Semiblind Channel Estimation for Space-time Coded WCDMA

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***Abstract*— A new semiblind channel estimation technique is proposed for space-time coded wideband CDMA systems using aperiodic and possibly multirate spreading codes. Using a decorrelating matched filter, the received signal is projected onto subspaces from which channel parameters and data symbols can be estimated jointly. Exploiting the subspace structure of the WCDMA signaling and the orthogonality of the space-time code, the proposed algorithm provides the least squares channel estimate in closed form. A new identifiability condition is established. The mean square error of the estimated channel is compared with the Cramér-Rao bound, and a bit error rate (BER) expression for the proposed algorithm is compared with differential schemes.**

I. Introduction

Future wireless systems will require high rate transmissions of multimedia data over time varying fading channels. This is especially the case for the downlink where a mix of voice, low rate data and possibly images are transmitted to mobile users. To provide reliable communications over a fading channel, diversity techniques in space and time are expected to play a critical role [1] [4]. A variety of space-time coding schemes have been proposed with multiple transmit antennas and a single or multiple receive antennas (*eg.* [2] [3] [5] [6]). Indeed, the third generation wireless standard supports base station transmit diversity using the WCDMA physical layer interface.

Many space-time techniques, the popular Alamouti scheme in particular, are designed for coherent detection where channel estimation is necessary. There is a substantial literature addressing the channel estimation issue for (space-time coded) multiple-input multiple-output (MIMO) systems, ranging from standard training based techniques that rely on pilot symbols in the data stream to blind and semiblind estimation where observations corresponding to data and pilots (if they exist) are used jointly. Noncoherent detection schemes for space-time coded systems have also been proposed based on differential or sequential decoding [7], [8] . These methods avoid the need for channel estimation by introducing structures in the transmitted symbol stream. Exploiting such structures, the receiver can recover the transmitted symbols directly but often pays a price in performance.

In this paper, we focus on space-time coded coherent long code WCDMA systems. The challenge of channel estimation in such a wideband system is twofold. First, the 3G WCDMA is a multirate system where the delay spread may exceed several symbol intervals causing severe multipath fading and intersymbol interference; the channel is a MIMO system with memory. Second, the increase in the number of channel parameters, due to the use of multiple antennas, makes the conventional training based scheme less reliable and more prone to multiaccess interference. Fortunately, WCDMA also offers code diversities that could be exploited in an estimation scheme.

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In this paper, we propose a semiblind channel estimation technique that requires no more than two pilot symbols per user per slot. Using a RAKE structure, our technique is compatible with standard receiver front-ends that suppress multiaccess interference and perform decoding for each user separately. The proposed algorithm exploits the subspace structure of the long code WCDMA transmission and the orthogonality of the Alamouti code. As a subspace technique, the proposed algorithm can obtain channel estimates quickly using only few symbols, which allows us to deal with rapidly fading channels.

The paper is organized as follows. The data model of a space-time coded long code CDMA system is described in Section II. A new blind channel estimation method based on decorrelation and an identifiability condition are proposed and several extensions are discussed in Section III. In Section IV, the performance of the proposed method is compared with the Cramér-Rao Bound (CRB) through Monte Carlo simulations and the bit error rate (BER) of the proposed method is compared with that of differential detection schemes.

II. Data Model

We consider a space-time coded WCDMA downlink with two transmit antennas and a single receive antenna, K synchronous users with aperiodic spreading sequences of spreading gain G, and slotted transmissions of M symbols per slot.

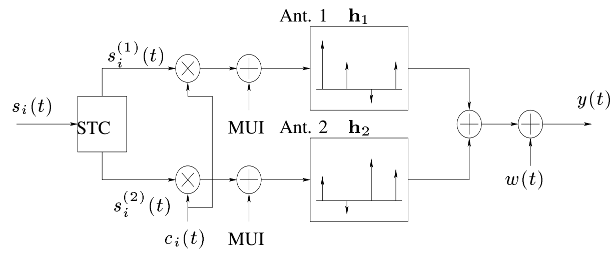

Fig. 1. Downlink Model with Space-time Coding (STC: Space-time Encoder).

Specifically, user i's data sequence for each transmitter is space-time encoded as [3]

$$s_{i,m+1}^{(1)} = -s_{i,m}^{(2)*}, \quad (1)$$

$$s_{i,m+1}^{(2)} = s_{i,m}^{(1)*}, \quad m = 1, 3, \ldots, M-1, \quad (2)$$

where $s_{i,m}^{(j)} \triangleq s_i^{(j)}(mT_s)$ is the data sequence of user i for transmit antenna j, T_s is the symbol interval, and $(\cdot)^*$ is complex conjugate. Each data sequence is spread by an aperiodic user specific code $c_i(t)$ and transmitted through the corresponding antenna. With $s_{i,m}$ denoting the symbol sequence of the i-th user, we have $s_{i,m}^{(1)} = s_{i,m}$.

Suppose that the channel for each transmit and receive antenna pair is modeled by a complex FIR filter with taps separated by multiples of chip intervals. The continuous-time channel impulse response of the path from transmitter j to the single receiver has the form

$$h_j(t) = \sum_{l=1}^{L_j} h_{jl}\delta(t - lT_c - d_jT_c),$$

where $\{h_{jl}\}$ are path gains, $T_c = T_s/G$ is the chip interval. We will assume that the channel doesn't change over a slot period and that delay d_j and channel order L_j are known[1]. L_j is often a design parameter. We can set $L_j = L, \quad j = 1, 2$, by taking the maximum of $\{L_j\}$. When the channel is sparse, it is more efficient to model the channel as separate clusters of fingers. In that case, we assume that the approximate locations of these clusters are known. Since any path gain can be zero, one can over-parameterize the channel to accommodate channel uncertainties.

We let the received signal $y(t)$ pass through a chip-matched filter and sample its output at the chip rate. The received noiseless signal vector $\mathbf{y}_{im}$ that corresponds to the mth symbol interval of user i is given by

$$\mathbf{y}_{im} = \mathbf{T}_{im}[\mathbf{h}_1 s_{im}^{(1)} + \mathbf{h}_2 s_{im}^{(2)}], \tag{3}$$

where $\mathbf{h}_j \triangleq [h_{j1}, h_{j2}, \cdots, h_{jL}]^T, j = 1, 2$, and $\mathbf{T}_{im}$ is the Toeplitz matrix whose first column is made of $(m-1)G + d_j$ zeros followed by the code vector $\mathbf{c}_{im}$—the mth segment of G chips of the spreading code of user i—and additional zeros that make the size of $\mathbf{y}_{im}$ the total number of chips of the entire M-symbol slot. The total received noiseless signal for user i is given by

$$\begin{aligned}
\mathbf{y}_i &= \sum_{m=1}^{M} \mathbf{T}_{im}[\mathbf{h}_1 s_{i,m}^{(1)} + \mathbf{h}_2 s_{i,m}^{(2)}] = \mathbf{T}_i(\mathbf{I}_M \otimes [\mathbf{h}_1 \quad \mathbf{h}_2])\mathbf{s}_i,\\
&\quad \mathbf{s}_i \triangleq [s_{i,1}^{(1)}, s_{i,1}^{(2)}, s_{i,2}^{(1)}, s_{i,2}^{(2)}, \cdots, s_{i,M}^{(2)}]^T,\\
&\quad \mathbf{T}_i \triangleq [\mathbf{T}_{i1}, \mathbf{T}_{i2}, \cdots, \mathbf{T}_{iM}],
\end{aligned} \tag{4}$$

where $\otimes$ is the Kronecker product and $\mathbf{T}_i$ is the code matrix of user i, and it has a special block shifting structure. Including all users, we have the complete matrix model

$$\begin{aligned}
\mathbf{y} &= [\mathbf{T}_1 \cdots \mathbf{T}_K]\mathrm{diag}(\mathbf{I}_M \otimes \mathbf{H}, \cdots, \mathbf{I}_M \otimes \mathbf{H})\mathbf{s} + \mathbf{w},\\
&= \mathbf{T}\mathcal{D}(\mathbf{H})\mathbf{s} + \mathbf{w},
\end{aligned} \tag{5}$$

where the overall $(MG \times LMK)$ code matrix $\mathbf{T}$ contains all code matrices, $\mathbf{H} \triangleq [\mathbf{h}_1 \ \mathbf{h}_2]$, and $\mathbf{s}$ all symbols for all K users. Matrix $\mathcal{D}(\mathbf{H})$ is block diagonal with $\mathbf{I}_M \otimes \mathbf{H}$ as the block element. The additive noise is denoted by $\mathbf{w}$.

We will make the following assumptions.

A1: The code matrix $\mathbf{T}$ is known.

A1': The code matrix $\mathbf{T}$ has full column rank.

A2: The noise vector is complex Gaussian $\mathbf{w} \sim \mathcal{N}(\mathbf{0}, \sigma^2\mathbf{I})$ with possibly unknown variance σ^2.

Assumption (A1) implies that the receiver knows the codes for all users (as well as the delay d_j and the maximum channel order L). For the downlink, the relative delays d_i and the number of multipaths L_i are the same for all the user signals. The assumption is especially valid for the multiple spreading code case.

Assumption (A1') is sufficient but not necessary for the channel to be identifiable and for the algorithm proposed in Section 3 to produce good estimates. This condition can of course be verified off line and is easy to satisfy for a relative large spreading gain by modeling K dominant users in the signal ($LK < G$) and other user in the additive noise.

[1]If d_j is approximately known, we can set $d_j = 0, \quad j = 1, 2$, and model all the paths.

III. Blind and Semiblind Channel Estimation

We propose a blind channel estimator based on decorrelation of user signals. The proposed method projects the received signal onto a subspace from which the channels of both transmitter-receiver pairs are estimated simultaneously using a low rank decomposition. Blind estimation is possible due to the unitary property of the space-time codes, especially the Alamouti code. The proposed method combines two consecutive symbols and eliminates the unknown symbols exploiting this unitary property. We assume that the channel and symbols are deterministic parameters.

A. Blind Algorithm

The decorrelating front end $\mathbf{T}^\dagger$ can be pre-computed or efficiently implemented using a state-space inversion technique that significantly reduces the complexity and storage requirement of the decorrelating receiver by exploiting the structure of the code matrix [10]. The output of the decorrelating matched filter is written in vector form as

$$\mathbf{z} = \mathbf{T}^\dagger\mathbf{y} = \mathrm{diag}(\mathbf{I} \otimes \mathbf{H}, \cdots, \mathbf{I} \otimes \mathbf{H})\mathbf{s} + \mathbf{n}, \tag{6}$$

where the $LKM \times 1$ vector $\mathbf{n} = \mathbf{T}^\dagger\mathbf{w}$ is now colored. Segmenting $\mathbf{z}$ to the subvectors corresponding to symbols $2n-1$, $2n$ of user i, we have

$$\begin{aligned}
\mathbf{z}_{i,2n-1} &= \mathbf{H}\begin{bmatrix} s_{i,2n-1} \\ -s_{i,2n}^* \end{bmatrix} + \mathbf{n}_{2n-1},\\
\mathbf{z}_{i,2n} &= \mathbf{H}\begin{bmatrix} s_{i,2n} \\ s_{i,2n-1}^* \end{bmatrix} + \mathbf{n}_{2n},
\end{aligned} \tag{7}$$

where $n = 1, 2, \ldots, M/2$, $\mathbf{z}_{im}$ is the $((i-1)M + m)$th L-dimensional subvector of $\mathbf{z}$. Collecting data for two consecutive coded symbols of user i gives

$$\mathbf{Z}_{in} \triangleq [\mathbf{z}_{i,2n-1} \ \mathbf{z}_{i,2n}] = \mathbf{H}\mathbf{S}_{in} + \mathbf{N}_n, \tag{8}$$

where $\mathbf{N}_n \triangleq [\mathbf{n}_{2n-1} \ \mathbf{n}_{2n}]$, and

$$\mathbf{S}_{in} \triangleq \begin{bmatrix} s_{i,2n-1} & s_{i,2n} \\ -s_{i,2n}^* & s_{i,2n-1}^* \end{bmatrix}. \tag{9}$$

Note that the symbol matrix $\mathbf{S}_{in}$ is unitary, i.e.,

$$\mathbf{S}_{in}\mathbf{S}_{in}^H = \mathbf{S}_{in}^H\mathbf{S}_{in} = \beta\mathbf{I}. \tag{10}$$

For the noiseless case, it is easily seen that multiplying $\mathbf{Z}_{in}$ by its Hermitian eliminates the unknown symbol which makes blind identification possible. The least squares estimator of the product $\mathbf{X}_i \triangleq \mathbf{H}\mathbf{S}_{in}$, which ignores the noise color, is given by

$$\arg\min_{\mathbf{H},\mathbf{S}_{in}} ||\mathbf{Z}_{in} - \mathbf{H}\mathbf{S}_{in}||_F^2,$$

which yields estimates of $\mathbf{H}$ and $\mathbf{S}_{in}$ (with an unknown unitary matrix factor) from the decomposition of $\mathbf{Z}_{in}$. More explicitly, let

$$\hat{\mathbf{R}}_i \triangleq \frac{1}{M}\sum_{n=1}^{M/2} \mathbf{Z}_{in}\mathbf{Z}_{in}^H, \tag{11}$$

and let the SVD of $\mathbf{R}_i$ be given as

$$\hat{\mathbf{R}}_i = \mathbf{U}_i\Sigma_i\mathbf{U}_i^H. \tag{12}$$

We obtain the least squares estimator as a low-rank approximation with the minimum Frobenius norm which is given by

$$\hat{\mathbf{H}} = \mathbf{U}\Sigma^{1/2}\mathbf{Q}, \tag{13}$$

where $\mathbf{Q}$ is an unknown 2×2 unitary matrix. The rotational ambiguity in the above estimate must be removed by either incorporating prior knowledge of the symbol, or by using pilot symbols.

With the estimated channel, we can perform blockwise maximum likelihood detection to obtain the symbol sequence. Rewriting eq.(7) gives

$$\begin{bmatrix} \mathbf{z}_{i,2n-1} \\ \mathbf{z}^*_{i,2n} \end{bmatrix} = \begin{bmatrix} \mathbf{h}_1 & -\mathbf{h}_2 \\ \mathbf{h}_2^* & \mathbf{h}_1^* \end{bmatrix} \begin{bmatrix} s_{i,2n-1} \\ s^*_{i,2n} \end{bmatrix} + \begin{bmatrix} \mathbf{n}_{2n-1} \\ \mathbf{n}^*_{2n} \end{bmatrix}. \tag{14}$$

Maximum likelihood estimates for symbol $s_{i,2n-1}$ and $s_{i,2n}$ are given by

$$\begin{bmatrix} \hat{s}_{i,2n-1} \\ \hat{s}^*_{i,2n} \end{bmatrix} = \mathcal{Q}\left(\frac{1}{\alpha}\begin{bmatrix} \hat{\mathbf{h}}_1^H & \hat{\mathbf{h}}_2^T \\ -\hat{\mathbf{h}}_2^H & \hat{\mathbf{h}}_1^T \end{bmatrix}\begin{bmatrix} \mathbf{z}_{i,2n-1} \\ \mathbf{z}^*_{i,2n} \end{bmatrix}\right), \tag{15}$$

where $\alpha = (||\mathbf{h}_1||^2 + ||\mathbf{h}_2||^2)$, superscripts T and H denote transpose and conjugate transpose, and $\mathcal{Q}$ is the quantization function which selects the symbol vector with minimum distance. The performance of this coherent detector with the proposed channel estimate is compared with differential schemes in Sec.IV.

We have so far assumed that the $(MG \times LMK)$ code matrix $\mathbf{T}$ has full column rank and therefore invertible from the left, i.e., $\mathbf{T}^\dagger\mathbf{T} = \mathbf{I}$. This assumption is usually valid for systems with large spreading gains or small delay spreads (need $G > LK$). Under this assumption, it is clear that each user's channel is identifiable up to a scaling factor. A single pilot symbol will be sufficient to remove the scalar ambiguity. When the spreading gain is small and the system is heavily loaded, $\mathbf{T}$ can be singular. We present a general identifiability condition for the proposed method. The condition is independent of the channel parameters, and can be checked easily off-line, and appropriate measures taken.

Proposition 1: Let $\tilde{\mathbf{T}}_{in} \triangleq [\mathbf{T}_{i,2n-1}\ \ \mathbf{T}_{i,2n}]$ be the matrix composed of two consecutive code matrices of user i for symbol $2n-1$, $2n$, and $\check{\mathbf{T}}_{in}$ the submatrix of $\mathbf{T}$ after removing $\tilde{\mathbf{T}}_{in}$. The channel $\mathbf{H}$ is identifiable if there exists an i and an n such that

$$R(\tilde{\mathbf{T}}_{in}) \bigcap R(\check{\mathbf{T}}_{in}) = \{\mathbf{0}\}. \tag{16}$$

where $R(\bullet)$ denotes the range space of a matrix.

Proof: If eq.(16) holds for some i and n, then the range space of $\mathbf{T}$ can be decomposed into the sum of tow subspaces, i.e., there exists a matrix $\mathbf{V}$ with rank($\mathbf{T}$) - rank($\tilde{\mathbf{T}}_{in}$) linearly independent columns such that

$$R([\tilde{\mathbf{T}}_{in}\ \ \mathbf{V}]) = R(\mathbf{T}).$$

Let $\mathcal{T} \triangleq [\tilde{\mathbf{T}}_{in}\ \ \mathbf{V}]$. We have, for the noiseless case,

$$\mathcal{T}^\dagger\mathbf{y} = \begin{bmatrix} * \\ \mathbf{h}_1 s_{i,2n-1} - \mathbf{h}_2 s^*_{i,2n} \\ \mathbf{h}_1 s_{i,2n} + \mathbf{h}_2 s^*_{i,2n-1} \\ * \end{bmatrix},$$

which implies that $\mathbf{H}$ is identifiable up to a rotational ambiguity. ■

Since eq.(16) needs to hold only for some n, the use of long codes makes the identifiability condition easier to satisfy.

B. Whitened Estimator and Extensions

Since the noise $\mathbf{n}_{im}$ after decorrelation is colored, a bias is introduced in estimation. We can apply whitening to remove the bias. The expectation of $\hat{\mathbf{R}}_i$ in eq. (11) is given by (assuming without loss of generality that $\beta = 1$ in eq.(10))

$$\begin{aligned} \mathbb{E}\{\hat{\mathbf{R}}_i\} &= \mathbf{H}_i\mathbf{H}_i^H + \sigma^2\Delta_i, \\ \Delta_i &= \frac{1}{M}\sum_{m=1}^{M}\Sigma_{im}, \end{aligned} \tag{17}$$

where Σ_{im} is the $(i-1)M+m$th diagonal block of $\mathbf{T}^\dagger(\mathbf{T}^\dagger)^H$ with size $L \times L$. The whitened estimator is given as

$$\hat{\mathbf{H}} = \Delta_i^{1/2}\mathbf{U}_i\Sigma_i^{1/2}\mathbf{Q}, \tag{18}$$

where $\Delta_i^{1/2}$ is the Cholesky factor of Δ_i and

$$\Delta_i^{-1/2}\hat{\mathbf{R}}_i\Delta_i^{-H/2} = \mathbf{U}_i\Sigma_i\mathbf{U}_i^H. \tag{19}$$

For the downlink case, all the user data experience the same channel. We can improve the estimator performance by exploiting this. A simple approach is to combine the outer product $\hat{\mathbf{R}}_i$.

$$\begin{aligned} \hat{\mathbf{R}} &= \frac{1}{K}\sum_{i=1}^{K}\hat{\mathbf{R}}_i = \frac{1}{KM}\sum_{i=1}^{K}\sum_{n=1}^{M/2}\mathbf{Z}_{in}\mathbf{Z}_{in}^H, \\ \Delta &= \frac{1}{K}\sum_{i=1}^{K}\Delta_i. \end{aligned} \tag{20}$$

This process further improves the performance by averaging out the noise.

In addition to the decorrelating front end, we can apply the same subspace technique to different front ends such as the conventional matched filter $\mathbf{T}^H$ or the regularized decorrelating front end which is given as

$$(\mathbf{T}^H\mathbf{T} + \sigma^2\mathbf{I})^{-1}\mathbf{T}^H. \tag{21}$$

As shown in eq.(21), the regularized decorrelating front end requires the estimation of noise power. For the case of conventional matched filter, the algorithm still gives better performance than training-based estimation. However, it exhibits the well known performance floor due to multiaccess interference.

C. Resolving the Rotational Ambiguity

The unknown unitary matrix $\mathbf{Q}$ in eq. (13) and eq. (18) can be solved using least squares, given pilot symbols. The problem of estimating $\mathbf{Q}$ is formulated as

$$\begin{aligned} \hat{\mathbf{Q}} &= \arg\min_{\mathbf{Q}\in\mathbb{C}^{2\times 2}} ||\mathbf{Z}_{ip} - \Delta_i^{1/2}\mathbf{U}_i\Sigma_i^{1/2}\mathbf{Q}\mathbf{S}_p||_F^2, \\ &= \arg\min_{\mathbf{Q}\in\mathbb{C}^{2\times 2}} ||\mathbf{Z}_{ip}\mathbf{S}_p^H - \beta\Delta_i^{1/2}\mathbf{U}_i\Sigma_i^{1/2}\mathbf{Q}||_F^2, \quad (22) \end{aligned} \tag{23}$$

under the constraint

$$\mathbf{Q}\mathbf{Q}^H = \mathbf{I},$$

where, for the example of two pilot symbols at slot front, $\beta = (|s_{i1}|^2 + |s_{i2}|^2)$ and the pilot related matrix $\mathbf{Z}_{ip}$ and $\mathbf{S}_p$ are given as

$$\mathbf{Z}_{ip} = [\mathbf{z}_{i1}, \mathbf{z}_{i2}], \quad \mathbf{S}_p = \begin{bmatrix} s_{i1} & s_{i2} \\ -s^*_{i2} & s_{i1} \end{bmatrix}. \tag{24}$$

The least square solution of eq. (22) can be obtained using singular value decomposition as [9]

$$\hat{\mathbf{Q}} = \mathbf{U}_Q \mathbf{V}_Q^H, \tag{25}$$

where

$$\mathbf{U}_Q \Sigma_Q \mathbf{V}_Q = \beta(\Delta_i^{1/2} \mathbf{U}_i \Sigma_i^{1/2})^H \mathbf{Z}_{ip} \mathbf{S}_p^H. \tag{26}$$

IV. Simulation Results

In this section, we present some simulation results. For channel estimation, MSE was used and our estimator was compared with the CRB using Monte Carlo runs. For symbol detection, BER was used as the performance measure and the BER of the proposed method was compared with that of the differential detection in [8].

We considered a downlink system with two transmit antennas and a single receive antenna. Single ($K = 1$) and four ($K = 4$) synchronous BPSK users with equal power were considered. The spreading codes for users were randomly generated with spreading gain $G = 32$ [12] and fixed throughout the Monte Carlo simulation for MSE and BER. Since our channel model is deterministic, the channel parameter was also fixed during the Monte Carlo runs. For the CRB calculation, the symbol sequence was fixed and for MSE and BER, symbol sequences were generated randomly for each Monte Carlo run.

The channel for each TX-RX pair had three fingers $L = 3$ (here $LK < G$). The coefficients are given by $\mathbf{h}_1 = [0.3766 + 0.1092i, 0.1569 - 0.0942i, 0.4384 - 0.7878i]$ and $\mathbf{h}_2 = [0.4388 + 0.6751i, -0.5052 + 0.0737i, -0.2891 + 0.0859i]$. The slot size was $M = 80$ and two pilot symbols were included at the beginning of the slot of each user. These pilot symbols were used to remove the rotational ambiguity of the blind estimator. The signal-to-noise ratio (SNR) is defined by $(||\mathbf{h}_1||^2 + ||\mathbf{h}_2||^2)GE_c/\sigma^2$ where E_c is the chip energy and σ^2 is the chip noise variance.

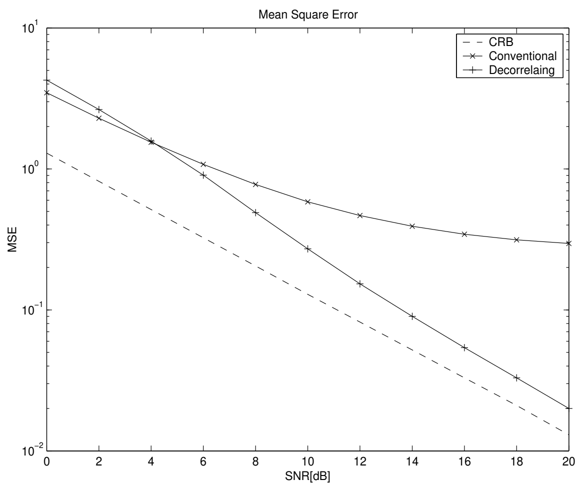

Fig. 2. Channel MSE vs. SNR. Two pilot symbols per user.

We compare the mean square error performance of the proposed channel estimator using the decorrelating front end (DRR) with the CRB and the conventional training based algorithm based on the standard matched filter. Due to the two pilot symbols inserted for rotational ambiguity resolution, we used the semi-blind CRB with a deterministic assumption on data symbols [11]. For the training based method, a least-squares channel estimate was obtained using data corresponding to the pilot symbols. Fig. 2 shows the MSE performance. We observe that as SNR increases, the conventional method has a performance deviation from the CRB due to multiaccess interference. The DRR, on the other hand, tracks the CRB closely at high SNR. The high MSE of DRR at low SNR is due to the noise enhancement by the decorrelating (zero-forcing) front end.

We evaluated BER performance for the coherent detector in Section III with decorrelating, regularized decorrelating(RDRR) front ends and the differential scheme in [8] using RDRR with true channel as the performance bound. Figure 3 shows the BER performance for

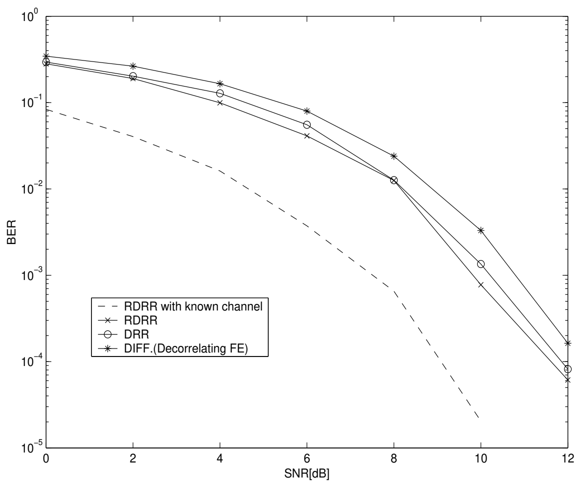

Fig. 3. BER vs. SNR. - Single User Case

a single user case. We observe an almost 2dB loss at BER of 10^{-3} due to channel estimation errors for the coherent detector. We observe that the difference between the proposed method and the differential detector is less than 0.5 dB over the whole BER range.

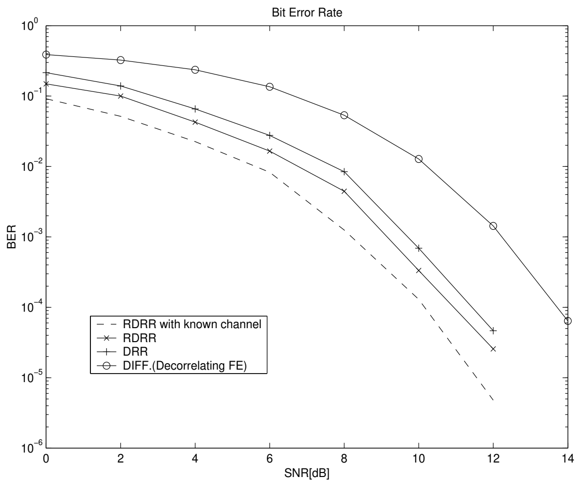

Fig. 4. BER vs. SNR. - Four User Case

Figure 4 shows the BER performance of the various detectors for four user case. The improvement of the proposed method over the differential scheme is pronounced. This is because the proposed method utilized all the user data constructively to estimate the downlink channel whereas the differential detection is performed individually.

V. Conclusion

We propose a new semi-blind channel estimation technique for space-time coded wideband CDMA systems. A new identifiability condition is established. The proposed method identifies the channel of each transmit-receive pair simultaneously exploiting the subspace structure of WCDMA signals and the orthogonality of space-time codes with a few pilot symbols. The performance of the proposed method is compared with that of differential schemes.

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