A Beamspace–Time Interference Canceling CDMA Receiver for Sectored Communications in a Multipath Environment

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***Abstract*—A beamspace–time (BT) receiver is proposed for interference suppression and multipath diversity reception in sectored wireless code division multiple access communications. The scheme involves two stages. First, a set of adaptive space–time diversity processors, in the form of beamformer-correlator pairs, is constructed which provides effective suppression of unwanted interference and reception of signals from a prescribed space–time region. Second, the output data obtained by these processors are maximum ratio combined to capture the signal multipath components coherently. The proposed BT receiver is blind in that no training signal is required. The only information required is the signature sequence, timing and a rough estimate of the angle of arrival of the signal for selecting the sector of interest.**

***Index Terms*—Antenna arrays, beamforming, CDMA, interference cancellation, RAKE receiver.**

I. INTRODUCTION

MULTIPLE ACCESS INTERFERENCE (MAI), multipath fading, and noise are limiting factors of a code division multiple access (CDMA) system. MAI causes the near–far effect and multipath induces intersymbol interference (ISI). Advanced signal processing algorithms have been developed to combat these problems. In particular, multi user detectors have been proposed which exhibit immunity to the near–far effect [1]. Also, the RAKE receiver [2] is introduced as a means of coherent combining of multipath signals. Conventional RAKE receivers are nonadaptive and nonblind in that the correlator bank is fixed, and a training sequence is used for channel estimation.

In general, the multipath signals arrive at the receiver with different angles of arrival (AOA). To exploit this spatial signature, an antenna array is employed. By using the extra degree of freedom offered by the antenna array, MAI and narrowband interference (NBI) that are spatially separated from the signal-of-interest (SOI) can be suppressed. A receiver that exploits both the temporal and spatial structures to coherently combine the multipath energy is referred to as the space–time (ST) RAKE receiver [3]–[6]. Adaptive interference cancellation can be incorporated as an option via criteria such as minimum mean-square-error (MMSE) [4], [5]. Most ST RAKE receivers work with the post-despread data obtained by a bank of fixed correlators, and require a training sequence for ST channel estimation.

Blind ST RAKE receivers have been proposed which do not require a training sequence. The pioneering work proposed in [3] exploits the distinctive structures of the pre-despread and post-despread spatial correlation matrices to estimate the spatial channels at different fingers of the SOI via the principal component (PC) eigenvalue decomposition. Beamforming is then performed at different fingers, followed by a post RAKE combining. In [4], the PC based scheme is extended to perform space–time joint channel estimation and postcombining. The receivers in [3], [4], [6] work with the outputs of a bank of fixed correlators, and are, thus, post-despread receivers. They are simple to implement but do not fully exploit the temporal degree of freedom available to handle strong MAI. As an alternative, a receiver which works directly with the chip-sampled data is called a pre-despread receiver, and can be regarded as an adaptive matched filter that performs despreading and MAI suppression simultaneously [7]. A popular blind pre-despread receivers is the minimum output energy (MOE) receiver [8]. It is similar in structure to the generalized sidelobe canceller (GSC) in array processing [9], and is shown to offer the performance of the MMSE receiver without the aid of a training sequence. However, blind adaptive receivers are sensitive to mismatch in the SOI's signature due to the signal cancellation effect [7]. To alleviate this, auxiliary processing is introduced [8]. The blind pre-despread receivers can be extended to the ST domain with an antenna array. In [10], an ST pre-despread receiver is proposed which consists of a novel adaptive algorithm for implementing adaptive matched filters in GSC form, a maximum likelihood AOA estimator for SOI location, and a pilot-assisted RAKE receiver for postcombining.

In this paper, a blind adaptive beamspace- (BT) pre-despread receiver is proposed for sectored CDMA communications. The receiver is designed with the following procedure. First, a set of ST diversity processors, in the form of beamformer-correlator (BC) pairs is constructed to collect the multipath signals in a prescribed angular sector and time duration, and to suppress MAI and NBI. This is done by performing adaptive nulling on a set of linearly constrained minimum variance (LCMV) ST processors "steered" to different look directions and delays. To avoid signal cancellation incurred with the mismatch of spatial–temporal signatures, a modified GSC is employed. Second, the outputs of these BC pairs corresponding to different look directions and delays are coherently combined to collect the multi-

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path energy. The beamformers and correlators together constitute a beamspace–time (BT) receiver which operates without the aid of a training signal. The only information required is a rough estimate of the AOA of the SOI for sector selection. Compared to conventional ST CDMA receivers [3]–[6], [10], in which spatial processing is done by direct beamforming, the proposed receiver performs "beamspace" beamforming in which array data are first pre-processed by a set of diversity beamformers encompassing an angular sector. This has two advantages. First, with sectorization, the capacity of the system can be potentially increased. Second, MAI and NBI from outside the sector can be suppressed adaptively, leading to improved reception quality for the SOI. As for adaptive correlators in GSC form, the proposed receiver is similar to that of [10]. A major distinction, however, is that the proposed method works with a temporal blocking matrix designed to eliminate all multipath signals instead of just a specific path. This can avoid signal cancellation incurred with coherent multipaths. In particular, a simple and effective partially adaptive implementation is suggested to reduce the adaptive weights dimension. Furthermore, with sectorization, only a rough AOA estimator (to determine the right sector) is needed for beamforming. In contrast, the method in [10] relies on ML AOA estimation which requires a small multipath angle spread. In summary, the proposed receiver is designed to combat strong interference by adaptive spatial preprocessing (diversity beamforming), to incorporate pre-despread temporal processing to enhance suppression of in-sector MAI, and to handle multipaths by spatial–temporal postprocessing [maximum ratio combining(MRC)].

The paper is organized as follows. Section II presents the data model for ST receivers. In Section III, a blind BT receiver using the LCMV criterion and GSC technique is developed. Section IV discusses the issues in implementation. Finally, Section V gives the simulation results and Section VI concludes the paper.

II. Data Model and ST RAKE Receivers

Suppose that there are Q active DS-CDMA users in the cell. The ith user's contribution to the received signal is given by

$$r_i(t) = \sqrt{P_i}\sum_k d_i(k)s_i(t-kT_s) \tag{1}$$

where
- P_i power;
- $d_i(k)$ kth transmitted information symbol;
- T_s symbol duration;
- $s_i(t)$ signature waveform given by

$$s_i(t) = \sum_{n=1}^{N} c_i[n]p(t-nT_c) \tag{2}$$

where
- $c_i[n]$ signature sequence (or spreading sequence) of the ith user;
- N spreading factor;
- $p(t)$ chip waveform;
- T_c chip duration.

Without loss of generality, we assume $c_i[n]$ to be complex valued. With a D-antenna array at the basestation, the reverse-link baseband complex data can be expressed in the vector form

$$\mathbf{x}(t) = \sum_{i=1}^{Q}\sqrt{P_i}\sum_{j=1}^{J}\alpha_{i,j}\mathbf{a}(\theta_{i,j})r_i(t-\tau_{i,j}) + \mathbf{i}_{NB}(t) + \mathbf{n}(t) \tag{3}$$

where
- J path number (assumed the same for all users);
- $\theta_{i,j}, \tau_{i,j}$ and $\alpha_{i,j}$ AOA, delay, and complex gain, respectively, of the jth path of the ith user;
- $\mathbf{a}(\theta)$ spatial signature vector (or steering vector) accounting for the gain/phase variation across the array aperture due to a source from θ;
- $\mathbf{i}_{NB}(t)$ narrowband interference vector;
- $\mathbf{n}(t)$ noise vector whose entries are independent, identically distributed (i.i.d.) complex Gaussian random variables with variance σ_n^2.

To fully exploit the temporal signature of the multipath channel, $\mathbf{x}(t)$ is chip matched filtered and then chip rate sampled at $t=(k-1)T_s+(n-1)T_c+T_c/2$ over the kth symbol duration, i.e., $n=1,\ldots,N+L-1$, where L is the number of RAKE fingers or the effective maximum delay spread of the channel. Assuming user 1 to be the desired signal, the pre-despread ST (pre-ST) data received at the array after chip-rate sampling can be written as a $D\times(N+L-1)$ matrix

$$\begin{aligned}\mathbf{X}(k) &= [\mathbf{x}_1(k),\mathbf{x}_2(k),\ldots,\mathbf{x}_{N+L-1}(k)]\\ &= \sqrt{P_1}\sum_{j=1}^{J}\alpha_{1,j}\mathbf{a}(\theta_{1,j})\mathbf{c}_{\tau_{1,j}}^T d_1(k) + \mathbf{I}(k) + \mathbf{N}(k)\\ &= \mathbf{H}_1 d_1(k) + \mathbf{I}(k) + \mathbf{N}(k)\end{aligned} \tag{4}$$

where
- $\mathbf{x}_n(k)$ $= \mathbf{x}(t)|_{t=(k-1)T_s+(n-1)T_c+T_c/2}$;
- $\mathbf{c}_{\tau_{1,j}}$ augmented temporal signature vector of the jth path of user 1;
- $(\cdot)^T$ transpose.

Depending on the delay, $\tau_{1,j}, \mathbf{c}_{\tau_{1,j}}$ is given by one of the columns of the $(N+L-1)\times L$ augmented temporal signature matrix

$$\mathbf{S}_1 = \begin{bmatrix} c_1[1] & 0 & \cdots & 0\\ \vdots & c_1[1] & \ddots & \vdots\\ c_1[N] & \vdots & \ddots & 0\\ 0 & c_1[N] & \ddots & c_1[1]\\ \vdots & \vdots & \ddots & \vdots\\ 0 & 0 & \cdots & c_1[N]\end{bmatrix} \tag{5}$$

$\mathbf{I}(k)$ is the pre-ST interference matrix consisting of ISI, MAI and NBI, and $\mathbf{N}(k)$ is the pre-ST white noise matrix. From (4), $\mathbf{H}_1$ is the composite pre-ST signature matrix of user 1 given by

$$\mathbf{H}_1 = \sqrt{P_1}\sum_{j=1}^{J}\alpha_{1,j}\mathbf{a}(\theta_{1,j})\mathbf{c}_{\tau_{1,j}}^T \tag{6}$$

A receiver for user 1 is designed to identify and remove $\mathbf{H}_1$ to retrieve the symbol $d_1(k)$ from $\mathbf{I}(k)$ and $\mathbf{N}(k)$.

A variation of the pre-ST data matrix suitable for the ST RAKE receiver is the $D \times L$ post-despread ST (post-ST) data matrix

$$\tilde{\mathbf{X}}(k) = \mathbf{X}(k)\mathbf{S}_1^* = \tilde{\mathbf{H}}_1 d_1(k) + \tilde{\mathbf{I}}(k) + \tilde{\mathbf{N}}(k) \quad (7)$$

where

$\tilde{\mathbf{H}}_1 =$ $\mathbf{H}_1\mathbf{S}_1^*$;
$\tilde{\mathbf{I}}(k) =$ $\mathbf{I}(k)\mathbf{S}_1^*$;
$\tilde{\mathbf{N}}(k) =$ $\mathbf{N}(k)\mathbf{S}_1^*$;
$(\cdot)^*$ complex conjugate.

Note that $\tilde{\mathbf{X}}(k)$ can be viewed as the time-compressed version of $\mathbf{X}(k)$. As a space domain counterpart, we can form the $M \times (N+L-1)$ pre-despread beamspace–time (pre-BT) data matrix as follows:

$$\hat{\mathbf{X}}(k) = \mathbf{W}^H\mathbf{X}(k) = \hat{\mathbf{H}}_1 d_1(k) + \hat{\mathbf{I}}(k) + \hat{\mathbf{N}}(k) \quad (8)$$

where

$\mathbf{W}$ $D \times M$ beamforming matrix;
$\hat{\mathbf{H}}_1 =$ $\mathbf{W}^H\mathbf{H}_1$;
$\hat{\mathbf{I}}(k) =$ $\mathbf{W}^H\mathbf{I}(k)$;
$\hat{\mathbf{N}}(k) =$ $\mathbf{W}^H\mathbf{N}(k)$;
$(\cdot)^H$ complex conjugate transpose.

With $M < D, \hat{\mathbf{X}}(k)$ can be viewed as the space-compressed version of $\mathbf{X}(k)$. Finally, we can form the $M \times L$ post-despread BT (post-BT) data matrix:

$$\bar{\mathbf{X}}(k) = \mathbf{W}^H\mathbf{X}(k)\mathbf{S}_1^* = \bar{\mathbf{H}}_1 d_1(k) + \bar{\mathbf{I}}(k) + \bar{\mathbf{N}}(k) \quad (9)$$

where

$\bar{\mathbf{H}}_1 =$ $\mathbf{W}^H\mathbf{H}_1\mathbf{S}_1^*$;
$\bar{\mathbf{I}}(k) =$ $\mathbf{W}^H\mathbf{I}(k)\mathbf{S}_1^*$;
$\bar{\mathbf{N}}(k) =$ $\mathbf{W}^H\mathbf{N}(k)\mathbf{S}_1^*$.

Various types of receivers can be developed based on different versions of the ST data. For example, the ST RAKE receiver combines the entries of the post-ST data matrix into an estimate of $d_1(k)$

$$d_1(k) = \operatorname{dec}\{\tilde{\mathbf{q}}^H\tilde{\mathbf{x}}(k)\} \quad (10)$$

where

$\tilde{\mathbf{q}}$ combining weight vector;
$\tilde{\mathbf{x}}(k)$ $DL \times 1$ vector obtained by concatenating the columns of $\tilde{\mathbf{X}}(k)$;
$\operatorname{dec}\{\cdot\}$ decision operator.

The weight vector can be chosen in accordance with the coherent or MMSE criterion [4]. For coherent combining, $\tilde{\mathbf{q}} = \tilde{\mathbf{h}}_1$, and for MMSE combining, $\tilde{\mathbf{q}} = \tilde{\mathbf{R}}^{-1}\tilde{\mathbf{h}}_1$, where $\tilde{\mathbf{R}} = E\{\tilde{\mathbf{x}}(k)\tilde{\mathbf{x}}^H(k)\}$ is the post-ST data correlation matrix, and $\tilde{\mathbf{h}}_1$ is the post-ST channel vector obtained by concatenating the columns of $\tilde{\mathbf{H}}_1$. Similarly, the BT RAKE receiver combines the entries of the post-BT data matrix into an estimate of $d_1(k)$

$$d_1(k) = \operatorname{dec}\{\bar{\mathbf{q}}^H\bar{\mathbf{x}}(k)\} \quad (11)$$

where $\bar{\mathbf{q}}$ is the combining weight vector and $\bar{\mathbf{x}}(k)$ is the $ML \times 1$ vector obtained by concatenating the columns of $\bar{\mathbf{X}}(k)$. For coherent combining, $\bar{\mathbf{q}} = \bar{\mathbf{h}}_1$, and for MMSE combining, $\bar{\mathbf{q}} = \bar{\mathbf{R}}^{-1}\bar{\mathbf{h}}_1$, where $\bar{\mathbf{R}} = E\{\bar{\mathbf{x}}(k)\bar{\mathbf{x}}^H(k)\}$ is the post-BT data correlation matrix, and $\bar{\mathbf{h}}_1$ is the post-BT channel vector obtained by concatenating the columns of $\bar{\mathbf{H}}_1$. Note that the channel vectors used in the above receivers need be estimated using a training sequence or by other blind methods. Working with a training sequence gives better channel estimation, but will inevitably decrease the system's efficiency. This prompts the development of blind receivers[6].

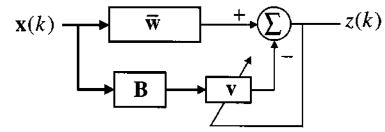

Fig. 1. Structure of GSC.

III. Proposed BT Receiver

In this section, a blind BT receiver suitable for sectored systems is developed with a two-stage procedure. First, an adaptive ST processor is designed for each beam-finger pair which transforms $\mathbf{X}(k)$ from the ST domain into a scalar. Second, the outputs of these ST processors are constructively combined to extract the SOI's symbols.

A. Construction of ST Processors

In sectored communications, the entire field-of-view of the receiver is divided into several angular sectors, each responsible for a subset of users [2]. Assume that the AOAs and delays of the multipaths of the SOI (user 1) are roughly known such that an angular sector Θ_S and a time duration T_D can be chosen to accommodate these multipaths. In order to effectively collect the multipath energy, a set of M diversity beams is formed with their patterns encompassing Θ_S, and a set of L diversity correlators is formed with their responses matched to the L fingers in T_D. This leads to a set of ML beamformer-correlator (BC) pairs, or ST processors, operating on the pre-ST data matrix. Let $\{\theta_m, m = 1, \ldots, M\}$ be the set of angles well representing Θ_S and denote as $\mathbf{a}_m = \mathbf{a}(\theta_m)$. Also let $\mathbf{s}_l$ be the lth column of $\mathbf{S}_1$. The ST processor transforms the pre-ST data matrix into a scalar output

$$\bar{z}_{m,l}(k) = \mathbf{a}_m^H\mathbf{X}(k)\mathbf{s}_l^* \quad (12)$$

with $m = 1, \ldots, M$ and $l = 1, \ldots, L$.

To ensure an effective suppression of strong interference, adaptive cancellation is performed for each of the ST diversity processors. An effective solution is to employ the scheme of GSC [9]. The GSC is an indirect but simpler implementation of the LCMV beamformer. Its concept, as depicted in Fig. 1, is to decompose the weight vector $\mathbf{w}$ into: $\mathbf{w} = \bar{\mathbf{w}} - \mathbf{B}\mathbf{v}$, where $\bar{\mathbf{w}}$ is the fixed weight vector and $\mathbf{B}$ is the blocking matrix which removes the signal before beamforming. The goal is to choose the adaptive weight vector $\mathbf{v}$ to cancel the interference. To apply the GSC concept for ST processors, some modifications should be made. First, instead of blocking signals with a specific direction or delay, the blocking matrix must remove signals from the entire Θ_S and T_D. Second, instead of using a different blocking matrix for each ST processor, the same matrix is shared by all of the ML beam-fingers. Let $\mathbf{v}_{m,l}$ and $\mathbf{g}_{m,l}$ be the spatial and temporal adaptive weight vectors, respectively. The GSC formulation leads to the following ST operation:

$$\begin{aligned} z_{m,l}(k) &= (\mathbf{a}_m - \tilde{\mathbf{B}}\mathbf{v}_{m,l})^H \mathbf{X}(k)(\mathbf{s}_l - \hat{\mathbf{B}}\mathbf{g}_{m,l})^* \\ &= h_{m,l}d_1(k) + i_{m,l}(k) + n_{m,l}(k) \end{aligned} \tag{13}$$

where $\tilde{\mathbf{B}}$ and $\hat{\mathbf{B}}$ are the spatial and temporal blocking matrices, respectively, and

$$h_{m,l} = (\mathbf{a}_m - \tilde{\mathbf{B}}\mathbf{v}_{m,l})^H \mathbf{H}_1(\mathbf{s}_l - \hat{\mathbf{B}}\mathbf{g}_{m,l})^* \tag{14}$$

$$i_{m,l}(k) = (\mathbf{a}_m - \tilde{\mathbf{B}}\mathbf{v}_{m,l})^H \mathbf{I}(k)(\mathbf{s}_l - \hat{\mathbf{B}}\mathbf{g}_{m,l})^* \tag{15}$$

$$n_{m,l}(k) = (\mathbf{a}_m - \tilde{\mathbf{B}}\mathbf{v}_{m,l})^H \mathbf{N}(k)(\mathbf{s}_l - \hat{\mathbf{B}}\mathbf{g}_{m,l})^* \tag{16}$$

Following the standard procedure of GSC, the adaptive weight vectors are determined by solving the MMSE problem

$$\begin{aligned} &\min_{\mathbf{v}_{m,l},\mathbf{g}_{m,l}} E\left\{|z_{m,l}(k)|^2\right\} \\ &\equiv E\left\{|(\mathbf{a}_m - \tilde{\mathbf{B}}\mathbf{v}_{m,l})^H \mathbf{X}(k)(\mathbf{s}_l - \hat{\mathbf{B}}\mathbf{g}_{m,l})^*|^2\right\} \end{aligned} \tag{17}$$

The ST processors in (13) utilizes the spatial and temporal degrees of freedom in a separate fashion. This is in contrast to receivers operating in a ST joint fashion. ST joint receivers are superior in terms processing dimension, but require a much higher complexity. For scenarios in which there are only few dominant interferers, ST separate receivers would be a more efficient solution.

The closed-form solution to (17) is not available, and an suboptimum solution can be obtained by keeping the spatial weight vector fixed and solving for the temporal weight vector, and vice versa. This leads to the development of the following alternate minimization procedure. First, the beamformer weight vector

$$\tilde{\mathbf{a}}_{m,l} = \mathbf{a}_m - \tilde{\mathbf{B}}\mathbf{v}_{m,l} \tag{18}$$

is initialized with a predetermined value (e.g., $\tilde{\mathbf{a}}_{m,l} = \mathbf{a}_m$), and the temporal adaptive weight vector is obtained in accordance with

$$\min_{\mathbf{g}_{m,l}} E\left\{|(\mathbf{s}_l - \hat{\mathbf{B}}\mathbf{g}_{m,l})^H \mathbf{X}^T(k)\tilde{\mathbf{a}}^*_{m,l}|^2\right\} \tag{19}$$

whose solution is given by

$$\mathbf{g}_{m,l} = \hat{\mathbf{R}}^{-1}_{m,l}\hat{\mathbf{r}}_{m,l} \tag{20}$$

where

$$\begin{aligned} \hat{\mathbf{R}}_{m,l} &= E\left\{\hat{\mathbf{B}}^H \mathbf{X}^T(k)\tilde{\mathbf{a}}^*_{m,l}\tilde{\mathbf{a}}^T_{m,l}\mathbf{X}^*(k)\hat{\mathbf{B}}\right\} \\ \hat{\mathbf{r}}_{m,l} &= E\left\{\hat{\mathbf{B}}^H \mathbf{X}^T(k)\tilde{\mathbf{a}}^*_{m,l}\tilde{\mathbf{a}}^T_{m,l}\mathbf{X}^*(k)\mathbf{s}_l\right\} \end{aligned} \tag{21}$$

The correlator weight vector

$$\hat{\mathbf{s}}_{m,l} = \mathbf{s}_l - \hat{\mathbf{B}}\mathbf{g}_{m,l} \tag{22}$$

with $\mathbf{g}_{m,l}$ given by (20), can suppress MAI with a fixed temporal signature. For time-varying NBI whose temporal signatures are not well defined, adaptive nulling in the spatial domain is more effective. The second step is then to fix $\hat{\mathbf{s}}_{m,l}$, and obtain the spatial adaptive weight vector in accordance with

$$\min_{\mathbf{v}_{m,l}} E\left\{|(\mathbf{a}_m - \tilde{\mathbf{B}}\mathbf{v}_{m,l})^H \mathbf{X}(k)\hat{\mathbf{s}}^*_{m,l}|^2\right\} \tag{23}$$

whose solution is given by

$$\mathbf{v}_{m,l} = \tilde{\mathbf{R}}^{-1}_{m,l}\tilde{\mathbf{r}}_{m,l} \tag{24}$$

where

$$\begin{aligned} \tilde{\mathbf{R}}_{m,l} &= E\left\{\tilde{\mathbf{B}}^H \mathbf{X}(k)\hat{\mathbf{s}}^*_{m,l}\hat{\mathbf{s}}^T_{m,l}\mathbf{X}^H(k)\tilde{\mathbf{B}}\right\} \\ \tilde{\mathbf{r}}_{m,l} &= E\left\{\tilde{\mathbf{B}}^H \mathbf{X}(k)\hat{\mathbf{s}}^*_{m,l}\hat{\mathbf{s}}^T_{m,l}\mathbf{X}^H(k)\mathbf{a}_m\right\} \end{aligned} \tag{25}$$

The above procedure can be iterated between (19) and (23) by substituting (24) in (18), and re-executing the steps in (19)–(25). Since both the spatial adaptive weight vector and temporal adaptive weight vector are adjusted to minimize the cost function of (17), and both are orthogonal to the desired signal, the minimization of cost function is solely due to the suppression of interference. In particular, at each alternate iteration, either the spatial or temporal weight vector will adjust itself to decrease the cost function with the other fixed. As a result, it can be sure that the cost function will reduce as iterations proceed. Nevertheless, the resulting solution may not lead to the global minimum of the cost function. However, as long as the degree of freedom of the adaptive weights are large enough compared to the number of interferers, the alternate procedure can always yield a converged solution that provides effective interference suppression. The proposed alternate iterative scheme is similar in principle to the iterative least square (ILS) problem described in [12]. Given $\tilde{\mathbf{a}}_{m,l}$ fixed, the minimization of cost function (17) is a least-squares problem for $\mathbf{g}_{m,l}$ to make the cost function approach zero. A similar situation holds for $\mathbf{v}_{m,l}$ with $\hat{\mathbf{s}}_{m,l}$ fixed.

Note that the above ST processors is developed based on the assumption of short spreading codes. Although the adaptive beamformers can be used with long codes, the adaptive correlators cannot. This is because that the adaptive correlators are designed to respond to fixed temporal signatures (composite spreading codes due to multipaths). Only with fixed temporal signatures can the adaptive correlators "recognize" and suppress strong MAI.

B. AOA Estimation and Sector Selection

To determine the working sector, some kind of preliminary location method can be used to obtain a coarse estimate the AOA of the SOI. The estimate need not be accurate as required in conventional LCMV beamforming but should lead to a right sector for the multipath arrivals. In this section, a simple technique of AOA estimation is proposed based on the same framework for adaptive beamforming.

For simplicity, we assume that the SOI is the the sector Θ_S. The AOA estimate of the lth path of the SOI is determined as the solution to the following spectral search problem

$$\max_{\theta} P_l(\theta) \equiv \mathbf{w}_l(\theta)^H \tilde{\mathbf{R}}_l \mathbf{w}_l(\theta) \tag{26}$$

where

$$\tilde{\mathbf{R}}_l = E\left\{\mathbf{X}(k)\mathbf{s}_l^*\mathbf{s}_l^T\mathbf{X}^H(k)\right\} \tag{27}$$

is the post-despread correlation matrix for the lth finger, and

$$\mathbf{w}_l(\theta) = [\mathbf{I} - \tilde{\mathbf{B}}(\tilde{\mathbf{B}}^H\tilde{\mathbf{R}}_l\tilde{\mathbf{B}})^{-1}\tilde{\mathbf{B}}^H\tilde{\mathbf{R}}_l]\mathbf{a}(\theta) \tag{28}$$

is the optimum beamforming weight vector steered to the look direction θ. It is the weight vector obtained from (23), with $\mathbf{a}_m$ and $\hat{\mathbf{s}}_{m,l}$ replaced by $\mathbf{a}(\theta)$ and $\mathbf{s}_l$, respectively. For any $\theta \in \Theta_S$, $\mathbf{w}_l(\theta)$ will produce a mainlobe at θ, and suppress interference outside Θ_S. Hence, by varying θ over Θ_S, a spectral peak can be observed in the spatial spectrum $P_l(\theta)$ corresponding to the AOA of the lth path. This is similar in principle to the Capon's location method [11], except that a different blocking scheme is used here to avoid signal cancellation. The same procedure can be executed for different sectors to locate all spectral peaks in the entire field-of-view. If multiple peaks are present in different sectors, then further identification is necessary to distinguish the true SOI from others.

C. Selection of Blocking Matrices

The selection of blocking matrices are based on two criteria: 1) to remove the signal in the data so as to avoid signal cancellation and 2) to leave a sufficient degree of freedom for interference suppression. For temporal blocking, since the processing gain is large and multipath timing is assumed known, $\hat{\mathbf{B}}$ can be constructed to remove signals with specific delays within the entire delay spread duration. On the other hand, since the number of antennas is small for spatial blocking, and the signal AOA is poorly defined, $\tilde{\mathbf{B}}$ should be chosen to provide a "fuzzy" blocking effect.

The temporal blocking matrix $\hat{\mathbf{B}}$ can be chosen to be a full rank $(N + L - 1) \times (N - 1)$ matrix whose columns are orthogonal to $\{\mathbf{s}_1, \mathbf{s}_2, \ldots, \mathbf{s}_L\}$, i.e., $\hat{\mathbf{B}}_l^H\mathbf{S}_1 = \mathbf{O}$. For a more reliable operation, extra vectors can be included to extend the blocking interval to a larger delay spread. This will help to avoid possible signal cancellation due to undetected multipath arrivals. For example, two extra vectors $\mathbf{s}_1^{(-)} = [c_1[2], c_1[3], \ldots, c_1[N], 0, \ldots, 0]^T$ and $\mathbf{s}_L^{(+)} = [0, \ldots, 0, c_1[1], c_1[2], \ldots, c_1[N-1]]^T$ can be included to extend by one chip duration at both ends of the blocking interval.

Let D' be the dimension of $\mathbf{v}_{m,l}$. The spatial blocking matrix $\tilde{\mathbf{B}}$ can be chosen to be a full rank $D \times D'$ matrix whose columns are orthogonal to a set of $D - D'$ steering vectors well representing the Θ_S. This is called the direction constrained method. An alternative scheme more suitable for small arrays is the eigenvector constrained (EC) design suggested in [14]. In this design, a dense set of angles $\theta_n, n = 1, \ldots, N_t$, is chosen to represent Θ_S. A matrix is then formed by

$$\mathbf{R}_a = \frac{1}{N_t}\sum_{n=1}^{N_t}\mathbf{a}(\theta_n)\mathbf{a}^H(\theta_n) = \sum_{i=1}^{D}\lambda_i\mathbf{e}_i\mathbf{e}_i^H \tag{29}$$

where $\{\lambda_i\}$ and $\{\mathbf{e}_i\}$ are the eigenvalues and eigenvectors of $\mathbf{R}_a$ in descending order. Depending on the array size and sector width, there will be $D - D'$ dominant modes (with large eigenvalues) in (29) which effectively represent Θ_S. The blocking matrix $\tilde{\mathbf{B}}$ can be chosen to consist of the remaining D' eigenvectors associated with those smaller eigenvalues

$$\tilde{\mathbf{B}} = [\mathbf{e}_{D-D'+1}, \mathbf{e}_{D-D'+2}, \ldots, \mathbf{e}_D] \tag{30}$$

Due to the orthogonality among $\mathbf{e}_i$s, the so constructed $\tilde{\mathbf{B}}$ should satisfy $\tilde{\mathbf{B}}^H\mathbf{a}(\theta) \approx 0$ for $\theta \in \Theta_S$. The EC method is shown to provide a better blocking effect due to its "fuzzy" nature stated earlier. The choosing of D' is a trade-off between the blocking effect and adaptive nulling. A small D' leads to better blocking, but poor nulling, and vice versa. A useful criterion that has been confirmed by numerical results is $D' \approx D/2$.

D. Maximum Ratio Combiner

It is assumed that, with adaptive ST processing, the dominant interference has been eliminated and the BC output data $z_{m,l}(k), m = 1, \ldots, M, l = 1, \ldots, L$, contain the SOI and colored noise only. This suggests that the MRC criterion can be employed to collect these components coherently to extract $d_1(k)$. Let $\mathbf{f}$ be the $ML \times 1$ weight vector that performs the combining

$$z_o(k) = \mathbf{f}^H\mathbf{z}_c(k) \tag{31}$$

where

$$\begin{aligned}\mathbf{z}_c(k) &= [z_{1,1}(k), \ldots, z_{M,1}(k), z_{1,2}(k), \ldots, z_{M,2}(k), \ldots,\\ &\qquad z_{1,L}(k), \ldots, z_{M,L}(k)]^T\\ &\approx \mathbf{h}_c d_1(k) + \mathbf{n}_c(k)\end{aligned} \tag{32}$$

is the BC output data vector associated with the kth symbol, with $\mathbf{h}_c$ and $\mathbf{n}_c(k)$ being the corresponding composite signature and noise vectors, respectively, given by

$$\mathbf{h}_c = [h_{1,1}, \ldots, h_{M,1}, h_{1,2}, \ldots, h_{M,2}, \ldots, h_{1,L}, \ldots, h_{M,L}]^T \tag{33}$$

$$\begin{aligned}\mathbf{n}_c(k) &= [n_{1,1}(k), \ldots, n_{M,1}(k), n_{1,2}(k), \ldots, n_{M,2}(k), \ldots,\\ &\qquad n_{1,L}(k), \ldots, n_{M,L}(k)]^T\end{aligned} \tag{34}$$

with $z_{m,l}(k), h_{m,l}$ and $n_{m,l}(k)$ given by (13), (14) and (16), respectively.

The MRC weight vector can be determined blindly as the solution to the following problem [6]:

$$\max_{\mathbf{f}} \frac{E\{|\mathbf{f}^H\mathbf{z}_c(k)|^2\}}{E\{|\mathbf{f}^H\mathbf{n}_c(k)|^2\}} \equiv \frac{\mathbf{f}^H\mathbf{R}_{z_c}\mathbf{f}}{\mathbf{f}^H\mathbf{R}_{n_c}\mathbf{f}} \tag{35}$$

where $\mathbf{R}_{z_c} = E\{\mathbf{z}_c(k)\mathbf{z}_c^H(k)\}$ and $\mathbf{R}_{n_c} = E\{\mathbf{n}_c(k)\mathbf{n}_c^H(k)\}$ are the BC output data and noise correlation matrices, respectively. The solution to (35) is well-known to be the principal generalized eigenvector $\{\mathbf{R}_{z_c}, \mathbf{R}_{n_c}\}$. The matrix $\mathbf{R}_{n_c}$ depends on the beamformer and correlator weight vectors, and can be shown to have the (i, j)th entry given by (see the Appendix)

$$[\mathbf{R}_{n_c}]_{(i,j)} = \sigma_n^2\tilde{\mathbf{a}}_{m_1,l_1}^H\tilde{\mathbf{a}}_{m_2,l_2}\hat{\mathbf{s}}_{m_1,l_1}^T\hat{\mathbf{s}}_{m_2,l_2}^* \tag{36}$$

where $i = (l_1 - 1)M + m_1$ and $j = (l_2 - 1)M + m_2$, with $1 \le m_1, m_2 \le M$ and $1 \le l_1, l_2 \le L$. The algorithm of the proposed BT receiver is summarized in Table I, and the overall system diagram is depicted in Fig. 2.

TABLE I
SUMMARY OF PROPOSED BT RECEIVER.

1. Determine angular sector Θ_S, look angles θ_m's, time duration T_D and GSC blocking matrices $\tilde{\mathbf{B}}$ and $\hat{\mathbf{B}}$.
2. Initialize with $\tilde{\mathbf{a}}_{m,l} = \mathbf{a}_m$, $m = 1, \ldots, M$, $l = 1, \ldots, L$.
3. Compute in parallel $\mathbf{g}_{m,l}$, $m = 1, \ldots, M$, $l = 1, \ldots, L$, according to (20).
4. Compute in parallel $\mathbf{v}_{m,l}$, $m = 1, \ldots, M$, $l = 1, \ldots, L$, according to (24).
5. Iterate between 3 and 4 if necessary.
6. Obtain $\mathbf{R}_{n_c}$ and compute $\mathbf{f}$ according to (35).

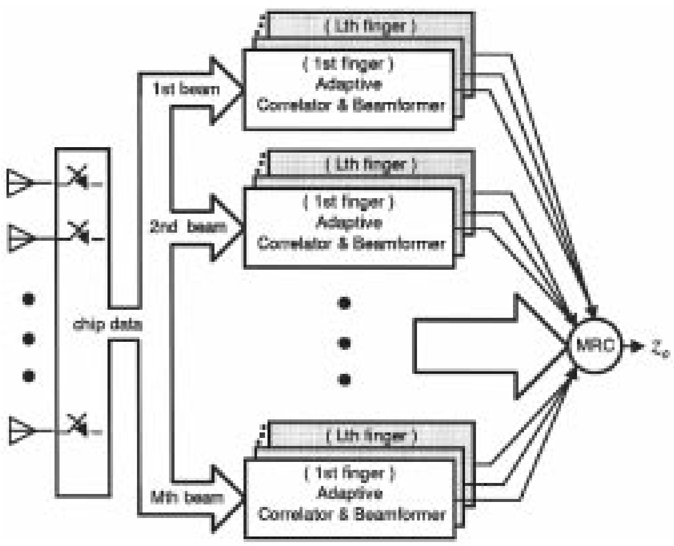

Fig. 2. Schematic diagram of proposed BT receiver.

IV. Implementation Issues

A. Numerical Stability

In the direct matrix inversion implementation, the computation of $\mathbf{g}_{m,l}$ in (20) and $\mathbf{v}_{m,l}$ in (24) involves the inversion of matrices $\hat{\mathbf{R}}_{m,l}$ and $\tilde{\mathbf{R}}_{m,l}$, respectively. Numerical instability may arise when there are few strong interferers present, leading to ill-conditioned correlation matrices. On the other hand, the convergence of weight vectors may be slow due to the residual signal not completely removed by the blocking matrices. To remedy these, pseudonoise terms $\eta_g\mathbf{I}$ and $\eta_v\mathbf{I}$ can be added to $\hat{\mathbf{R}}_{m,l}$ and $\tilde{\mathbf{R}}_{m,l}$, respectively, to alleviate the sensitivity problem [15]. The pseudonoise has the effect of deemphasizing the strong interference and masking the residual signal and can help to improve signal reception [15], [16]. It should be chosen large enough to handle the ill-condition problem, but not too large to distort the original signal scenario. A suitable choice which proves effective is such that η_g and η_v are equal to a small fraction (e.g., 0.1–0.3) of the dominant signal power in $\hat{\mathbf{R}}_{m,l}$ and $\tilde{\mathbf{R}}_{m,l}$, respectively [16]. In practice, a convenient estimate of the dominant signal power is the largest eigenvalue of the corresponding correlation matrix.

B. Partially Adaptive Implementation

The degree of freedom for the adaptive correlators is on the order of N, which can be quite large in wideband applications. A large degree of freedom requires a high computational complexity and is likely to incur poor convergence behaviors [9]. To alleviate this, the partially adaptive (PA) methods can be applied to reduce the dimension of $\mathbf{g}_{m,l}$s. Partial adaptivity can be achieved by either working with a reduced-dimension data vector or with a reduced-size blocking matrix [17], [18]. Here, a simple approach is suggested based on the principle of maximum cross correlation [19].

Let $\hat{\mathbf{B}}_P$ be the $(N+L-1)\times N'$ reduced-size blocking matrix and $\hat{\mathbf{B}}$ be the $(N+L-1)\times(N-1)$ full-size blocking matrix satisfying $\hat{\mathbf{B}}^H\mathbf{S}_1 = 0$ and $\hat{\mathbf{B}}^H\hat{\mathbf{B}} = \mathbf{I}$. The maximum cross correlation criterion dictates that $\hat{\mathbf{B}}_P$ should be composed of N' vectors that maximize the magnitude of the cross correlation between the upper and lower branch data of the temporal GSC described by (19). That is, $\hat{\mathbf{B}}_P$ is chosen according to:

$$\max_{\hat{\mathbf{B}}_P} \|E\left\{\hat{\mathbf{B}}_P^H\mathbf{X}^T(k)\tilde{\mathbf{a}}_{m,l}^*\tilde{\mathbf{a}}_{m,l}^T\mathbf{X}^*(k)\mathbf{s}_l\right\}\| \quad \text{subject to: } \hat{\mathbf{B}}_P^H\mathbf{S}_1 = 0 \quad \text{and} \quad \hat{\mathbf{B}}_P^H\hat{\mathbf{B}}_P = \mathbf{I} \tag{37}$$

Optimum solution to (37) requires a time consuming iterative procedure if N' is large [19]. As a simplified alternative, $\hat{\mathbf{B}}_P$ can be obtained with the N' orthonormal columns of $\hat{\mathbf{B}}$ that give the largest cross correlation values. Note that the cross correlation vector in (37) can be readily obtained as a subvector of $\hat{\mathbf{r}}_{m,l}$ given in (21) without extra computation. Finally, the PA correlator weight vector is given by (20)-(21), with $\hat{\mathbf{B}}$ replaced by $\hat{\mathbf{B}}_P$.

C. Recursive Computation of Weight Vectors

To avoid matrix inversion, the GSC can be implemented in a time-recursive fashion using stochastic gradient algorithms such as LMS [13]. This leads to recursive formulation of the solutions to (19) and (23), respectively

$$\begin{aligned}\mathbf{g}_{m,l}&(k+1)\\ &= \mathbf{g}_{m,l}(k) + \mu_g\left[(\mathbf{s}_l - \hat{\mathbf{B}}\mathbf{g}_{m,l}(k))^H\mathbf{X}^T(k)\tilde{\mathbf{a}}_{m,l}^*(k)\right]^*\\ &\quad\times\hat{\mathbf{B}}^H\mathbf{X}^T(k)\tilde{\mathbf{a}}_{m,l}^*(k)\end{aligned} \tag{38}$$

$$\begin{aligned}\mathbf{v}_{m,l}&(k+1)\\ &= \mathbf{v}_{m,l}(k) + \mu_v\left[(\mathbf{a}_m - \tilde{\mathbf{B}}\mathbf{v}_{m,l}(k))^H\mathbf{X}(k)\hat{\mathbf{s}}_{m,l}^*(k)\right]^*\\ &\quad\times\tilde{\mathbf{B}}^H\mathbf{X}(k)\hat{\mathbf{s}}_{m,l}^*(k)\end{aligned} \tag{39}$$

where μ_g and μ_v are the adaptation stepsizes. Clearly, (38) and (39) are coupled and should be executed in an alternate fashion.

On the other hand, the generalized eigenvector required for the computation of the MRC weight vector in (35) can be also obtained via a time-recursive algorithm. One such example is the recently developed projection approximation subspace tracking (PAST) algorithm [20], which is shown to exhibit global convergence and requires a computational complexity of the order ML (BT dimension) per iteration. The PAST algorithm was originally proposed for tracking multiple

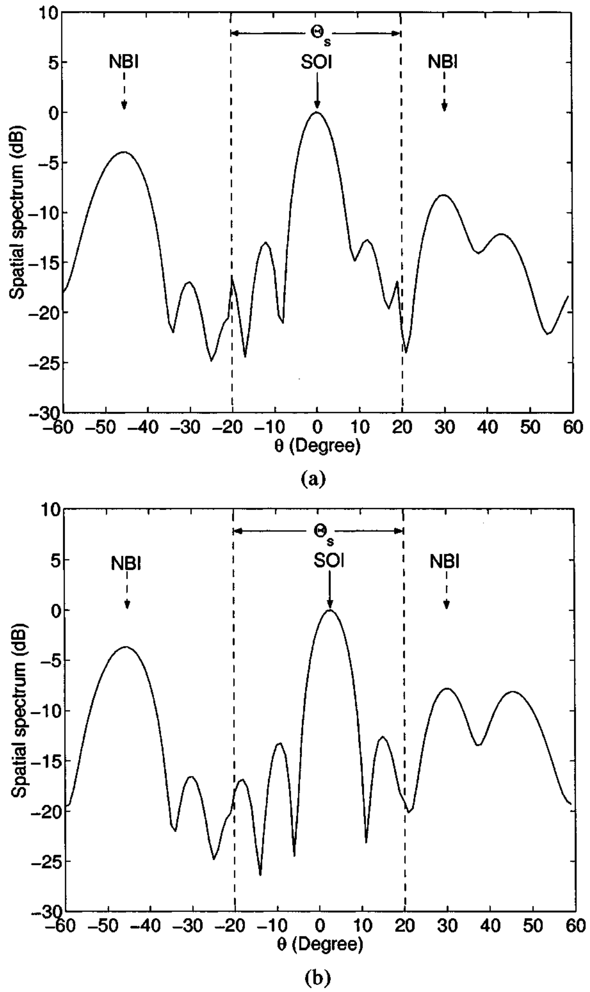

Fig. 3. Spatial spectrum of AOA estimation for (a) finger 1 and (b) finger 3, obtained with $Q = 10$ users and $\mathrm{NFR} = 0$ dB.

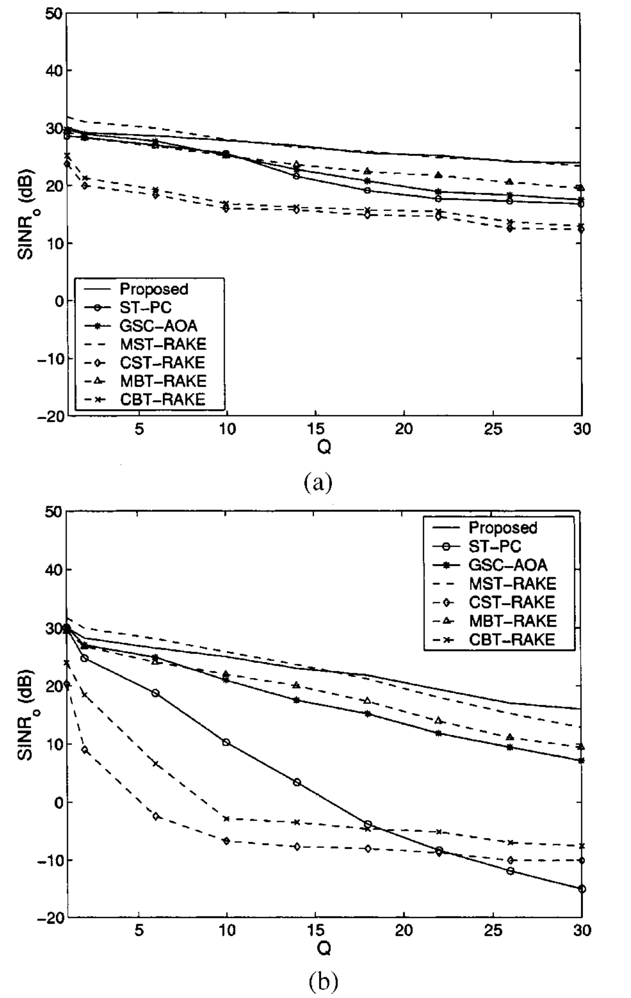

Fig. 4. Evaluation of system capacity with (a) $\mathrm{NFR} = 0$ dB and (b) $\mathrm{NFR} = 20$ dB.

eigenvectors, and can be easily modified for the tracking of a single generalized eigenvector as dictated by (35) [21].

D. Complexity and Performance

The tradeoff between system performance and computational complexity is an issue depending on the number of antennas D and processing gain N of the system. In general, more antennas and a larger processing gain will provide better interference suppression but lead to high computational complexity and poor convergence behaviors. However, the spatial and temporal dimensions (D and N) do not affect the complexity of the eigenvector computation in MRC because the MRC weight vector has a size of ML, regardless of D and L. For batch processing, the major computations in the proposed algorithm involves the inversion of $\hat{\mathbf{R}}_{m,l}^{-1}$ with size $(N-1)\times(N-1)$ in (20), inversion of $\tilde{\mathbf{R}}_{m,l}^{-1}$ with size $D' \times D'$ in (24), and a principal eigenvector computation with size $ML \times ML$ in (35). The overall complex is about $O(N^3 + D'^3 + M^3L^3)$. For comparison, the ST post-despread receiver in [4] requires about $O(D^3L^3)$, and the ST pre-despread receiver in [10] requires about $O(N^3+D+L)$ for batch processing. The higher complexity of the proposed receiver and that in [4] is the price paid for not using a training sequence for channel estimation.

For recursive processing as described in the previous section, the complexity for each iteration is about $O(N + D' + ML)$ for the proposed receiver, $O(DL)$ for the receiver in [4], and $O(N + D^3)$ for the receiver in [10]. The proposed receiver can be further simplified by incorporating partial adaptivity in temporal processing to reduce N to a smaller N'. As long as N' is large compared to the number of strong interferers, the performance of the receiver will be virtually unaffected.

In the flat-fading case, the multipath delay spread is smaller than the chip period such that the receiver would require only a single finger to work with. As a result, only a single adaptive correlator is needed for each diversity beam. On the other hand, in time-invariant channels without Doppler shift, it will be easier for the adaptive algorithms to converge and keep track of SOI. In addition, since the fading gain remains constant at each beam-finger pair, the MRC weights need not be computed frequently. This should alleviate the load for eigenvector computation.

V. Computer Simulations

Here, the performance of the proposed BT receiver is evaluated using a linear array of $D = 12$ identical antennas uniformly spaced by a $1/\sqrt{3}$ wavelength. The interantenna spacing was chosen for the field-of-view $[-60°, 60°]$. The working sector

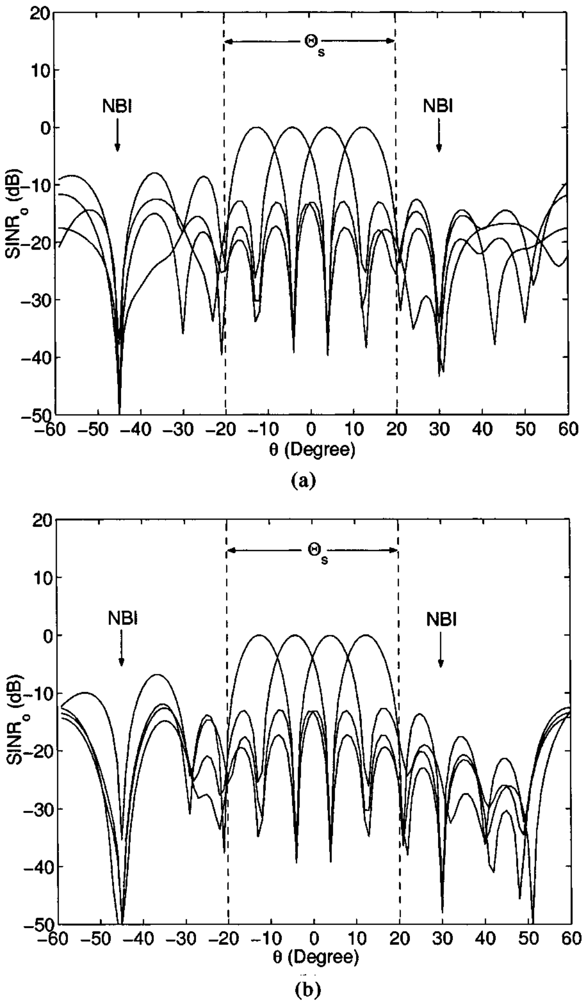

Fig. 5. Patterns of diversity beams for (a) finger 1 and (b) finger 3, obtained with $Q = 5$ users and $\mathrm{NFR} = 0$ dB.

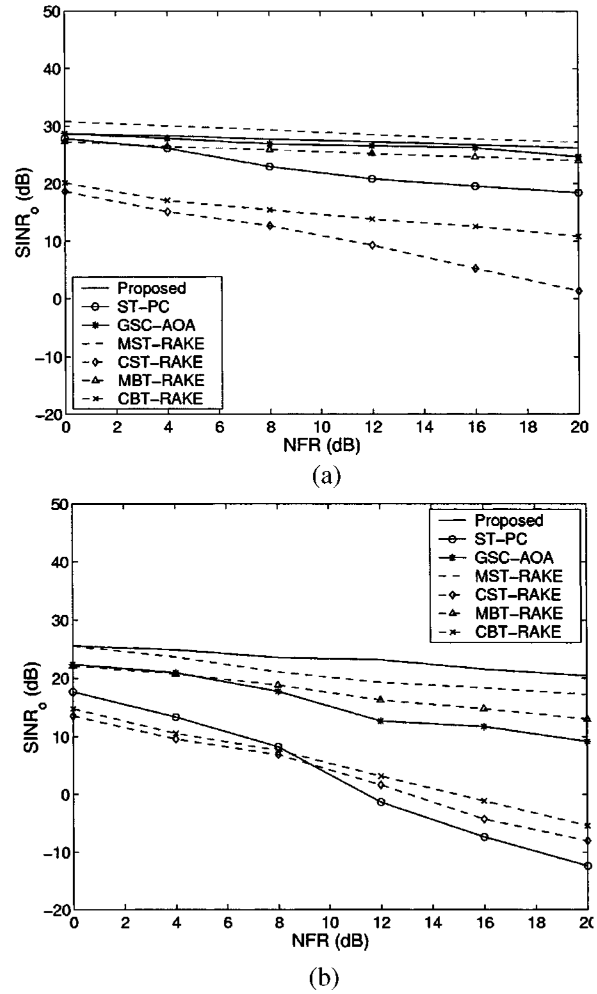

Fig. 6. Evaluation of near–far resistance with (a) $Q = 5$ and (b) $Q = 25$ users.

was $\Theta_S = [-20°, 20°]$, and $M = 4$ diversity beams were formed at look directions $\{-12.45°, -4.15°, 4.15°, 12.45°\}$. The spatial blocking matrix $\tilde{\mathbf{B}}$ was constructed with $D' = 6$ using the eigenvector constrained method. It is assumed that the SOI was in Θ_S, and the MAI were uniformly distributed in the entire field-of-view, with each user having the same multipath angle spread of $10°$. For all users, $J = 4$ paths were generated with delays chosen from $\{0, T_c, 2T_c, 3T_c\}$, and fading gains being i.i.d. unit variance complex Gaussian random variables. All CDMA signals were BPSK data modulated and spread with the Gold code of length $N = 31$. In addition to the MAI, there were two equal power BPSK NBIs arriving from 30° and − 45°, with bit rate being 0.8 times that of the CDMA signals. Finally, the number of fingers was $L = 4$, and the temporal blocking matrix $\hat{\mathbf{B}}$ was chosen, according to Section III.C, to be the 34×28 matrix whose columns formed an orthogonal complement to $[\mathbf{s}_1^{(-)}, \mathbf{s}_1, \mathbf{s}_2, \mathbf{s}_3, \mathbf{s}_4, \mathbf{s}_4^{(+)}]$.

As a performance index, the output SINR (in) is defined as

$$\mathrm{SINR}_o = 10\log_{10}\frac{\text{output power of signal in} z_o(k)}{\text{output power of (MAI+NBI+noise) in} z_o(k)}$$

and the input SNR (in dB) is defined as

$$\mathrm{SNR}_i = 10\log_{10}\frac{P_1}{\sigma_n^2}$$

The near–far-ratio (NFR) is the ratio of the MAI power to signal power before despreading, and the NBI-to-signal-ratio (NSR) is the ratio of the NBI power to signal power before despreading. For all but one result in the simulations, N_s symbols were used to obtain sample estimates of $\hat{\mathbf{R}}_{m,l}, \hat{\mathbf{r}}_{m,l}, \tilde{\mathbf{R}}_{m,l}, \tilde{\mathbf{r}}_{m,l}$ and $\mathbf{R}_{z_c}$, and a total of 50 Monte-Carlo runs were executed to obtain an average SINR$_o$. In direct matrix inversion weight vector computation, the pseudo noise powers η_g and η_v were chosen to be one fifth of the largest eigenvalue of $\hat{\mathbf{R}}_{m,l}$, and $\tilde{\mathbf{R}}_{m,l}$, respectively. For performance comparison, we included the results obtained with the blind receivers proposed in [4] (ST-PC) and [10] (GSC-AOA), respectively, and the nonblind coherent ST RAKE (CST-RAKE), MMSE ST RAKE (MST-RAKE), coherent BT RAKE (CBT-RAKE), and MMSE BT RAKE (MBT-RAKE) receivers described in Section II. For nonblind receivers, the post-ST and post-BT channel vectors were estimated using the same N_s symbols as the training sequence. For CBT-RAKE and MBT-RAKE receivers, the beamforming matrix was constructed using the four steering vectors $\{\mathbf{a}_1, \mathbf{a}_2, \mathbf{a}_3, \mathbf{a}_4\}$ asso-

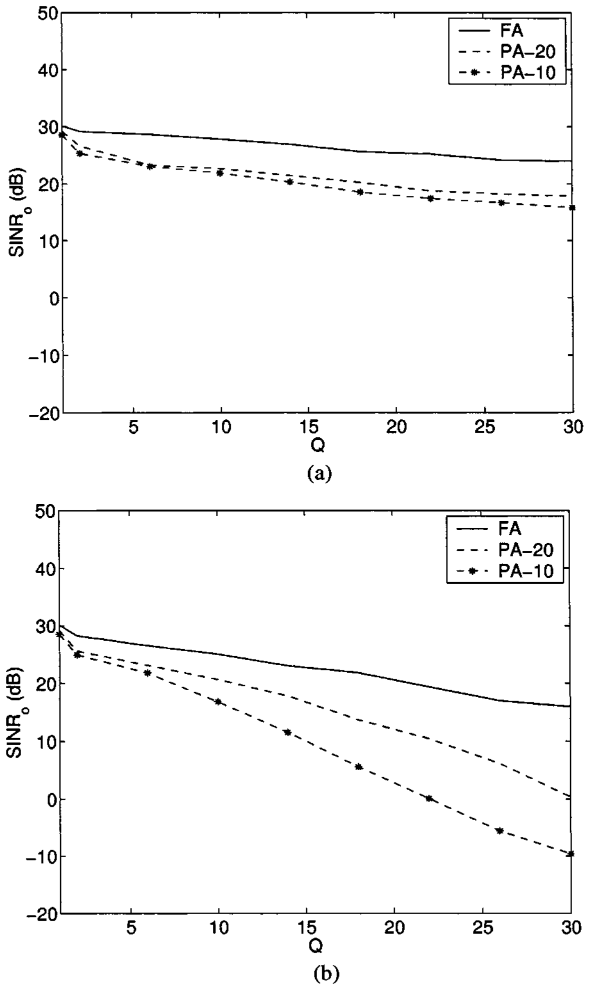

Fig. 7. Evaluation of efficacy of partial adaptivity with (a) $\mathrm{NFR} = 0$ dB and (b) $\mathrm{NFR} = 20$ dB.

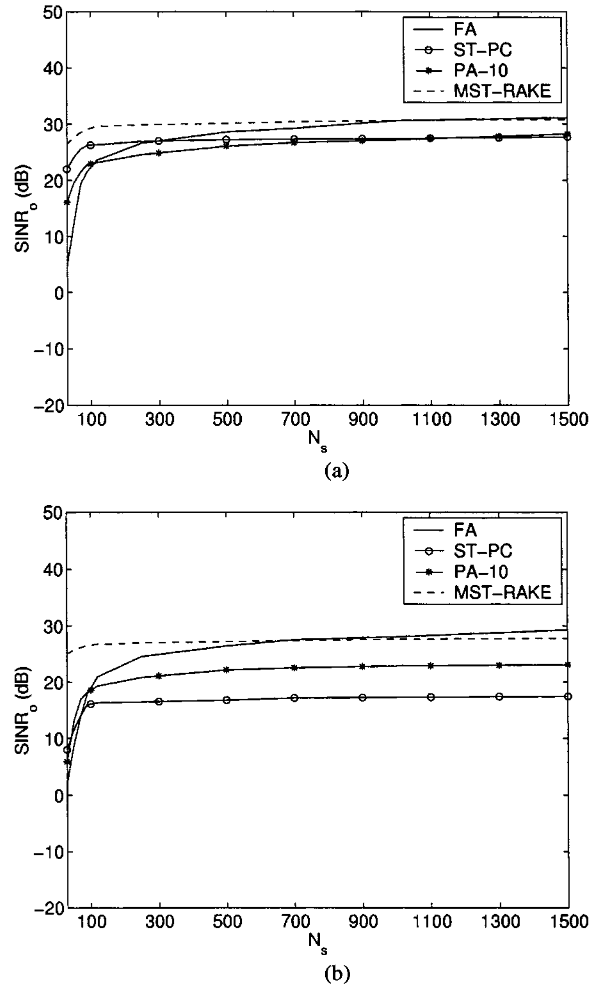

Fig. 8. Evaluation of convergence behavior with $Q = 5$ users, and (a) $\mathrm{NFR} = 0$ dB and (b) $\mathrm{NFR} = 20$ dB.

ciated with the look directions of the diversity beams. Unless otherwise mentioned, the following "standard" parameters will be used throughout the section: $\mathrm{SNR}_i = 0$ dB, $\mathrm{NSR} = 20$ dB, and $N_s = 500$.

To demonstrate the efficacy of AOA estimation described in Section III-B, the spatial spectrum were computed for the three sectors $[-60°, -20°]$, $[-20°, 20°]$, and $[20°, 40°]$. Fig. 3 shows the result for the first and third fingers ($l = 1, 3$), with $Q = 10$ users and $\mathrm{NFR} = 0$ dB. The peak in the working sector gives the AOA of the SOI, and those in other sectors are "false peaks" due to NBI and can be easily detected by the system.

In the first set of simulations, the system capacity is evaluated with different user numbers Q. The resulting output SINR curves are plotted in Fig. 4 for $\mathrm{NFR} = 0$ and {\hbox{20}} dB. As observed, the proposed BT receiver performs reliably for a wide range of Q, even outperforming the MST-RAKE receiver with a moderately large Q and high NFR. This indicates that the adaptive beamformers and correlators have successfully eliminated the strong NBI and MAI, respectively. The CST-RAKE, CBT-RAKE and ST-PC receivers totally fail with $\mathrm{NFR} = 20$ dB and a large Q due to the lack of interference suppression or poor channel estimation. Finally, the GSC-AOA blind receiver performs reliably but exhibits a certain degradation compared to the proposed receiver with a large Q. This is observed to be due to the residual NBI not effectively suppressed by the temporal GSC. In Fig. 5, the patterns of the diversity beams for the first and third fingers ($l = 1, 3$) are plotted for the case $Q = 5$ users and $\mathrm{NFR} = 0$ dB. The mainlobes and deep nulls confirm that the adaptive beamformers can effectively collect the in-sector signals and suppress the strong NBI.

In the second set of simulations, the near–far resistance of the proposed BT receiver is evaluated with different NFR values. Fig. 6 shows the results obtained with $Q = 5$ and 25 users. As observed, with $Q = 25$, the CST-RAKE, CBT-RAKE and ST-PC receivers receivers fail again, and the MST-RAKE receiver loses its advantage due to ineffective interference suppression. On the contrary, the BT receiver achieves excellent near–far resistance by successfully canceling the MAI and NBI simultaneously. The GSC-AOA receiver performs well with $Q = 5$, in which case the temporal GSC has a sufficient degree of freedom to handle the NBI.

In the third set of simulations, the efficacy of partial adaptivity is demonstrated through system capacity evaluation. Fig. 7 shows the results obtained with the fully adaptive (FA) BT receiver, partially adaptive BT receiver with $N' = 10$ (PA-10), and partially adaptive BT receiver with $N' = 20$ (PA-20), for

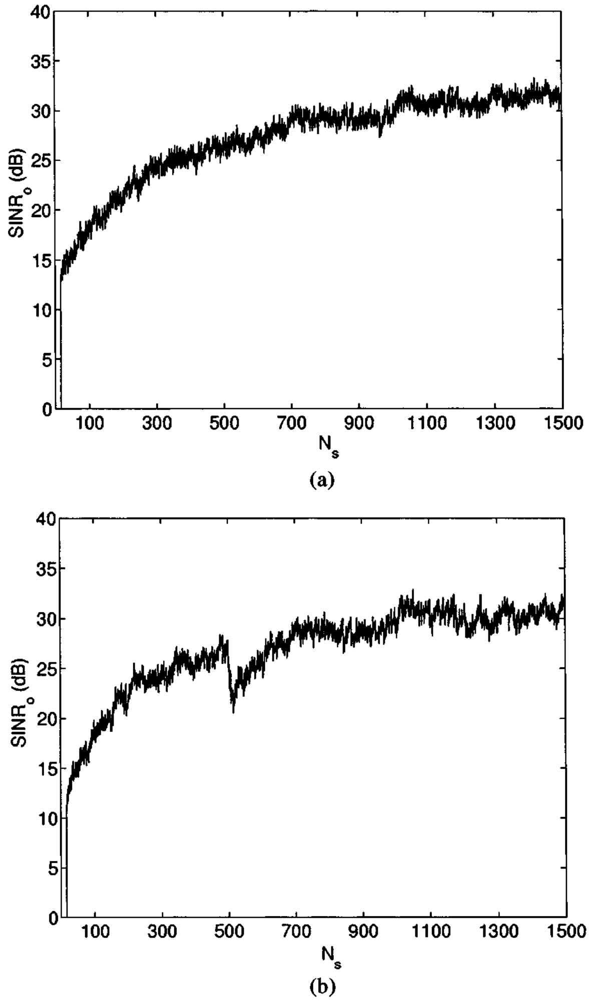

Fig. 9. Evaluation of of recursive algorithms with $Q = 5$ users and $\mathrm{NFR} = 0$ dB. (a) NBIs AOA fixed and (b) NBIs AOA changed at 500th iteration.

$\mathrm{NFR} = 0$ dB and 20 dB. As expected, the PA receivers perform reliably as compared to the FA receiver with a small Q, and degrade as Q increases due to the exhaustion of the temporal degree of freedom for interference suppression.

In the fourth set of simulations, the convergence behaviors of the proposed FA BT receiver, PA-10 receiver, ST-PC receiver, and MST-RAKE receiver were evaluated by varying the data sample size N_s. The resulting output SINR are plotted in Fig. 8 with $Q = 5$, and $\mathrm{NFR} = 0$ and 20 dB. As observed, the output SINR increases as N_s increases for all receivers. The MMSE receiver converges fastest due to the use of a training sequence. The PA-10 receiver converges faster than the FA receiver, confirming the well-known fact that adaptive filters of smaller size converge faster.

Finally, to demonstrate the effectiveness of the recursive algorithms for weight adaptation and eigenvector computation, we replaced direct matrix inversions by the formulae given in (38) and (39), and direct eigenvector computation by the PAST algorithm. The adaptation stepsizes were chosen as $\mu_g = \mu_w = 10^{-6}$. Fig. 9(a) shows the resulting learning curve of the FA BT receiver, with $\mathrm{NFR} = 0$ dB and $Q = 5$. To show the tracking capability of the algorithm, we repeated the same simulation but deliberately changed the AOAs of the two NBIs to 40° and $-50°$, respectively, at the 500th iteration. The learning curves given in Fig. 9(b) confirm that the recursive receiver successfully adjusted its weights to suppress the NBI.

VI. CONCLUSION

A blind adaptive beamspace–time receiver has been proposed for sectored CDMA wireless communications. The proposed receiver is designed with a two-stage procedure. First, a set of adaptive space–time diversity processors, in the form of beamformer-correlator pairs, is constructed which provides effective suppression of interference and reception of signals from a prescribed space–time region. Second, the output data obtained from these processors are maximum ratio combined to capture the signal multipath components coherently. For an efficient implementation, recursive algorithms are employed for weight vector adaptation and eigenvector computation, and a simple partially adaptive solution is given based on the maximum cross correlation principle. The proposed receiver can be operated without the aid of a training signal. The only information required is the signature sequence, timing and a rough estimate of the angle of arrival of the signal for selecting the sector of interest. Compared to the conventional space–time receivers, the proposed beamspace receiver enhances the the SINR by suppressing interference via adaptive nulling, and increases system capacity by sectorization. From simulation results, it is shown that the proposed blind receiver is near-far resistant, and performs reliably in an overloaded system.

APPENDIX
Derivation of entries of $\mathbf{R}_{n_c}$

Let $i = (l_1 - 1)M + m_1$ and $j = (l_2 - 1)M + m_2$. From (34), the (i, j)th entry of $\mathbf{R}_{n_c}$ is the cross correlation between $n_{l_1,m_1}(k)$ and $n_{l_2,m_2}(k)$:

$$\begin{aligned}[\mathbf{R}_{n_c}]_{(i,j)} &= E\{\tilde{\mathbf{a}}_{m_1,l_1}^H \mathbf{N}(k)\hat{\mathbf{s}}_{m_1,l_1}^* \hat{\mathbf{s}}_{m_2,l_2}^T \mathbf{N}^H(k)\tilde{\mathbf{a}}_{m_2,l_2}\} \\ &= \tilde{\mathbf{a}}_{m_1,l_1}^H E\{\mathbf{N}(k)\hat{\mathbf{s}}_{m_1,l_1}^* \hat{\mathbf{s}}_{m_2,l_2}^T \mathbf{N}^H(k)\}\tilde{\mathbf{a}}_{m_2,l_2} \\ &= \tilde{\mathbf{a}}_{m_1,l_1}^H \left(\sum_{l=1}^{N+L-1} \sum_{n=1}^{N+L-1} [\hat{\mathbf{s}}_{m_1,l_1}^*]_l [\hat{\mathbf{s}}_{m_1,l_1}]_n \right. \\ &\qquad \left. \times\ E\{\mathbf{n}_l(k)\mathbf{n}_n^*(k)\} \right) \tilde{\mathbf{a}}_{m_2,l_2}\end{aligned}$$

where $[\hat{\mathbf{s}}_{m_1,l_1}]_n$ is the nth entry of $\hat{\mathbf{s}}_{m_1,l_1}$, and $\mathbf{n}_l(k)$ is the lth column of $\mathbf{N}(k)$. Using the fact that $\mathbf{n}_l(k)$ are i.i.d. random vectors with the same variance σ_n^2, we have

$$E\{\mathbf{n}_l(k)\mathbf{n}_n^*(k)\} = \sigma_n^2 \mathbf{I}\delta[l - n]$$

where $\delta[n]$ is the Kronecker delta function. This gives

$$\begin{aligned}[\mathbf{R}_{n_c}]_{(i,j)} &= \sigma_n^2 \tilde{\mathbf{a}}_{m_1,l_1}^H \left(\sum_{n=1}^{N+L-1} [\hat{\mathbf{s}}_{m_1,l_1}^*]_l [\hat{\mathbf{s}}_{m_1,l_1}]_n \right) \tilde{\mathbf{a}}_{m_2,l_2} \\ &= \sigma_n^2 \tilde{\mathbf{a}}_{m_1,l_1}^H \tilde{\mathbf{a}}_{m_2,l_2} \hat{\mathbf{s}}_{m_1,l_1}^T \hat{\mathbf{s}}_{m_2,l_2}^*.\end{aligned}$$

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