Iterative Multiuser Detection Using Antenna Arrays and FEC on Multipath Channels

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***Abstract*—This paper investigates a multiple-access communication receiver system that receives coded data modulated using either direct-sequence code division multiple access or narrow-band binary phase shift keying, with an antenna array in a multipath propagation environment. We describe an iterative receiver that improves the initial estimates from the antenna array, and therefore reduces the multiple access interference. Simulation results show that the bit error rate performance approaches that obtained when only one user's signal is incident on the array. This occurs even with a large number of users in comparison to the product of the spreading gain and array size.**

***Index Terms*— Array signal processing, channel coding, code division multiple access, iterative methods, multiuser channels, turbo codes.**

I. Introduction

WITH capacity constraints becoming a problem in current mobile communication systems, researchers are looking for ways of enhancing system performance. A topic of interest is antenna arrays and array signal processing, which has been studied extensively by the signal-processing community [1]–[3]. Antenna arrays allow spatial separation of users even within the same frequency band; this technique is also known as space division multiple access (SDMA). SDMA has been shown to increase the capacity of systems by reducing the cochannel interference between users that are spatially separated. We define SDMA to be the exploitation of spatially separated users by an antenna array with the antenna elements close enough together to experience the same multipath and fading effects but with independent thermal noise.

Spatial diversity results to date have detailed techniques that use SDMA for narrow-band [4]–[6] and direct-sequence code division multiple access (DS-CDMA) [7], [8]. A number of papers now have recognized the fact that spatial and temporal interference[1] can be combined and viewed as intersymbol interference (ISI). This is discussed in [9]–[13], where the exponentially complex maximum likelihood sequence estimation solution is attained by the use of the Viterbi equalizer [14].

Another method of removing the ISI is the use of null steering [1], [9]. This technique is analogous to the linear operation of decorrelation in DS-CDMA [15], which is known to enhance the noise variance of the detector output and not reach the optimal Gaussian noise performance. Separation of users when no spatial separation exists is only possible by using some other form of diversity such as code separation with DS-CDMA [7] or time or frequency separation by using system management [16] to reallocate time or frequency slots.

We utilize a unified system model for two different channels, one with SDMA, DS-CDMA, and multipath and another with only SDMA. We show that spatial separation can be modeled as a complex spreading code in a similar manner as a spreading code in DS-CDMA. We incorporate multipath into the model and show that it yields a channel model with memory. We assume that each user first encodes and then interleaves their information sequences prior to transmission.

Low-complexity interference cancelation receiver solutions have been discussed in [17]–[20], where the solutions are applicable to asynchronous user channels where there is no need for a causal channel and the associated whitening filter. We extend this low-complexity interference canceler design work [20] to the unified channel model and demonstrate how this iterative structure can be used for DS-CDMA, SDMA, narrow-band binary phase shift keying (BPSK), and multipath to achieve optimal (Gaussian noise channel) performance. The optimal receiver is prohibitively complex [21], for a realistic number of users ($>$10), and this complexity increases exponentially with the number of users and channel code memory. We instead utilize a simpler design that is linear in complexity, with respect to the number of users. This technique uses concepts from turbo coding [22] and a low-complexity receiver also independently discovered in [23].

This paper is organized as follows. In Section II, we discuss the system model. Section III describes the receiver, MAP decoding, and iterative process, with Section IV showing the complexity of the receiver/decoder. Sensitivity to parameter errors is discussed in Section V before simulation results are presented in Section VI. Conclusions are drawn in Section VII. Throughout this paper, scalars are lower case, vectors are underlined lower case, and matrices are underlined upper case. The symbols $\underline{0}_L, \otimes, (\cdot)^\top$, and $(\cdot)^H$ are the zero vector of length L, the Kronecker tensor product [24],

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[1] Temporal interference is assumed here to indicate interference that occurs from more than one user's transmitting in the same bandwidth at the same time

transposition, and complex-conjugate transposition operators respectively.

II. System Model

We shall employ a model similar to that described in [25]–[27], where we assume that K users each transmit T symbols $\{d_{k,t} \in \mathcal{D}\colon t = 1, \ldots, T\}$, where $\mathcal{D} = \{+1, -1\}$ as used for BPSK modulation. The spreading code for user k at time t is defined as $\{\underline{s}_{k,t} \in \{-1/\sqrt{N},\ 1/\sqrt{N}\}^N\colon t = 1, \ldots, T\}$, where the length of the spreading code is N. Note that our model permits the spreading code to change from symbol interval to symbol interval. We shall further assume that the multipath impulse response of the channels over which the users transmit to the base station are invariant during the T symbol intervals. This condition is enforced by low vehicle speeds, high signalling rate, and short packet lengths T. The sequences transmitted across the channel, $\{d_{k,t} \in \mathcal{D}\colon t = 1, \ldots, T\}$, are the result of encoding and subsequent interleaving of an information sequence. We assume that a method is available to decode a noise corrupted version of the channel sequence $\{d_{k,t} \in \mathcal{D}\colon t = 1, \ldots, T\}$ to obtain *a posteriori* likelihoods for the channel symbols and the information symbols. Typically, an encoder/decoder pair would consist of a convolutional encoder and a MAP decoder [28].

An important assumption we shall make is that the delay of the multipath taps is externally supplied. This would be normally done via a sliding correlator with dither control [29]. Tap p of a total of P paths per user is therefore represented by a complex amplitude $\alpha_{k,p}$. It has a known delay $\tau_{k,p} \in (0, N)$. We model the use of an L element antenna array at the receiver by defining for each user k, multipath p, and antenna element l, a set of L antenna element phases $\phi_{k,p,l}$. Note that we allow multipath components to arrive from different directions.

Let us sample the down-converted signal at each of the receiver antennas at one sample per chip interval and collect all of the samples into one stream $\underline{e}$. We may now write

$$\underline{e} = \underline{A}_S \underline{A}_A \underline{A}_M \underline{d} + \underline{n} = \underline{A}\underline{d} + \underline{n}$$

where $\underline{A} = \underline{A}_S \underline{A}_A \underline{A}_M$. The constituent matrices are similar to those discussed in [27] and are now described.

$\underline{A}_S$, $(T+1)NL$ by $TLPK$. This contains information pertaining to the spreading codes and the multipath delays (all of which are assumed known at the receiver). It has column j

$$\begin{bmatrix} \underline{0}_{((t-1)N+\tau_{k,p})L+l} \\ (\underline{s}_{k,t}) \otimes [1, \underline{0}_{L-1}^\top]^\top \\ \underline{0} \end{bmatrix} \in R^{(T+1)NL,1}$$

where j, k, t, p, and l are uniquely related by

$$j = (((t-1)K + k - 1)P + p - 1)L + l.$$

Each column of this matrix contains a spreading code, and when applied via the inner product operation, a spreading code correlation results. For time t there are LPK such correlators, one for each antenna, multipath, and user.

$\underline{A}_A$, $TLPK$ by TPK. This matrix describes the phases of each user's multipath component at each antenna element over time. It has columns j

$$\begin{bmatrix} \underline{0}_{(j-1)L} \\ e^{j\phi_{k,p,u}} \\ \vdots \\ e^{j\phi_{k,p,L}} \\ \underline{0}_{(TPK-j)L} \end{bmatrix} \in C^{TLPK,1}$$

where j, k, t, and p are uniquely related by

$$j = ((t-1)K + k - 1)P + p.$$

Each column of this matrix is a set of phases that when applied via the inner product operator result in a beamforming operation. There are TKP such beamformers, one for each path. Given an angle of arrival $\theta_{k,p}$ at a $\lambda/2$ spaced uniform linear array, the phases are determined according to $\phi_{k,p,l} = \pi(l-1)\cos(\theta_{k,p})$. Here λ denotes the carrier frequency wavelength. Note that there is no time reference for the symbols $\alpha, \phi, \theta, \tau$, as we assume they are stationary for each frame interval (but reselected each frame interval).

$\underline{A}_M$, TPK by TK. The received amplitude and phase of each of the TKP multipath components as they would be measured on a single antenna are included in this matrix. It has column j

$$\begin{bmatrix} \underline{0}_{(j-1)P} \\ \alpha_{k,1} \\ \vdots \\ \alpha_{k,P} \\ \underline{0}_{(TK-j)P} \end{bmatrix} \in C^{TPK,1}$$

where j, k, and t are uniquely related by

$$j = (t-1)K + k.$$

Each column of this matrix contains the impulse response of the multipath channel. Its application via the inner product operation results in a RAKE combiner. There are TK such combiners, one for each symbol transmitted by each user.

III. The Iterative Receiver

The receiver structure developed here follows the methods described in [18], [20], [19] and [23]. The differences are that we implement the interference canceler in the spread domain (which substantially reduces complexity), that is, we perform the interference cancelation at the chip rate, not at the coded bit rate.

The receiver, which is iterative, takes the output of the channel and performs a combination of soft interference cancelation and matched filtering, as shown in Fig. 1.

The interference canceler takes *a priori* probabilities on the channel symbols of every user, $\{P(d_{k,t})\colon t = 1, \ldots, T, k = 1, \ldots, K\}$. These are then employed to construct a soft estimate of the symbol by taking the expectation

$$\tilde{d}_{k,t} = \sum_{d \in \mathcal{D}} P(d_{k,t} = d)\, d. \tag{1}$$

The interference cancelation operation is then performed in the spread domain as follows:

$$\underline{x} = \underline{A}^H \underline{e}_k + \underline{z}_k = \underline{A}^H(\underline{e} - \underline{A}\underline{\tilde{d}}) + \underline{W\tilde{d}} \tag{2}$$

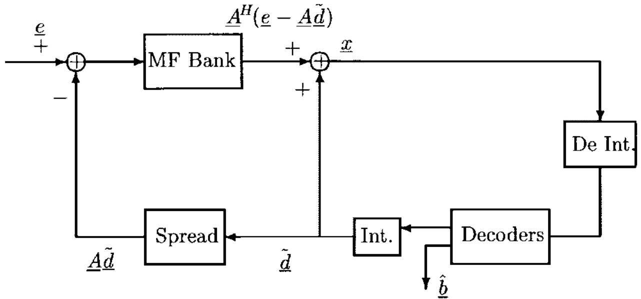

Fig. 1. Spread domain IC/decoder structure.

where $\underline{W} = \text{diag}(\underline{A}^H\underline{A})$ and $\underline{e}_k$ is the signal $\underline{e}$ with contributions from other users removed. Here a particular element of $\underline{x}$ corresponding to a particular symbol $d_{k,t}$ is

$$x_{k,t} = \underline{A}^H[k,t](\underline{e} - \underline{A}\tilde{\underline{d}}) + w_{k,t}\tilde{d}_{k,t} \tag{3}$$

where $\underline{A}[k,t]$ is column $j = (t-1)K + k$ of $\underline{A}$ and is the channel response of user k in symbol interval t.

The method used to produce the interference canceler outputs can vary. In [20], the channel output was matched filtered then used by the interference canceler to produce the MAP decoder metrics. The problem with this method is that the channel needs to be generated by computing the cross-correlation matrix $\underline{A}^H\underline{A} = \underline{H}$, which is a dominating complexity term. We instead employ the spread interference canceler solution. This approach is similar to a technique used in successive interference cancelation [30], where data estimates are constructed and then respread.

As suggested by (2), the vector $\underline{x}$ can be treated as a noise-perturbed version of $\underline{d}$, i.e., $x_{k,t} = d_{k,t} + z_{k,t}$, where $z_{k,t}$ is Gaussian with variance σ_k^2 [20] enabling the generation of the *posteriori* probabilities $P(d_{k,t})$ since $x_{k,t} \sim N(d, \sigma_k^2)$.

The variance σ_k^2 of the zero mean Gaussian noise sequence $\{z_{k,1}, \ldots, z_{k,T}\}$ is

$$\sigma_k^2 = \text{E}\left\{(x_{k,t} - d_{k,t})^2\right\} \tag{4}$$

but since $d_{k,t}$ is not available at the receiver, we instead form

$$\sigma_k^2 \approx \text{E}\{(x_{k,t} - \tilde{d}_{k,t})^2\}. \tag{5}$$

We expect this to be a good approximation since when $\tilde{d}_{k,t}$ is an inaccurate estimate, it is likely that the noise will be dominated by signals from other users. The likelihoods generated from $x_{k,t}$ are then reordered, for each user, according to their respective interleavers and passed to a decoder for each user, for the constraints employed in their respective channel code encoders. We choose to use a different random interleaver for each user; however, tests have shown little difference in results with the same interleaver for each user. The interleaver is used to break up correlations between decoding and interference cancelation, for similar reasons to turbo decoding [22]. The interleaving function (f_i) allows us to write

$$p(d_{k,t} = d) = f_i(p(d_{k,t} = d|\underline{x})) \tag{6}$$

which output is used to generate $\tilde{d}_{k,t}$.

As shown in (6) the output of the decoders is a sequence of *a posteriori* probabilities of the channel symbols for each user. The channel symbol probabilities are then interleaved according to (6) and converted to a soft estimate using (1) before being passed to the interference canceler for the next iteration.

TABLE I
COMPLEXITY OF ITERATIVE RECEIVER/DECODER

Procedure	*Complexity*
MAP Decoder (MAP)	$I(M_sP_s(10 + 1/(2R))$ $-2M_s + 1/R)$
Interference Canc.(IC)	$I(LN + P)/R$
Channel Calc. (ACHAN)	$IPNL/R$
MF Bank Rx.(MF BANK)	$IPNL/R$

After a sufficient number of iterations have been executed, hard decisions on the information bits can be performed by the single-user decoders to determine the receiver's estimate of the data sent for the entire block for every user, where $\hat{\underline{b}}$ represents this output over all users and for the entire block size.

IV. COMPLEXITY ANALYSIS

In this section, we study the complexity of the iterative receiver/decoder. We use the measure of floating-point operation (FLOP) per information bit to define a real multiply, add, multiply-accumulate, or subtract function. We use the spread interference canceler receiver design as discussed in Section III; however, we alter the size of the system to discuss the complexity of a more realistic receiver.

We determine the number of floating-point operations required per information bit transmitted for the interference canceler (IC), MAP decoders (MAP), MF bank receiver (MF BANK), and spreading operation (ACHAN). Here we assume that the interleaver/deinterleaver complexity is just a lookup table process, and its complexity is insignificant.

We assign the variable $M_s = 2^\nu$ to be the number of states in the channel decoder, where ν is the memory of the convolutional code. We also set the number of users to equal K, the number of paths out of each state in the channel code to equal P_s, and the rate of the FEC code to equal R. The variable P equals the total number of multipaths, L is the number of antenna elements, and I represents the number of iterations of the receiver/decoder.

Table I shows the complexity per user in terms of these variables. Fig. 2 shows the complexity of the six components of the system with respect to the code memory (ν), as the complexity of each component is independent of the number of users. We set the processing gain $N = 25$, the number of IC/decoder iterations to $I = 4$, the paths out of each decoder state to $P_s = 2$, the code rate $R = 0.5$, the number of users $K = 45$, and the number of multipath components to $P = 5$. The total complexity for $L = 2$ antennas is the sum of all these components and is shown as "TOTAL $L = 2$." This system is similar in configuration (but with five times more spreading and five times more users) to that tested in Section VI-B. For comparison, we include the complexity of a conventional receiver (CONV), which we assume consists of a RAKE receiver and a Viterbi decoder.

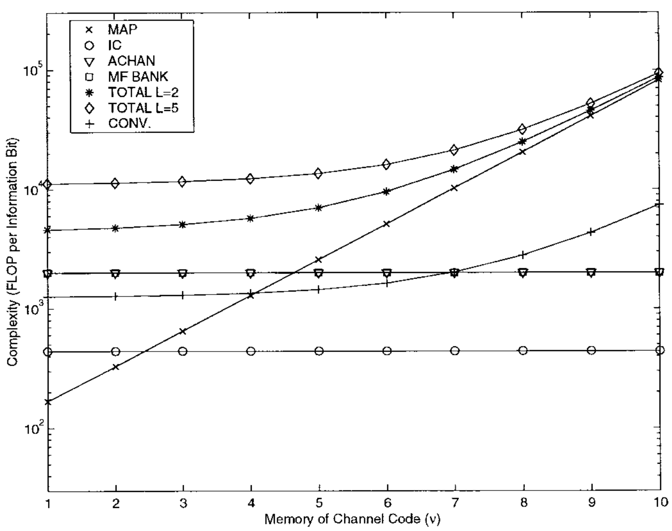

Fig. 2. Complexity in FLOP per information bit versus channel code memory.

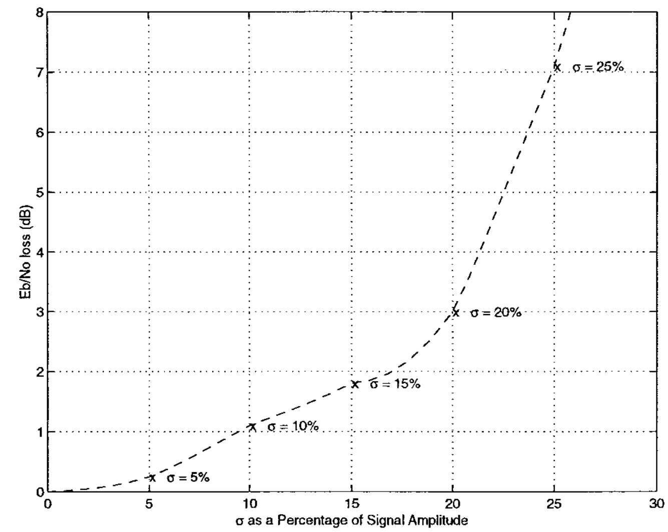

Fig. 3. Performance degradation of iterative receiver with complex amplitude error.

For code memory greater than five, the complexity is dominated by the MAP decoders, and this complexity increases exponentially with the memory of these decoders. This complexity result shows that for realistic code memories of eight as used in IS-95 [31], the complexity of the receiver is dominated by the MAP decoder.

We also selected the case where the number of antenna elements equals $L = 5$, which is shown as "TOTAL $L = 5$." With five antenna elements, the complexity is still dominated by the MAP decoder. Efficient methods of implementing the MAP decoders, not the interference cancelation process, are therefore necessary to make this receiver/decoder more realizable.

For $K = 45$ users, the complexity of the MF bank receivers and one iteration of the MAP decoding operation (approximately equivalent to the conventional design approach) is shown in Fig. 2 as "CONV." The total complexity of our proposed system compared to CONV is approximately 13 times (per information bit). This increase in complexity is very small considering the increase in performance that this system provides over the conventional system.

Last, we note that the estimation and tracking of important channel parameters is not included in this complexity analysis. The complexity analysis also assumes real value operations throughout.

V. Sensitivity Analysis

In this section, we study the sensitivity of the iterative receiver/decoder to various parameter estimation errors. This work enables us to determine under what conditions the receiver operates well and under what conditions it does not. It also shows whether the receiver will be robust enough for "real world" conditions. We model the additive parameter errors as Gaussian distributed and show performance results as a function of the standard deviation of this error distribution.

A. Phase and Amplitude

The first parameters we will test are phase and amplitude errors. The system we test has a processing gain of $N = 5, K = 9$ users, $L = 2$ antenna elements, and rate 1/2 code with generator polynomials $G[5_8, 7_8]$. The coded frame size is $T = 200$ and $I = 4$ iterations of the receiver. We show the results by determining the E_b/N_0 loss compared the case when there are no parameter errors. The receiver estimated amplitude and phase are computed using

$$\alpha' = \alpha + n_a$$

where the true amplitude and phase is α. The amplitude and phase error (n_a) is complex and distributed $\sim\mathcal{N}(0, \sigma_a^2)$ for the real and imaginary axis, where σ_a is a percentage of the amplitude. In Fig. 3, we can see that for a standard deviation of 20%, we achieve a loss of about 3 dB. These results therefore show that the receiver is very insensitive to parameter errors.

B. Direction of Arrival

Another important parameter is the direction of arrival (DOA). Here we have chosen to show results for no spreading $N = 1, L = 5$ antenna elements, $I = 4$ iterations, and $K = 4$ users. We show the performance degradation with respect to the standard deviation of the DOA error in Fig. 4. Here the receiver estimate of the DOA is computed using

$$\theta' = \theta + n_\theta \tag{6}$$

where θ is the true DOA and the error in the DOA (n_θ) is distributed $\sim\mathcal{N}(0, \sigma_\theta^2)$.

The lobe width for $L = 5$ at the 3-dB point is approximately 20° at boresight. For a receiver that can tolerate parameter errors, we would expect degradation to begin when the 3-dB point is reached, as this is when the MF signal energy begins to substantially decrease. For this example, the 3-dB point occurs at 3.3°. This is approximately where we begin to see large performance degradation in Fig. 4. The receiver performance therefore only degrades when the MF output signal energy degrades, not when the receiver's estimate of the DOA is in error. This is therefore a good result for this receiver.

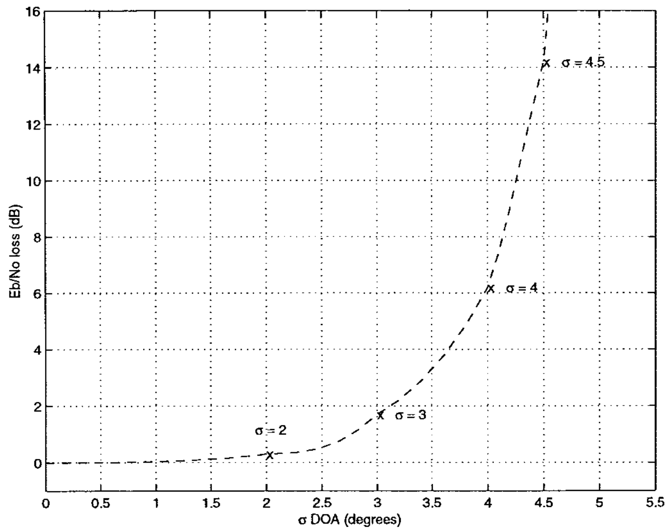

Fig. 4. Performance degradation of iterative receiver with DOA error.

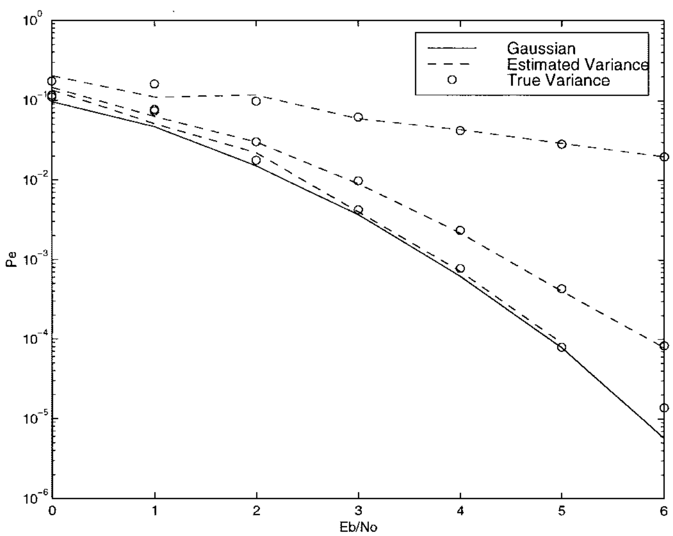

Fig. 5. Performance of iterative receiver with true noise variance calculation.

C. Noise Variance

All the results in this paper assume that we can compute the noise variance output from the interference canceler by using the actual data sent $d_{k,t}$. In a real system, however, we will not have access to the true data value, and in this case we will have to compute the noise variance using estimates of the data, namely, $\tilde{d}_{k,t}$.

The results for $K = 9$ users, a processing gain of $N = 5, T = 200$ coded frame size, and $L = 2$ antenna element system using (5) to compute the noise variance are shown in Fig. 5 (estimated variance). It can be seen that the receiver performance on nearly every iteration is very close to the case when the data sent is known using (4) (true variance). This may, however, not be the case for all system configurations.

VI. Numerical Results

In this section, we show performance of our system under two scenarios. This is to highlight the power of the technique

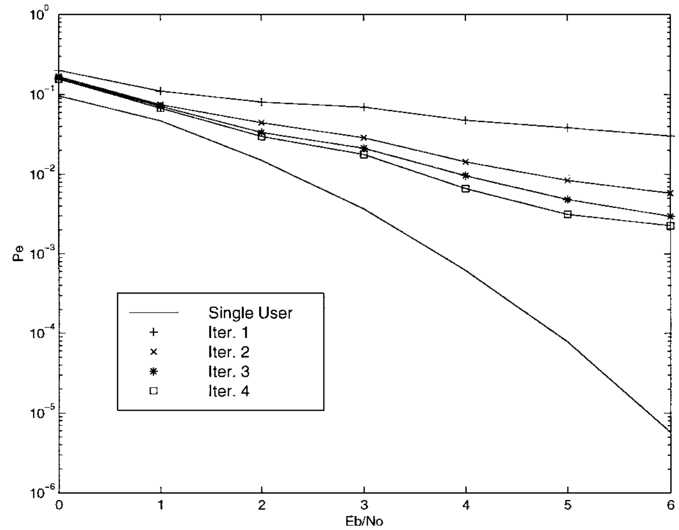

Fig. 6. Performance of iterative receiver with antennas only (minang $= 0°$).

and the adaptability of the SDMA technique to situations with and without spreading and with and without multipath effects. We assume perfect power control, that is, constant and equal power at the receiver for all users at all times.

In Section VI-A, we show results for a narrow-band BPSK system with $L = 5$ antenna elements, $K = 4$ users, $I = 4$ iterations, and no multipath. In Section VI-B, we show performance results for an $L = 2$ antenna system, $I = 4$ iterations, a spreading length of $N = 5, P = 2$ multipath channel taps, and $K = 9$ users.

For both Sections VI-A and VI-B, each user is encoded with a nonrecursive, nonsystematic, rate 1/2, four-state convolutional code with generator polynomial [5_8,7_8]. The decoders assume that the initial state and final state of the code is known; therefore the codes are perfectly terminated. Both systems also utilize random interleaving, which is independent for each user following the encoder functions. As the results show, the performance of these highly spectrally efficient systems, approaches that obtained for channels with no ISI, or only a single user. Note that the antenna gain and multipath gain is normalized so that the system performances between configurations can be compared. This gain provides a 10 $\log_{10} L$ improvement due to the antennas and up to a 10 $\log_{10} P$ improvement due to the multipath diversity gain (depending on the received amplitudes of the different paths).

A. Narrow-Band BPSK with Antennas

In the following simulation result, we show the performance using a universal linear antenna (ULA) as the receive antenna with $L = 5$ elements and $K = 4$ users. The antenna configuration is as described in Section II, with a coded bit frame size of $T = 200$. We set the minimum angle of separation between users to 0° initially, and we set the sector size to $2\pi/3$. The users DOA are selected from a uniform distribution over the sector $((\pi/6) < \theta < (5\pi/6))$ radians every frame T and are assumed to be known exactly. In this result, we assume a fully interleaved, flat Rayleigh fading channel.

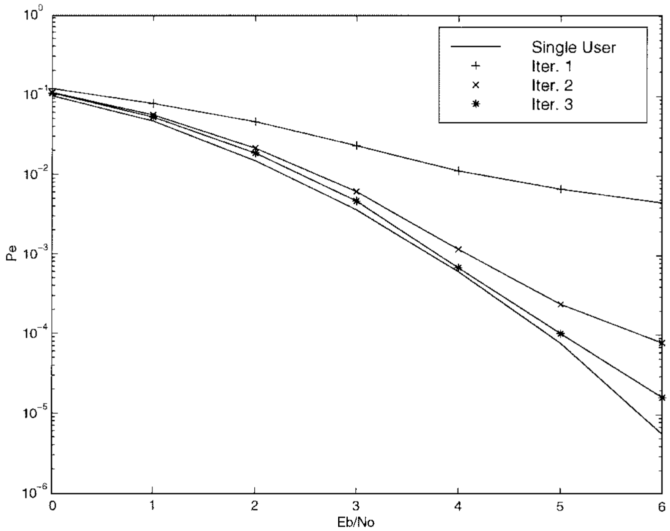

Fig. 7. Performance of iterative receiver with antennas only (minang $= 10^\circ$).

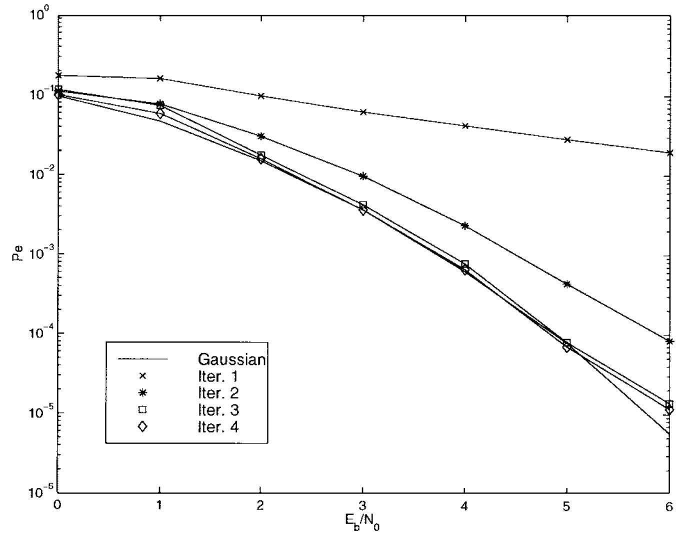

Fig. 8. Multipath combining receiver performance with multipath, DS/CDMA, and antennas.

The results in Fig. 6 show the performance over four iterations with the "single user" performance representing the case when there is only one user, i.e., the no interference case. This poor performance is due to two reasons. First, two users' DOA's can be the same for the whole frame interval, therefore providing no spatial separation. Second, the DOA for each user is fixed for the entire frame size, and the interference is stationary and therefore highly correlated. This is unlike DS/CDMA with random spreading codes where the interference power is time variant, bit to bit.

The reason for the poor performance is due to the lack of separation that occurs in some frame intervals. To achieve better separation, we include a constraint that the DOA's of each user must be greater than 10° apart (half the 3-dB beamwidth). This could be easily achieved by simple system management software in the base station where an interfering user is handed off to another time or frequency slot [26]. This constraint ensures that the users maintain partial spatial separation.

With this DOA constraint, our new result is shown in Fig. 7 with performance very close to single user (no interference case) after three iterations. We make no attempt to determine the optimum minimum angle of separation that still yields "good performance." We instead highlight the fact that as the minimum angle of separation increases the interference decreases. As the minimum angle increases, so too does the amount of system management required to hand off interfering users to new time or frequency slots. A tradeoff therefore exists between system management complexity and receiver performance. This result in Fig. 7 shows that with some realistic constraints on the antenna beamwidth and the separation of each user we can achieve a reuse within cell [26] of four with $L = 5$ antenna elements. Note that the first iteration is equivalent to a space-time MF with a soft input Viterbi decoder.

Although for this system configuration we assume no multipath, in systems such as GSM there can be delay spreads of up to five channel bits [32]. These could be easily included in the model as ISI in a similar way to the DS/CDMA channel. Another important point is that a small number of antenna elements are used yielding a beam width at the 3-dB point (from boresight) of 20°. Although this may be thought as a disadvantage due to collection of energy from other users that have up to 10° spatial separation, it can also be an advantage. Channel measurement results show angle spreads of up to $\theta_{\rm BW} = 30^\circ$ [26] for hilly macro cell environments. Our small ULA would therefore collect most of the energy from all these multipaths, while ULA's with a large number of elements would need separate beams to pick up each multipath component.

In the configuration tested, we make no attempt to determine the channel parameters, which include DOA, delay, and Doppler spreads. It has been suggested [26] that it may be difficult to estimate these parameters for narrow-band systems. The reasons given are lack of tight power control, the amount of Doppler, and multipath effects.

B. Multipath, DS/CDMA and Antennas

In this section, we show the performance of the receiver with a combination of multipath, DS/CDMA, and SDMA. We allow two multipath taps $(P = 2)$ uniformly distributed over three DS/CDMA chip intervals, we utilize an antenna array with $L = 2$ elements, and we use a processing gain of $N = 5$ for a $K = 9$ user system and a frame size of $T = 200$. We maintain sectorization to a sector size of $2\pi/3$ radians and remove the minimum angle between users restriction. Due to the spreading code separation, it is now possible to reduce interference between users with similar DOA's.

The result we show is using the receiver as discussed in Section III. The performance of the system is illustrated in Fig. 8 and shows excellent results very close to the single user (no ISI) performance case. This shows that we have successfully mitigated the multiple access interference and ISI due to the multipath channel.

In Fig. 8, there is a slight amount of flooring that occurs at $E_b/N_0 = 6$ dB. We believe this is because of the additional correlations due to the chip synchronous system we have modeled. The interference power reduces by one-third (for square pulses) if asynchronous chips are used [33]. Also, other results [20] indicate that with asynchronous chips, no flooring is evident at a probability of information bit error of 10^{-5}.

For both results, we make no attempt to estimate channel parameters. Due to power control and the processing gain of the DS/CDMA channel estimation, parameter estimation errors may be less critical to this system configuration [26] and potentially make this a much more realizable design than the narrow-band solution previously discussed.

VII. Summary

In this paper, we have incorporated the modeling of a receive antenna array and multipath into our asynchronous bit DS/CDMA channel model. This allows us to test our receiver on a channel that is similar to that proposed for IMT-2000 [34]. This design also discusses reuse within cell (SDMA) improvements that could be applied to GSM [35]. We describe the iterative receiver and show how we perform space- and time-matched filtering. We discuss spread domain interference cancellation and show how this can be used to minimize receiver complexity.

The computational load of our new receiver is important, and we show that the complexity per user is independent of the number of users. For realistic systems the dominant complexity component is the MAP channel decoders. This is the desired result, as we have moved the dominating complexity component from the interference mitigation task to the decoding task.

The receiver we have described relies on channel parameter information. We studied the sensitivity of the receiver to phase, amplitude, and DOA errors and found that it is very tolerant of such errors. We performed these tests by providing the receiver with a perturbed value, which was Gaussian distributed. We found that for a spatially separated BPSK system, DOA errors up to the standard deviation of the beamwidth could be tolerated with less than 3-dB loss. We also found for a DS-CDMA system that standard deviations in the amplitude and phase of the receiver could be as high as 20% of the amplitude before a 3-dB loss in performance was encountered.

In this paper, we have illustrated the performance, complexity, and sensitivity of our receiver design. The applications of this technique to GSM [35] to increase capacity is demonstrated in Section VI-A. The potential for capacity improvements in IS-95A or third-generation systems is demonstrated in Section VI-B. An important point to note is that the overall gain shown is achieved by the mitigation of the ISI. The additional antenna diversity gain of 10 $\log_{10} L$ plus a gain of up to 10 $\log_{10} P$ is also achieved, where L is the number of antenna elements and P is the number of multipath components.

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