Blind Adaptive MIMO Receivers for CDMA Systems with Space-Time Block-Codes and Low-Cost Algorithms

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Abstract-In this paper we present low-complexity blind multi-input multi-output (MIMO) adaptive linear multiuser receivers for direct sequence code division multiple access (DS-CDMA) systems using multiple transmit antennas and spacetime block codes (STBC) in multipath channels. A spacetime code-constrained constant modulus (CCM) design criterion based on constrained optimization techniques and low-complexity stochastic gradient (SG) adaptive algorithms are developed for estimating the parameters of the space-time linear receivers. The receivers are designed by exploiting the unique structure imposed by both spreading codes and STBC. A blind space-time channel estimation scheme for STBC systems based on a subspace approach is also proposed along with an efficient SG algorithm. Simulation results for a downlink scenario assess the receiver structures and algorithms and show that the proposed schemes achieve excellent performance, outperforming existing methods.

I. INTRODUCTION

The high demand for performance and capacity in wireless networks has motivated the development of a large number of signal processing and communications techniques for utilizing these resources efficiently. Recent results on information theory have shown that it is possible to achieve high spectral efficiency [1] and to make wireless links more reliable [2], [3] through the deployment of multiple antennas at both transmitter and receiver. Space-time coding (STC) is a set of techniques that provides an effective way of exploiting spatial and temporal transmit diversity, resulting in more reliable wireless links [2], [3]. Because it is more cost-effective to deploy multiple antennas at the base station rather than at the mobile terminal, several space-time coding techniques such as space-time block codes (STBC) [2], [3] and differential space-time codes [4], have been recently investigated for improving the quality of wireless downlink connections. Among these techniques, STBC has become very popular as it offers maximum diversity gain based on linear processing at the receiver and due to the simplicity in which it can be adopted

in several communications technologies such as the widely adopted CDMA systems.

In CDMA systems, the designer usually has to deal with multiple access interference (MAI), which arises due to lack of orthogonality between the user signals at the receiver, and intersymbol interference (ISI), which is a result of multiple propagation paths. Interference suppression in multiantenna systems is more challenging than in single-antenna ones because the former is subject to self-interference. The problem of receiver design for DS-CDMA systems using multiple transmit antennas and STBC has been considered in several recent works [5]-[10]. Maximum likelihood solutions [5] that consider both STBC and multiuser systems are very complex and this calls for suboptimal and low-complexity solutions. Since original STBCs were originally designed for flat channels, they may experience performance degradation under multipath, where their diversity gain capabilities can be considerably reduced. In order to design CDMA systems with STBC in multipath scenarios an interesting and promising approach is to combine the transmit streams separation and interference suppression tasks into one single processing stage. This enables the use of low-complexity multiuser receivers and design algorithms.

In this work we present blind adaptive linear MIMO receivers based on the code-constrained constant modulus (CCM) design for DS-CDMA systems using multiple transmit antennas and STBC in multipath channels. We consider a multi-antenna system with STBC and exploit the unique code structure of space-time coding to derive efficient blind receivers based on the CCM design and develop low-complexity stochastic gradient (SG) algorithms. To blindly estimate the channel, we describe a subspace approach that exploits the STBC structure present in the received signal and derive an adaptive SG channel estimator. The only requirement for the receivers is the knowledge of the signature sequences for the

desired user.

This paper is organized as follows. Section II briefly describes the space-time DS-CDMA communication system model. The space-time linearly constrained receivers based on the CCM design criterion are presented in Section III. A subspace framework for blind channel estimation of DS-CDMA signals with STBC is presented in Section IV. Section V is devoted to the derivation of adaptive SG algorithms for both receiver and channel parameter estimation. Section VI presents and discusses the simulation results and Section VII gives the concluding remarks of this work.

II. SPACE-TIME DS-CDMA SYSTEM MODEL

Let us consider the downlink of a symbol synchronous QPSK DS-CDMA system with K users, N chips per symbol, N_t antennas at the transmitter, N_r antennas at the receiver and L_p propagation paths. For simplicity, we assume that the transmitter (Tx) employs only $N_t = 2$ antennas and adopts Alamouti's STBC scheme [2], even though other STBC can also be deployed. According to this scheme, depicted in Fig. 1, during the (2i-1)th symbol interval, where *i* is an integer, two symbols $b_k(2i-1)$ and $b_k(2i)$ drawn from the same constellation are transmitted from Tx1 and Tx2, respectively, and during the next symbol interval, $-b_k(2i)$ and $b_k^*(2i)$ are transmitted from Tx1 and Tx2, respectively. In addition, each user is assigned a different spreading code for each Tx, which may be constructed from a single spreading code s_k as $[\mathbf{s}_k^T, \mathbf{0}^T]^T$ and $[\mathbf{0}^T, \mathbf{s}_k^T]^T$, respectively, a scheme proposed for UMTS W-CDMA or $[\mathbf{s}_k^T, \mathbf{s}_k^T]^T$ and $[\mathbf{s}_k^T, -\mathbf{s}_k^T]^T$, respectively, an approach adopted in the IS-2000 standard [6].



Fig. 1. Proposed space-time system: schematic of the kth user (a) Transmitter and (b) Receiver .

In this context, the baseband signal transmitted by the k-th active user to the base station is given by

$$x_{k}^{1}(t) = A_{k} \sum_{i=1}^{P} b_{k}^{1}(i) s_{k}^{1}(t-2iT) - b_{k}^{2}(i) s_{k}^{2}(t-(2i-1)T)$$
(1)

$$x_k^2(t) = A_k \sum_{i=1}^{k} b_k^2(i) s_k^2(t-2iT) + b_k^1(i) s_k^1(t-(2i-1)T)$$
(2)

where P is the packet length, $b_k(i) \in \{\pm 1 \pm j\}$ is the *i*th symbol for user k with $j^2 = -1$, the real valued spreading

waveform and the amplitude associated with user k are $s_k^{1,2}(t)$ and A_k , respectively. The spreading waveforms are expressed by $s_k^{1,2}(t) = \sum_{i=1}^N a_k(i)\phi(t-iT_c)$, where $a_k(i) \in \{\pm 1/\sqrt{N}\}$, $\phi(t)$ is the chip waverform, T_c is the chip duration and $N = T/T_c$ is the processing gain. Assuming that the receiver is synchronised with the main path, the coherently demodulated composite received signal for receive antenna m is

$$r_m(t) = \sum_{k=1}^{K} \sum_{l=0}^{L_p - 1} h_{l,m}^1(t) x_{k,1}(t - \tau_{l,m}^1) + h_{l,m}^2(t) x_{k,2}(t - \tau_{l,m}^2) + n(t)$$
(3)

where $h_{l,m}^{1,2}(t)$ and $\tau_{l,m}^{1,2}$ are, respectively, the channel coefficients for the *m*th receive antenna, transmit antenna $n_t = 1, 2$ and their corresponding delay associated with the *l*-th path. Assuming that $\tau_{l,m}^{1,2} = lT_c$, $h_{l,m}^{1,2}(i) = h_{l,m}^{1,2}(iT_c)$, the channel is constant during two symbol intervals and the spreading codes are repeated from symbol to symbol, the received signal r(t) at antenna *m* after filtering by a chip-pulse matched filter and sampled at chip rate over two consecutive symbols yields the *M*-dimensional received vectors

$$\mathbf{r}(2i-1) = \sum_{k=1}^{K} A_k b_k (2i-1) \mathbf{C}_k^1 \mathbf{h}_m^1 + A_k b_k (2i) \mathbf{C}_k^2 \mathbf{h}_m^2 + \eta_{k,m} (2i-1) + \mathbf{n}_m (2i-1) \mathbf{r}(2i) = \sum_{k=1}^{K} A_k b_k^* (2i-1) \mathbf{C}_k^2 \mathbf{h}_m^2 - A_k b_k^* (2i) \mathbf{C}_k^1 \mathbf{h}_m^1 + \eta_{k,m} (2i) + \mathbf{n}_m (2i) i = 1, \dots, P, \qquad m = 1, \dots, N_r$$
(4)

where $M = N + L_p - 1$, $\mathbf{n}_m(i) = [n_1(i) \dots n_M(i)]^T$ is the complex Gaussian noise vector with $E[\mathbf{n}_m(i)\mathbf{n}_m^H(i)] = \sigma^2 \mathbf{I}$, where $(.)^T$ and $(.)^H$ denote transpose and Hermitian transpose, respectively, E[.] stands for ensemble average, the amplitude of user k is A_k , the channel vector for users transmitted at Tx n_t $(n_t = 1, 2)$ and received at Rx m is $\mathbf{h}_m^{n_t} = [h_{m,0}^{n_t} \dots h_{m,L_p-1}^{n_t}]^T$, $\boldsymbol{\eta}_m(i)$ is the intersymbol interference at Rx m and the $M \times L_p$ convolution matrix $\mathbf{C}_k^{n_t}$ contains onechip shifted versions of the signature sequence for user k and transmit antenna n_t given by $\mathbf{s}_k = [a_k(1) \dots a_k(N)]^T$. Let us now organize the received data in (4) into a single received vector within the *i*th symbol interval at the *m*th receive antenna (Rx) as described by

$$\mathbf{y}_{m}(i) = \sum_{k=1}^{K} A_{k} b_{k}(2i-1) \mathcal{C}_{k} \mathcal{H}_{m}(i) + A_{k} b_{k}(2i) \bar{\mathcal{C}}_{k} \mathcal{H}_{m}^{*}(i) + \boldsymbol{\eta}_{k}(i) + \mathbf{n}(i) = \sum_{k=1}^{K} \mathbf{x}_{k}(i) + \bar{\mathbf{x}}_{k}(i) + \boldsymbol{\eta}_{k}(i) + \mathbf{n}(i)$$
(5)

where

$$\boldsymbol{\mathcal{C}}_{k} = \begin{bmatrix} \mathbf{C}_{k}^{1} & \mathbf{0} \\ \mathbf{0} & \mathbf{C}_{k}^{2} \end{bmatrix}, \quad \boldsymbol{\bar{\mathcal{C}}}_{k} = \begin{bmatrix} \mathbf{0} & \mathbf{C}_{k}^{2} \\ -\mathbf{C}_{k}^{1} & \mathbf{0} \end{bmatrix}, \quad \mathbf{C}_{k}^{1,2} \in \boldsymbol{\mathcal{R}}^{M}$$

$$\boldsymbol{\mathcal{H}}_{m}(i) = \begin{bmatrix} \mathbf{h}_{k,m}^{1} \\ \mathbf{h}_{k,m}^{2} \end{bmatrix}, \quad \boldsymbol{\eta}_{k}(i) = \begin{bmatrix} \boldsymbol{\eta}_{k,1}(2i-1) \\ \boldsymbol{\eta}_{k,2}(2i) \end{bmatrix}, \quad (7)$$

$$\mathbf{n}(i) = \begin{bmatrix} \mathbf{n}_1(2i-1) \\ \mathbf{n}_2(2i) \end{bmatrix}$$
(8)

The $2M \times 1$ received vectors $\mathbf{y}_m(i)$ are linearly combined with the $2M \times 2$ parameter matrix $\mathbf{W}_{k,m}$ of the desired user k of the *m*th antenna at the receiver end to provide the soft estimates

$$\mathbf{z}_{m}(i) = \mathbf{W}_{k,m}^{H}(i)\mathbf{y}_{m}(i) = [z_{k,m}(2i-1), z_{k,m}(2i)]^{T} \quad (9)$$

By collecting the soft estimates $\mathbf{z}_m(i)$ at each receive antenna, the designer can also exploit the spatial diversity at the receiver as given by

$$\mathbf{z}(i) = \sum_{m=1}^{N_r} \boldsymbol{\alpha}_l(i) \mathbf{z}_l(i)$$
(10)

where $\alpha(i) = diag(\alpha_{l,1}(i), \alpha_{l,2}(i))$ are the gains of the combiner at the receiver, which can be made equal to each other leading to Equal Gain Combining (ECG) or proportional to the channel parameters in accordance with Maximal Ratio Combining (MRC) [27].

III. SPACE-TIME LINEARLY CONSTRAINED RECEIVERS BASED ON THE CCM DESIGN CRITERION

Consider the 2*M*-dimensional received vector at the *m*th receiver $\mathbf{y}(i)$, the $2M \times 2L_p$ constraint matrices \mathcal{C}_k and $\overline{\mathcal{C}}_k$ that were defined in (6) and the $2L_p \times 1$ space-time channel vector $\mathcal{H}_m(i)$ with the multipath components of the unknown channels from Tx1 and Tx2 to the *m*th antenna at the receiver. The space-time linearly constrained receiver design according to the CCM criterion corresponds to determining an FIR filter matrix $\mathbf{W}_{k,m}(i) = \left[\mathbf{w}_{k,m}(i)^T, \bar{\mathbf{w}}_{k,m}(i)^T\right]^T$ with dimension $2M \times 2$ that provides an estimate of the desired symbol at the *m*th antenna of the receiver as given by

$$\hat{\mathbf{b}}_{k}(i) = \operatorname{sgn}\left(\Re\left[\mathbf{W}_{k,m}^{\mathrm{H}}(\mathbf{i})\mathbf{y}_{m}(\mathbf{i})\right]\right) + \operatorname{j}\operatorname{sgn}\left(\Im\left[\mathbf{W}_{k,m}^{\mathrm{H}}(\mathbf{i})\mathbf{y}_{m}(\mathbf{i})\right]\right) \quad \mathbf{x}_{k}(i) = (11)$$

where $sgn(\cdot)$ is the signum function, $\Re(.)$ selects the real component, $\Im(.)$ selects the imaginary component and $\mathbf{W}_{k,m}(i)$ is optimized according to the CM cost functions

$$J_{CM}(\mathbf{w}_{k,m}) = E\left[(|\mathbf{w}_{k,m}^{H}(i)\mathbf{y}_{m}|^{2} - 1)^{2} \right]$$
(12)

$$J_{CM}(\bar{\mathbf{w}}_{k,m}) = E\left[\left(|\bar{\mathbf{w}}_{k,m}^{H}(i)\mathbf{y}_{m}|^{2} - 1\right)^{2}\right]$$
(13)

subject to the set of constraints described by

$$\boldsymbol{\mathcal{C}}_{k}^{H}\mathbf{w}_{k,m}(i) = \nu \ \boldsymbol{\mathcal{H}}_{m}(i), \quad \bar{\boldsymbol{\mathcal{C}}}_{k}^{H}\mathbf{w}_{k,m}(i) = \nu \ \boldsymbol{\mathcal{H}}_{m}^{*}(i) \quad (14)$$

where ν is a constant to ensure the convexity of (12) and (13). Our approach is to consider the parameter vector design problem in (12) and (13) through the optimization of the two filters $\mathbf{w}_{k,m}(i)$ and $\bar{\mathbf{w}}_{k,m}(i)$ in a simultaneous fashion. The optimization of each parameter vector aims to suppress the interference and estimate the symbol transmitted by a given transmit antenna. The expressions for the filters of the spacetime CCM linear receiver are given by

$$\mathbf{w}_{k,m}(i) = \mathbf{R}_{k,m}^{-1}(i) \Big[\mathbf{d}_{k,m}(i) - \mathcal{C}_k (\mathcal{C}_k^H \mathbf{R}_{k,m}^{-1}(i) \mathcal{C}_k)^{-1} \\ \times \left(\mathcal{C}_k^H \mathbf{R}_{k,m}^{-1}(i) \mathbf{d}_{k,m}(i) - \nu \ \mathcal{H}_m(i) \right) \Big]$$
(15)

$$\bar{\mathbf{w}}_{k,m}(i) = \bar{\mathbf{R}}_{k,m}^{-1}(i) \Big[\bar{\mathbf{d}}_{k,m}(i) - \bar{\mathcal{C}}_{k} (\bar{\mathcal{C}}_{k}^{H} \bar{\mathbf{R}}_{k,m}^{-1}(i) \bar{\mathcal{C}}_{k})^{-1} \\ \times \left(\bar{\mathcal{C}}_{k}^{H} \bar{\mathbf{R}}_{k,m}^{-1}(i) \bar{\mathbf{d}}_{k,m}(i) - \nu \ \mathcal{H}_{m}^{*}(i) \right) \Big]$$
(16)

The expressions in (15) and (16) are not closed form as they depend on the previous values of the estimators and require a complexity $O((2M)^3)$ to invert the matrices. Note also that (15) and (16) assume the knowledge of the space-time channel parameters. However, in applications where multipath is present these parameters are not known and thus channel estimation is required. In the next section, we present a method to blindly estimate channels exploiting the STBC structure.

IV. SPACE-TIME CHANNEL ESTIMATION

In this section, we present a framework that exploits the signature sequences of the desired user and the unique structure of STBC for blind channel estimation. Let us consider the received vector $\mathbf{y}_m(i)$ at the *m*th Rx, its associated $2M \times 2M$ covariance matrix $\mathbf{R}_m = E[\mathbf{y}_m(i)\mathbf{y}_m^H(i)]$, the space-time $2M \times 2L_p$ constraint matrices \mathcal{C}_k and $\overline{\mathcal{C}}_k$ given in (6), the space-time channel vector $\mathcal{H}_m(i)$ and from (5) we have that the *k*th user space-time coded received signals without ISI and noise are given by

$$\mathbf{x}_{k}(i) = A_{k}b_{k}(2i-1)\mathcal{C}_{k}\mathcal{H}_{m}(i), \quad \bar{\mathbf{x}}_{k}(i) = A_{k}b_{k}(2i)\bar{\mathcal{C}}_{k}\mathcal{H}_{m}^{*}(i)$$
(17)

Let us perform singular value decomposition (SVD) on the space-time $JM \times JM$ covariance matrix \mathbf{R}_m . By neglecting the ISI, we have:

$$\mathbf{R}_{m} = \sum_{k=1}^{K} E[\mathbf{x}_{k}\mathbf{x}_{k}^{H}] + E[\bar{\mathbf{x}}_{k}\bar{\mathbf{x}}_{k}^{H}] + \sigma^{2}\mathbf{I}$$

$$= [\mathbf{V}_{s} \ \mathbf{V}_{n}] \begin{bmatrix} \mathbf{\Lambda}_{s} + \sigma^{2}\mathbf{I} & \mathbf{0} \\ \mathbf{0} & \sigma^{2}\mathbf{I} \end{bmatrix} [\mathbf{V}_{s} \ \mathbf{V}_{n}]^{H}$$
(18)

where V_s and V_n are the signal and noise subspaces, respectively. Since the signal and noise subspaces are orthogonal, we have the following conditions

$$\mathbf{V}_{n}^{H}\mathbf{x}_{k}(i) = \mathbf{V}_{n}^{H}\boldsymbol{\mathcal{C}}_{k}\boldsymbol{\mathcal{H}}_{m}(i) = \mathbf{0}$$
(19)

$$\mathbf{V}_{n}^{H}\bar{\mathbf{x}}_{k}(i) = \mathbf{V}_{n}^{H}\bar{\boldsymbol{\mathcal{C}}}_{k}\boldsymbol{\mathcal{H}}_{m}^{*}(i) = \mathbf{0}$$
(20)

and hence we have

$$\boldsymbol{\Omega} = \boldsymbol{\mathcal{H}}_m(i)^H \boldsymbol{\mathcal{C}_k}^H \mathbf{V}_n \mathbf{V}_n^H \boldsymbol{\mathcal{C}}_k \boldsymbol{\mathcal{H}}_m(i) = 0 \qquad (21)$$

$$\bar{\boldsymbol{\Omega}} = \boldsymbol{\mathcal{H}}_m^T(i) \bar{\boldsymbol{\mathcal{C}}}_k^H \mathbf{V}_n \mathbf{V}_n^H \bar{\boldsymbol{\mathcal{C}}}_k \boldsymbol{\mathcal{H}}_m^*(i) = 0$$
(22)

From the conditions above and taking into account the conjugate symmetric properties induced by STBC [9], it suffices to consider only Ω , which allows the recovery of $\mathcal{H}_m(i)$ as the eigenvector corresponding to the smallest eigenvalue of the matrix $\mathcal{C}_k^H \mathbf{V}_n \mathbf{V}_n^H \mathcal{C}_k$, provided \mathbf{V}_n is known. In this regard, $\mathcal{H}_m(i)$ belongs to the null space of $\mathcal{C}_k^H \mathbf{V}_n \mathbf{V}_n^H \mathcal{C}_k$ and is a linear combination of all eigenvectors corresponding to eigenvalue zero. If $\mathcal{C}_k^H \mathbf{V}_n \mathbf{V}_n^H \mathcal{C}_k$ has multiple zero-valued eigenvalues, we choose the eigenvalue with smallest index as the solution. To avoid the SVD on \mathbf{R}_m and overcome the need for determining the noise subspace rank that is necessary to obtain \mathbf{V}_n , we resort to the following approach.

Lemma: Consider the SVD on \mathbf{R}_m as in (18), then we have:

$$\lim_{p \to \infty} (\mathbf{R}_m / \sigma^2)^{-p} = \mathbf{V}_n \mathbf{V}_n^H$$
(23)

Proof: Using the decomposition in (18) and since $I + \Lambda_s / \sigma^2$ is a diagonal matrix with elements strictly greater than unity, we have the following limit as $p \to \infty$:

$$(\mathbf{R}_m/\sigma^2)^{-p} = [\mathbf{V}_s \ \mathbf{V}_n] \begin{bmatrix} (\mathbf{I} + \mathbf{\Lambda}_s/\sigma^2)^{-p} & \mathbf{0} \\ \mathbf{0} & \mathbf{I} \end{bmatrix} [\mathbf{V}_s \ \mathbf{V}_n]^H$$
$$\rightarrow \quad [\mathbf{V}_s \ \mathbf{V}_n] \begin{bmatrix} \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{I} \end{bmatrix} [\mathbf{V}_s \ \mathbf{V}_n]^H = \mathbf{V}_n \mathbf{V}_n^H.$$
(24)

To blindly estimate the space-time channel of user k at the mth antenna of the receiver we propose the following optimization:

$$\hat{\mathcal{H}}_{m}(i) = \arg\min_{\mathcal{H}} \quad \mathcal{H}^{H} \mathcal{C}_{k}^{H} \hat{\mathbf{R}}_{m}^{-p}(i) \mathcal{C}_{k} \mathcal{H}$$
(25)

subject to $||\hat{\mathcal{H}}_{m}(i)|| = 1$, where p is an integer, $\hat{\mathbf{R}}_{m}(i)$ is an estimate of the covariance matrix $\mathbf{R}_{m}(i)$ and whose solution is the eigenvector corresponding to the minimum eigenvalue of the $JL_{p} \times JL_{p}$ matrix $\mathcal{C}_{k}^{H}\hat{\mathbf{R}}_{m}^{-p}(i)\mathcal{C}_{k}$ that can be obtained using SVD. The performance of the estimator can be improved by increasing p even though our studies reveal that it suffices to use powers up to p = 2 to obtain a good estimate of $\mathbf{V}_{n}\mathbf{V}_{n}^{H}$. For the space-time block coded CCM receiver design, we employ the matrix $\mathbf{R}_{k,m}$ instead of \mathbf{R}_{m} to avoid the estimation of both \mathbf{R}_{m} and $\mathbf{R}_{k,m}$, and which shows no performance loss as verified in our studies. The computational complexity of the proposed space-time estimator in (25) is $O((2L_{p})^{3})$.

V. BLIND ADAPTIVE SG ALGORITHMS FOR RECEIVER AND CHANNEL PARAMETER ESTIMATION

Here we describe SG algorithms for the blind estimation of the channel and the parameter vector $\mathbf{w}_{k,m}$ of the proposed space-time linear receivers using the CCM criterion.

A. Space-Time Constrained Constant Modulus SG Algorithm

To derive an CCM-SG algorithm let us transform the constrained optimization problem given by (12)-(14) into unconstrained problems in the form of the Lagrangians

$$\mathcal{L}(\mathbf{w}_{k,m}) = (|z_{k,m}[i]|^2 - 1)^2 + 2\Re \Big[(\mathcal{C}_k^H \mathbf{w}_{k,m}[i] - \nu \mathcal{H}_m[i])^H \boldsymbol{\lambda}_1 \Big]$$
(26)
$$\mathcal{L}(\bar{\mathbf{w}}_{k,m}) = (|\bar{z}_{k,m}[i]|^2 - 1)^2 + 2\Re \Big[(\bar{\mathcal{C}}_k^H \bar{\mathbf{w}}_{k,m}[i] - \nu \mathcal{H}_m^*[i])^H \boldsymbol{\lambda}_2 \Big]$$
(27)

where $z_{k,m}[i] = \mathbf{w}_{k,m}^{H}[i]\mathbf{y}_{m}[i]$, $\bar{z}_{k,m}[i] = \bar{\mathbf{w}}_{k,m}^{H}[i]\mathbf{y}_{m}[i]$ and λ_{1} and λ_{2} are vectors of Lagrange multipliers. SG solutions to (26) and (27) can be obtained by taking the gradient terms of (26) and (27) with respect to $\mathbf{w}_{k,m}(i)$ and $\bar{\mathbf{w}}_{k,m}(i)$ which yields the following parameter estimators:

$$\mathbf{w}_{k,m}[i+1] = \mathbf{\Pi}_k(\mathbf{w}_{k,m}[i] - \mu_w e_{k,m}(i) z_{k,m}^*[i]) + \mathbf{C}_k(\mathbf{C}_k^H \mathbf{C}_k)^{-1} \mathcal{H}_m[i]$$
(28)

$$\bar{\mathbf{w}}_{k,m}[i+1] = \bar{\mathbf{\Pi}}_k(\bar{\mathbf{w}}_{k,m}[i] - \mu_w \bar{e}_{k,m}(i) \bar{z}^*_{k,m}[i]) + \bar{\mathbf{C}}_k(\bar{\mathbf{C}}^H_k \bar{\mathbf{C}}_k)^{-1} \bar{\boldsymbol{\mathcal{H}}}^*_m[i]$$
(29)

where $e_{k,m}[i] = (|z_{k,m}[i]|^2 - 1)$, $\bar{e}_{k,m}[i] = (|\bar{z}_{k,m}[i]|^2 - 1)$, $\Pi_k = \mathbf{I} - \mathcal{C}_k (\mathcal{C}_k^H \mathcal{C}_k)^{-1} \mathcal{C}_k^H$, $\bar{\Pi}_k = \mathbf{I} - \bar{\mathcal{C}}_k (\bar{\mathcal{C}}_k^H \bar{\mathcal{C}}_k)^{-1} \bar{\mathcal{C}}_k^H$. The proposed SG algorithm has complexity $O(4ML_p)$ in comparison with that of $O((2M)^3)$, as required by the expressions in (15) and (16).

B. Blind Space-Time SG Channel Estimation Algorithm

In order to estimate the space-time channel and avoid the SVD on $\mathcal{C}_k^H \mathbf{R}_k^{-1}[i] \mathcal{C}_k$, we compute the estimates $\mathbf{\Omega}_m[i] =$

 $\mathbf{C}_{k}^{H} \hat{\mathbf{\Psi}}_{k}[i]$, where $\hat{\mathbf{\Psi}}_{k}[i]$ is an estimate of the matrix $\mathbf{R}_{m}^{-1}[i]\mathbf{C}_{k}$. The estimate $\hat{\mathbf{\Psi}}_{k}[i]$ is obtained with the following recursion:

$$\hat{\boldsymbol{\Psi}}_{m}[i] = \alpha \hat{\boldsymbol{\Psi}}_{m}[i-1] + \mu_{h} \Big(\hat{\boldsymbol{\Psi}}_{k}[i-1] - \mathbf{y}_{m}[i] \mathbf{y}_{m}^{H}[i] \hat{\boldsymbol{\Psi}}_{m}[i-1] \Big)$$
(30)

where $\hat{\Psi}_m(0) = C_k$ and $0 < \alpha < 1$. To estimate the spacetime channel and avoid the SVD on $C_k^H \mathbf{R}_k^{-1}[i] C_k$, we employ the variant of the power method introduced in [21]

$$\hat{\mathcal{H}}_m(i) = (\mathbf{I} - \gamma(i)\mathbf{\Omega}_m[i])\hat{\mathcal{H}}_m(i-1)$$
(31)

where $\gamma(i) = 1/tr[\Gamma_{k,m}(i)]$ and we make $\hat{\mathcal{H}}_m(i) \leftarrow \hat{\mathcal{H}}_m(i)/||\hat{\mathcal{H}}_m(i)||$ to normalize the channel. The proposed space-time channel estimation algorithm has complexity $O(4ML_p)$ and shows excellent performance.

VI. SIMULATIONS

We evaluate the bit error rate (BER) performance of the proposed blind space-time block coded receivers based on the CCM design, the proposed space-time channel estimation method in terms of mean squared error (MSE) performance and the SG adaptive algorithms. We also compare the proposed space-time CCM linear receivers and blind channel estimation algorithms with some previously reported techniques, namely, the constrained minimum variance (CMV) with a single antenna [11] and with STBC [9] and the subspace receiver of Wang and Poor without [24] and with STBC [10]. The DS-CDMA system employs randomly generated spreading sequences of length N = 32, employs one or two transmit antennas with the Alamouti STBC [2] and employs one receive antenna with MRC. All downlink channels assume that $L_p = 6$ as an upper bound. We use three-path channels with relative powers $p_{l,m}^{1,2}$ given by 0, -3 and -6 dB, with path-spacing given by a discrete uniform random variable between 1 and 2 chips. The sequence of channel coefficients for each transmit antenna $n_t = 1,2$ and each receive antenna m = 1,2 is $h_{l,m}^{n_t}(i) = p_{l,m}^{n_t} \alpha_{l,m}^{n_t}(i) \ (l = 0, 1, 2, \dots),$ where $\alpha_{l,m}^{n_t}(i)$, is obtained with Clarke's model [27].

In the first scenarios, shown in Figs. 2 and 3, we evaluate the BER convergence performance of the proposed SG algorithms for both receiver and channel parameter estimation. We consider a system with 8 users, the power level distribution among the interferers follows a log-normal distribution with associated standard deviation of 3 dB and the desired user power level corresponds to the signal-to-noise-ratio (SNR) defined by $SNR = E_b/N_0 = 15$ dB. After 1500 symbols, 6 additional users enter the system and the power level distribution among interferers is loosen with associated standard deviation being increased to 6 dB. The results, shown in Fig. 2, indicate that the proposed CCM-based SG algorithms outperform the existing blind algorithms (CMV-based and Subspace) and approach the performance of supervised algorithms. The results for channel estimation, depicted in Fig. 3, show that the proposed STBC-based SG algorithm which equips both CCM-based and CMV-based schemes outperforms the singleantenna method of [21] and approaches the highly complex subspace approach of [10].



Fig. 2. BER performance versus number of received symbols .



Fig. 3. MSE channel estimation performance versus number of symbols .

The BER performance versus SNR and number of users is illustrated in Fig. 4. In these scenarios, we considered data packets of P = 1500 symbols, 2 transmit antennas, 1 receive antenna and measured the BER after 200 independent transmissions. The plots indicate that the proposed STBC-CCM receiver achieves the best performance, followed by the subspace receiver with STBC of [10] and the STBC-CMV. A substantial capacity increase and performance improvement is verified for the schemes with multiple antennas.

VII. CONCLUSIONS

We have presented in this work blind adaptive linear multiuser receivers for DS-CDMA systems using multiple



Fig. 4. BER performance versus (a) E_b/N_0 and (b) number of users.

transmit antennas and space-time block codes (STBC) in multipath channels. We considered a CCM design criterion based on constrained optimization techniques and described low-complexity SG adaptive algorithms for estimating the parameters of the linear receivers. The receiver was designed in order to exploit the unique structure imposed by both spreading codes and STBC. We also developed a blind spacetime channel estimation scheme for STBC systems based on the subspace approach along with an efficient SG algorithm for channel estimation. Numerical results for a downlink scenario have shown that the proposed techniques outperform existing schemes in realistic scenarios.

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