A Space-Time Channel Estimator and Single-User Receiver for Code-Reuse DS-CDMA Systems

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***Abstract*—In this paper, a blind asynchronous single-user code-reuse direct sequence code division multiple access (CDMA) array receiver is proposed for the uplink. By assigning each short PN-code more than once, code reuse allows the number of active users to be increased beyond the spreading gain. The proposed receiver is based on a blind single-code multipath joint space–time channel estimation technique that utilizes the concept of the spatio–temporal array manifold, in conjunction with a novel preprocessor, to deal with the multipath problem. From the estimated space–time channel parameters of a particular active code, the subset of parameters of a specific co-code user is then identified, and a single-user receiving weight vector is finally formed. The proposed approach is a subspace type method, and therefore, it is "near–far" resistant. Furthermore, in contrast to existing receivers such as the Space–Time Decorrelating Detector, the proposed receiver weight vector is tolerant to partial channel estimation errors and the incomplete estimation of channel parameters. The theoretical framework is supported by computer simulation studies.**

***Index Terms*—Antenna arrays, CDMA, code-reuse, interference suppression, multipath channels, space-time processing, vector channels.**

NOMENCLATURE

A, a	Scalar.
$\mathbf{a}$, $\boldsymbol{a}$	Column vector.
$\mathbf{A}$	Matrix.
$\mathbf{0}_N$	Column vector of N zeros.
$\mathbf{1}_N$	Column vector of N ones.
$\mathbf{O}_{N\times M}$	Matrix of zeros (size $N \times M$).
$\mathbf{I}_N$	N dimensional identity matrix.
$\Re e(a)$	Real part of a.
$\Im m(a)$	Imaginary part of a.
DOA	Direction-of-arrival.
TOA	Time-of-arrival.
$\lceil a \rceil$	Round a up to nearest integer.
$row_N(\mathbf{A})$	Row N of matrix $\mathbf{A}$.
$\exp(\mathbf{a})$	Element-by-element exponential.
$\otimes$	Kronecker product.
$\mathcal{L}[\mathbf{A}]$	Subspace spanned by columns of $\mathbf{A}$.
$\mathcal{L}^{\perp}[\mathbf{A}]$	Complement of $\mathcal{L}[\mathbf{A}]$.
$P[\mathbf{A}]$	Projection operator on $\mathcal{L}[\mathbf{A}]$ i.e., $P[\mathbf{A}] = \mathbf{A}(\mathbf{A}^H\mathbf{A})^{-1}\mathbf{A}^H$.
$P^{\perp}[\mathbf{A}]$	Projection operator on $\mathcal{L}^{\perp}[\mathbf{A}]$ i.e., $P^{\perp}[\mathbf{A}] = \mathbf{I} - P[\mathbf{A}]$.

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I. INTRODUCTION

IT HAS been shown that code-division multiple access (CDMA) offers significant promise in meeting the growing demand for data rate and capacity in wireless communication systems [1]. The long code approach (adopted in IS-95) uses PN-code sequences that span many data symbol periods, and this allows a large number of users to be included in the system. The conventional long code receivers, such as the matched filter [2] or the RAKE receiver [3], treat multiuser interference as additive white Gaussian noise, such that each user can be considered to be functioning in a single-user channel in the presence of only noise. The capacity of such systems is then limited by the total interference power and not by the number of available codes. In a high data rate system, the time dispersive nature of the communication channel introduces severe intersymbol interference (ISI), and this substantially deteriorates the performance of these conventional receivers. Nonlinear interference cancellation techniques based on several iterative stages of decision feedback have been proposed [4], [5]; however, the use of such methods requires complete and accurate knowledge of the amplitudes, phases, and delays of all users, and hence, their implementation may be impractical. As a consequence, linear interference cancellation techniques based on subspace type methods are generally preferred.

Subspace-type channel estimation techniques may be considered as the most powerful type of technique, outperforming (especially in severe near–far situations) the conventional "correlation type" channel estimation methods. However, such techniques (which originally were derived for direction-finding array systems [6], [7]) are only applicable to the short coded system in which each user occupies a finite number of dimensions in signal subspace. Recently, these techniques have been employed in DS-CDMA initially for single antenna systems (single path [8] and multipath [9]) and then for antenna arrays operating in multipath environments [10]–[15]. In [16], the concept of the spatio–temporal manifold is used in a non-CDMA environment, whereas in [10], it is employed to propose a subspace-type CDMA technique that is capable of identifying/estimating jointly the coherent multipath delays and DOAs of the desired signals in the presence of multiple access interference (MAI), where the coherency between paths is removed by a one–dimensional (1-D) temporal smoothing procedure employed in a transformed domain. In [11] and [12], a reduced dimension space–time RAKE receiver for DS-CDMA channels has been proposed, where a joint estimation procedure of the delays and DOA of the dominant paths is developed based on the estimated space–time channel response vector of the desired user. This is a subspace-type approach employing the two–dimensional (2-D) unitary ESPRIT algorithm as the estimator of the delays and DOAs, jointly. However, this approach restricts the multipath delay spread to be a fraction of the data symbol period so that the RAKE fingers can be roughly located. In [13], an integrated space–time multiuser array receiver has been proposed for asynchronous DS-CDMA that is robust to errors in the estimation of multipath arrival times.

Subspace-based techniques have also been employed for diffused space–time CDMA frequency selective channels. For instance, in [14], a multidimensional subspace technique has been proposed to jointly estimate the channel parameters and the degree of channel diffusion under the assumption that the distribution function of the space–time parameters are known. In [15], the channel estimation technique proposed in [14] for space–time dispersive channels has been adapted to synchronous DS-CDMA based on two semi-blind spatio–temporal channel estimation techniques.

The subspace-type channel estimation methods discussed thus far (excluding [14]) consider the case where each user is assumed to have a unique short PN-code sequence, and consequently, the maximum number of users (capacity) is constrained by the short code spreading gain. Several single-antenna systems have also been proposed for increasing user capacity through code reuse, where each short PN-code sequence is assigned simultaneously to more than one users.

For instance, in the sequence sharing approach of [17], each user is assigned a subset of orthogonal sequences, such that a particular sequence may be shared by more than one user. The system, hence, needs fewer orthogonal sequences than one that does not employ sequence sharing. The inevitable problem of sequence collision is overcome by coordinating users' start times through basestation to mobile feedback. The sequences are used in conjunction with the TOAs at the basestation receiver to distinguish the signals of different users. An alternative approach, which is proposed in [18], is also based on temporal processing. In this approach, the same orthogonal spreading sequence is assigned to a group of users, with the orthogonal sequence being overlaid with a different PN-code sequence for each user in the group. At the base station, detection is then performed in two separate stages, first for the PN-code sequence and then for the orthogonal sequence. The process is then repeated two or more times to improve the receiver's decisions. This "multiple stage spreading" approach assumes perfect time synchronization amongst all mobile users. Both of these temporal only solutions assume a nondispersive (multipath free) channel.

It is well known that in general, if an antenna array system is used at the base station, communication system performance can be improved and a significant capacity gain achieved. However, the single antenna code-reuse techniques discussed thus far do not exploit the spatial aspect of the channel. In [19], a code-reuse technique is proposed that employs an antenna array, such that users with the same code are separated via spatial-only processing. The approach employs an array receiver in conjunction with a code-reuse strategy. This code-reuse strategy exploits the higher probability of occurrence of symbol synchronized code-reuse interferers among the interfering users than do symbol synchronized co-code interferers to the desired user.

The problem of estimating the channel parameters of the co-code users is not adequately addressed in any of the code-reuse methods discussed thus far. In addition, it is well known that joint space–time processing is superior to performing spatial and temporal processing separately [20]. A joint space–time channel estimation technique applicable to a code-reuse system has been proposed in [14]; however, this approach is for diffuse sources where the rays for a particular user are spatially and temporally diffused, arriving in clusters such that the mean and the variance of each cluster can be estimated. However, in this paper, it will be assumed that the received signal arrives through a number of distinct paths (due to sparse scatterers), which is a commonly used assumption especially in macrocellular channels [21]. In addition, if the system employs a code-reuse strategy, where the process of code assignment by the base station is eliminated (since each user may effectively choose a code randomly from the set of available codes), then this may lead to an increase in system capacity as well as to a simplified system design.

In this paper, a new space–time approach to code-reuse is proposed. The approach uses a novel joint space–time preprocessor to deal multipath problem and a MuSiC-type cost function to blindly estimate the space–time channel parameters of a single active code corresponding to a group of co-code users. The proposed procedure exploits the structure of the spatio–temporal manifold vectors, which are defined herein. Channel parameter estimation can be performed given only knowledge of the active PN-code sequence and the locations of the antennas in the base station array. From the estimated space–time channel parameters of a particular active code, the subset of parameters of a specific co-code user is then identified using a "correlation analysis assignment" method. A new single-user receiver weight vector that combines despreading and interference cancellation is also proposed. This weight vector is tolerant to partial channel estimation errors and incomplete estimation of channel parameters. This is a subspace-type method and is, therefore, "near–far" resistant. It is a technique that has superresolution capabilities but requires knowledge of the array manifold, which implies that the array should be properly calibrated [22], [23].

The organization of the paper is as follows. In Section II, the array DS-CDMA mobile vector channel is modeled using the concept of the spatio–temporal manifold vectors. Based on this model, in Section III, a new subspace type (i.e., near–far resistant) single-user receiver structure is proposed, facilitating a code-reuse approach. In Section IV, the proposed receiver is simulated and compared in terms of performance and assumptions with the decorrelating detector [24] and the space-time (ST)-RAKE receiver [25]. Mathematical descriptions of these benchmark receivers are also given and adapted to the notation employed throughout the paper. The simulations are carried out in cases of complete and incomplete channel estimation, whereas the case of partial channel estimation errors is also considered. Finally, in Section V, the paper is concluded.

II. Channel Model

In an M-user asynchronous quaternary phase-shift keying (QPSK) DS-CDMA communication system, the baseband tran-

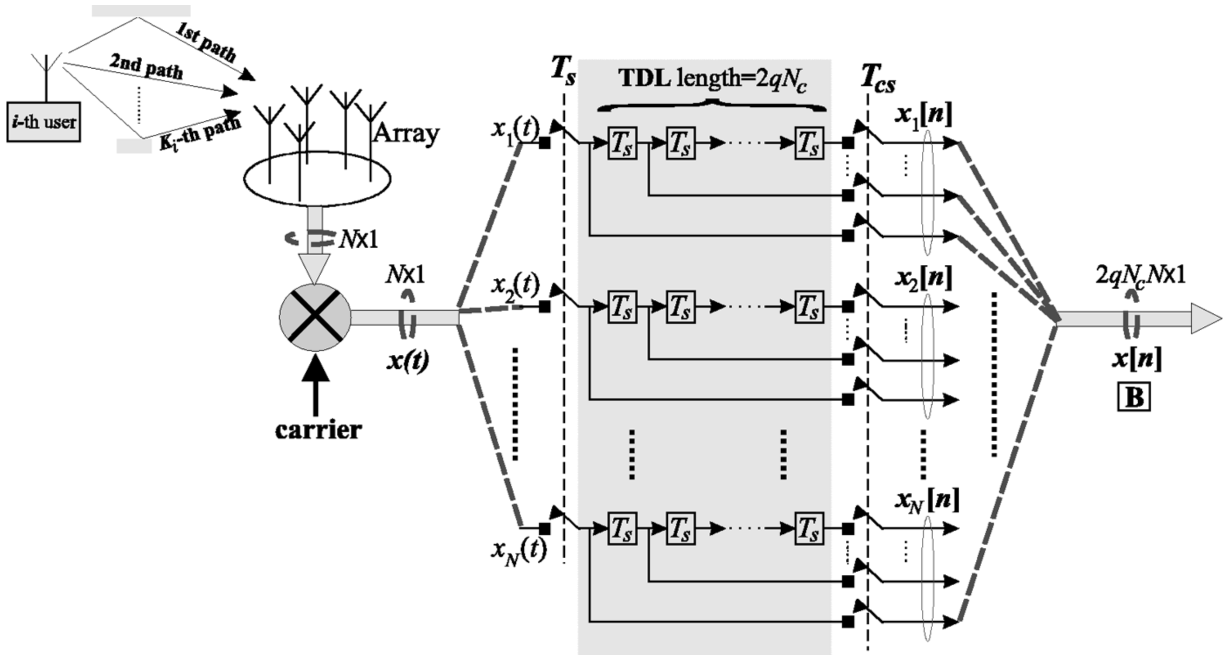

Fig. 1. Front-end of the proposed array system.

smitted signal of the ith user can be modeled as

$$m_i(t) = \sum_{n=-\infty}^{\infty} a_i[n] c_{i,\,\mathrm{PN}}(t - nT_{cs}) \tag{1}$$

with $nT_{cs} \le t < (n+1)T_{cs}$, where $\{a_i[n],\ \forall n \in \mathcal{N}\}$ is the ith user's sequence of channel symbols $(\pm 1 \pm j)/\sqrt{2}$ of period T_{cs} such that $\mathcal{E}\{a_i[m]a_j^*[n]\} = \delta_{ij}\delta_{mn}$. In (1), $c_{i,\,\mathrm{PN}}(t)$ denotes one period of the PN waveform associated with the ith user, i.e.,

$$c_{i,\,\mathrm{PN}}(t) = \sum_{k=0}^{N_c-1} \alpha_i[k] c(t - kT_c) \tag{2}$$

where $\{\alpha_i[k] \in \pm 1\}$ corresponds to the ith user's PN-code sequence of period $N_c = T_{cs}/T_c$, $c(t)$ denotes the chip pulse-shaping waveform of duration T_c, and $kT_c \le t < (k+1)T_c$. Let us consider that the ith transmitted signal arrives at the base station through K_i distinct paths (sparse scatterers). Furthermore, let us assume that the jth path of the ith user arrives at the array reference point from the direction $\{\theta_{ij},\, \phi_{ij}\}$, with θ_{ij} and ϕ_{ij} representing the azimuth and elevation, respectively, with channel propagation parameters β_{ij} and τ_{ij}. Note that τ_{ij} represents the path delays, whereas β_{ij} denotes the complex path coefficient and models the effects of path losses, shadowing and random phase shifts due to reflection. In addition, the coefficients $\beta_{ij},\ \forall (i,\, j)$ encompass the effects of the phase offset between the modulating/demodulating carriers, as well as the transmitter powers. It should be noted that this paper focuses on the estimation of the multipath delays and directions and not on the complex path coefficients that can be obtained after this estimation process using, for instance, [25]. Finally, it is assumed that the delay spread is in the region of a channel symbol period.

Based on the previous environment, the baseband continuous-time received signal-vector $\boldsymbol{x}(t) \in \mathcal{C}^N$ due to the M users at the antenna array can be modeled as

$$\begin{aligned} \boldsymbol{x}(t) &= [x_1(t),\, x_2(t),\, \ldots,\, x_N(t)]^T \\ &= \sum_{i=1}^{M} \sum_{j=1}^{K_i} \beta_{ij} \mathbf{s}_{ij} m_i(t - \tau_{ij}) + \mathbf{n}(t) \end{aligned} \tag{3}$$

where $\mathbf{n}(t) \in \mathcal{C}^N$ is the continuous-time complex white Gaussian noise vector. In (3), the vector $\mathbf{s}_{ij} = \mathbf{s}(\theta_{ij},\, \phi_{ij}) \in \mathcal{C}^N$ is the array manifold vector of the jth path of the ith user, which is defined as

$$\mathbf{s}_{ij} = \exp(-j[\boldsymbol{r}_1,\, \boldsymbol{r}_2,\, \ldots,\, \boldsymbol{r}_N]^T \mathbf{k}_{ij}) \tag{4}$$

where $\boldsymbol{r}_k \in \mathcal{R}^{3\times 1}$ is the location vector of the kth antenna element (in half wavelengths of the carrier frequency), and

$$\mathbf{k}_{ij} = \pi[\cos(\theta_{ij})\cos(\phi_{ij}),\, \sin(\theta_{ij})\cos(\phi_{ij}),\, \sin(\phi_{ij})]^T$$

is the wavenumber vector, with the azimuth angles θ_{ij} measured with respect to the x-axis. Without any loss of generality, ϕ_{ij} is assumed equal to zero for every $(i,\, j)$. This implies that all users are located on the $(x,\, y)$-plane.

The N-dimensional received signal-vector $\boldsymbol{x}(t)$ is then sampled with a period $T_s = T_c/q$ (oversampling if $q > 1$) and passed through a bank of N tap-delay lines (TDLs), each of length $2qN_c$. Upon concatenation of the outputs of the TDLs, the $(2qN_cN)$-dimensional discretized signal vector thus formed is

$$\boldsymbol{x}[n] = [\boldsymbol{x}_1[n]^T,\, \boldsymbol{x}_2[n]^T,\, \ldots,\, \boldsymbol{x}_N[n]^T]^T. \tag{5}$$

This is illustrated in Fig. 1, where $\boldsymbol{x}_k[n]$ is the output from the kth antenna's TDL.

Note, however, that due to the multipath delay spread, the content of each TDL contains contributions from not only the current but also the previous and next symbols. To model such contributions, the spatio–temporal manifold vector due to the jth path of the ith user is modeled as

$$\mathfrak{h}_{ij} = \mathbf{s}_{ij} \otimes \mathbf{J}^{l_{ij}} \mathfrak{c}_{\mathbf{i}} \tag{6}$$

where $l_{ij} = \lceil \tau_{ij}/T_s \rceil$ is the discretized equivalent of the delay of the jth path of the ith user, with $\lceil \bullet \rceil$ being the roundup to nearest integer operator, and $\mathbf{s}_{ij}$ is the array manifold vector of the jth path of the ith user. In (6), the matrix $\mathbf{J}$ is the shift operator matrix, which is defined as

$$\mathbf{J} = \begin{bmatrix} 0 & 0 & \cdots & 0 & 0 \\ 1 & 0 & \cdots & 0 & 0 \\ 0 & 1 & \cdots & 0 & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & \cdots & 1 & 0 \end{bmatrix} = \begin{bmatrix} \mathbf{0}^T_{2qN_c-1} & 0 \\ \mathbf{I}_{2qN_c-1} & \mathbf{0}_{2qN_c-1} \end{bmatrix} \tag{7}$$

and is used to model the path delay effect. In particular, every time the matrix $\mathbf{J}$ (or $\mathbf{J}^T$) operates on a column vector, the contents of the vector are downshifted (or upshifted) by one position, with zeros being added to the top (or bottom) of the vector, that is, $\mathbf{J}^l \mathfrak{c}_i$ is a downshifted version of $\mathfrak{c}_i$ by l elements, and $(\mathbf{J}^T)^l \mathfrak{c}_i$ is an upshifted version of $\mathfrak{c}_i$ by l elements. Finally, the vector $\mathfrak{c}_i \in \mathcal{R}^{2qN_c}$ is defined as

$$\mathfrak{c}_i = \sum_{k=0}^{N_c-1} \alpha_i[k] \mathbf{J}^{kq} \boldsymbol{c} \tag{8}$$

where the vector $\boldsymbol{c}$ is the oversampled chip pulse-shaping waveform $c(t)$ extended with zeros, i.e., $\boldsymbol{c} = [c(0), c(T_s), \ldots, c((q-1)T_s), \mathbf{0}^T_{2qN_c-q}]^T$. If a rectangular chip pulse-shaping waveform is employed, then $\boldsymbol{c}$ is simplified to $\boldsymbol{c} = [\mathbf{1}^T_q, \mathbf{0}^T_{2qN_c-q}]^T$, which, if $q = 1$ (i.e., no oversampling), simplifies (8) to

$$\mathfrak{c}_i = \Bigg[\underbrace{\alpha_i[0], \alpha_i[1], \ldots, \alpha_i[N_c-1]}_{i\text{th user's PN-code}}, \mathbf{0}^T_{N_c} \Bigg]^T.$$

It is important to point out that in a code-reuse system, the ith user will share its PN-code sequence $\{\alpha_i[k], 0 \le k < N_c\}$ with a number of other users such that $\mathfrak{c}_i$ is common to all members of the group of co-code mobile users.

We will assume, without loss of generality, that the first user is the desired user. It can now be shown that $\boldsymbol{x}[n]$ can be expressed as a function of the desired user's nth (current) channel symbol $a_1[n]$ and the "Spatio–Temporal ARray (STAR) manifold vectors" $\mathfrak{h}_{ij}$, as well as the contributions of the previous and next channel symbols, as

$$\begin{aligned} \boldsymbol{x}[n] = &\sum_{i=1}^{M_c} (a_i[n]\mathbf{H}_i\boldsymbol{\beta}_i + a_i[n-1]\mathbf{H}_{i,\,\text{prev}}\boldsymbol{\beta}_i \\ &\qquad + a_i[n+1]\mathbf{H}_{i,\,\text{next}}\boldsymbol{\beta}_i) \\ &+ \sum_{i=M_c+1}^{M} (a_i[n]\mathbf{H}_i\boldsymbol{\beta}_i + a_i[n-1]\mathbf{H}_{i,\,\text{prev}}\boldsymbol{\beta}_i \\ &\qquad + a_i[n+1]\mathbf{H}_{i,\,\text{next}}\boldsymbol{\beta}_i) + \mathbf{n}[n] \end{aligned} \tag{9}$$

with

$$\begin{cases} \mathbf{H}_{i,\,\text{prev}} = (\mathbf{I}_N \otimes (\mathbf{J}^T)^{qN_c})\mathbf{H}_i \\ \mathbf{H}_{i,\,\text{next}} = (\mathbf{I}_N \otimes \mathbf{J}^{qN_c})\mathbf{H}_i \\ \boldsymbol{\beta}_i = [\beta_{i1}, \beta_{i2}, \ldots, \beta_{iK_i}]^T \\ \text{and } \mathbf{n}[n] \text{ representing the noise effects} \end{cases}$$

where $\mathbf{H}_i = [\mathfrak{h}_{i1}, \mathfrak{h}_{i2}, \ldots, \mathfrak{h}_{iK_i}]$ is the matrix with columns of the K_i "STAR" manifold vectors of the ith user. In (9), the first summation involves the effects of all M_c co-code users, including the desired (first) user, and can be rewritten as

$$\begin{aligned} &\underbrace{\mathbf{a}_1[n]\mathbf{H}_1\boldsymbol{\beta}_1}_{\text{desired}} + \underbrace{\sum_{i=2}^{M_c} \mathbf{a}_i[n]\mathbf{H}_i\boldsymbol{\beta}_i}_{\text{CCI}} \\ &\quad \underbrace{+ \sum_{i=1}^{M_c} (\mathbf{a}_i[n-1]\mathbf{H}_{i,\,\text{prev}}\boldsymbol{\beta}_i + \mathbf{a}_i[n+1]\mathbf{H}_{i,\,\text{next}}\boldsymbol{\beta}_i)}_{\text{ISI}} \end{aligned} \tag{10}$$

where the first term is the desired term, the second term represents the co-code interference (CCI) to the desired signal, and the third term represents the ISI effects (due to the previous and next data symbols of all co-code users including the desired user). It is clear that the "STAR" manifold vectors associated with the K_1 paths of the desired (first) user are linearly combined by the fading coefficient vector $\boldsymbol{\beta}_1$ (i.e., $\mathbf{H}_1\boldsymbol{\beta}_1$), and this will make these paths indistinguishable in their contribution to the signal subspace associated with the covariance matrix $\mathbf{R}_{xx}$ of the received signal $\boldsymbol{x}[n]$. The same is of course valid for the CCI. This looks like the well-known "coherence problem" of subspace-based estimation techniques, and as a consequence, it is not possible to estimate the desired users spatio–temporal channel parameters (path delays and directions) using signal subspace techniques such as MuSiC [6]. In the following section, an approach to overcome this problem is presented.

III. Channel Estimator and Receiver

Initially, let us define the matrix $\mathbf{C}_1$

$$\mathbf{C}_1 = [\mathbf{J}^1\mathfrak{c}_1, \mathbf{J}^2\mathfrak{c}_1, \ldots, \mathbf{J}^\ell\mathfrak{c}_1, \ldots, \mathbf{J}^{qN_c}\mathfrak{c}_1] \tag{11}$$

which is related to the first code $\mathfrak{c}_1$ [see (8)], employed by the desired (first) user and its co-code partners (that is, users 2 to M_c).

We now define what we call the preprocessor matrix $\mathbf{P}_\ell \in \mathcal{R}^{2qN_cN \times 2qN_cN}$ as

$$\mathbf{P}_\ell = \mathbf{I}_N \otimes P^\perp[\mathbf{C}_{1\ell}] \tag{12}$$

where $\mathbf{C}_{1\ell}$ is formed from the matrix $\mathbf{C}_1$ by removing its ℓth column. Applying the preprocessor $\mathbf{P}_\ell$ to the signal $\boldsymbol{x}[n]$ given by (9), i.e.,

$$\boldsymbol{y}_\ell[n] = \mathbf{P}_\ell \boldsymbol{x}[n] \tag{13}$$

implies that the preprocessor $\mathbf{P}_\ell$ is applied directly on the matrices $\mathbf{H}_i$, $\mathbf{H}_{i,\,\text{prev}}$ and $\mathbf{H}_{i,\,\text{next}}$ for every i from 1 to M (i.e., for every user). However, its effects on $\mathbf{H}_i$ are different from those on $\mathbf{H}_{i,\,\mathrm{prev}}$ and $\mathbf{H}_{i,\,\mathrm{next}}$ for the desired user (i.e., $i = 1$) and its co-code partners (i.e., $2 \leq i \leq M_c$). Analysis of the effects of the preprocessor $\mathbf{P}_\ell$ on the columns of the matrix $\mathbf{H}_i$ shows that (14), shown at the bottom of the page, is valid for users $1 \leq i \leq M_c$ but not valid for MAI and ISI contributions, which are "transformed" rather than "simplified."

Equation (14) means, in nonmathematical terms, that for the desired (first) user and its co-code partners, there will be no contributions to the preprocessed signal vector $\boldsymbol{y}_\ell[n]$ other than those from signals arriving through paths with delay equal to ℓT_s (if any such paths exist). This is because all columns of $\mathbf{H}_i$, for $1 \leq i \leq M_c$, except those corresponding to paths of delay equal to ℓT_s, will lie in the null space of the preprocessor matrix $\mathbf{P}_\ell$.

To emphasize this point, let us express the signal vector $\boldsymbol{y}_\ell[n]$ in a similar fashion to (9), as in (15), shown at the bottom of the page, and let us focus on its first line, in conjunction with a representative example of five co-code users ($M_c = 5$). Let us assume that only the second path ($j = 2$) of the desired user ($i = 1$) as well as the first path ($j = 1$) of one of the desired users co-code partners ($i = 4$, say) arrive with a delay equal to $3T_s$ (i.e., $l_{12} = l_{41} = 3$). In this case, $\ell = 3$, and the first line of (15) becomes

$$\underbrace{\mathbf{a}_1[n]\beta_{12}\mathbf{P}_3\mathfrak{h}_{12}}_{\text{desired term}} + \underbrace{\mathbf{0} + \mathbf{0} + \mathbf{a}_4[n]\beta_{41}\mathbf{P}_3\mathfrak{h}_{41} + \mathbf{0}}_{\text{CCI}} \tag{16}$$

indicating that by using this preprocessor $\mathbf{P}_3$, all paths of the second, third, and fifth users have been removed, whereas the coherence-like problem associated with the desired user (as well as with the fourth user) has been eliminated.

It is clear from the previous discussion that after the preprocessor (transformation), only the multipath rays (with delay ℓT_s) of the desired user and its co-code partners are described [see (6) and (13)] as $\mathbf{P}_\ell(\mathbf{s}(\theta) \otimes \mathbf{J}^\ell \mathfrak{c}_1)$. The locus of all vectors

$$\{\mathbf{P}_\ell(\mathbf{s}(\theta) \otimes \mathbf{J}^\ell \mathfrak{c}_1); \quad \forall\, \theta \in \Omega\}, \text{ where } \Omega = \text{ parameter space}$$

is a 1-D continuum (i.e., a curve) lying in an $2qN_cN$-dimensional complex space $\mathcal{C}^{2qN_cN}$ (observation space of the signal $\boldsymbol{y}_\ell[n]$). This curve is the transformed manifold-curve of the desired user and its co-code partners for this specific delay ℓT_s and does not include any contribution from the transformed ISI or MAI manifold vectors (which are described by different equations, i.e., belong to different manifold curves). The proposed algorithm exploits this property (using the received data) by aiming to find a "signal subspace" to which the transformed spatio–temporal manifold vectors, for delay ℓT_s, also belong. Therefore, *the intersection* of the manifold curve with the "overall signal-subspace" of $\mathbf{R}_\ell = \mathcal{E}\{\boldsymbol{y}_\ell[n]\boldsymbol{y}_\ell^H[n]\}$ will provide only the desired signal components.

Indeed, by forming the covariance matrix $\mathbf{R}_\ell$ of the preprocessed signal $\boldsymbol{y}_\ell[n]$ and then by partitioning the observation space of the $\mathbf{R}_\ell$ into the "signal-subspace" and "noise-subspace," it is clear from the above modeling and discussion that the "signal subspace" will contain the contributions of those paths with delays equal to ℓT_s as well as the transformed ISI and MAI. For example, in (16), the signal subspace of the covariance matrix $\mathbf{R}_3$ of the signal $\boldsymbol{y}_3[n]$ will include the vectors $\mathbf{P}_3\mathfrak{h}_{12}$ and $\mathbf{P}_3\mathfrak{h}_{41}$, as well as the contributions of the transformed ISI and transformed MAI. This implies that in the signal subspace, the desired term and the CCI term will be unresolvable from the transformed ISI and MAI. However, if we define the transformed "STAR" manifold curve for $\ell = 3$ as the locus of all manifold vectors $\mathbf{P}_3(\mathbf{s}(\theta) \otimes \mathbf{J}^3\mathfrak{c}_1)$, $\forall\, \theta$, then only the vectors $\mathbf{P}_3\mathfrak{h}_{12}$ and $\mathbf{P}_3\mathfrak{h}_{41}$ will belong to this manifold curve, as well as to the signal subspace. It is then clear through this representative example that by searching the transformed "STAR" manifold curve associated with the PN-code of the desired user, its intersection with the signal subspace will provide only the directional parameters of any existing path (associated with the desired user and its co-code partners) having delay equal to $3T_s$. If there is no path from the co-code set of users with delay equal to $3T_s$, then no intersection will exist. Fig. 2 shows a representation of the signal and noise subspaces and how the transformed "STAR" manifold curve

$$\mathbf{P}_\ell\mathbf{H}_i = \begin{cases} [\mathbf{0}, \mathbf{0}, \ldots, \mathbf{0}, \mathbf{P}_\ell\mathfrak{h}_{ij}\mathbf{0}, \ldots, \mathbf{0}, \mathbf{0}], & \text{if } l_{ij} = \ell \text{ (for a single path)} \\ [\mathbf{0}, \mathbf{0}, \ldots, \mathbf{0}, \mathbf{0}\mathbf{0}, \ldots, \mathbf{0}, \mathbf{0}], & \text{if } l_{ij} \neq \ell, \quad \forall\, j \end{cases} \tag{14}$$

$$\begin{aligned} \boldsymbol{y}_\ell[n] = {} & \underbrace{\mathbf{a}_1[n]\mathbf{P}_\ell\mathbf{H}_1\boldsymbol{\beta}_1 + \sum_{i=2}^{M_c}\mathbf{a}_i[n]\mathbf{P}_\ell\mathbf{H}_i\boldsymbol{\beta}_i}_{\text{transformed desired \& CCI}} \\ & + \underbrace{\sum_{i=1}^{M_c}(\mathbf{a}_i[n-1]\mathbf{P}_\ell\mathbf{H}_{i,\,\mathrm{prev}}\boldsymbol{\beta}_i + \mathbf{a}_i[n+1]\mathbf{P}_\ell\mathbf{H}_{i,\,\mathrm{next}}\boldsymbol{\beta}_i)}_{\text{transformed ISI}} \\ & + \underbrace{\sum_{i=M_c+1}^{M}(\mathbf{a}_i[n]\mathbf{P}_\ell\mathbf{H}_i\boldsymbol{\beta}_i + \mathbf{a}_i[n-1]\mathbf{P}_\ell\mathbf{H}_{i,\,\mathrm{prev}}\boldsymbol{\beta}_i + \mathbf{a}_i[n+1]\mathbf{P}_\ell\mathbf{H}_{i,\,\mathrm{next}}\boldsymbol{\beta}_i)}_{\text{transformed MAI}} + \mathbf{P}_\ell\mathbf{n}[n] \end{aligned} \tag{15}$$

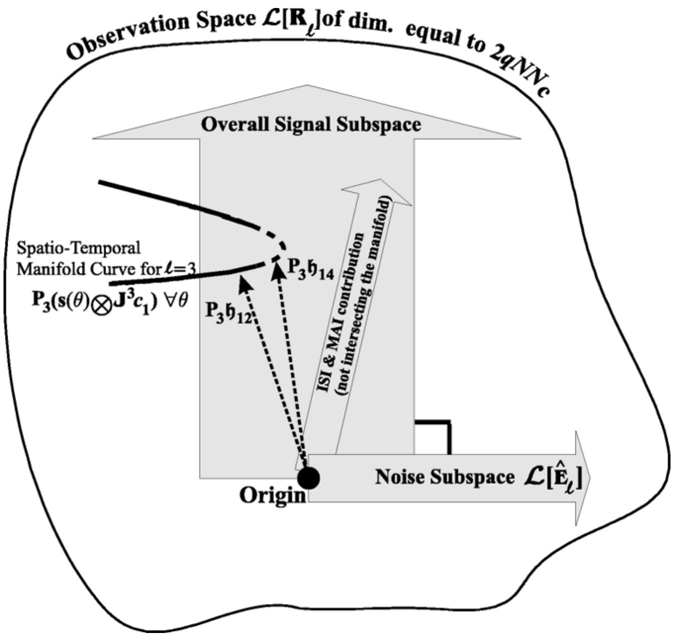

Fig. 2. Subspaces for the representative example.

will only intersect the signal subspace at certain points for the example discussed above.

Thus, in general terms, if there is a bank of preprocessors $\mathbf{P}_\ell$ for $\ell = 1, 2, \ldots, qN_c$, then the above-described procedure is translated to a number of 1-D searches over azimuth direction θ for every delay $1 \le \ell \le qN_c$, or, in an equivalent way, to a 2-D MuSiC-type cost function that is based on the "STAR" manifold vector concept and can be expressed as

$$\xi(\ell, \theta) = \frac{1}{(\mathbf{P}_\ell(\mathbf{s}(\theta) \otimes \mathbf{J}^\ell \mathfrak{c}_1))^H P\left[\hat{\mathbf{E}}_\ell\right](\mathbf{P}_\ell(\mathbf{s}(\theta) \otimes \mathbf{P}^\ell \mathfrak{c}_1))} \quad \forall\, \ell, \theta. \tag{17}$$

In (17), $P[\hat{\mathbf{E}}_\ell]$ is the projection operator of the estimated noise subspace $\mathcal{L}[\hat{\mathbf{E}}_\ell]$ obtained by the eigendecomposition of the ℓth preprocessed spatio–temporal covariance matrix $\mathbf{R}_\ell = \mathcal{E}\{\boldsymbol{y}_\ell[n]\boldsymbol{y}_\ell^H[n]\}$. The cost function $\xi(\ell, \theta)$ would be expected to be approximately zero for delays and directions where paths do not exist and would be expected to give large peaks for delays and directions at which paths do exist.

In summary, the action of the ℓth preprocessor of the parallel bank can then be stated as follows. For a particular active code shared by the desired user and its co-code partners, the ℓth preprocessor acts to null the current bit interference arising from this set of users due to paths arriving with delay not equal to ℓT_s. In terms of the signal subspace, the preprocessor action can be considered to be the conversion of a single dimension contribution (for each co-code user) arising due to the coherence-like problem to either a single dimensional contribution (for each co-code user), corresponding to a path of delay ℓT_s, or to a null vector if no path of delay ℓT_s exists. Each of these single-dimension contributions (if any such contributions exist) to the ℓth preprocessed space–time covariance matrix signal subspace will be of the form of a modified "STAR" manifold vector. Thus, a peak search of the spectrum obtained by the evaluation of (17) will provide the set of angle ϑ and delay $\breve{\mathfrak{l}}$ estimates

$$\Xi = \left\{ \left(\vartheta_1, \breve{\mathfrak{l}}_1\right), \left(\vartheta_2, \breve{\mathfrak{l}}_2\right), \ldots, \left(\vartheta_{K_{\text{est}}}, \breve{\mathfrak{l}}_{K_{\text{est}}}\right) \right\}$$

where K_{est} is the total number of estimated paths (given by the total number of peaks in the spectrum) corresponding to the desired (first) user and its co-code partners.

It should be noted that in this approach, there is no information requirement beyond the active code of the desired user and its co-code partners. This is sufficient to achieve for each possible delay a single contribution, with estimatable parameters, in the signal subspace for each of the co-code users.

The proposed system is illustrated in Fig. 3, which shows the described parallel bank of preprocessors ($\mathbf{P}_\ell$ for $1 \le \ell \le qN_c$) alongside the peak search of the spectrum obtained through the evaluation of (17), which will provide the set of parameters Ξ. From this set, in the code-reuse system, the subset Ξ_1 of Ξ (i.e., $\Xi_1 \subseteq \Xi$) of channel parameters associated with the first co-code user need to be identified. To achieve this, a weight matrix $\mathbf{W}$ is constructed in order to be employed by a "correlation analysis assignment" block. This matrix $\mathbf{W} \in \mathcal{C}^{2qN_cN \times K_{\text{est}}}$ has columns that are functions of the "STAR" manifold vectors $\mathfrak{h}_k = \mathbf{s}(\vartheta_k) \otimes \mathbf{J}^{\breve{\mathfrak{l}}_k}\mathfrak{c}_1$ associated with the set of parameters Ξ. In particular, its kth column is described (a derivation can be found in the Appendix) as

$$\mathbf{w}_k = \beta_k \mathbf{R}_{\breve{\mathfrak{l}}_k}^{\dagger} \mathbf{P}_{\breve{\mathfrak{l}}_k} \mathfrak{h}_k \tag{18}$$

where $\mathbf{R}_{\breve{\mathfrak{l}}_k} = \mathbf{P}_{\breve{\mathfrak{l}}_k}\mathbf{R}_{xx}\mathbf{P}_{\breve{\mathfrak{l}}_k}$, and $\dagger$ denotes a pseudo inverse-type operator.

Based on $\mathbf{W}$ operating on a received signal frame of L channel symbols, the matrix $\mathbf{Z} \in \mathcal{C}^{K_{\text{est}} \times L}$ is formed as

$$\mathbf{Z} = \mathbf{W}^H[\boldsymbol{x}[n], \boldsymbol{x}[n+1], \ldots, \boldsymbol{x}[n+L-1]] \tag{19}$$

implying that each column of the matrix $\mathbf{W}$ can be seen as a "pseudo" single path receiver weight vector that could be used to recover the symbols from the user to whom this single path corresponds. Consequently, this implies that each row of the matrix $\mathbf{Z}$ contains the decision variables for a frame of L symbols from the output of each of the K_{est} single path receiver weight vectors. If these decision variables are highly correlated (i.e., based on the correlation matrix of $\mathbf{Z}$), then the associated parameters from the set Ξ should belong to the same user. We define this process as the "correlation analysis assignment." The assignment process can be summarized as saying that those outputs of the single path receiver weight vectors that are highly correlated must belong to transmissions from the same user, and hence, those single-path weight vectors must have been formed from estimated channel parameters that can be identified to belong to the same user. Having assigned all the identified paths amongst the co-code users and with $\Xi_1 = \{(\hat{\theta}_{1j}, \hat{l}_{1j}); \quad \forall j \in [1, \ldots, \hat{K}_1]\}$ denoting the subset of the desired user, the actual single-user receiver for the desired user can be formed as

$$\boldsymbol{w}_1 = \sum_{j=1}^{\hat{K}_1} \hat{\beta}_{1j} \mathbf{R}_{\hat{l}_{1j}}^{\dagger} \mathbf{P}_{\hat{l}_{1j}} \hat{\mathfrak{h}}_{1j} \tag{20}$$

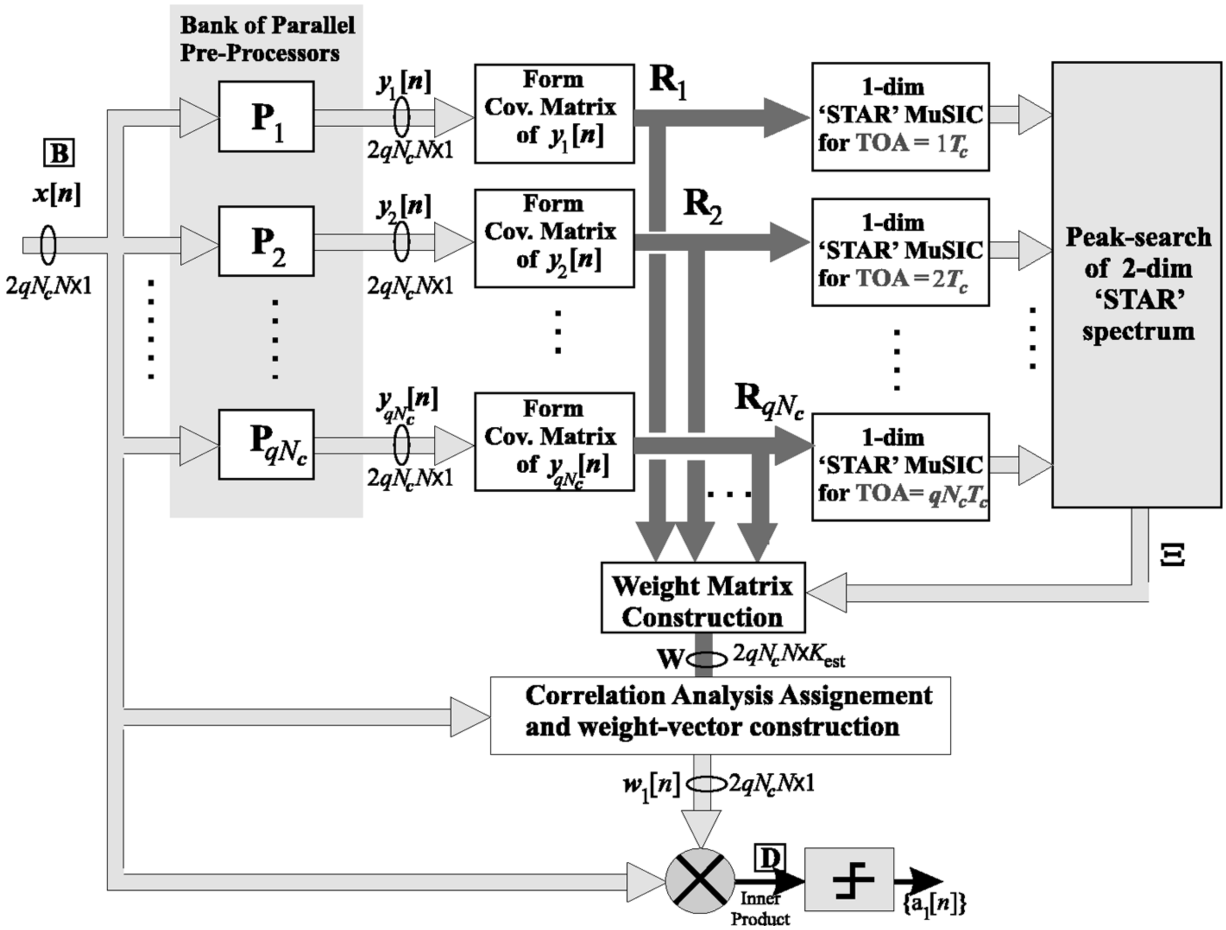

Fig. 3. Structure of the proposed code-reuse array receiver.

where $\hat{\mathfrak{h}}_{1j} = \mathbf{s}(\hat{\theta}_{1j}) \otimes \mathbf{J}^{\hat{l}_{1j}} \mathfrak{c}_1$. The desired (first) user's receiver weight vector could then be followed by a decision device. For example, the nth channel symbol decision can be taken as

$$\hat{\mathbf{a}}_1[n] = \text{sign}\left(\Re\text{e}\left\{\boldsymbol{w}_1^H \boldsymbol{x}[n]\right\}\right) + j \cdot \text{sign}\left(\Im\text{m}\left\{\boldsymbol{w}_1^H \boldsymbol{x}[n]\right\}\right). \tag{21}$$

Some comments relating to the proposed algorithm are as follows.

- The receiver represented by (20) is formed on a path by path basis and can be formed for any number of estimated paths. It is not necessary that every existent path for a particular user be identified, and in this way, the proposed receiver is robust to incomplete channel estimation.
- Should the estimated parameters of a particular identified path be erroneous, the "correlation analysis assignment" stage of the proposed algorithm will leave that path unassigned, and hence, the proposed algorithm is robust to the presence of partial channel estimation errors.

IV. Simulation Results

A number of computer simulations are now presented to show the performance of the proposed methods. For the purpose of simulation, we propose two code reuse strategies.

Strategy- 1: The active PN codes are assigned such that no code is assigned $(C+1)$ times until all other codes have been assigned at least $C(\geq 1)$ times. The code-reuse factor (CRF) can then be defined as CRF= $\lceil M/N_A \rceil$, where N_A is the number of active codes.

Strategy- 2: Each of the M users selects at random a code from the set of N_A active codes such that CRF= $\lceil \mathcal{E}\{M/N_A\} \rceil$. Such a strategy removes the requirement for code allocation by the base station and, hence, can be used to simplify the overall design of the communication system.

For the purposes of simulation, the spatio–temporal channel parameters are taken from statistical distributions that arise from an experimentally verified macrocellular multipath channel model [21] that assumes a sparse scattering environment. The path directions are taken from a Gaussian distribution with a standard deviation of 20° and a mean value that is uniformly distributed between 40° and 140°. The path delays are assumed to be exponentially distributed. The path coefficients are ordered such that the strongest coefficients are associated with the shortest paths, that is, $|\beta_{i1}| \geq |\beta_{i2}| \cdots \geq |\beta_{iK_i}| \; \forall i$.

For the purpose of performance comparison with the proposed receiver weight vector, a number of benchmark receiver weight vectors are used in the simulations. Using the notation employed throughout this paper, these vectors can be defined as follows.

ST-RAKE:

$$\boldsymbol{w}_{1,\,\text{RAKE}} = \mathbf{H}_1 \boldsymbol{\beta}_1 \tag{22}$$

ST-Decorrelating Detector:

$$\boldsymbol{w}_{1,\,\mathrm{DEC}} = \mathrm{row}_2 \Big([\mathbf{H}_{1,\,\mathrm{prev}}\boldsymbol{\beta}_1,\, \mathbf{H}_1\boldsymbol{\beta}_1,\, \mathbf{H}_{1,\,\mathrm{next}}\boldsymbol{\beta}_1 \; \mathbf{H}_{2,\,\mathrm{prev}}\boldsymbol{\beta}_2,\, \mathbf{H}_2\boldsymbol{\beta}_2,\, \mathbf{H}_{2,\,\mathrm{next}}\boldsymbol{\beta}_2,\, \ldots \; \mathbf{H}_{M,\,\mathrm{prev}}\boldsymbol{\beta}_M,\, \mathbf{H}_M\boldsymbol{\beta}_M,\, \mathbf{H}_{M,\,\mathrm{next}}\boldsymbol{\beta}_M]^{\dagger} \Big)^H \quad (23)$$

ST-Decorr (Limited Information):

$$\boldsymbol{w}_{1,\,\mathrm{DEC\text{-}LIM}} = \mathrm{row}_2 \Big(\Big[\overline{\mathbf{H}}^{(k)}_{1,\,\mathrm{prev}}\boldsymbol{\beta}_1,\, \overline{\mathbf{H}}^{(k)}_1\boldsymbol{\beta}_1,\, \overline{\mathbf{H}}^{(k)}_{1,\,\mathrm{next}}\boldsymbol{\beta}_1 \; \overline{\mathbf{H}}^{(k)}_{2,\,\mathrm{prev}}\boldsymbol{\beta}_2,\, \overline{\mathbf{H}}^{(k)}_2\boldsymbol{\beta}_2,\, \overline{\mathbf{H}}^{(k)}_{2,\,\mathrm{next}}\boldsymbol{\beta}_2,\, \ldots \; \overline{\mathbf{H}}^{(k)}_{M,\,\mathrm{prev}}\boldsymbol{\beta}_M,\, \overline{\mathbf{H}}^{(k)}_M\boldsymbol{\beta}_M,\, \overline{\mathbf{H}}^{(k)}_{M,\,\mathrm{next}}\boldsymbol{\beta}_M \Big]^{\dagger} \Big)^H \quad (24)$$

where $\overline{\mathbf{H}}_i^{(k)} = [\mathfrak{h}_{i1}, \mathfrak{h}_{i2}, \ldots, \mathfrak{h}_{i(K_i-k)}, \mathbf{O}_{2qN_cN\times k}]$, that is, $\overline{\mathbf{H}}_i^{(k)}$ is the matrix $\mathbf{H}_i$ with the last k (with $k < K_i$) columns replaced by zeros. This is equivalent to assuming that the weakest k paths have not been estimated (i.e., the receiver is formed with limited information). The ST-RAKE receiver has been chosen as a lower benchmark since it is the optimum single user receiver under the condition that multiuser interference is treated as additive white Gaussian noise and is also close to the type of receiver that is to be used in existing commercial CDMA systems (IS-95 and some UMTS). The decorrelating detector serves as a good upper benchmark since it has been shown to have a performance that is close to that of the optimal multiuser detector proposed by Verdu [24]. The decorrelating detector is also considered in the presence of limited information (i.e., when the single weakest path remains unestimated, $k = 1$) to satisfy the requirement of a benchmark against which the performance of the proposed receiver can be compared in the presence of only limited channel information.

Simulation 1—Analysis of the Proposed Channel Parameter Estimation Technique: Here, a representative example of the proposed channel parameter estimation technique is given when a uniform linear array of five antennas ($N = 5$) with half-wavelength spacing is used. Table I provides the simulation parameters, whereas Table II gives the unknown channel parameters associated with the desired user and its three co-code partners. These channel parameters represent one realization of the statistical channel parameter distributions described in the introduction of Section IV. The experimental received signal $\boldsymbol{x}[n]$ is formed over 500 QPSK symbols with the channel parameters assumed to remain fixed over this observation interval of 500 symbols.

The input signal-to-noise ratio (SNR) associated with the desired (first) user is assumed equal to 10 dB. However, in order to create a near–far scenario, all the interfering users (including co-code interferers) are assumed equal powered, and the signal-to-interference ratio is assumed to be -30 dB (near–far ratio: NFR $= 30$ dB). Fig. 4 presents the spectrum of (17) based on the "STAR" manifold vector and obtained using the proposed method, whereas Table III gives the set Ξ of the estimated (DOA, TOA) parameters.

TABLE I
SIMULATION PARAMETERS

Parameter	Value
No. of Users, M	40
No. of Antennas, N	5
Paths per User, K_i $\forall i$	6
Processing Gain, N_c	31
PN-codes	Gold Sequences
Number of Active Codes	12

TABLE II
DESIRED AND CO-CODE USERS SPACE–TIME CHANNEL PARAMETERS

	1^{st} User (Desired)			2^{nd} User		
	TOA	DOA	Path Coefficients	TOA	DOA	Path Coefficients
Path	l_{1j}	θ_{1j}	β_{1j} ($\Vert\boldsymbol{\beta}_1\Vert = 1$)	l_{2j}	θ_{2j}	β_{2j} ($\Vert\boldsymbol{\beta}_2\Vert^2 = 1000$)
$j=1$	$1T_c$	91°	$-0.3019 - j\,0.5287$	$1T_c$	142°	$+5.7939 + j23.4891$
$j=2$	$3T_c$	104°	$-0.4325 - j\,0.3558$	$4T_c$	81°	$+7.8624 - j13.4327$
$j=3$	$8T_c$	139°	$-0.1543 + j\,0.3299$	$8T_c$	63°	$-10.7869 + j6.7995$
$j=4$	$11T_c$	112°	$-0.0527 + j\,0.2640$	$11T_c$	73°	$-0.9610 + j2.5814$
$j=5$	$15T_c$	149°	$+0.1829 - j\,0.1610$	$15T_c$	67°	$-0.9961 + j0.8978$
$j=6$	$25T_c$	131°	$+0.2224 - j\,0.0418$	$23T_c$	73°	$+0.6151 - j0.2906$
	3^{rd} User			4^{th} User		
	TOA	DOA	Path Coefficients	TOA	DOA	Path Coefficients
Path	l_{3j}	θ_{3j}	β_{3j} ($\Vert\boldsymbol{\beta}_3\Vert^2 = 1000$)	l_{4j}	θ_{4j}	β_{4j} ($\Vert\boldsymbol{\beta}_4\Vert^2 = 1000$)
$j=1$	$2T_c$	83°	$+23.7113 + j3.5921$	$2T_c$	116°	$+19.3972 + j11.5491$
$j=2$	$5T_c$	78°	$-17.8917 + j9.7625$	$4T_c$	19°	$+15.8028 - j13.4504$
$j=3$	$7T_c$	72°	$-1.7619 - j1.6855$	$6T_c$	60°	$-1.9658 + j1.3654$
$j=4$	$18T_c$	62°	$+1.5597 - j0.1198$	$9T_c$	73°	$-4.1206 - j4.2266$
$j=5$	$25T_c$	54°	$+0.8572 + j0.4416$	$13T_c$	103°	$+0.1180 - j3.3415$
$j=6$	$26T_c$	55°	$-0.1452 + j0.3335$	$18T_c$	78°	$-0.1297 + j2.8207$

From Fig. 4 and Table III (first three columns), it can be seen that the channel parameters (delay and direction) of all 24 paths of the four co-code users are correctly estimated. The subset of parameters Ξ_1 of Ξ (i.e., $\Xi_1 \subseteq \Xi$) associated with desired user remain to be identified. These can be found by forming the $\mathbf{W}$ matrix and applying this to the received signal (corresponding to a frame of 500 symbols). This will provide the matrix $\mathbf{Z}$. Four columns of the correlation matrix of the rows of $\mathbf{Z}$ corresponding to the four co-code users of the code $\mathfrak{c}_1$ are shown in Table III. By finding the values that are greater than a threshold of $\gamma = 0.95$ (say), it is clear that the first, fifth, 12th, 15th, 18th, and 23rd sets of estimated parameters belong to the desired user (see 1st user's column). Note also from Table III (correlation analysis results) that the maximum values of the last three columns will identify the parameters associated with the remaining three co-code users.

Let us now consider the hypothetical case that there exists a channel parameter estimation error in one of the estimated paths. For example, let us say that all channel parameters for all paths other than the first identified path [which belongs to the desired (first) user and has true channel parameters (91°, $1T_c$)] have been estimated correctly. If the estimation error is directional of

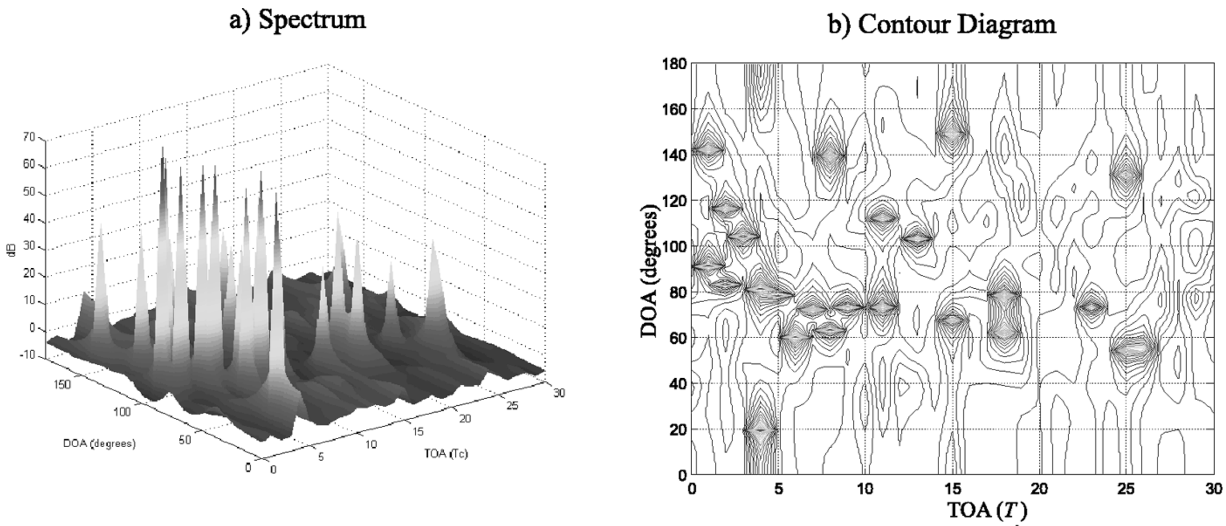

Fig. 4. Two-dimensional spectrum (and the associated contour diagram) that provides the parameters of the desired user and its co-code partners.

TABLE III
ESTIMATED SET (Ξ) AND CORRELATION ANALYSIS

	Ξ		Correlation Analysis Results			
	TOA, $\breve{l}_j$	DOA, ϑ_j	1^{st} user	2^{nd} user	3^{rd} user	4^{th} user
1^{st}	$1T_c$,	91°	**1.00**	0.03	0.00	−0.02
2^{nd}	$1T_c$,	142°	0.03	**1.00**	−0.04	0.08
3^{rd}	$2T_c$,	83°	0.00	−0.04	**1.00**	0.05
4^{th}	$2T_c$,	116°	−0.02	0.08	0.05	**1.00**
5^{th}	$3T_c$,	104°	**0.97**	−0.01	0.00	−0.02
6^{th}	$4T_c$,	19°	−0.04	0.00	0.00	**0.98**
7^{th}	$4T_c$,	81°	0.00	**0.98**	−0.03	0.00
8^{th}	$5T_c$,	78°	0.01	−0.04	**0.98**	0.00
9^{th}	$6T_c$,	60°	−0.03	0.08	0.00	**0.99**
10^{th}	$7T_c$,	72°	0.01	−0.04	**0.99**	0.00
11^{th}	$8T_c$,	63°	0.03	**0.99**	−0.03	0.08
12^{th}	$8T_c$,	139°	**0.97**	0.02	−0.01	−0.02
13^{th}	$9T_c$,	73°	−0.03	0.08	0.00	**0.99**
14^{th}	$11T_c$,	73°	0.02	**0.99**	−0.04	0.08
15^{th}	$11T_c$,	112°	**0.97**	0.02	−0.01	−0.02
16^{th}	$13T_c$,	103°	−0.03	0.08	0.00	**0.99**
17^{th}	$15T_c$,	67°	0.02	**0.99**	−0.04	0.08
18^{th}	$15T_c$,	149°	**0.96**	0.02	−0.01	−0.03
19^{th}	$18T_c$,	62°	0.01	−0.04	**0.99**	0.05
20^{th}	$18T_c$,	78°	−0.03	0.07	0.05	**0.99**
21^{st}	$23T_c$,	73°	0.00	**0.99**	−0.04	0.08
22^{nd}	$25T_c$,	54°	0.00	−0.04	**0.99**	0.00
23^{rd}	$25T_c$,	131°	**0.96**	0.01	0.00	−0.02
24^{th}	$26T_c$,	55°	0.01	−0.03	**0.97**	0.01

$+5°$, say, (i.e., $\hat{\vartheta}_1 = 96°$ rather than 91°), then the "correlation analysis assignment" will provide almost identical results to those shown in Table III, with the only difference being that the first element of the first user's column will be 0.55 rather than 1. In fact, the correlation with the identified paths of the desired (first) user is only close to unity if there is no estimation error. Therefore, the path identified with errors in its estimated channel parameters will not been recognized as a path associated with the desired (first) user or any of its co-code partners. The fact that the path with estimation errors remains unassigned means that it will not effect the formation of the proposed receiver weight vector $\boldsymbol{w}_1$ for the desired (first) user, and in this way, the proposed receiver is robust to partial channel estimation errors.

Simulation 2—Analysis of Variation of Numbers of Users: Simulation results are now presented to show the variation in $\mathrm{SNIR}_{\mathrm{out}}$ with an increasing number of users. The simulations assume that channel estimation has already been performed. The benchmark receivers (ST-RAKE receiver, space–time decorrelating detector) are evaluated for the case of complete channel information, such that the receivers are based on all the required channel information assumed known with complete accuracy [see (22) and (23), respectively]. In addition, the space–time decorrelating detector is considered for the case of incomplete channel estimation where the weakest of the K_i paths [hence, (24) with $k = 1$] of each user is assumed to be unestimated. The proposed receiver weight vector is evaluated for three cases. That is, it is assumed that the desired user's channel parameters have been estimated correctly for only

1) the single strongest (shortest) path, i.e., $\hat{K}_1 = 1$;
2) the three strongest (shortest) paths, i.e., $\hat{K}_1 = 3$;
3) the single weakest (longest) path, i.e., $\hat{K}_1 = 1$.

The simulation parameters are as shown in Table I. The interfering users are of power NFR = 20 dB higher than the desired user (to create a severe near–far scenario), and the desired user SNR = 20 dB. The channel parameters of all users are taken from the statistical distributions discussed above. The results shown are obtained as an average over 40 simulation runs for 500 channel symbols each, where for each simulation run the active codes ($N_A = 12$) are taken randomly from the complete set of 31 codes. Figs. 5 and 6 plot the output SNIR against the

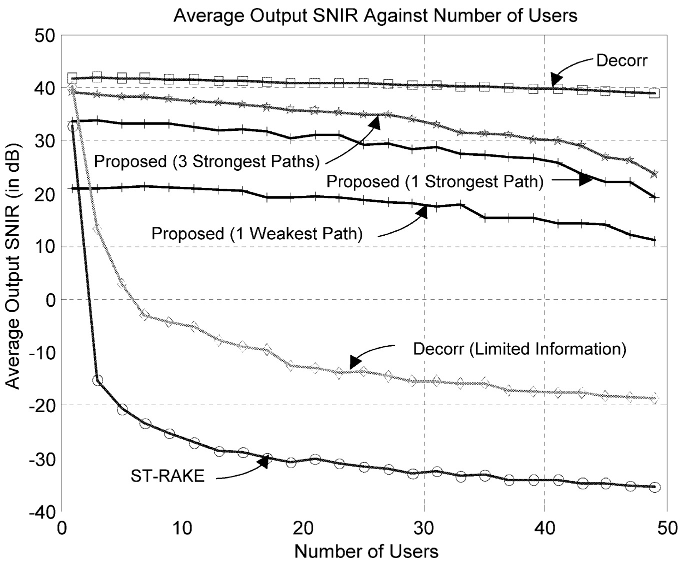

Fig. 5. $\mathrm{SNIR}_{\mathrm{out}}$ as a function of the number of users for code reuse Strategy 1.

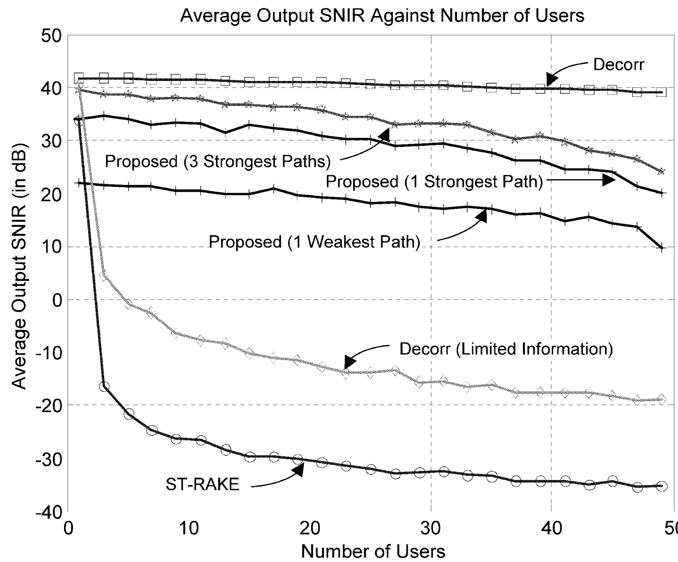

Fig. 6. $\mathrm{SNIR}_{\mathrm{out}}$ as a function of the number of users for code reuse Strategy 2.

number of users for the benchmark and proposed receivers in the cases of code-reuse Strategy 1 (see Fig. 5) and Strategy 2 (see Fig. 6). From Fig. 5, it can be seen that the proposed receiver weight vectors maintain an acceptably interference free output as the number of users is increased, even in cases where each active PN-code is used by as many as four uncoordinated users simultaneously. It can also be seen that in the case where only the single weakest path can be identified for the desired user, on average, a proposed receiver can still be formed whose output can be used to make reliable channel symbol decisions, and this shows that the proposed receiver weight vector is robust against incomplete channel estimation. The figure also shows the deterioration that plagues the space–time decorrelating detector in the case where the receiver is formed with accurate but limited channel information. Fig. 6 shows that in the case that the system has no control over code allocation, such that each user selects at random a code from the active set (Strategy 2), the proposed method performs similarly to the case when the base station controls code allocation (Strategy 1).

Simulation 3—Comparison of Decision Variables for Different Receivers: In Fig. 7, the decision variables (at point D of Fig. 3) are plotted for the benchmark and proposed receiver weight vectors for the environment described in Simulation 1. The results shown in Fig. 7 represent one realization over 500 QPSK channel symbols. From this figure, it is obvious that the space–time decorrelating detector provides the best results [Fig. 7(a)], but its performance deteriorates dramatically [Fig. 7(b)] if only limited information is available. The performance of the RAKE receiver is the worst [Fig. 7(c)], whereas the proposed procedure provides a performance close to the decorrelating receiver, even if three paths [Fig. 7(d)] or only the strongest path [Fig. 7(e)] is estimated. For illustrative purposes, Fig. 7(f) presents the decision variables when only the weakest path is used in the proposed approach.

Simulation 4—Demonstration of Near–Far Resistance: To demonstrate the near–far resistance of the proposed scheme, the output SNIR as a function of the interfering user power is evaluated for the environment described in Simulation 2. With SNR $= 20$ dB for the desired user, the powers of the interfering user varies from 0 to 60 dB. The results shown are obtained as an average over 40 simulation runs for 500 channel symbols each, where for each simulation run, the $N_A = 12$ active codes are taken randomly from the complete set of 31 codes. From Figs. 8 and 9, it can be seen that for both proposed code reuse strategies, the proposed receiver shows near–far resistance. Again, it is interesting to observe the case of the limited space–time decorrelating detector. From the figures, it can be clearly seen that there is a rapid deterioration in this receiver's performance as the NFR is increased. The figures demonstrate that the proposed receiver weight vector unlike the limited space–time decorrelating detector is near–far resistant in face of limited channel information.

From Simulations 2 and 4, it can be seen that the suggestion that code assignment by the base station can be eliminated is justified since there is little difference in observed system performance under the two proposed code-reuse strategies.

V. Conclusions

A single-code multipath channel estimator and single-user receiver weight vector have been proposed to facilitate code reuse, such that each active PN-code can be used simultaneously by a number of users. The proposed receiver is proven to be near–far resistant and robust against incomplete channel estimation and partial channel estimation errors. The proposed method uses joint spatio–temporal processing and would therefore be expected to have a performance that is superior to that which can be achieved by existing code reuse techniques that implement spatial and temporal processing independently with the separation of co-code users being carried out solely in either domain. The proposed approach has been shown to handle co-code users so long as there exists a single estimatable path that is separated in at least one domain (space or time) for each user in a group of co-code users. It has also been shown that code assignment by the base station can be eliminated since there is little difference in observed system performance if each user is

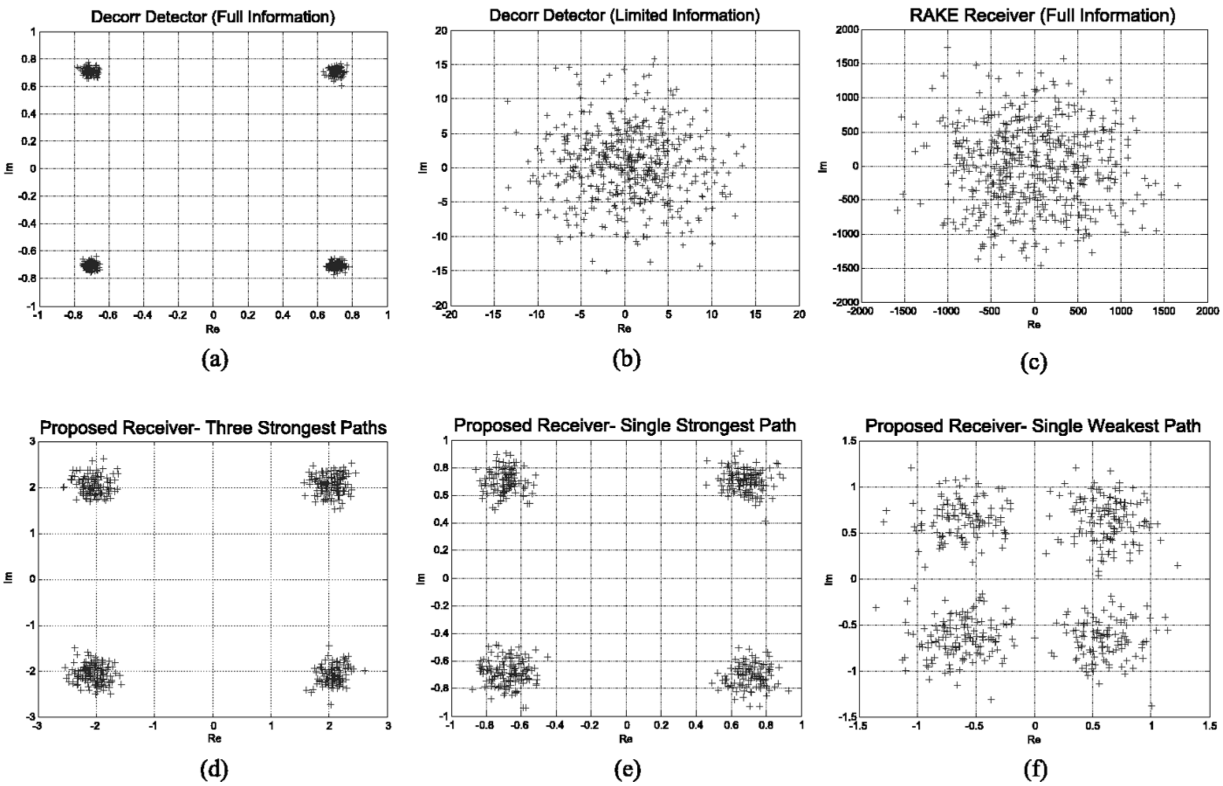

Fig. 7. Decision variables for 500 channel symbols for a single simulation run.

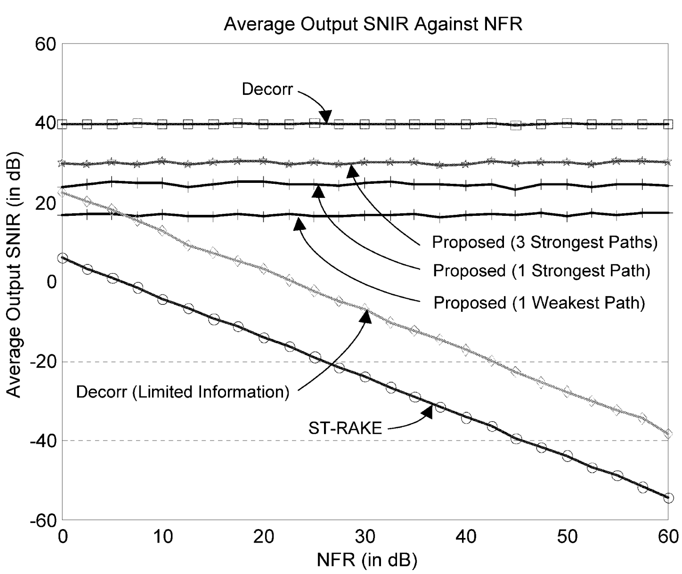

Fig. 8. $\mathrm{SNIR}_{\mathrm{out}}$ as the NFR increases for code reuse Strategy 1.

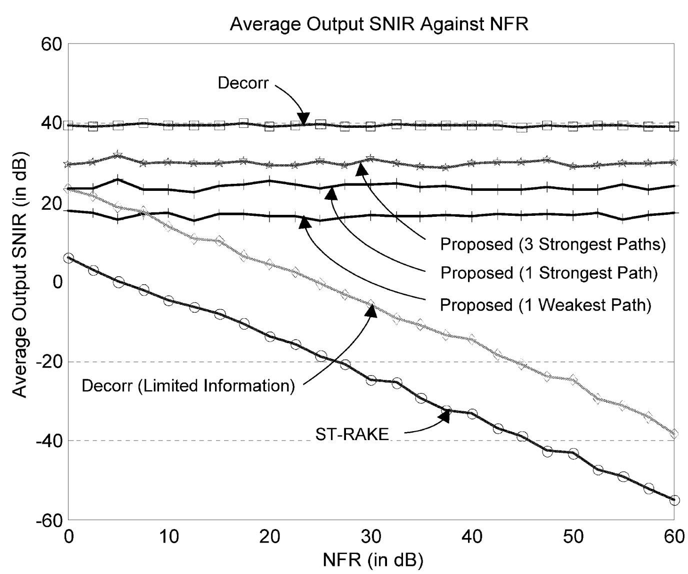

Fig. 9. $\mathrm{SNIR}_{\mathrm{out}}$ as the NFR increases for code reuse Strategy 2.

allowed to choose at random its own PN-code sequence, rather than using a sequence assigned by the base station.

APPENDIX
DERIVATION OF (18)

In the case of the single path receiver, only the estimated parameters of one path (the kth path, say) will be known. Thus, the information available to form a receiver weight vector is limited to β_k (single path estimated complex path coefficient) and $\mathfrak{h}_k = \mathbf{s}(\vartheta_k) \otimes \mathbf{J}^{\breve{\mathfrak{l}}_k} \mathfrak{c}_1$, with $(\vartheta_k, \breve{\mathfrak{l}}_k) \in \mathbf{\Xi}$ denoting the kth path's estimated direction and delay, respectively, where

$$\mathbf{\Xi} = \left\{ \left(\vartheta_1, \breve{\mathfrak{l}}_1\right), \left(\vartheta_2, \breve{\mathfrak{l}}_2\right), \ldots, \left(\vartheta_{K_{\mathrm{est}}}, \breve{\mathfrak{l}}_{K_{\mathrm{est}}}\right) \right\}.$$

The PN-code $\mathfrak{c}_1$ being used is also known; however, in the code reuse system, this is shared by a number of users such that the kth identified path could belong to anyone of the co-code users. In other words, $(\vartheta_k, \breve{\mathfrak{l}}_k)$ corresponds to the ith user's jth path parameters (θ_{ij}, l_{ij}), where i could be anyone of $i \in$

$(1, \ldots, M_c)$, with M_c denoting the number of co-code users. That is

$$\left(\vartheta_k,\, \check{\mathfrak{l}}_k\right) = (\theta_{ij},\, l_{ij}) \tag{25}$$

$$\Rightarrow \underbrace{\mathbf{s}(\vartheta_k) \otimes \mathbf{J}^{\check{\mathfrak{l}}_k}\mathfrak{c}_1}_{=\mathfrak{h}_k} = \underbrace{\mathbf{s}(\theta_{ij}) \otimes \mathbf{J}^{l_{ij}}\mathfrak{c}_1}_{=\mathfrak{h}_{i,j}}. \tag{26}$$

Using the proposed spatio–temporal preprocessor $\mathbf{P}_\ell$ with $\ell = \check{\mathfrak{l}}_k (= l_{ij})$, a complex weight vector $\mathbf{w}_k \in \mathcal{C}^{2qN_cN}$ can be formed such that

$$\begin{aligned} d_k[n] &= \mathbf{w}_k^H \boldsymbol{y}_{\check{\mathfrak{l}}_k}[n] \\ &= \mathbf{w}_k^H \mathbf{P}_{\check{\mathfrak{l}}_k} \boldsymbol{x}[n] \end{aligned} \tag{27}$$

where $d_k[n]$ is the ith user's decision device input for the nth symbol period, and $\boldsymbol{x}[n]$ is the nth symbol period discretized received signal [see (9)]. The difference between the desired symbol $g_k[n] = \mathbf{a}_i[n]$ and $d_k[n]$ can be expressed as an error

$$\begin{aligned} \epsilon_k[n] &= g_k[n] - \mathbf{w}_k^H \mathbf{P}_{\check{\mathfrak{l}}_k} \boldsymbol{x}[n] \\ \Rightarrow \epsilon_k^2[n] &= g_k^2[n] - 2\mathrm{Re}\left(g_k^*[n]\mathbf{w}_k^H \mathbf{P}_{\check{\mathfrak{l}}_k} \boldsymbol{x}[n]\right) \\ &\quad + \mathbf{w}_k^H \mathbf{P}_{\check{\mathfrak{l}}_k} \boldsymbol{x}[n]\boldsymbol{x}[n]^H \mathbf{P}_{\check{\mathfrak{l}}_k} \mathbf{w}_k \end{aligned} \tag{28}$$

where $\mathbf{P}_{\check{\mathfrak{l}}_k} = \mathbf{P}_{\check{\mathfrak{l}}_k}^H$ has been used. Taking the expectation of both sides

$$\begin{aligned} \mathcal{E}\left\{\epsilon_k^2[n]\right\} = \mathcal{E}\left\{g_k^2[n]\right\} - 2\mathrm{Re}\left(\mathbf{w}_k^H \mathbf{P}_{\check{\mathfrak{l}}_k} \boldsymbol{\rho}_{gx}\right) \\ + \mathbf{w}_k^H \mathbf{P}_{\check{\mathfrak{l}}_k} \mathbf{R}_{xx} \mathbf{P}_{\check{\mathfrak{l}}_k} \mathbf{w}_k \end{aligned} \tag{29}$$

where $\mathbf{R}_{xx}$ is the spatio–temporal covariance matrix, and the cross covariance vector is

$$\boldsymbol{\rho}_{gx} = \mathcal{E}\left\{g_k^*[n]\boldsymbol{x}[n]\right\} = \mathbf{H}_i \boldsymbol{\beta}_i. \tag{30}$$

The objective in a least mean square sense is to minimize the expected value of the squared error. Differentiating (29) with respect to the weight vector $\mathbf{w}_k$ and then setting this equal to zero for minimization, we have

$$\frac{\partial}{\partial \mathbf{w}_k}\left(\mathcal{E}\left\{\epsilon_k^2[n]\right\}\right) = -\mathbf{P}_{\check{\mathfrak{l}}_k}\boldsymbol{\rho}_{xg} + \mathbf{P}_{\check{\mathfrak{l}}_k}\mathbf{R}_{xx}\mathbf{P}_{\check{\mathfrak{l}}_k}\mathbf{w}_k = 0 \tag{31}$$

$$\begin{aligned} \Rightarrow \mathbf{w}_k &= \left(\mathbf{P}_{\check{\mathfrak{l}}_k}\mathbf{R}_{xx}\mathbf{P}_{\check{\mathfrak{l}}_k}\right)^\dagger \mathbf{P}_{\check{\mathfrak{l}}_k}\boldsymbol{\rho}_{xg} \\ &= \left(\mathbf{P}_{\check{\mathfrak{l}}_k}\mathbf{R}_{xx}\mathbf{P}_{\check{\mathfrak{l}}_k}\right)^\dagger \mathbf{P}_{\check{\mathfrak{l}}_k}\mathbf{H}_i\boldsymbol{\beta}_i. \end{aligned} \tag{32}$$

However, using (14) and taking into account the relationship of (25) and (26), the term $\mathbf{P}_{\check{\mathfrak{l}}_k}\mathbf{H}_i\boldsymbol{\beta}_i$ is simplified to $\beta_k\mathbf{P}_{\check{\mathfrak{l}}_k}\mathfrak{h}_k$, providing (18), i.e.,

$$\mathbf{w}_k = \beta_k \mathbf{R}_{\check{\mathfrak{l}}_k}^\dagger \mathbf{P}_{\check{\mathfrak{l}}_k}\mathfrak{h}_k$$

where $\mathbf{R}_{\check{\mathfrak{l}}_k} = \mathbf{P}_{\check{\mathfrak{l}}_k}\mathbf{R}_{xx}\mathbf{P}_{\check{\mathfrak{l}}_k}$.

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