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Blind Adaptive Decorrelating RAKE (DRAKE) Downlink Receiver for Space-Time Block Coded Multipath CDMA

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A downlink receiver is proposed for space-time block coded CDMA systems operating in multipath channels. By combining the powerful RAKE receiver concept for a frequency selective channel with space-time decoding, it is shown that the performance of mobile receivers operating in the presence of channel fading can be improved significantly. The proposed receiver consists of a bank of decorrelating filters designed to suppress the multiple access interference embedded in the received signal before the space-time decoding. The new receiver performs the space-time decoding along each resolvable multipath component and then the outputs are diversity combined to obtain the final decision statistic. The proposed receiver relies on a key constraint imposed on the output of each filter in the bank of decorrelating filters in order to maintain the space-time block code structure embedded in the signal. The proposed receiver can easily be adapted blindly, requiring only the desired user's signature sequence, which is also attractive in the context of wireless mobile communications. Simulation results are provided to confirm the effectiveness of the proposed receiver in multipath CDMA systems.

Keywords and phrases: multiuser detection, space-time coding, blind adaptive receivers, code division multiple access.

1. INTRODUCTION

Space-time coding [1, 2, 3], multiuser detection [4], and RAKE combining [5, 6, 7, 8] are powerful techniques that can offer significant performance gains in mobile communications environments and many emerging wireless systems adopt some combination of these techniques. For example, the application of RAKE receivers to multiuser systems has been studied extensively, prompted by the fact that 3G technology is based on the wideband code division multiple access (CDMA) concept and thus it must contend with frequency-selective channels.

Of course, RAKE combining allows the receiver to extract the useful signal energy embedded in the signals that arrive via different paths in such channels. Thus, the RAKE receiver provides one way of providing frequency diversity at the receiver [5] in CDMA systems. On the other hand, recent advances in space-time coding exploit the spatial diversity inherent in fading channels [1, 2, 3]. Most space-time coding schemes are based on the assumption of a frequency flat fading channel. Under these conditions, the introduction of space-time coding into multiuser systems seems to provide significant performance gains just as they do in single-user channels [9, 10, 11, 12, 13, 14].

However, as mentioned above, most next generation wireless systems will be wideband communications systems, and thus they must contend with frequency selective channel fading. Transmit diversity schemes for wideband CDMA systems have recently been proposed in [15, 16, 17, 18], for instance. In future wireless systems, the downlink is expected to present a performance bottleneck due to the asymmetric nature of the system traffic, and thus it is specifically important to consider transmit diversity schemes at the base station [12, 19]. It is also conceivably easier to implement multiple antennas at a base station than at a mobile station due to space, power, and other considerations.

These observations motivate us to consider space-time coding for the downlink of multiple access systems operating in multipath environments. A natural way to exploit the space-time coding advantage in multipath environments is to employ an RAKE combining technique on top of the space-time decoding. In [16], for example, a receiver based on RAKE combining was proposed for a space-time spreading (STS) transmit diversity system in order to improve the downlink performance in the presence of multipath. In this paper, we propose a blind adaptive downlink scheme for a space-time block coded multipath CDMA system. Similar blind adaptive downlink receivers for flat fading channels have also been considered in [12]. Space-time multiuser detection for space-time block coded CDMA has been considered in [20] with a blind adaptive implementation. The blind adaptive space-time multiuser detector proposed in [20] assumes multiple receiver antennas whereas we limit our attention to only single receiver antenna systems. Analytical expressions for BER performance of linear multiuser detection based receivers in space-time block coded CDMA can also be found in [20].

In this paper, we assume that the base station employs space-time block coding. The proposed receiver operates by first applying a decorrelating RAKE (DRAKE) filter [6] to the received signal and then performing space-time decoding along each path component separately, thus exploiting space-time coding gains along each path. The DRAKE filter zero forces the interpath interference and mitigates the multiple access interference (MAI) according to a minimum mean square error (MMSE)-like criterion. Multipath combining is then performed after space-time decoding. The proposed scheme can easily be adapted blindly, meaning that the only information required for its successful operation at the receiver is the desired user's signature sequence. This makes the new receiver an attractive candidate for downlink mobile receivers in future wideband wireless communication systems.

This presentation is organized as follows. In Section 2, we present our signal model and the system description. Next, in Sections 3 and 4, we derive the batch-mode receiver for a frequency selective CDMA channel and the blind adaptive receiver implementation, respectively. Finally, in Section 5, we give performance results of the proposed receiver architecture for a representative space-time block code and some typical system environments.

2. SIGNAL MODEL

Consider a system with K simultaneous users. We assume that the base station consists of N_T transmitter antennas and the mobile is equipped with N_R receiver antennas. For the sake of simplicity, we will assume that $N_T = 2$ and $N_R = 1$ although generalization to larger numbers of transmit and receive antennas is straightforward. The BPSK information symbol sequence $\{d_k(i)\}_{i=0}^{\infty}$ of user k, for $k = 1, \ldots, K$, is first encoded by a space-time block encoder. For this $N_T = 2$ case, the space-time block code that achieves full transmit diversity is given by the well-known Alamouti scheme [1]. If two consecutive input symbols of user k to the space-time encoder are denoted by $d_k(i)$ and $d_k(i+1)$, then according to the Alamouti scheme, during the first symbol period the symbols $d_k(i)$ and $d_k(i+1)$ are transmitted simultaneously from the first and second antennas, respectively. During the second symbol period, the symbols $-d_k^*(i+1)$ and $d_k^*(i)$ are transmitted from the first and second antennas, respectively, where $*$ denotes complex conjugation. If we denote the space-time block encoder function by $F : \mathscr{C}^{1\times N_T} \to \mathscr{C}^{N_T\times N_T}$, we may write this coding scheme in matrix form as below, where the columns and rows correspond to space (transmit antennas) and time dimensions, respectively,

$$F\big(\big[d_k(i)\, d_k(i+1)\big]\big) = \begin{bmatrix} d_k(i) & d_k(i+1) \\ -d_k^*(i+1) & d_k^*(i) \end{bmatrix}. \quad (1)$$

The space-time encoder output of each user is next modulated by the user's spreading sequence and transmitted from the two transmit antennas at the base station. Note that the two symbols transmitted from the two antennas, corresponding to a particular user, are modulated by the same spreading sequence, that is, each user is assigned only one spreading code. The spreading waveform $s_k(t)$ of user k is assumed to be of the form

$$s_k(t) = \sum_{j=0}^{N-1} c_k(j)\varphi(t - jT_c), \quad (2)$$

where N denotes the processing gain of the CDMA system and $\mathbf{c}_k = [c_k(0), \ldots, c_k(N-1)]^T$ denotes the kth user's spreading code sequence. In (2), T_c denotes the chip period, $c_k(j) \in \{+1/\sqrt{N}, -1/\sqrt{N}\}$, and $\varphi(t)$ is the normalized chip waveform.

We may assume that we are interested in detecting the signals for user k. The continuous time received signal at the mobile k can be written as

$$r_k(t) = \sum_{k'=1}^{K} \sum_{n_T=1}^{N_T} x_{k',n_T}(t) \star f_{k,n_T}(t) + n(t), \quad (3)$$

where $\star$ denotes the convolution operation and $n(t)$ is a complex white Gaussian noise process with zero mean and variance 1/2 per dimension. The function $f_{k,n_T}(t)$ in (3) denotes the composite channel response corresponding to transmitter antenna n_T at the base station and receiver k, which can be written as

$$f_{k,n_T}(t) = \sum_{l=1}^{L} h_{k,n_T,l}\,\delta\big(t - (l-1)T_c\big). \quad (4)$$

In the above, L is the number of paths present in the channel between transmitter antenna n_T of the base station and the receiver antenna of user k, $h_{k,n_T,l}$ is the fading coefficient of the path l, for $l = 1, \ldots, L$, and $\delta(t)$ is the Dirac delta function. Also, $x_{k',n_T}(t)$ in (3) is given as

$$x_{k',n_T}(t) = \sum_{i=0}^{B-1} b_{k',n_T}(i)s_{k'}(t - iT), \quad (5)$$

where B is the received frame size and $b_{k',n_T}(i)$ denotes the transmitted symbol of user k' on antenna n_T, for $n_T = 1, 2$, during the transmission interval $[iT, (i+1)T]$.

Note that, in order to minimize the notational complexity in (4), we have assumed that the number of multipaths in the channels between the receiver k and different transmit antennas at the base station is constant and equal to L, and that multipath delays are always multiples of the chip interval. Again, it is straightforward to modify the above model in order to let the number of multipaths depend on the transmit antenna index. Also, in a practical implementation, it is not difficult to include different values for the multipath delays at the expense of some extra notational complexity. In fact, we may get rid of the constraint of multipath delays being an integral multiple of the chip interval and include the effect of fractional correlation of the chip waveforms in the fading coefficient $h_{k,n_T,l}$. Thus, our model is general enough to absorb these generalizations and yet is simple enough to avoid unnecessary notational complexity in the current discussion.

Next, we make the following assumption for the sake of simplicity. We assume that the maximum delay spread is $(L-1)T_c$ and the symbol period is defined as $T=(N+L-1)T_c$. In that case, if we assume synchronous transmission at the base station, then all the delayed replicas of the transmitted signals will be received at the receiver of user k during the same symbol period, resulting in no intersymbol interference. This is equivalent to the assumption that the actual spreading waveform is obtained by appending $L-1$ zeros to the tail of the spreading code $\mathbf{c}_k$. This is a reasonable approximation for channels with delay spreads on the order of a few chip intervals. Let N' be the modified signature sequence length, that is,

$$N' = N + L - 1. \tag{6}$$

The signal $r_k(t)$ is first chip matched filtered and then sampled at the receiver for user k. The output of the chip matched filter corresponding to the ith symbol period is given as, for $j = 0, 1, \ldots, N'-1$,

$$r_k(i,j) = \int_0^\infty r_k(t)\varphi(t - iT - jTc)\,dt. \tag{7}$$

The resulting chip-rate-sampled output can be collected to form a vector of length N' as

$$\begin{aligned}\mathbf{r}_k(i) &= \mathbf{C}_k\left[\mathbf{h}_{k,1}b_{k,1}(i) + \mathbf{h}_{k,2}b_{k,2}(i)\right] \\ &\quad + \sum_{k'\neq k}^{K} \mathbf{C}_{k'}\left[\mathbf{h}_{k,1}b_{k',1}(i) + \mathbf{h}_{k,2}b_{k',2}(i)\right] + \mathbf{n}(i),\end{aligned} \tag{8}$$

where $\mathbf{n}(i)$ is a complex Gaussian noise vector such that $\mathbf{n}(i) \sim \mathcal{N}(\mathbf{0}, \mathbf{I}_{N'})$, the $N' \times L$ matrix $\mathbf{C}_{k'}$, for $k' = 1, 2, \ldots, K$, is defined as

$$\mathbf{C}_{k'} = \begin{bmatrix} c_{k'}(0) & 0 & \cdots & 0 \\ c_{k'}(1) & c_{k'}(0) & \cdots & 0 \\ \vdots & \vdots & \cdots & \vdots \\ c_{k'}(N-1) & c_{k'}(N-2) & \cdots & c_{k'}(N-L) \\ 0 & c_{k'}(N-1) & \cdots & c_{k'}(N-L+1) \\ \vdots & \vdots & \cdots & \vdots \\ 0 & 0 & \cdots & c_{k'}(N-1) \end{bmatrix} \tag{9}$$

and the L-vector of fading coefficients of the different multipaths between transmit antenna n_T, for $n_T = 1, 2$, and mobile k is defined as $\mathbf{h}_{k,n_T} = [h_{k,n_T,1}, h_{k,n_T,2}, \ldots, h_{k,n_T,L}]^T$, where $h_{k,n_T,l}$ is the path gain of the lth path between transmit antenna n_T and the receiver for user k. Defining the $L \times N_T$ matrix $\mathbf{H}_k$ of fading coefficients as

$$\mathbf{H}_k = \begin{bmatrix}\mathbf{h}_{k,1} & \mathbf{h}_{k,2}\end{bmatrix} = \begin{bmatrix} h_{k,1,1} & h_{k,2,1} \\ h_{k,1,2} & h_{k,2,2} \\ \vdots & \vdots \\ h_{k,1,L} & h_{k,2,L} \end{bmatrix} \tag{10}$$

and the N_T length transmit symbol vector of user k' at time instant i as $\mathbf{b}_{k'}(i) = [b_{k',1}(i), b_{k',2}(i)]^T$, we may write the chip-rate sampled output (8) corresponding to the symbol time i as

$$\mathbf{r}_k(i) = \mathbf{C}_k\mathbf{H}_k\mathbf{b}_k(i) + \sum_{k'\neq k}^{K} \mathbf{C}_{k'}\mathbf{H}_k\mathbf{b}_{k'}(i) + \mathbf{n}(i). \tag{11}$$

Suppose that the receiver processes received chip matched signals in blocks of size N_T, corresponding to the space-time code word length. Then, the two consecutive received signals during the mth received space-time code block correspond to the two consecutive information symbols $d_k(mN_T)$ and $d_k(mN_T+1)$ of the desired user k, for $m = 0, 1, 2, \ldots$. Thus, without loss of generality, from now on, we will assume that the symbol index i is of the form $i = mN_T$ for some $m = 0, 1, 2, \ldots$ and simply refer to the two received signals during the mth block as $\mathbf{r}_k(i)$ and $\mathbf{r}_k(i+1)$ (i.e., by assuming that $i = mN_T$, for $m = 0, 1, 2, \ldots$, we may refer to the mth space-time code block simply as ith block without causing any confusion). In this case, the two transmit symbol vectors $\mathbf{b}_{k'}(i)$ and $\mathbf{b}_{k'}(i+1)$, corresponding to the symbol times i and $i+1$, are given by $[d_{k'}(i), d_{k'}(i+1)]^T$ and $[-d_{k'}^*(i+1), d_{k'}^*(i)]^T$, respectively. Suppose that the receiver generates a set of N_T decision statistics, $\mathbf{y}_k^1(i), \mathbf{y}_k^2(i), \ldots, \mathbf{y}_k^{N_T}(i)$, corresponding to the N_T received symbols, $\mathbf{r}_k(i), \mathbf{r}_k(i+1), \ldots, \mathbf{r}_k(i+N_T-1)$, during the ith space-time code block. In our case of $N_T = 2$, these are defined as

$$\mathbf{y}_k^1(i) = \mathbf{r}_k(i), \qquad \mathbf{y}_k^2(i) = \mathbf{r}_k^*(i+1). \tag{12}$$

Then it is easily seen that we can write $\mathbf{y}_k^1(i)$ and $\mathbf{y}_k^2(i)$ as

$$\mathbf{y}_k^1(i) = \mathbf{C}_k\mathbf{H}_k\mathbf{b}_k(i) + \sum_{k'\neq k}^{K} \mathbf{C}_{k'}\mathbf{H}_k\mathbf{b}_{k'}(i) + \mathbf{n}_1, \tag{13}$$

$$\mathbf{y}_k^2(i) = \mathbf{C}_k\tilde{\mathbf{H}}_k\mathbf{b}_k(i) + \sum_{k'\neq k}^{K} \mathbf{C}_{k'}\tilde{\mathbf{H}}_k\mathbf{b}_{k'}(i) + \mathbf{n}_2, \tag{14}$$

where

$$\tilde{\mathbf{H}}_k = \begin{bmatrix}\mathbf{h}_{k,2}^* & -\mathbf{h}_{k,1}^*\end{bmatrix} = \begin{bmatrix} h_{k,2,1}^* & -h_{k,1,1}^* \\ h_{k,2,2}^* & -h_{k,1,2}^* \\ \vdots & \vdots \\ h_{k,2,L}^* & -h_{k,1,L}^* \end{bmatrix} \tag{15}$$

and the two noise vectors $\mathbf{n}_1 = \mathbf{n}(i)$ and $\mathbf{n}_2 = \mathbf{n}^*(i+1)$ are independent $\mathcal{N}(\mathbf{0}, \mathbf{I}_{N'})$ vectors. Note also that, we always have the property that the lth row of $\mathbf{H}_k$ is orthogonal to the lth row of $\tilde{\mathbf{H}}_k$, that is, $\boldsymbol{\alpha}_{k,l}^H \tilde{\boldsymbol{\alpha}}_{k,l} = 0$ for $l = 1, \dots, L$ and $k = 1, \dots, K$, where $\boldsymbol{\alpha}_{k,l}^H$ and $\tilde{\boldsymbol{\alpha}}_{k,l}^H$ denote the lth row of $\mathbf{H}_k$ and $\tilde{\mathbf{H}}_k$, respectively.

3. DECORRELATING RAKE-BASED RECEIVER FOR SPACE-TIME CODED CDMA

In this section, we develop the structure of the proposed two-stage receiver for multipath CDMA systems. The proposed system is an extension of the DRAKE receiver to space-time coded frequency selective channels [6].

In the presence of multipath, in order to take advantage of the diversity offered by the multipath channel, the receiver needs to be able to coherently combine the useful signal energy present in each of the paths in an optimal way. The RAKE receiver operates by resolving the paths at the receiver and combining them in order to improve the signal-to-interference-plus-noise ratio (SINR). We may employ the same strategy of resolving the paths at the space-time receiver. However, it is important to ensure that the application of the RAKE receiver to the received signal does not destroy the structure of the space-time code embedded in the signal. The first stage of the proposed receiver consists of a bank of L filters with weight vectors $\mathbf{w}_{k,l}$, for $l = 1, \dots, L$, so that the output of the lth branch filter is given by

$$z_{k,l}(i) = \mathbf{w}_{k,l}^H \mathbf{y}_k(i). \tag{16}$$

In order to preserve the space-time code structure at the output of each branch, we impose the constraint that

$$\mathbf{C}_k^H \mathbf{w}_{k,l} = \mathbf{e}_l, \tag{17}$$

where $\mathbf{e}_l$ is a vector of length L having all zeros except a single one at the lth position. Then, combining (13), (16), and (17), we can write the outputs of the lth branch filter

$$\begin{aligned}
z_{k,l}^1(i) &= \mathbf{w}_{k,l}^H \mathbf{y}_k^1(i) \\
&= \mathbf{e}_l^H \mathbf{H}_k \mathbf{b}_k(i) + \sum_{k' \neq k}^{K} \mathbf{w}_{k,l}^H \mathbf{C}_{k'} \mathbf{H}_k \mathbf{b}_{k'}(i) + \mathbf{w}_{k,l}^H \mathbf{n}_1 \\
&= \boldsymbol{\alpha}_{k,l}^H \mathbf{b}_k(i) + \boldsymbol{\eta}_{k,l}^1(i), \\
z_{k,l}^2(i) &= \mathbf{w}_{k,l}^H \mathbf{y}_k^2(i) \\
&= \mathbf{e}_l^H \tilde{\mathbf{H}}_k \mathbf{b}_k(i) + \sum_{k' \neq k}^{K} \mathbf{w}_{k,l}^H \mathbf{C}_{k'} \tilde{\mathbf{H}}_k \mathbf{b}_{k'}(i) + \mathbf{w}_{k,l}^H \mathbf{n}_2 \\
&= \tilde{\boldsymbol{\alpha}}_{k,l}^H \mathbf{b}_k(i) + \boldsymbol{\eta}_{k,l}^2(i),
\end{aligned} \tag{18}$$

where, as before, $\boldsymbol{\alpha}_{k,l}^H$ denotes the lth row of $\mathbf{H}_k$, and $\boldsymbol{\eta}_{k,l}^1(i)$ and $\boldsymbol{\eta}_{k,l}^2(i)$ denote the residual MAI plus noise terms at the output of the lth branch filter. Hence, the output SINR of the lth branch is then given by

$$\text{SINR}_l = \frac{E\{|\boldsymbol{\alpha}_{k,l}^H \mathbf{b}_k(i)|^2\}}{E\{(\boldsymbol{\eta}_{k,l}^1(i))^H \boldsymbol{\eta}_{k,l}^1(i)\}} = \frac{|\boldsymbol{\alpha}_{k,l}|^2}{E\{(\boldsymbol{\eta}_{k,l}^1(i))^H \boldsymbol{\eta}_{k,l}^1(i)\}}. \tag{19}$$

Thus, maximizing the SINR at the output of the branch filter l is the same as minimizing the output variance $E\{(\boldsymbol{\eta}_{k,l}^1(i))^H \boldsymbol{\eta}_{k,l}^1(i)\}$. However, since the output signal power is fixed at $|\boldsymbol{\alpha}_{k,l}|^2$ due to the constraint (17), this in turn is achieved by minimizing the output energy $E\{|z_{k,l}(i)|^2\}$ given by

$$E\{|z_{k,l}(i)|^2\} = \mathbf{w}_{k,l}^H \mathbf{R}_{yy} \mathbf{w}_{k,l}, \tag{20}$$

where the autocorrelation matrix $\mathbf{R}_{yy}$ is given by

$$\mathbf{R}_{yy} = E\{\mathbf{y}_k(i)\mathbf{y}_k^H(i)\}. \tag{21}$$

In effect, this means that we are employing a minimum mean output energy (MMOE) multiuser detector to suppress the MAI while the constraint (17) nulls the interpath interference caused by the frequency selective fading process. Note that, in order to keep the notation simple in (21), we have dropped the superscripts from $\mathbf{y}_k^{n_T}(i)$ for $n_T = 1, 2, \dots, N_T$. Thus, maximizing the output SINR at each branch reduces to the minimization of the cost function $\mathbf{w}_{k,l}^H \mathbf{R}_{yy} \mathbf{w}_{k,l}$ and the filter design problem for the lth branch reduces to the following constrained optimization problem. For $l = 1, \dots, L$,

$$\mathbf{w}_{k,l} = \arg\min_{\mathbf{w}} \mathbf{w}_{k,l}^H \mathbf{R}_{yy} \mathbf{w}_{k,l} \quad \text{subject to } \mathbf{C}_k^H \mathbf{w}_{k,l} = \mathbf{e}_l. \tag{22}$$

It is a straightforward exercise to show that the solution of this problem is

$$\mathbf{w}_{k,l} = \mathbf{R}_{yy}^{-1} \mathbf{C}_k \left(\mathbf{C}_k^H \mathbf{R}_{yy}^{-1} \mathbf{C}_k\right)^{-1} \mathbf{e}_l \tag{23}$$

and the output of the filter bank, for $n_T = 1, 2$, is simply given as

$$\mathbf{z}_k^{n_T}(i) = \mathbf{W}_k^H \mathbf{y}_k^{n_T}(i) = \left(\mathbf{C}_k^H \mathbf{R}_{yy}^{-1} \mathbf{C}_k\right)^{-1} \mathbf{C}_k^H \mathbf{R}_{yy}^{-1} \mathbf{y}_k^{n_T}(i), \tag{24}$$

where $\mathbf{w}_{k,l}$ is the lth column of the $N' \times L$ matrix $\mathbf{W}_k$, for $l = 1, \dots, L$.

At the next stage of the receiver, the space-time decoding is performed along each branch in two symbol blocks. Observing from (18) that the space-time code property is preserved at the output of each branch, we may form the vector

$$\begin{aligned}
\mathbf{z}_{k,l}(i) &= \begin{bmatrix} z_{k,l}^1(i) \\ z_{k,l}^2(i) \end{bmatrix} \\
&= \begin{bmatrix} \boldsymbol{\alpha}_{k,l}^H \\ \tilde{\boldsymbol{\alpha}}_{k,l}^H \end{bmatrix} \mathbf{b}_k(i) + \begin{bmatrix} \eta_{k,l}^1(i) \\ \eta_{k,l}^2(i) \end{bmatrix} \\
&= \mathbf{H}_{k,l} \mathbf{b}_k(i) + \boldsymbol{\eta}_{k,l}(i),
\end{aligned} \tag{25}$$

where we have defined the 2×2 matrix ($N_T \times N_T$, in general) $\mathbf{H}_{k,l} = [\boldsymbol{\alpha}_{k,l}^H; \tilde{\boldsymbol{\alpha}}_{k,l}^H]$ and the 2-vector (N_T-vector, in general) of noise term $\boldsymbol{\eta}_{k,l}(i) = [\eta_{k,l}^1(i) \;\; \eta_{k,l}^2(i)]^T$. Note that, the matrix $\mathbf{H}_{k,l}$ satisfies the property

$$\mathbf{H}_{k,l}^H \mathbf{H}_{k,l} = |\boldsymbol{\alpha}_{k,l}|^2 \mathbf{I}_2 = \left(|h_{k,1,l}|^2 + |h_{k,2,l}|^2\right) \mathbf{I}_2. \tag{26}$$

Thus, we still have the advantage of simple linear decoding of the space-time block code along each branch filter as

$$\tilde{\mathbf{z}}_{k,l}(i) = \mathbf{H}_{k,l}^H \mathbf{z}_{k,l}(i) = \left(|h_{k,1,l}|^2 + |h_{k,2,l}|^2\right)\mathbf{b}_k(i) + \tilde{\boldsymbol{\eta}}_{k,l}(i), \quad (27)$$

where we have defined

$$\tilde{\boldsymbol{\eta}}_{k,l}(i) = \mathbf{H}_{k,l}^H \boldsymbol{\eta}_{k,l}(i). \quad (28)$$

Note that $\eta_{k,l}^1(i)$ and $\eta_{k,l}^2(i)$ are no longer independent due to the MAI component included in them. This also results in correlated components in the noise vector $\tilde{\boldsymbol{\eta}}_{k,l}(i)$. Thus, it is no longer optimal to decode the components of $\tilde{\mathbf{z}}_{k,l}(i)$ separately. However, in order to keep the receiver complexity to a minimum, we will ignore these correlations in the noise components and decode $\tilde{\mathbf{z}}_{k,l}(i)$ componentwise, as in a single-user channel; that is, we will assume that the space-time decoder output decouples the effect of $d_k(i)$ and $d_k(i+1)$.

Finally, the receiver combines the space-time decoder outputs from all the branches in order to form the final decision variable. Due to the above assumption of decoupled space-time decoder output, the decision variable for $d_k(i)$ has to take into account only the L-vector of branch outputs defined as

$$\begin{aligned}
\tilde{\mathbf{z}}_k^1(i) &= \begin{bmatrix} \tilde{z}_{k,1,1}(i) \\ \tilde{z}_{k,2,1}(i) \\ \vdots \\ \tilde{z}_{k,L,1}(i) \end{bmatrix} \\
&= \begin{bmatrix} |\boldsymbol{\alpha}_{k,1}|^2 \\ |\boldsymbol{\alpha}_{k,2}|^2 \\ \vdots \\ |\boldsymbol{\alpha}_{k,L}|^2 \end{bmatrix} d_k(i) + \begin{bmatrix} \tilde{\eta}_{k,1,1}(i) \\ \tilde{\eta}_{k,2,1}(i) \\ \vdots \\ \tilde{\eta}_{k,L,1}(i) \end{bmatrix} \\
&= \mathbf{g}_k d_k(i) + \mathbf{u}_{k,1}(i),
\end{aligned} \quad (29)$$

where $\tilde{z}_{k,l,m}(i)$ and $\tilde{\eta}_{k,l,m}(i)$ denote the mth component of the vectors $\tilde{\mathbf{z}}_{k,l}(i)$ and $\tilde{\boldsymbol{\eta}}_{k,l}(i)$, respectively. Similarly, the decision variable for $d_k(i+1)$ needs only to be based on the vector of outputs

$$\begin{aligned}
\tilde{\mathbf{z}}_k^2(i) &= \begin{bmatrix} \tilde{z}_{k,1,2}(i) \\ \tilde{z}_{k,2,2}(i) \\ \vdots \\ \tilde{z}_{k,L,2}(i) \end{bmatrix} \\
&= \begin{bmatrix} |\boldsymbol{\alpha}_{k,1}|^2 \\ |\boldsymbol{\alpha}_{k,2}|^2 \\ \vdots \\ |\boldsymbol{\alpha}_{k,L}|^2 \end{bmatrix} d_k(i+1) + \begin{bmatrix} \tilde{\eta}_{k,1,2}(i) \\ \tilde{\eta}_{k,2,2}(i) \\ \vdots \\ \tilde{\eta}_{k,L,2}(i) \end{bmatrix} \\
&= \mathbf{g}_k d_k(i+1) + \mathbf{u}_{k,2}(i),
\end{aligned} \quad (30)$$

where we have introduced the notation $\mathbf{g}_k = [|\boldsymbol{\alpha}_{k,1}|^2, \dots, |\boldsymbol{\alpha}_{k,L}|^2]^T$ and $\mathbf{u}_{k,n_T}(i) = [\tilde{\eta}_{k,1,n_T}(i), \dots, \tilde{\eta}_{k,L,n_T}(i)]^T$, for $n_T = 1, 2$. It should be noted that the noise vectors $\mathbf{u}_{k,n_T}$, for $n_T = 1, 2$, are vectors of correlated noise components.

In the final stage of the receiver, the space-time decoded outputs are multipath combined to form the final decision variable. The optimal combining scheme is given as $\hat{z}_k = \mathbf{R}_{\tilde{\mathbf{z}}_k,\tilde{\mathbf{z}}_k}^{-1} \tilde{\mathbf{z}}_k$ [6]. However, we may instead employ maximal ratio combining (MRC) given as

$$\hat{z}_k(i) = \mathbf{g}_k^H \tilde{\mathbf{z}}_k^1(i), \qquad \hat{z}_k(i+1) = \mathbf{g}_k^H \tilde{\mathbf{z}}_k^2(i). \quad (31)$$

The multipath combined outputs $\hat{z}_k(i)$ and $\hat{z}_k(i+1)$, for $i = 0, 1, \dots$, are used either to make a final decision on the information symbols by passing through a quantizer, or to feed into a channel decoder as the soft inputs if the transmitter also performs any additional channel coding. We will not consider any channel coding in our system, however, and will thus base our performance analysis on raw bit error rates (BER).

4. BLIND ADAPTIVE IMPLEMENTATION

The first stage of the above proposed scheme requires knowledge of the signature sequences of all the interfering users. However, similarly to the scheme proposed in [6] for the DRAKE receiver, we may also easily adapt this receiver blindly, requiring only knowledge of the signature sequence of the desired user, namely $\mathbf{c}_k$. In this section, we outline a blind adaptive process based on the least mean squares (LMS) [21, 22] algorithm.

The LMS algorithm adapts the filter coefficients for each branch of the first stage of DRAKE receiver as

$$\mathbf{w}_{k,l}(i+1) = \mathbf{w}_{k,l}(i) - \mu \nabla_{\mathbf{w}_{k,l}}(J(\mathbf{w}_{k,l})), \quad (32)$$

where $J(\mathbf{w}_{k,l}) = \mathbf{w}_{k,l}^H \mathbf{R}_{\mathbf{yy}} \mathbf{w}_{k,l}$ is the cost function to be minimized. Noting that,

$$\nabla_{\mathbf{w}_{k,l}}(J(\mathbf{w}_{k,l})) = 2\mathbf{R}_{\mathbf{yy}} \mathbf{w}_{k,l} \quad (33)$$

and approximating the covariance matrix by its instantaneous value $\mathbf{y}\mathbf{y}^H$, we have

$$\nabla_{\mathbf{w}_{k,l}}(J(\mathbf{w}_{k,l})) = 2\mathbf{y}\mathbf{y}^H \mathbf{w}_{k,l}. \quad (34)$$

In order to make sure that the constraint $\mathbf{C}_k^H \mathbf{w}_{k,l} = \mathbf{e}_l$ is satisfied at each iteration, we may initialize the LMS algorithm with

$$\mathbf{w}_{k,l}(0) = (\mathbf{C}_k^H)^{\dagger} \mathbf{e}_l, \quad (35)$$

where $\mathbf{A}^{\dagger}$ denotes the Moore-Penrose generalized inverse of the matrix $\mathbf{A}$, and then perform the weight adaptation in the subspace orthogonal to the range of $\mathbf{C}_k$. Since the projection matrix on to the column space of $\mathbf{C}_k$ is $\mathbf{P}_C = \mathbf{C}_k(\mathbf{C}_k^H \mathbf{C}_k)^{-1}\mathbf{C}_k^H$, the component of the gradient vector along the subspace orthogonal to the constraint subspace is simply given as $\mathbf{P}_C^{\perp} = \mathbf{I} - \mathbf{P}_C$, and thus the blind LMS weight adaptation algorithm becomes, for $l = 1, \dots, L$,

$$\mathbf{w}_{k,l}(i+1) = \left(\mathbf{I} - \mu \mathbf{C}_k \left(\mathbf{C}_k^H \mathbf{C}_k\right)^{-1} \mathbf{C}_k^H\right) \mathbf{y}\mathbf{y}^H \mathbf{w}_{k,l}(i) \quad (36)$$

with the initialization

$$\mathbf{w}_{k,l}(0) = \left(\mathbf{C}_k^H\right)^{\dagger} \mathbf{e}_l. \quad (37)$$

Thus, the proposed receiver provides a simple and effective way to exploit the frequency diversity offered by the multipath channel and at the same time to take full advantage of the space-time coding diversity. With knowledge of only the desired user's spreading code, we may implement the blind adaptive algorithm to mitigate the MAI. It should be noted that still the knowledge of channel fading coefficients of the desired user are required for the successful decoding of the space-time block code, and thus the proposed scheme is *blind* only in the sense that it does not require the knowledge of the signature sequences of the interfering users.

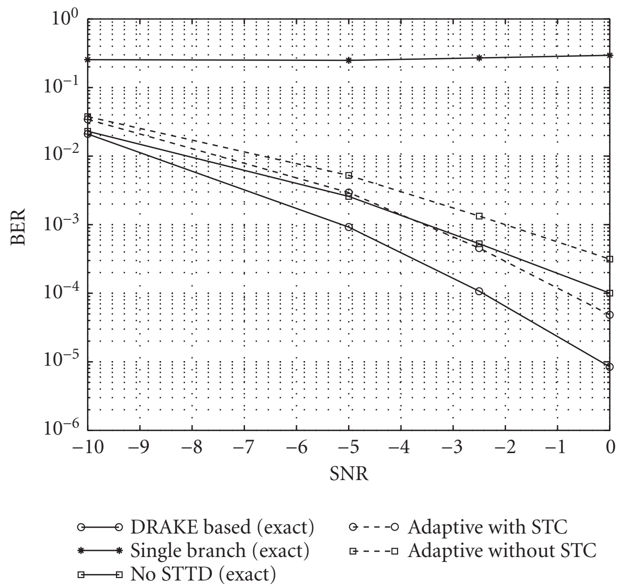
As we will see in the next section, the performance of the blind algorithm is still much better than that of a conventional system in which either only a single branch space-time decoder or a RAKE receiver without space-time coding is used.

The complexity of the LMS algorithm is of order $\mathcal{O}(N + L - 1)$ per update, which will usually be dominated by the processing gain N. The filtering operation in each branch requires another $N+L-1$ multiplications and a single addition. The space-time decoding requires only $N_T = 2$ multiplications and N_T additions per branch. The implementation of MRC requires only L multiplications and additions. Hence, the total complexity per bit of the proposed receiver using MRC is $\mathcal{O}(N+L-1+L(N+L-1+N_T)) \sim \mathcal{O}(LN+LN_T+L^2)$.

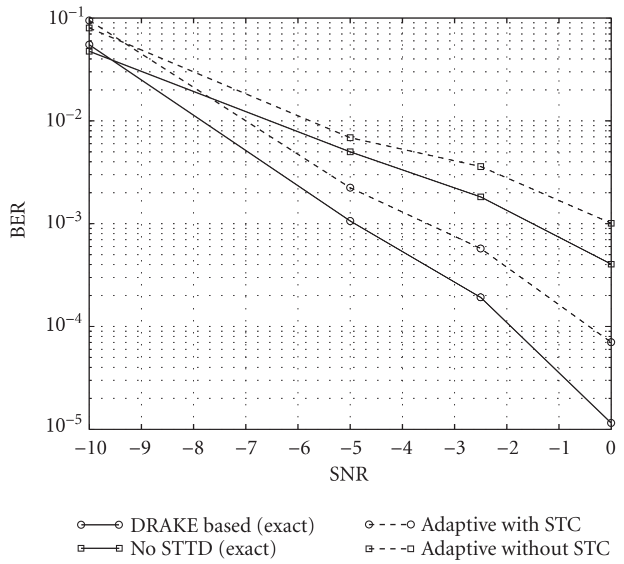
5. SIMULATION RESULTS

Plots (a) and (b) in Figure 1 show the BER performance of the proposed receiver for a BPSK system with $K = 4$ users and $L = 3$ paths per user. The processing gain of the CDMA system is $N = 16$ (random spreading codes were used in these simulations). Figure 1a corresponds to the BER averaged over 10 random code simulations. Figure 1b shows the performance of the best code set out of those 10 random spreading code sets. The fading is assumed to be quasi-static Rayleigh fading where the fading coefficients are constant for a block of 100 symbols and then change independently to new values. The space-time code used is the Alamouti code [1] with $N_T = 2$.

In both figures, we have also included the performance of a similar system without space-time coding but with the same multipath combining receiver. From Figure 1, it is clear that, when the SNR is sufficiently high, the proposed receiver with space-time coding can offer improved performance in multipath channels. Figure 1 also shows the BER performance obtained with the adaptive implementation of the receiver. It should be noted that in the simulations we did not attempt to optimize the step-size parameter of the LMS algorithm.



(a)



(b)

Figure 1: BER performance versus E_b/N_0 (in dB) of the DRAKE-based space-time detector. $N_T = 2$, $K = 4$, $L = 3$, and $N = 16$ in all cases. (a) Average. (b) Best.

Figure 1a also includes the performance of the space-time coded system which does not exploit the multipath components present in the received signal. This corresponds to a receiver with only a single branch filter. Clearly, the single branch filter fails to offer any acceptable performance level since it considers all the paths and the signal components from different transmit antennas embedded in those paths as pure interference, and thus the SINR at the output of the filter is extremely poor. Of course, increasing all the users'

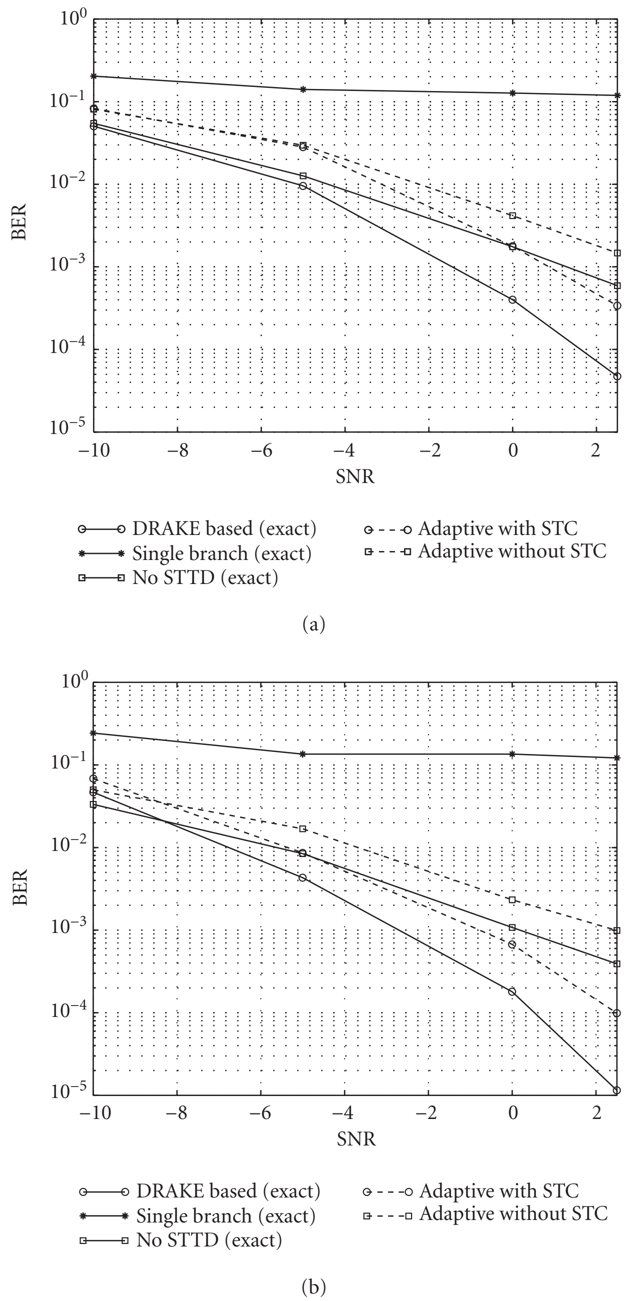


Figure 2: BER performance versus E_b/N_0 (in dB) of the DRAKE-based space-time detector. $N_T = 2$, $K = 8$, $L = 2$, and $N = 16$ in all cases. (a) Average. (b) Best.

transmit powers does not help in this case and thus the performance of the single branch filter is insensitive to the SNR.

Figure 2 gives similar performance results for a system with $K = 8$ users and $L = 2$ paths per user. Again, we observe that the proposed receiver maintains improved performance over a system without space-time coding as before.

It is clear from these simulation results that the proposed scheme offers significant performance improvement with only linear complexity at the downlink receiver as long as the number of paths in the channel is not too large and the SNR is sufficiently high.

6. CONCLUSIONS

We have proposed a downlink receiver for space-time block coded CDMA systems in multipath channels. The new receiver exploits both the useful signal energy contained in the multipath replicas of the received signal and the diversity offered by the space-time coding process. We have shown that this type of receiver can offer significant performance improvement on the downlink in the presence of fading. An LMS-based adaptive implementation has also been proposed, which requires the knowledge of only the desired user's spreading code. Our simulation results show that the new receiver can offer significant performance gain over systems without space-time coding, without significant increase in the computational complexity, as long as the number of paths in the channel is not too large and the SNR is sufficiently high.

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