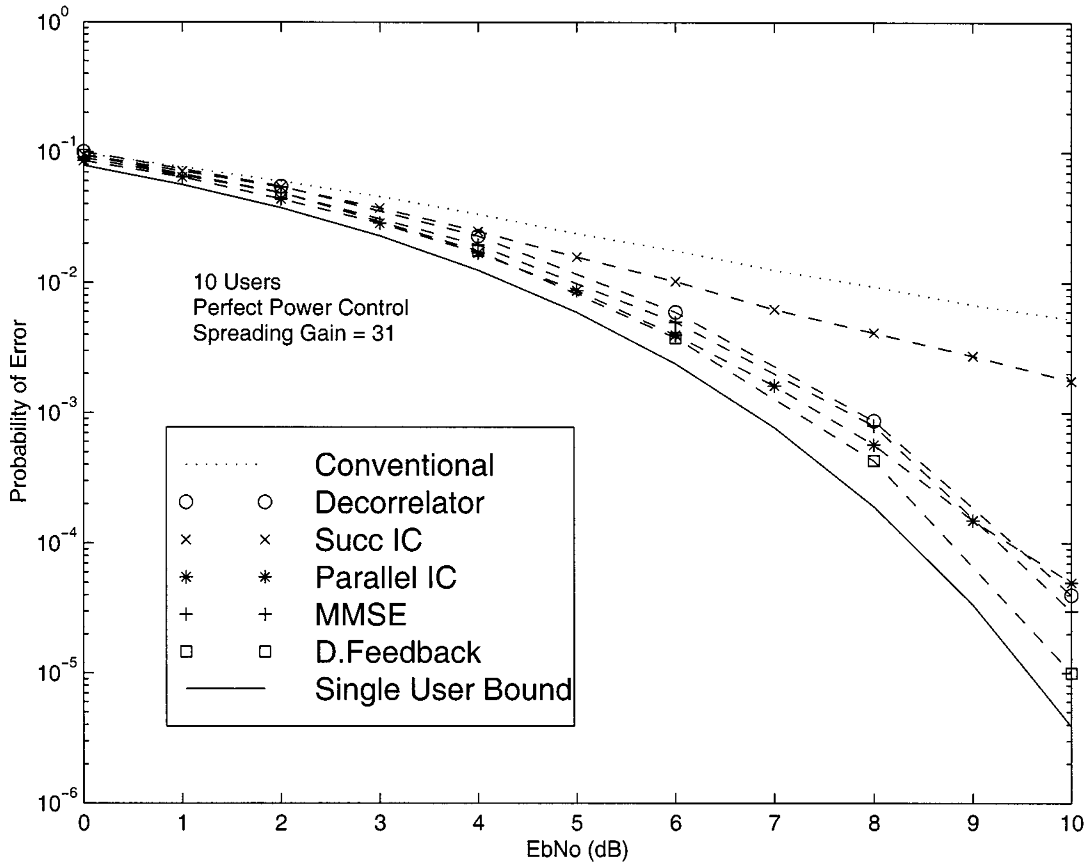

Fig. 2. BER versus E_b/N_o with Perfect Power Control (ten users and processing gain $= 31$).

near–far resistance with only two stages of cancellation [30]. While the decision feedback detector will also require channel estimation (which we bypassed for this simulation) it performs very well in near–far scenarios due to the successive cancellation nature of its decision feedback operation which benefits from disparate powers.

B. Performance in Rayleigh Fading

The performance for each of the receiver structures in flat Rayleigh fading, with perfect channel information, is presented in Fig. 4. The channel is assumed flat with a single path experiencing Rayleigh fading. Again we find significant improvement over the conventional receiver with each of the receivers providing nearly equivalent performance. Although prior studies indicate that successive cancellation significantly outperformed parallel cancellation, these results were obtained for the full cancellation approach [31]. In our work we use partial cancellation as opposed to full cancellation, which results in improved bit-error-rate performance for the parallel canceller [32]. The theoretical performance of each of the receivers in flat Rayleigh fading is presented in Fig. 5. These curves were generated by using expressions found in [14] for successive cancellation, [20] for parallel cancellation, [33] for the decorrelator, [34] for MMSE, and [35] for the decorrelating decision feedback receiver. The performance of the cancellation approaches both assumed a single bit estimate for the channel. That is a channel estimate based on the decision statistics from a single bit. As a result of channel estimation, we can see that the performance of the interference cancellers degrades compared to the linear and decision feedback approaches. In fact, an irreducible error floor is seen due to estimation errors. While the nonlinear approaches were superior in AWGN and equivalent in Rayleigh fading with perfect channel estimation, when channel estimation is a factor, the linear approaches provide superior performance.

The performance of the receivers was then compared for practical amplitude and phase estimation. It is assumed that the fading is sufficiently slow (on the order of tens of bits) that the phase and amplitudes can be tracked with sufficient accuracy. In these simulations the subtractive interference cancellation receivers generate their channel estimates from a moving average over ten consecutive soft decision statistics. The remaining receivers use in phase and quadrature estimates provided by a decorrelator.

Fig. 6 illustrates the performance of the multiuser receivers with amplitude and phase estimation. The subtractive interference receivers experience some degradation due to imperfect cancellation caused by the amplitude and phase mismatch. For the remaining receivers, the amplitude and phase estimates provided by the decorrelator are satisfactory and do not reduce performance significantly. Channel estimation mismatches, however reduce the advantage of the decision feedback receiver over the decorrelator [36]. Fig. 6 confirms the importance of proper channel estimation for adequate interference mitigation in subtractive interference cancellation receivers. Note again that for channel estimation an error floor is introduced for the subtractive cancellation

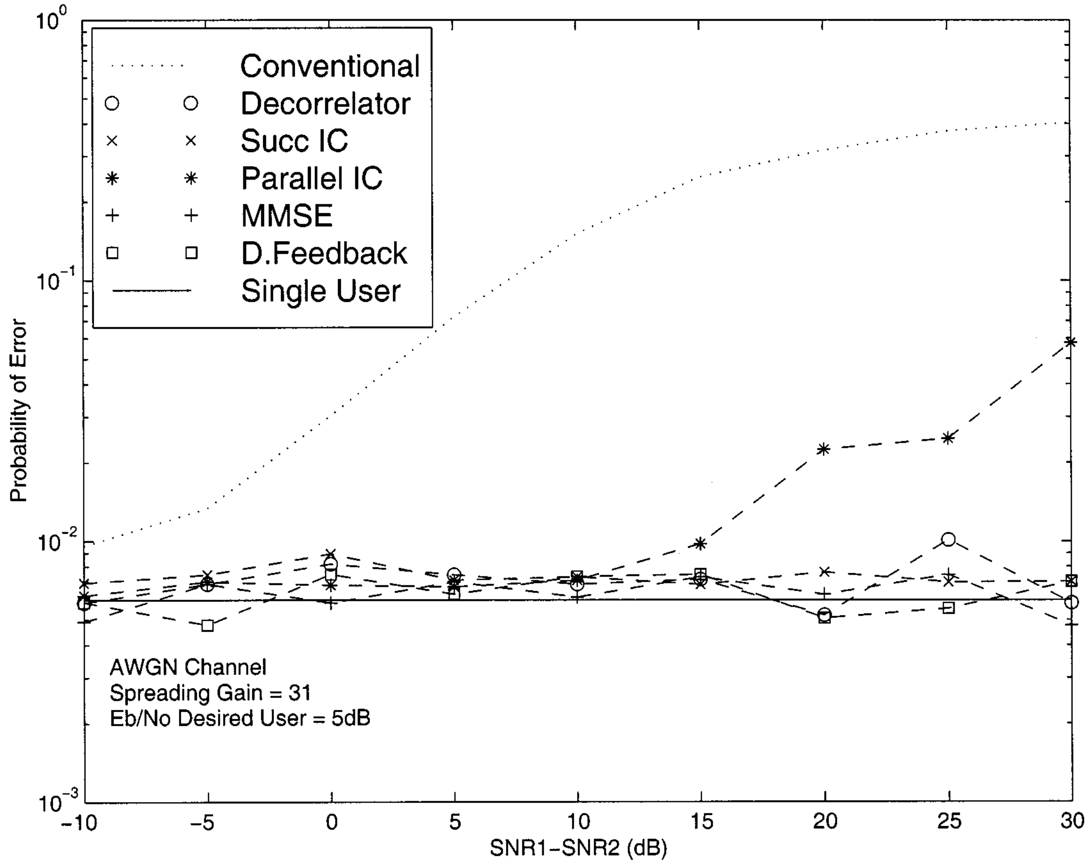

Fig. 3. Performance degradation in Near–Far Channels ($\overline{E_b/N_o} = 5$ dB and processing gain $= 31$).

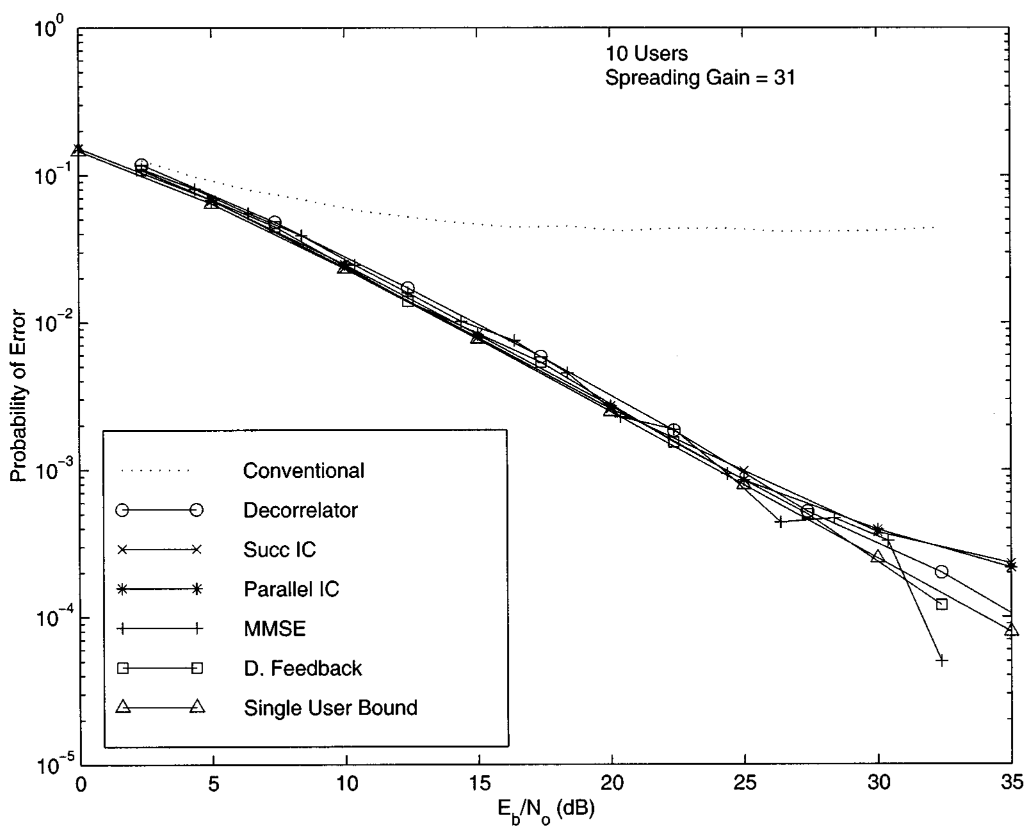

Fig. 4. BER versus E_b/N_o for Flat Rayleigh Fading with Perfect Channel Estimation (ten users, processing gain $= 31$).

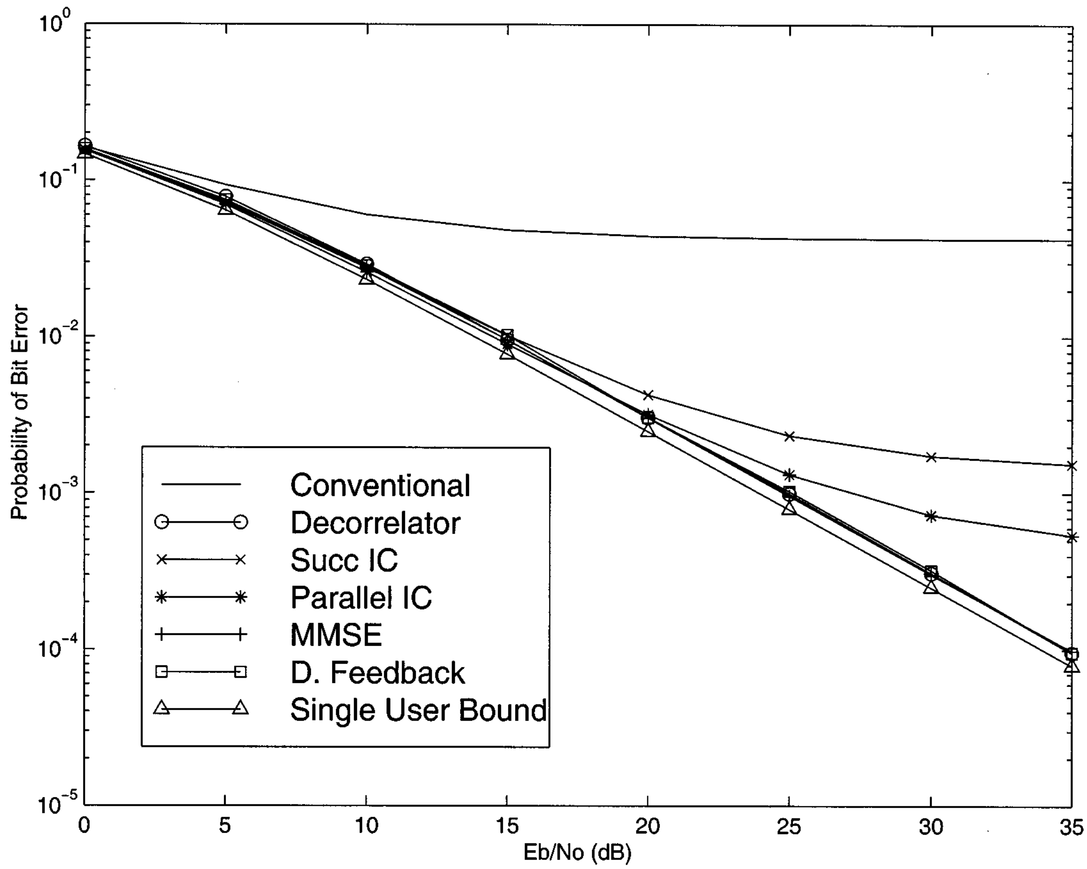

Fig. 5. Theoretical BER versus E_b/N_o for Flat Rayleigh Fading with Channel Estimation (ten users, processing gain $= 31$, 1 symbol average channel estimate).

approaches. We also note the similarity between the simulation and theoretical performance. In our simulations, the coherence time of the channel is assumed to be much larger than the interval used to create the channel estimates. Consequently, the effects of Doppler spread are not analyzed.

Results for two-ray frequency selective Rayleigh fading are presented in Fig. 7. The receivers each use Rake receivers to combine the multipath using maximal ratio combining. The channel parameters are taken from measurement data presented in [37] and represent a strong main path and a fairly weak second path, i.e., $\sigma_1 = 0.93$ and $\sigma_2 = 0.28$. The single user bound for maximal ratio combining for independent and unequal fading amplitudes was computed according to [38]. Maximum ratio combining occurs after the linear transformation for the decorrelator and the MMSE receivers. For the subtractive interference cancellation receivers (i.e., PIC, SIC and Decision Feedback) rake combining occurs after interference subtraction.

The results in Fig. 7 show the difference between the decorrelating receivers and the cancellation techniques. Although the additional paths provide diversity, they also present additional MAI which degrades the estimates of the channel gains, affecting the cancellation receivers more profoundly. All show significant increases over the conventional receiver.

C. Delay Estimate Errors

In CDMA systems, detection of the transmitted data is accomplished with a demodulator driven by a synchronized replica of the signature sequence. A receiver is perfectly synchronized when the local version of the spreading code is fully aligned with the spreading code embedded in the received signal. The assumption up to this point has been that the delays of each path for each user are known exactly (i.e., $\hat{\tau}_k = \tau_k$). In a realistic system some delay estimation error will be present. Thus we examine the effect of delay estimation errors on the performance of each of the multiuser receivers. The estimation error is assumed to be Gaussian with some standard deviation in chips (or fraction of a chip). Since the number of samples per chip is finite, we can only simulate a finite number of possible delay errors. Therefore while we generate a continuous delay estimation error, the actual error is taken as the generated error rounded up to the nearest sample. The effects of delay estimation error on the performance of each of the receivers studied for $E_b/N_o = 8$-dB, $N = 31$, $K = 20$ in a perfect power control AWGN channel are shown in Fig. 8. As expected there comes a point when the attempted removal of interference becomes no longer useful and even harmful. We also see that the performance degrades very rapidly for delay estimation errors. For an error standard deviation of only one-tenth of a chip, performance is degraded by almost an order of magnitude. The successive interference canceller degrades gracefully, agreeing with the results of [39]. The performance of the multistage receiver agrees with the analysis in [40]. Fig. 9 shows the degradation in near–far performance with timing errors. There are two interferers, one with equal power to the desired user, and one with a power which varies from

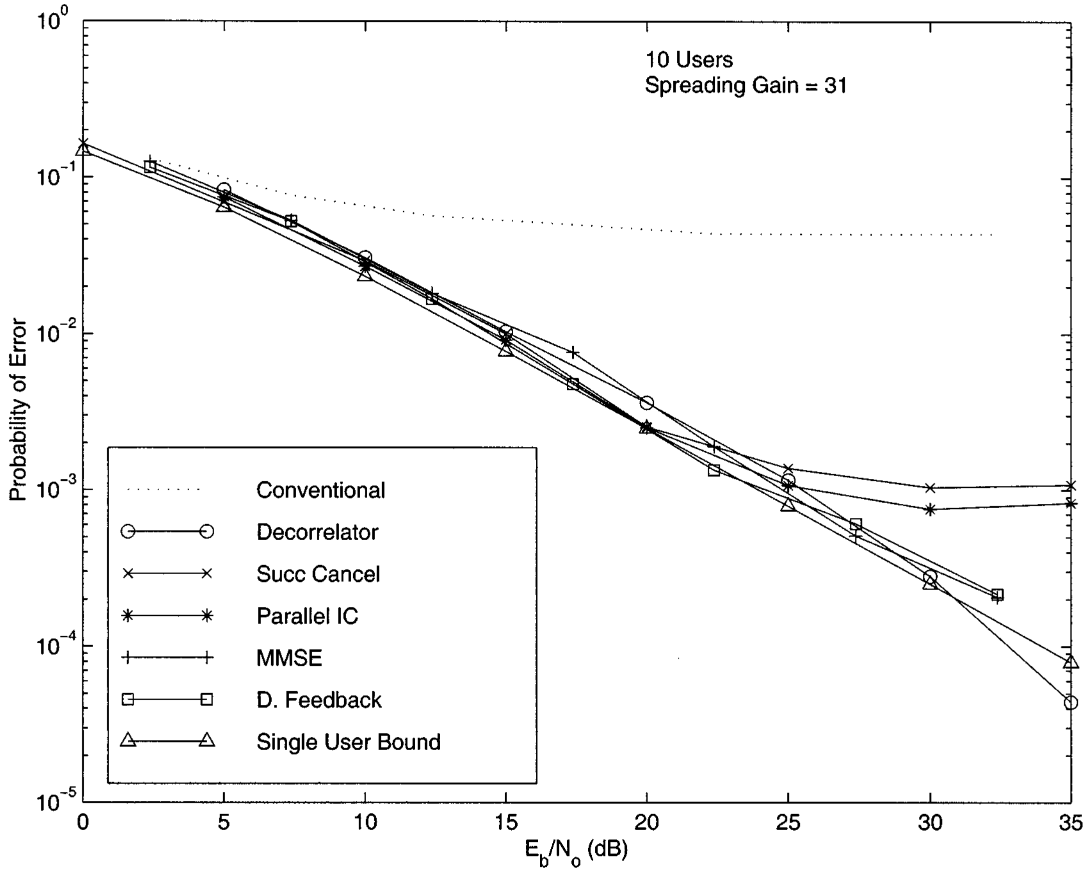

Fig. 6. BER versus E_b/N_o for Flat Rayleigh Fading with Amplitude and Phase Estimation (ten users, processing gain $= 31$, ten symbol average channel estimate).

10-dB below the desired user to 30-dB above. These results suggest that synchronization errors lead to inaccurate interference cancellation and intersymbol interference, thus causing degradation of the near–far robustness and BER performance, a behavior noticed for the linear receivers in [41]. Also note that the linear receivers are no longer near–far resistant when timing errors are present. These results show that timing errors are more critical for multiuser receivers than they are for matched filters since they not only lose correlation energy, but also perform increasingly inaccurate cancellation. The combination of these two effects makes timing more critical. Thus, timing will be a significant design issue for multiuser implementation.

V. Computational Complexity

The computational complexity of the detection scheme used in a system is vital to for both implementation and simulation. Receiver structures with high computational complexity require extremely high speed processors for implementation as well as extremely long run times for simulation. Thus, we wish to examine the computational complexity of each of the detection schemes presented here. In the following examination of computational requirements we will define complexity as the number of real single floating point operations (with no parallelism) required per bit decision and will define this quantity in terms of the number of users K, the frame length N_b, the number of paths tracked (Rake fingers) L, the spreading factor N, the number of samples per chip N_s, and the number of stages for the multistage receiver S.

A. The Decorrelator

The decorrelator detection can be broken down into four distinct operations:

1) calculating the matched filter outputs;
2) multiplication of the matched filter outputs by $\mathcal{R}^{-1}$;
3) creating the decision metrics;
4) creating the inverse of the correlation matrix.

It can be shown that the total number of flops[3] required per frame is

$$C_{\text{decor}} = N_b LK(2NN_s + N_b LK + 5) + 2LK(KL-1)NN_s + \frac{2}{3}(N_b LK)^3. \quad (39)$$

This leads to computational complexity per bit decision of

$$C_{\text{decor}}(b) = LK(2NN_s + N_b LK + 5) + \frac{2LK(KL-1)NN_s}{N_b} + \frac{2}{3}N_b^2(LK)^3. \quad (40)$$

The previous expressions assume direct matrix inversion. In [42], it is shown that due to the sparsity of the matrix $\mathcal{R}$, the

[3]We define a floating point operation as any multiply or add. More complex operations as division are considered multiple operations.

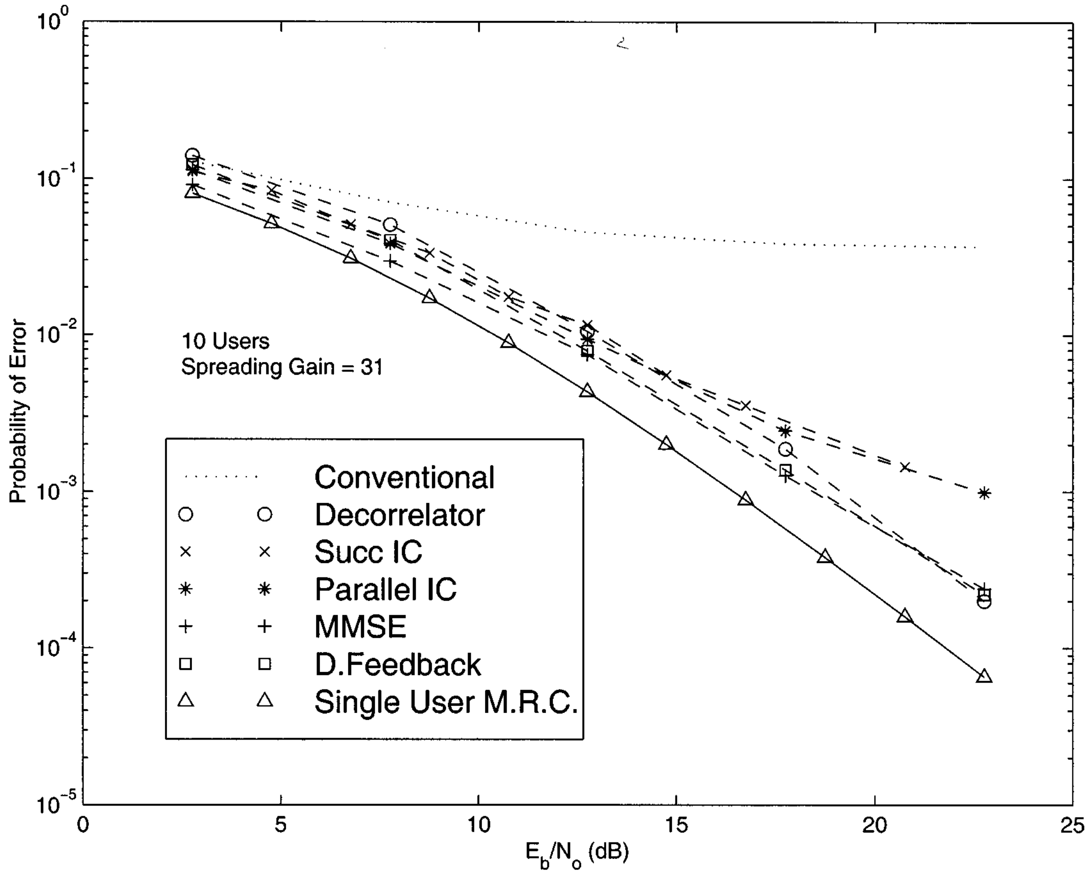

Fig. 7. BER versus E_b/N_o for Frequency Selective Rayleigh Fading (ten users, processing gain $= 31$, $\sigma_1 = 0.93$ and $\sigma_2 = 0.28$, 1 symbol channel estimate).

decorrelating detector can be determined with a complexity of order $N(KL)^3$. Iterative approaches for determining the outputs of the linear detector without explicit matrix inversion have been reported in [43]. These detectors, based on the conjugate gradient and steepest descent techniques for solving a linear matrix equation, have a computational complexity of order $N(KL)^2$ per iteration.

Additional reductions in the computational complexity are also possible if the matrix inversion need not be done every frame. The rate of update depends entirely on the rate at which users enter and leave the system, or timing information changes, provided short spreading codes are used. If long pseudo-random codes are used, the correlation matrix will change every symbol.

B. Multistage Parallel Interference Cancellation

As indicated previously the parallel cancellation approach can be implemented using either wideband or narrowband signals. While the theoretical performance of the two are identical, the computational and memory requirements of the two are significantly different. For the wideband approach the spread signals of each user are regenerated and cancelled at each stage, whereas in the narrowband approach the cancellation is performed using the channel gain and data estimates along with the cross-correlation information. This cancellation does not require regeneration of the wideband signals since cancellation is performed directly on the correlator outputs.

In the case of the wideband implementation we must generate matched filter outputs for each user at each stage as well as regenerating and cancelling each path of each user in stages 2 through S. In addition, stage S requires the maximal ratio combining of the individual paths to create the final decision statistic. It can be shown that the required number of flops per frame is

$$C_{\text{Wideband}} = N_b KL(S(6NN_s + 7) - 4NN_s - 1) \tag{41}$$

leading to a complexity per bit decision of

$$C_{\text{Wideband}}(b) = KL(S(6NN_s + 7) - 4NN_s - 1). \tag{42}$$

While the narrowband approach does not require signal regeneration it does require that the correlation matrix of the users be generated and stored in memory. This leads to a complexity per frame of

$$\begin{aligned} C_{\text{Narrowband}} = KLN_b[S(4KL+5) + 2NN_s - 4LK] \\ + 2KL(KL-1)NN_s \end{aligned} \tag{43}$$

and a complexity per bit decision of

$$\begin{aligned} C_{\text{Narrowband}}(b) = KL[S(4KL+5) + 2NN_s - 4LK] \\ + \frac{2KL(KL-1)NN_s}{N_b}. \end{aligned} \tag{44}$$

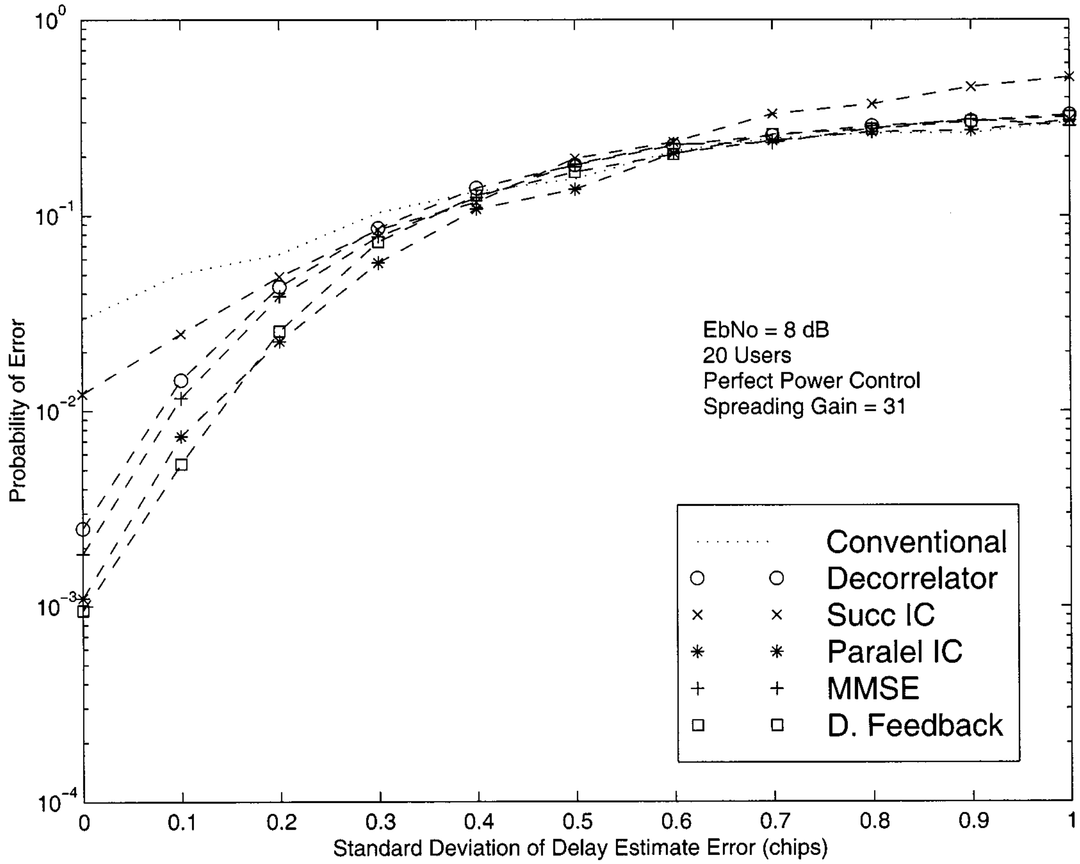

Fig. 8. Effect of Timing Errors (Delay Estimate Errors) on System Capacity in AWGN Channel with Perfect Power Control ($E_b/N_o = 8$ dB, Processing gain $= 31$, $K = 20$).

Again, the narrowband implementation assumed short spreading codes which repeat every few symbols. If long codes are used, the wideband approach is preferred.

C. Successive Interference Cancellation

Similar to the parallel cancellation scheme, the successive cancellation scheme incorporates estimation of the users, regeneration in the wideband case and cancellation. However, in this scheme each user is estimated and cancelled only once. Still, users must be ranked each frame in order of received user powers. It can be shown that this leads to a computational complexity per frame in the wideband case of

$$N_b L[K(8NN_s + 12) - 6NN_s - 7] + K(2N_b + \log_2(K) + 1). \quad (45)$$

This is equivalent to a computational complexity per bit decision of

$$L[K(8NN_s + 12) - 6NN_s - 7] + \frac{K}{N_b}(2N_b + \log_2(K) + 1). \quad (46)$$

In the narrowband case, we require

$$\left(N_b L\left[2NN_sK + 5K + 8\sum_{k=1}^{K-1} k + (K-1)(5 + 2NN_s)\right]\right) + 2KLNN_s(KL-1) + 2KN_b + K + K\log_2(K) \quad (47)$$

floating point operations per frame and

$$L\left[2NN_sK + 5K + 8\sum_{k=1}^{K-1} k + (K-1)(5 + 2NN_s)\right] + \frac{2KLNN_s(KL-1) + K + K\log_2(K)}{N_b} + 2K \quad (48)$$

flops per bit decision. This of course assumes average power ordering as implemented in this paper. If reliability order is used, the computations would increase by a factor of nearly $N^2/3$.

D. Decorrelating Decision Feedback

Implementation of a decision feedback detector involves estimation of the received energies, sorting of the users, computation of the matched filter bank outputs, generation of the forward filter using a Cholesky decomposition of the matrix $\mathcal{R}$, an inversion of the resulting triangular matrix, forward/feedback filtering, and computation of the decision statistics.

It can be shown that the number of flops required per frame is

$$C_{D\,\text{Feedback}} = 2N_bLKNN_s + \frac{1}{3}(N_bLK)^3 + \frac{9}{2}(N_bLK)^2 - \frac{8}{3}N_bLK + K\log_2(K) + C_{\text{Energy Est.}} \quad (49)$$

The complexity due to estimation of the received energies is included as a separate term since there are different strategies that can be used for this operation.

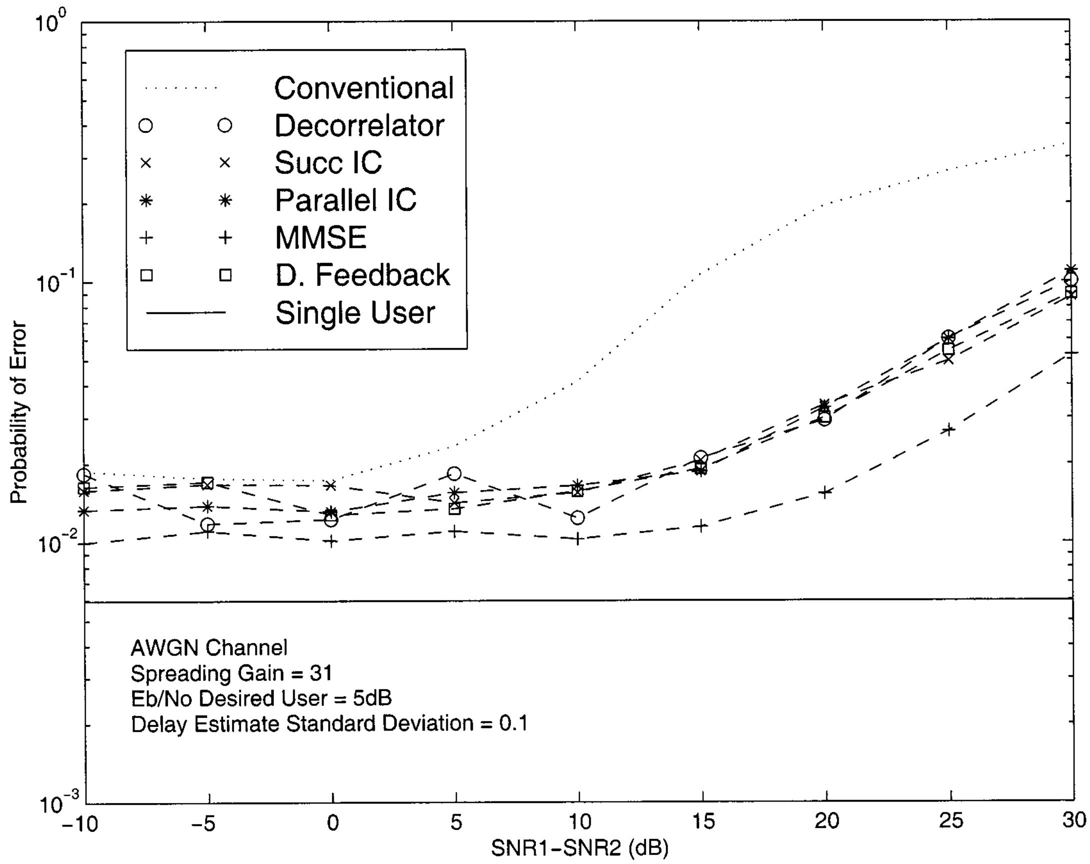

Fig. 9. Effect of Timing Errors (Delay Estimate Errors) on Near–Far performance ($E_b/N_o = 8$ dB, Processing gain $= 31$, $K = 20$).

The complexity per bit decision is given by

$$C_{D\,\text{Feedback}}(b) = 2LKNN_s + \frac{1}{3}N_b^2(LK)^3 + \frac{9}{2}N_b(LK)^2 - \frac{8}{3}LK + \frac{K\log_2(K) + C_{\text{Energy Est.}}}{N_b}. \quad (50)$$

We do not compute the computational complexity of the direct implementation of the MMSE because it will be certainly larger than the decorrelator since not only does it require computation of the cross-correlation matrix and a matrix inversion, but it also requires energy estimation. The computational complexity of the MMSE receiver based on the LMS algorithm is very attractive however, only requiring on the order of $(8NN_s + 1)K$ for computation of the decision statistics and weight adaptation.

An additional note is that the parallel cancellation scheme allows much of its computational burden to be split among separate processors while the successive scheme does not. This is particularly true of the regenerative scheme which requires only that the actual cancellation be performed successively. All other computations can be split among KL processors where K is the number of users and L is the number of paths being tracked. The decorrelator and MMSE schemes may also exploit some amount of parallelism.

E. Results

The preceding equations were used to compare the computational complexity of the different receivers as well as the different possible implementations. Fig. 10 shows the number of floating point operations required per bit decision for the multistage receiver implemented in both wideband (solid line) and narrowband (dashed line), the successive cancellation receiver in both wideband (A) and narrowband (B) as well as the decorrelator. The results are shown for $N = 31$, $N_s = 4$, $L = 2$, $N_b = 100$, and $S = 3$. Due to the matrix inversion operation, direct implementation of the decorrelator is extremely computationally intensive. In this plot, the curve for the decorrelator corresponds to an estimated lower bound for the reduced complexity iterative technique of [43], assuming eight iterations. For the subractive interference cancellation receivers, the narrowband implementations evidence computational advantages over the wideband approach. The parallel cancellation scheme is more expensive than the successive scheme for both the narrowband and wideband approaches, although the advantage is not nearly as significant for the narrowband approach. The maximum computational requirements for a single processor in the parallel cancellation approach are also shown in Fig. 10 (curves marked by '□'), showing that the maximum complexity required for a single processor is significantly less than the successive scheme which offers no complexity reduction due to parallelism. The computational complexity of the decorrelating decision feedback was not plotted since it will vary with the approach utilized to create the energy estimates. However, its computational complexity would be similar to the decorrelator for reasonable energy estimation algorithms, such as matched filtering.

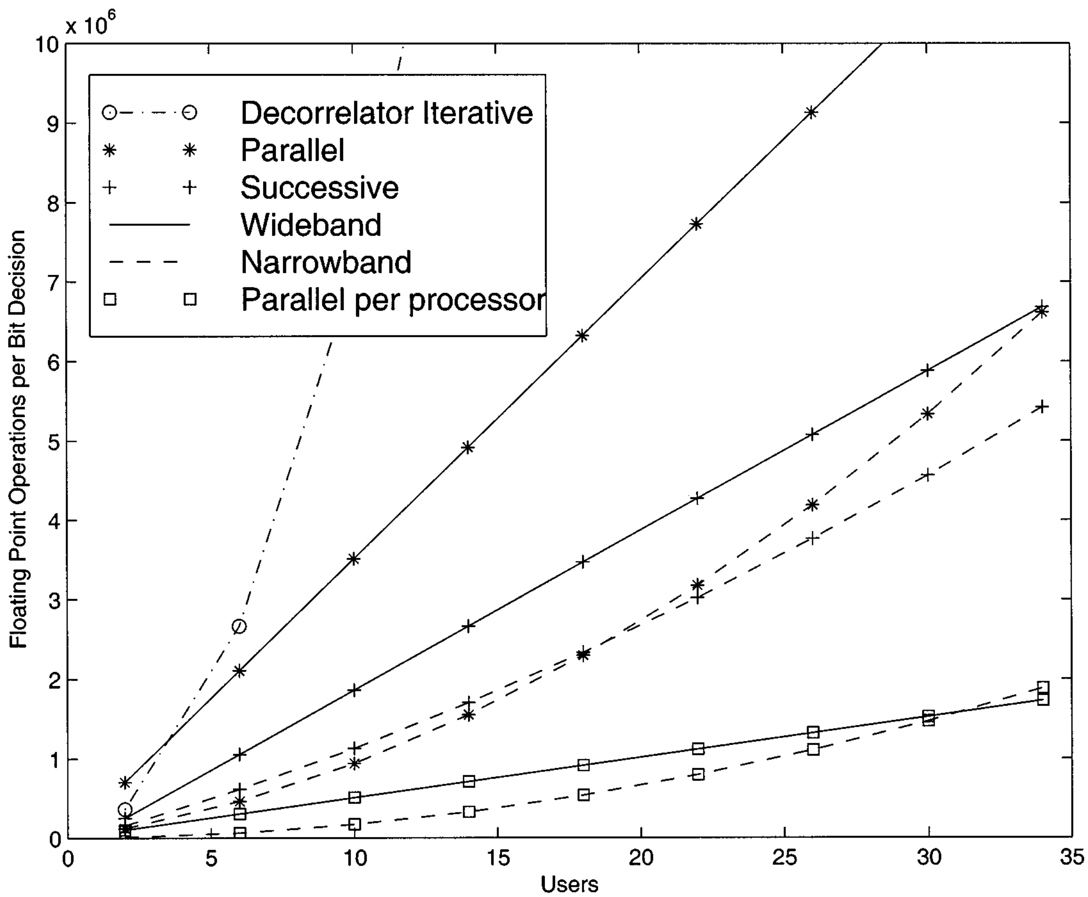

Fig. 10. Computational Requirements for Multiuser Receiver Implementations (Processing Gain $= 31$, Samples per Chip $= 4$, Paths $= 2$, Frame Size $= 100$ bits).

VI. Multiuser Detection in Multi-Cell Scenarios

In single-cell cases, ideal multiuser detection creates an equivalent of a single-user channel, removing all interference other than background noise. In the multi-cell case, one possible scenario is that the multiuser detector focuses only on interference from users in the same cell while neglecting the out-of-cell interferers. An upper limit to the capacity improvements achievable with multiuser detection in multi-cell systems can be computed assuming the multiuser detector's ability to cancel all the interference within the cell. Using the popularly accepted value of 0.6 for the ratio of the ratio of intra-cell to out-of-cell interference results in an estimated improvement on the order of 2.7 times over conventional CDMA systems.

A more favorable scenario would occur if multiuser detection is not restricted to interferers within the cell. In this case, significant advantages are obtained by also taking into account the strongest out-of-cell interferers. In [44], analytical and simulation results estimate that selective interference cancellation of strong out-of-cell interferers can potentially result in capacity increases of 20%–40%. Robustness to the Near–Far problem is another important motivation for adopting multiuser receivers in cellular systems.

VII. Noncoherent Implementations

The receiver structures to this point have focused on coherent reception. While some researchers are investigating coherent reception using pilot symbols [45], the speed of phase fluctuation in the mobile channel often discourages coherent reception. Thus it is desirable to investigate multiuser techniques which do not require a coherent phase reference. The most straightforward extension to noncoherent multiuser reception is the implementation of the preceding receivers using differential modulation. In this section we investigate the performance of the decorrelator, parallel cancellation, and successive cancellation employing differential detection. Additionally we show that the cancellation approaches extend to M-ary orthogonal signaling.

A. Differential Detection

1) The Decorrelator: Differential decoding requires the modulation symbols to be transmitted so that information is transmitted via changes in the phase of the signal rather than the absolute phase. It follows that the decision statistic of the conventional receiver is

$$Z_{i,k} = y^I_{i,k} * y^I_{i-1,k} + y^Q_{i,k} * y^Q_{i-1,k}. \tag{51}$$

For a differential decorrelator [46], the transform $\mathcal{R}^{-1}$ would first be applied to the in-phase and quadrature arms prior to differential detection. The result is a straightforward extension to the coherent decorrelator. The simulated performance of the differential decorrelator in an AWGN channel with $K = 10$ users, spreading gain $N = 31$, and perfect power control is presented

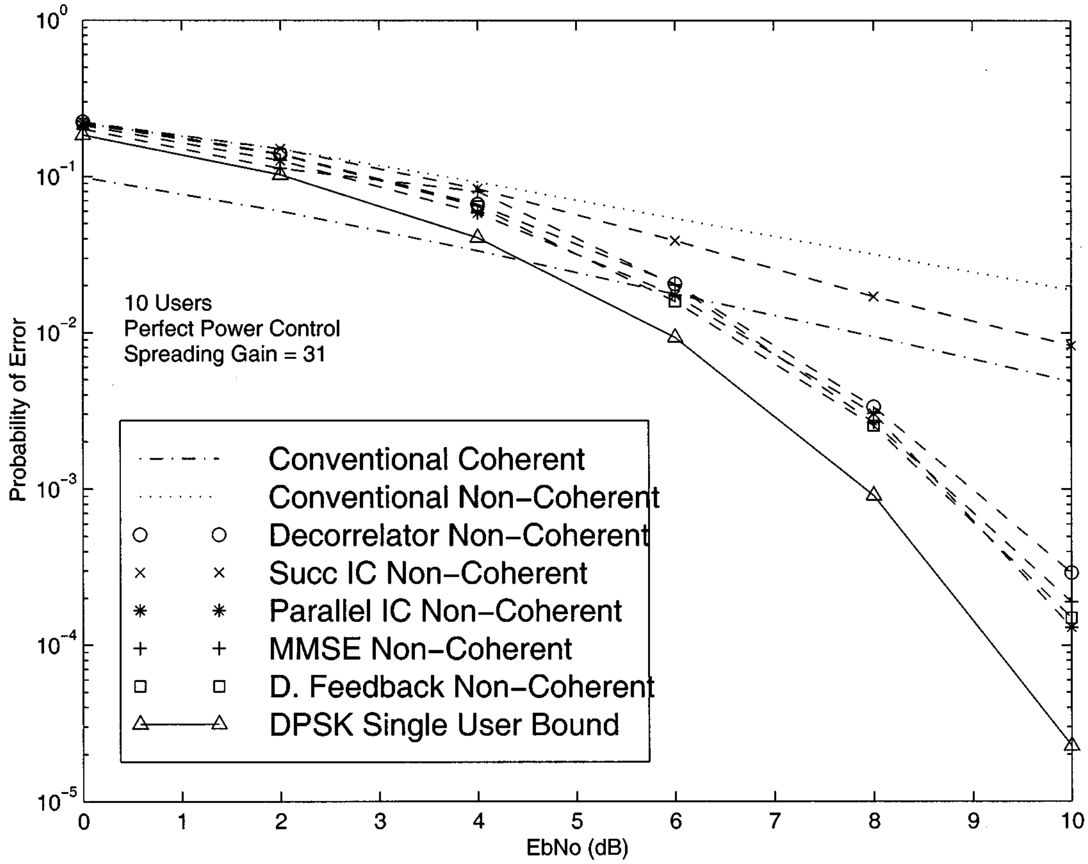

Fig. 11. BER versus E_b/N_o with Perfect Power Control using Differential Detection (ten users and processing gain $= 31$).

in Fig. 11. The coherent conventional receiver is also shown for reference. We can see that there is an expected reduction in performance due to the noncoherent reception, but otherwise the decorrelator shows the types of gains over the conventional noncoherent receiver as present in the coherent case. Performance in flat Rayleigh fading is shown in Fig. 12 where $K = 10$, spreading gain $N = 31$, and $\sigma = 0.93$.

2) MMSE: Operation of the differentially coherent MMSE receiver, parallels that of the differential decorrelator. Fig. 11 shows that as in the coherent case, the MMSE receiver outperforms the decorrelator. Perfect channel estimation is assumed for the MMSE receiver however. Performance in flat Rayleigh fading is illustrated in Fig. 12. Notice that for AWGN and Rayleigh fading, in the low SNR region the MMSE outperforms the decorrelator. The performance at high SNR's is basically equivalent.

3) Parallel Cancellation: For noncoherent parallel cancellation we follow the same approach as coherent cancellation with the exception that cancellation occurs in both in-phase and quadrature channels separately. The estimate used for cancellation is simply the in-phase or quadrature channels correlated with a synchronous copy of the spreading code. These estimates will contain inherent phase information for the cancellation process. That is the expected value of $\int r_i(t)a(t-\tau_k)\,dt$ is $\sqrt{P_k}b\cos(\theta_k)$. A similar result is true for the quadrature channel. Once these estimates of the $I\&Q$ interferers are removed, the $I\&Q$ channels are again passed through a matched filter for re-estimation. Once cancellation is complete (i.e., all s stages), the resulting in-phase and quadrature estimates are combined according to (51) to form the final decision statistic. The performance of the differential version of parallel cancellation is shown in Fig. 11 for an AWGN channel with $K = 10$ users, spreading gain $N = 31$, and perfect power control. Performance improvements over the conventional receiver are similar to the coherent case. Results for flat Rayleigh fading are presented in Fig. 12, showing the degradation due to the additional noise involved in the cancellation process. One bit estimates were assumed for the cancellation approaches since we are assuming that the channel does not permit long averaging times. In the coherent case we know the phase explicitly while in the current case we are implicitly cancelling utilizing an estimate of the phase. This additional noise degrades the accuracy of the cancellation. Thus the performance is weaker than the noncoherent decorrelator which has no such degradation.

4) Successive Cancellation: Successive cancellation using differential encoding is similar to the parallel case with the exception that users are cancelled only once and in descending order of received power levels. Intrinsic phase estimates provided by the in-phase and quadrature decision statistics. Fig. 11 shows that with perfect power control the differential successive canceller exhibits poor performance when compared with the other receivers. In Rayleigh fading, the successive canceler benefits from diverse received and its performance is similar to that of the partial parallel interference cancellation receiver.

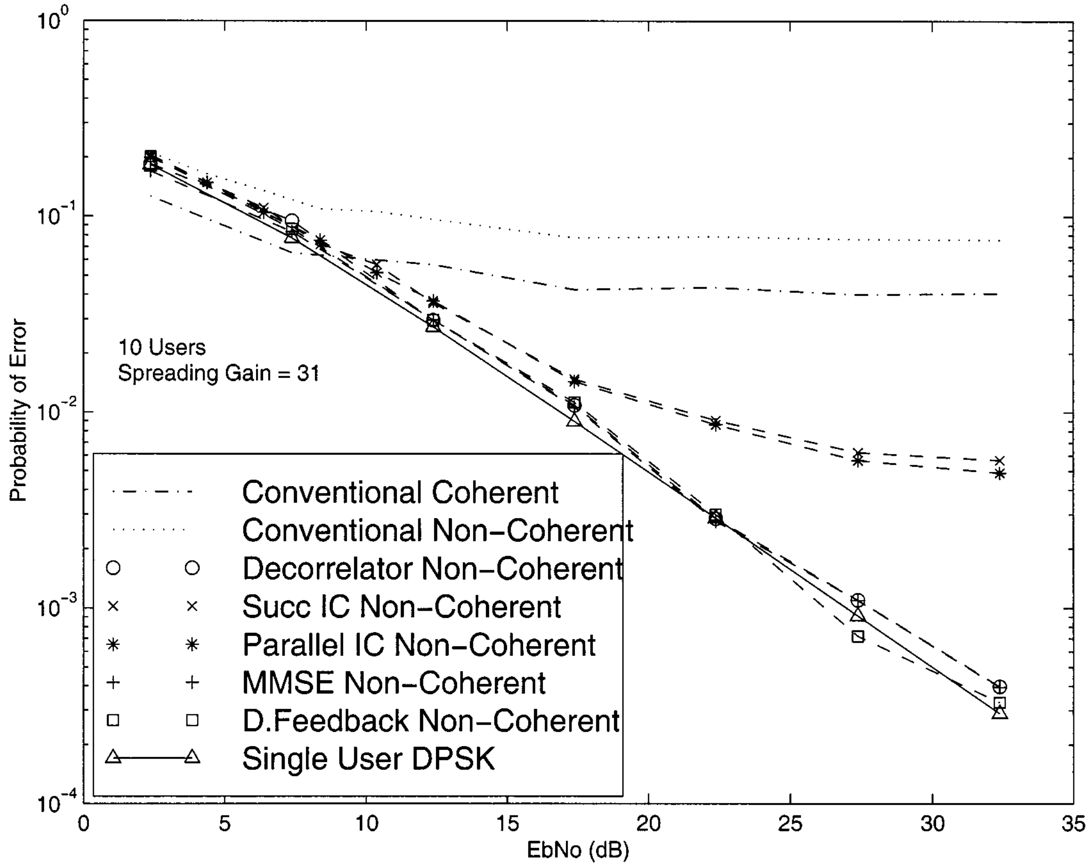

Fig. 12. BER versus E_b/N_o for Flat Rayleigh Fading (ten users, processing gain $= 31$).

5) Decision Feedback: For the Decision Feedback receiver, the feedforward filter is applied to the in-phase and quadrature matched filter outputs. The feedback filter uses the forward filtered soft decision statistics z_I and z_Q, which have inherent phase information, to provide an estimate for $\exp(j\Theta)\hat{\mathbf{b}}$ in (36) and (37) (recall that the bits have been differentially encoded at the transmitter). Figs. 11 and 12 show that in AWGN and flat Rayleigh fading channels, the decision feedback receiver approaches the single user bound for DPSK.

6) Near/Far and Delay Estimation Errors: The performance degradation in near–far channels of the differential receivers is illustrated in Fig. 13. In this case, we have three users in an AWGN channel. The probability of error for users 1 and 2 is plotted as the power of user 3 is varied from 10-dB below the signal power of the users of interest to 30-dB above. As in the coherent case, the parallel interference cancellation approach is not strictly near–far resistant. However, this figure indicates that all the receivers display significant robustness to the near–far problem for reasonable power differences. The near–far performance of the multistage approach can be improved by performing selective cancellation, that is, not canceling those weak users for which the estimates are unreliable.

Fig. 14 shows the degradation experienced in the presence of delay estimation errors. The results for the differentially coherent receivers are analogous to those observed in the coherent case, where a delay estimation error of $0.3T_c$ or larger eliminates the advantages of multiuser detection.

B. M-Ary Orthogonal Signaling

Another method of noncoherent reception in use today in M-ary orthogonal signaling [47], which is used in IS-95. Note that differential detection, addressed in the previous section, is a special case of noncoherent orthogonal modulation when it is considered over two bit intervals. In this section, in order to gauge the potential performance improvements afforded by multiuser detection, the performance of a system with some of the key features of today's system augmented with a multiuser detector will be studied. For practical implementation of multiuser detection, subtractive interference cancellation approaches are preferable since nonlinear modulation is used which all but prohibits linear multiuser detection techniques. This means that subtractive cancellation methods are the only practical multiuser detection methods for IS-95 systems. For this reason we present simulations comparing successive and parallel cancellation. The performance of successive cancellation using this modulation format has been discussed in [31]. M-ary Walsh codes are used to represent $\log 2(M)$ symbols to provide a spreading of $M/\log 2(M)$. This may alternatively be considered as a form of block coding. Additionally, the signal is transmitted over both in-phase and quadrature channels and envelope detection can be performed. The two channels are spread with separate $I\&Q$ short spreading codes before being spread with a user-defined long code. By applying the two spreading codes an additional spreading gain is achieved.

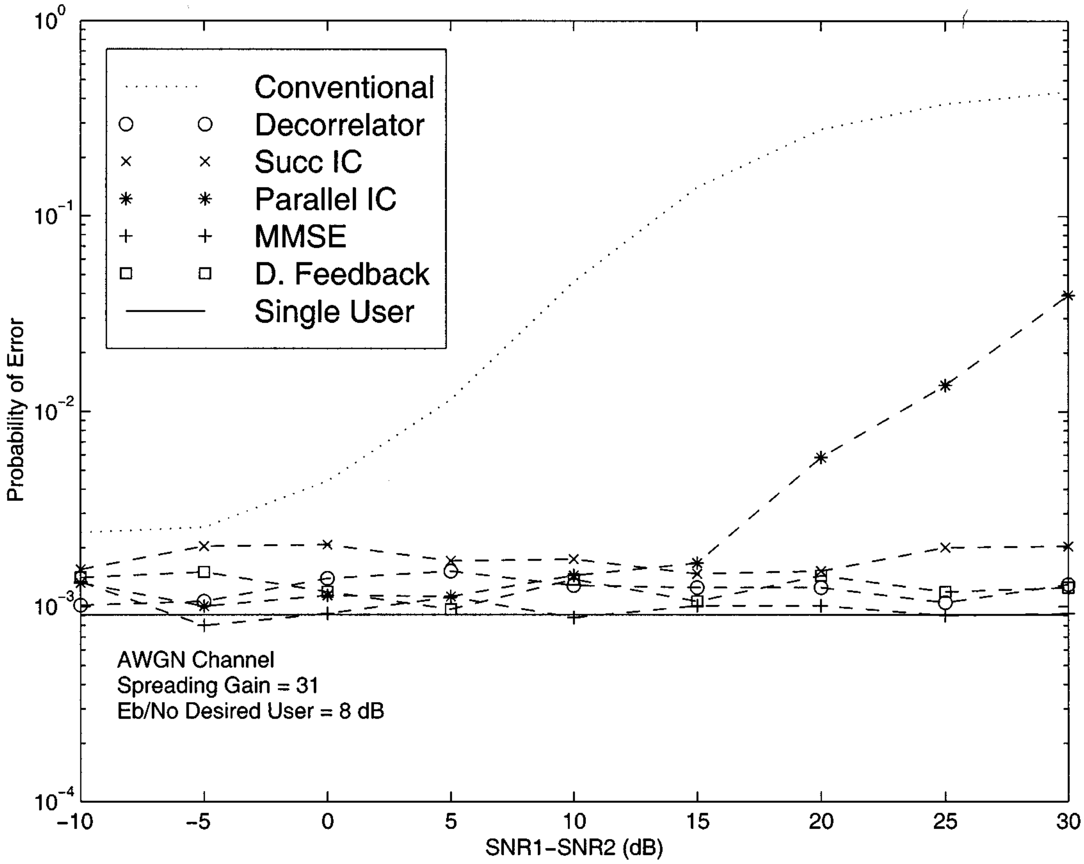

Fig. 13. Performance degradation of noncoherent receivers in Near–Far Channels ($\overline{E_b/N_o} = 8$ dB and processing gain $= 31$).

At the receiver, the incoming signal is mixed down with both an in-phase and a quadrature channel without a phase reference and is despread with the appropriate long user code as well as the associated I or Q short code. This provides four sets of decision statistics at the receiver after despreading and taking the Walsh transform [48]

$$\begin{aligned} Z_{I,I}^m &= \int r(t)\cos(\omega_c t)W_m(t)a_k(t-\tau_k)a_I(t-\tau_k)\,dt \\ Z_{I,Q}^m &= \int r(t)\cos(\omega_c t)W_m(t)a_k(t-\tau_k)a_Q(t-\tau_k)\,dt \\ Z_{Q,I}^m &= \int r(t)\sin(\omega_c t)W_m(t)a_k(t-\tau_k)a_I(t-\tau_k)\,dt \\ Z_{Q,Q}^m &= \int r(t)\sin(\omega_c t)W_m(t)a_k(t-\tau_k)a_Q(t-\tau_k)\,dt \end{aligned} \quad (52)$$

where $r(t)$ is the received signal, $a_k(t)$ is the individual user's long spreading code, $a_I(t)$ and $a_Q(t)$ are the $I\&Q$ short codes and $W_m(t)$ is the mth Walsh code. By combining these using $Z = (Z_{I,I} + Z_{Q,Q})^2 + (Z_{I,Q} - Z_{Q,I})^2$ we can determine the most likely transmitted Walsh function.

Using the estimate of the Walsh function, the known timing information and the four estimates given above, we can estimate the signals in the in-phase and quadrature arms and perform cancellation, i.e.,

$$\begin{aligned} r_I^k(t) = r_I(t) &- \sum_{j\neq k} Z_{I,I}^m W_m(t-\tau_k)a_k(t-\tau_k)a_I(t-\tau_k) \\ &- \sum_{j\neq k} Z_{I,Q}^m W_m(t-\tau_k)a_k(t-\tau_k)a_Q(t-\tau_k) \end{aligned} \quad (53)$$

and

$$\begin{aligned} r_Q^k(t) = r_Q(t) &- \sum_{j\neq k} Z_{Q,I}^m W_m(t-\tau_k)a_k(t-\tau_k)a_I(t-\tau_k) \\ &- \sum_{j\neq k} Z_{Q,Q}^m W_m(t-\tau_k)a_k(t-\tau_k)a_Q(t-\tau_k). \end{aligned} \quad (54)$$

Once cancellation has been performed, re-estimation can be performed as described above using the new $I\&Q$ signals for each user. The parallel and successive schemes are similar with the difference being that users are cancelled only once in the order of descending powers in the successive scheme. Whereas, in the parallel scheme, estimation will occur in parallel and the cancellation process will occur S times.

The performance of 64-ary signaling in AWGN is shown for ten users in perfect power control in Fig. 15 for both successive and parallel cancellation. The long and short codes have a chip rate which is four times that of the Walsh chip rate providing a spreading gain of $(64/log_2(64))4 = 42.7$. While the successive cancellation provides some gain, parallel cancellation ($s = 3$) shows well over a magnitude of improvement for $E_b/N_o \geq$6-dB. This figure also includes curves for BPSK and DPSK parallel and successive interference cancellation receivers with random signature sequences and a spreading gain of 43. The M-ary receivers provide better performance than the DPSK receivers. At high SNR's, the noncoherent orthogonal interference cancellation receivers outperform their BPSK

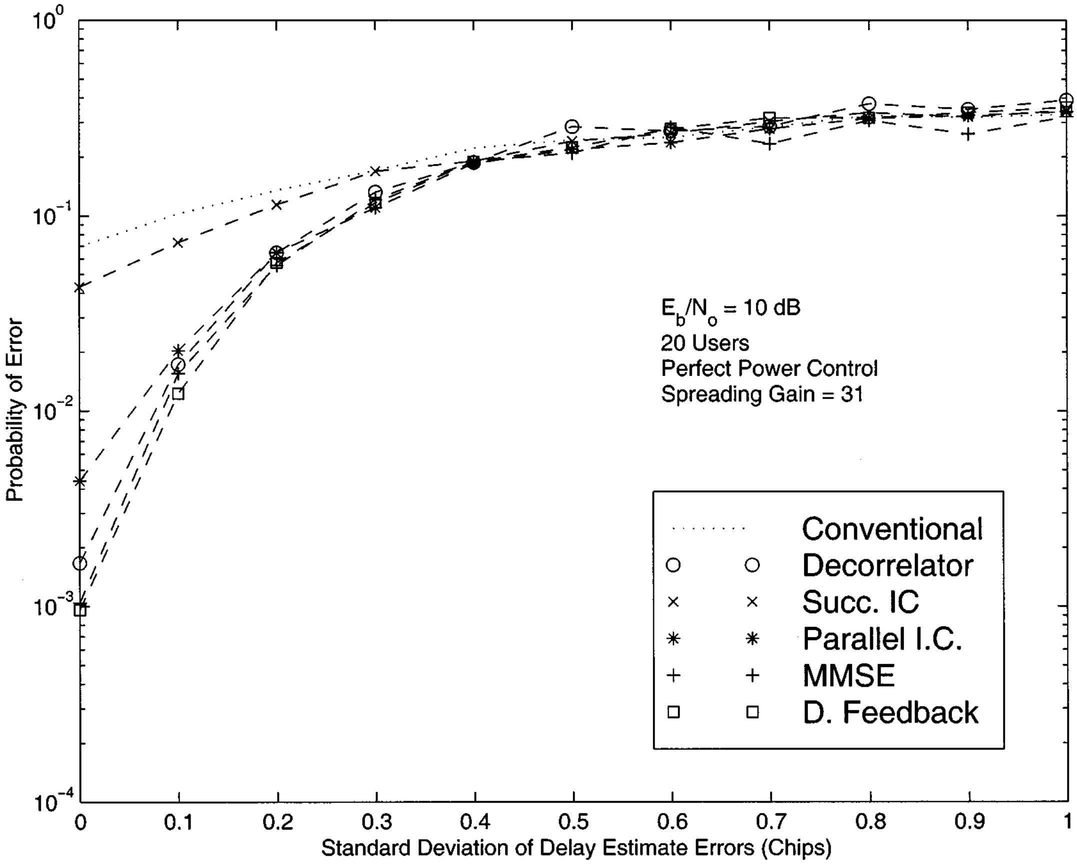

Fig. 14. Effect of Timing Errors (Delay Estimate Errors) on System Capacity in AWGN Channel with Perfect Power Control for Noncoherent Receivers ($E_b/N_o = 8$ dB, Processing gain $= 31$, $K = 20$).

counterparts due to the combination of orthogonal modulation and block coding in the M-ary system, and the lack of coding in the BPSK systems.

The previous simulation results also suggest that pseudo random CDMA systems such as IS-95 can also benefit from joint detection at the base-station as code-on-pulse systems do (in code-on-pulse systems, the spreading sequences have a period equal to one symbol).

VIII. Conclusions and Future Research

In this paper, we have provided a description of five multiuser receivers proposed for CDMA systems. The issues discussed include detection theory, complexity, capacity, and near–far resistance. Additionally, noncoherent versions of these receivers were also presented.

We have shown that in an AWGN channel with perfect or imperfect power control multiuser receivers provide significant gains over the conventional matched filter. We found that for perfect power control, the multistage parallel cancellation receiver, decorrelator, MMSE and decorrelating decision feedback receivers significantly outperform the successive cancelation approach. However, as the power control error is increased, the performance of successive cancellation rivals that of the decorrelator, while the parallel cancellation receiver shows a slow degradation. All are significantly more robust to energy disparity than the conventional receiver.

When channel estimation is not an issue, nonlinear multiuser detection techniques proved to provide the best performance. However, in fading when channel estimation is performed linear receivers gave better performance since explicit channel estimation is not needed for interference mitigation only for detection. If noncoherent detection is performed, channel estimation is not required at all. Thus, in situations where channel tracking is difficult the linear detectors are preferred. However, if channel estimation performance is adequate and the desired BER's are in the range of 10^{-2}, then the performance of the detectors are nearly identical.

In terms of complexity, it was shown that the decorrelator (and thus the MMSE and decorrelating DF structures) was extremely sensitive to the amount of update required for the matrix inverse coefficients. If a new matrix inversion must be performed at regular intervals, the complexity becomes a major impediment to implementation. However, if simple coefficient updates will provide sufficient dynamics, the scheme is the least computationally intensive. Between the two cancellation schemes, successive is overall less computationally intensive, but the parallel scheme is more flexible, allowing slower processors in parallel to perform the computation. The successive scheme provides no such flexibility. Also, cancellation approaches which avoid regeneration of the wideband signal can provide significant computational savings at the cost of

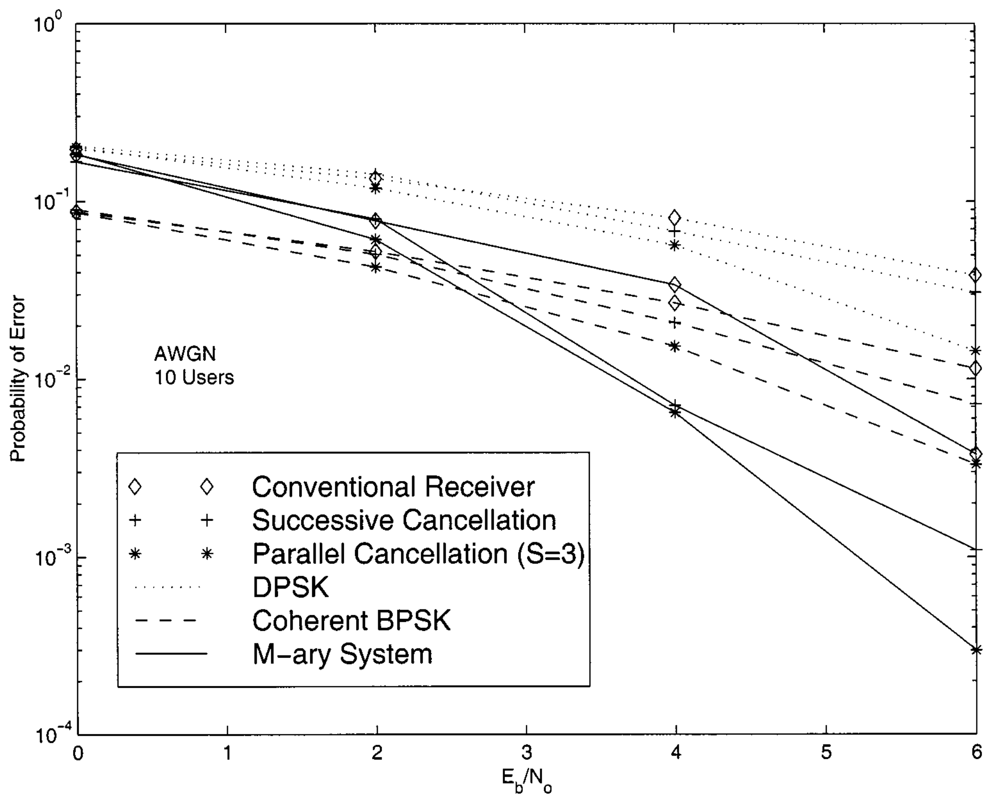

Fig. 15. BER versus E_b/N_o BER performance for Multistage Cancellation, Successive Cancellation and Conventional Reception in AWGN for M-ary Orthogonal Signaling (ten users, $M = 64$, Spreading Gain $= 42.7$), DPSK and BPSK (Spreading Gain $= 43$).

memory storage requirements. The exact amount of savings achieved will depend on the rate of update required for the correlation matrix.

Timing misalignment was shown to be crucial but not fatal to each of the five receivers. The decorrelator, MMSE, and DF receivers appear to be more sensitive to timing misalignment, although none of the structures can withstand timing errors above $0.3T_c$ where T_c is the chip period.

Another issue in implementation is the modulation scheme used as well as the use of long or short spreading codes. If nonlinear modulation is used, the linear approaches are not valid. Additionally, if long spreading codes are used the complexity of the linear approaches is prohibitive leading the system designer to prefer subtractive cancellation methods.

It is clear that the multiuser receiver, if made economically feasible, can greatly increase the attractiveness of CDMA for cellular and PCS. Future research should focus on specific implementation issues such as synchronization, robustness to frequency offsets, efficient implementation for the necessary algorithms, the effect of narrowband interference, as well as others.

APPENDIX
CONFIDENCE INTERVALS FOR SIMULATION COMPARISON

Monte Carlo simulation is simply a sequence of Bernoulli trials where one performs multiple independent experiments and computes an ensemble average by counting the number of successes (or errors in our context) and dividing by the number of trials [49]. The true probability of error of a system p is estimated with the sample mean, i.e.,

$$\hat{p} = \frac{1}{n}\sum_{i=1}^{n} g(\hat{b}_i) \tag{55}$$

where n is the number of observed symbols and $g(\hat{b}_i)$ is the error estimator

$$g(\hat{b}_i) = \begin{cases} 1 & \hat{b}_i \neq b_i \\ 0 & \hat{b}_i = b_i \end{cases} \tag{56}$$

Thus, assuming that all symbols will occur with equal probability, $\hat{p} = m/n$ where m is the number of occurrences or errors. By the law of large numbers as $n \to \infty$, $\hat{p}$ converges to p. However, to allow $n \to \infty$ would require infinite processing time. Since we must run less than infinitely long processes, we would like to know how close our estimation is to the true BER. This can be accomplished exactly by realizing that for a given n, $m\hat{p}$ is binomially distributed [49]. One powerful and simple approximation is to assume that the $\hat{p}$ is Gaussian distributed, which is appropriate for large n. Using this approximation one can create a confidence interval of the form [49]

$$\begin{aligned} P\Bigg\{ &\frac{n}{n+d_\alpha^2}\left[\hat{p} + \frac{d_\alpha^2}{2n} - d_\alpha\left(\frac{\hat{p}(1-\hat{p})}{n} + \frac{d_\alpha^2}{2n}\right)^{1/2}\right] \\ &\le p \le \frac{n}{n+d_\alpha^2}\left[\hat{p} + \frac{d_\alpha^2}{2n} + d_\alpha\left(\frac{\hat{p}(1-\hat{p})}{n} + \frac{d_\alpha^2}{2n}\right)^{1/2}\right]\Bigg\} \\ &= 1-\alpha \end{aligned} \tag{57}$$

where d_α is chosen such that

$$\frac{1}{\sqrt{2\pi}} \int_{-d_\alpha}^{d_\alpha} e^{-t^2/2}\, dt = 1 - \alpha. \tag{58}$$

Using this approximation it can be shown that a 95% confidence interval is equivalent to $(1.8\hat{p}, 0.55\hat{p})$ when $n = 10/p$. Using this as a guide, for large n our simulations would require a minimum error count of $n = 10$. Thus, we typically run simulations such that approximately 10–50 errors would occur for the lowest BER desired. Since in our simulations the same number of bits were run for all the test points in a given set, this resulted in order of magnitude increases in the number of observed errors at high (10^{-1}) BER's. At higher BER's the bounds are much tighter due to the larger number of observed errors.

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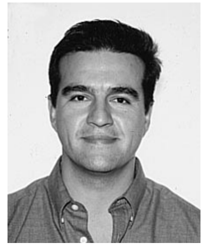

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