DIFFUSED CHANNEL FRAMEWORK FOR BLIND SPACE-TIME DS-CDMA RECEIVER

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Abstract - In this paper, a blind space-time receiver is proposed to handle point and diffused sources for asynchronous multipath DS-CDMA systems. The receiver is based on a computationally efficient subspace-type algorithm for its joint space-time channel estimation which is insusceptible to near-far problems. The coherency between the sources is removed by a novel temporal smoothing technique operating in the transformed domain. Unlike many conventional DS-CDMA receivers, the proposed formulation and approach is applicable even with the presence of co-code interferers. Furthermore it is robust to channel estimation errors in the event of any unidentified (incomplete) or erroneous (incorrect) channel parameter.

I. INTRODUCTION

Traditional sensor array processing frequently assumes a simplistic pictorial of the multipath environment which is modeled as an assemblage of point sources. However in a typical wireless communication system, particularly in an urban or suburban setup, the signal transmitted into the channel suffers multiple reflections, diffraction and scattering which will inevitably result in a diffusion of its signal component. This hence leads to a performance degradation when conventional processing techniques, based on point sources scenario, are applied. Thus it is paramount to adopt a generalised diffusion framework in order to achieve an effective system design.

In this work an asynchronous blind space-time receiver is developed for DS-CDMA systems, applicable in both point and/or diffused multipath environment. Its channel estimation technique is derived from the efficient near-far resistant PADE (polarisation-angle-delay estimation) algorithm [1, 2] employed in diversely polarised arrays to provide its joint angle and delay estimation. Such adaption of algorithm from diversely polarised arrays has been utilised in [3] to obtain the direction of arrival estimates, but it excludes the temporal dimension in its estimation operation; hence the estimation algorithm in [2] is attempted in [4], but its estimation procedure is not as compact and efficient and can only be completed after a few tens of observation intervals (bursts), making it inappropriate in the implementation of a receiver. However in this paper, a more computationally efficient near-far resistant channel estimator is proposed which is then being employed and integrated as the front-end of a novel blind space-time receiver. The resulted receiver is robust against any unidentifiable or erroneous channel parameter resulted in the estimation process. It is optimum with respect to the SIR criterion and is able to achieve (asymptotically) complete interference cancellation. Computer simulation studies demonstrate the effectiveness of the proposed approach and it is shown to perform just as well even when the signal components are co-located in either one/two of the space, time or code domains.

II. SIGNAL MODEL

Consider an M-user asynchronous DS-CDMA system, with the modulating information signal associated with the \( i \)th user given as a function of the symbol index \( n \) and chip index \( p \) as

\[
m_i(t) = \sum_{n=\infty}^{\infty} a_i[n] \sum_{p=0}^{N_c-1} a_i[p] c(t - nT_{cs} - pT_c) \quad (1)
\]

where \( \{a_i[n]\} \in \mathbb{C} \) is the \( i \)th user’s data symbol, \( T_{cs} \) is the channel symbol period, \( \{a_i[p]\} \in \mathbb{C} \), \( p = 0, 1, \ldots, N_c - 1 \) corresponds to the \( i \)th user's pseudo-noise spreading sequence of period \( N_c = T_{cs}/T_c \), and \( c(t) \) denotes the chip pulse-shaping waveform of duration \( T_c \).

Suppose the transmitted signal due to the \( i \)th user arrives at the base station, employing an array of \( N \) antenna sensors, via \( K_i \) scattering clusters [5] as shown in Fig. 1. The complex baseband signal vector pertaining to the \( j \)th cluster, consisting of \( L_{ij} \) scatterers, can therefore be modelled as a superposition of all the scatterers’ contribution within the cluster, expressed as

\[
s_{ij}(t) = \sum_{k=1}^{L_{ij}} \beta_{ijk} S(\theta_{ijk} + \tau_{ijk}) m_i(t - \tau_{ijk}) \quad (2)
\]

where \( \beta_{ijk} \), \( \theta_{ijk} \), \( \tau_{ijk} \) denotes respectively the complex fading coefficient, angular perturbation about the cluster's nominal direction \( \theta_{ijk} \), and time delay of the \( k \)th scatterer from the \( j \)th cluster corresponding to the \( i \)th user. The manifold vector

### NOTATION

- \( A \) : Scalar
- \( A \) : Vector
- \( A \) : Matrix
- \( \exp(A) \) : Elemental exponential
- \( A^b \) : Elemental power
- \( (\cdot)^T \) : Transpose
- \( (\cdot)^* \) : Complex conjugate
- \( (\cdot)^H \) : Conjugate transpose
- \( \lfloor \cdot \rfloor \) : Roundup to integer
- \( \otimes \) : Kronecker product

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\( S(\theta) = \exp(-j \cdot [r_1 \cdot \frac{\theta}{\sqrt{\rho}}, \ldots, \frac{\theta}{\sqrt{\rho}}] \cdot \mathbf{k}) \) is the array response vector with \( [r_1, \ldots, r_M] \) defining the array geometry and \( \mathbf{k} \) is the wavenumber vector pointing towards the azimuth direction \( \theta \). The scatterers are grouped into clusters in the space and time domains such that their angular and delay spread are smaller than the spatial and temporal resolution of the employed system. The delay differences from the cluster's nominal time delay \( \tau_{ij} \) are below the system sampling interval \( T_s = T_c/q \) (with \( q \) being the oversampling factor) which are temporally unresolvable and negligible \( \{ \tau_{ijk} \approx \tau_{ij}, \quad k = 1, 2, \ldots, L_{ij} \} \). Additionally, the maximum angular perturbation \( \sigma_{ij} = 2\max_k [\tilde{\theta}_{ijk}] \) within the cluster is smaller than the system's resolution limit such that its scatterers' contribution to the signal subspace is one-dimensional. Therefore, by defining the first derivative of the array response vector \( \tilde{S}(\theta) = \partial S(\theta)/\partial \theta \), the signal vector in (2) can be remodelled based on the first-order Taylor approximation as follows

\[
\tilde{s}_{ij}(t) = \sum_{k=1}^{L_{ij}} \beta_{ijk} \{ \tilde{S}(\theta_{ij}) + \tilde{\theta}_{ijk} \tilde{S}(\theta_{ij}) \} m_i(t - \tau_{ij}) = \kappa_{ij} \{ \tilde{S}(\theta_{ij}) + \varphi_{ij} \tilde{S}(\theta_{ij}) \} m_i(t - \tau_{ij}) \tag{3}
\]

where \( \kappa_{ij} = \sum_{k=1}^{L_{ij}} \beta_{ijk} \) and \( \varphi_{ij} = \sum_{k=1}^{L_{ij}} (\beta_{ijk} \tilde{\theta}_{ijk}) \)/\( \sum_{k=1}^{L_{ij}} \beta_{ijk} \) represent the cluster's aggregate fading coefficient and the weighted perturbation factor respectively. Notice that in the case of line-of-sight scenario with no angular spreading, the above framework in (3) can still be utilised since \( \varphi_{ij} = 0 \) for zero angular perturbation.

By denoting \( \mathbf{A}_j = \{ \tilde{S}(\theta_{ij}) + \varphi_{ij} \tilde{S}(\theta_{ij}) \} \), and letting \( \mathbf{A}_i = [ \mathbf{A}_{i1}, \mathbf{A}_{i2}, \ldots, \mathbf{A}_{iK_i} ] \), \( \mathbf{\kappa}_i = [ \kappa_{i1}, \kappa_{i2}, \ldots, \kappa_{iK_i} ] \) and \( \mathbf{m}_i(t) = [ m_i(t - \tau_{i1}), m_i(t - \tau_{i2}), \ldots, m_i(t - \tau_{iK_i}) ] \), the net baseband vector representation of the received signal in the presence of additive isotropic white Gaussian noise with double sided power spectral density \( N_0/2 \) can therefore be shown to be given by

\[
\mathbf{x}(t) = \mathbf{A}_i \cdot \text{diag}(\mathbf{\kappa}_i) \cdot \mathbf{m}(t) + \mathbf{n}(t) \tag{4}
\]

where \( \mathbf{n}(t) \) is the complex white Gaussian noise vector and

\[
\mathbf{A}_i = [ \mathbf{A}_{i1}, \mathbf{A}_{i2}, \ldots, \mathbf{A}_{iM} ] \tag{5}
\]

\[
\mathbf{\kappa}_i = [ \mathbf{\kappa}_{i1}, \mathbf{\kappa}_{i2}, \ldots, \mathbf{\kappa}_{iM} ] \tag{6}
\]

\[
\mathbf{m}(t) = [ \mathbf{m}_1(t), \mathbf{m}_2(t), \ldots, \mathbf{m}_M(t) ]^T \tag{7}
\]

Taking the clusters' delay to lie within the range \( [0, T_s) \), and the \( N \)-dimensional signal vector \( \mathbf{x}(t) \), received from the sensors output, is then sampled at a rate of \( 1/T_s \) and passed through a bank of \( N \) tapped-delay lines (TDL), each of length \( 2qN_c \). Upon concatenating the contents of the TDLs, the \( 2qN_c \)-dimensional discretised signal vector is thus formed and read for every \( T_s \) with the \( n \)-th observation interval represented as

\[
\mathbf{z}[n] = [ \mathbf{z}_1[n]^T, \mathbf{z}_2[n]^T, \ldots, \mathbf{z}_N[n]^T ]^T \tag{8}
\]

where \( \mathbf{z}_n[n] \) is the \( 2qN_c \)-dimensional discretised output frame from the \( k \)-th TDL.

To model the inclusion of the discretised temporal dimension, the manifold vector \( \mathbf{A}_j \) due to the \( j \)-th cluster of the \( i \)-th user is modified to form the spatio-temporal array (STAR) manifold vector \( \mathbf{h}_{ij} \) expressed as

\[
\mathbf{h}_{ij} = \mathbf{A}_{ij} \otimes \mathbf{J}^{ij} \mathbf{\xi}_j \tag{9}
\]

where \( \mathbf{J} \) is the array's cluster's delay; the matrix \( \mathbf{J} \) (or \( \mathbf{J}^T \)) is a \( 2qN_c \times 2qN_c \) time down-shift (or up-shift) operator matrix given as follows

\[
\mathbf{J} = \begin{bmatrix} \mathbf{J}^{2qN_c-1} & 0 \\ 0 & \mathbf{J}^{2qN_c-1} \end{bmatrix} \tag{10}
\]

and \( \mathbf{\xi}_j \) is related to the \( j \)-th user's PN-code sequence defined as

\[
\mathbf{\xi}_j = \sum_{p=0}^{N_c-1} \alpha_j[p] \cdot \mathbf{J}^p \mathbf{\xi}_j \tag{11}
\]

with the vector \( \mathbf{c} \) being the oversampled chip-level pulse shaping function \( c(t) \) padded with zeros at the end, i.e.

\[
\mathbf{\xi}_j = [ c(0), c(T_s), \ldots, c((q-1)T_s), \mathbf{J}^T(2qN_c-1) ] \tag{12}
\]

Now by taking user 1 as the desired user of interest, the net spatio-temporal discretised signal vector \( \mathbf{z}[n] \) can thus be written as

\[
\mathbf{z}[n] = \mathbf{R}_1 \mathbf{G}_1 \mathbf{a}_1[n] + \sum_{i=2}^{l_i} \mathbf{H}_i \mathbf{G}_i \mathbf{a}_i[n] + \sum_{i=U_1+1}^{M} \mathbf{R}_i \mathbf{G}_i \mathbf{a}_i[n] + \mathbf{n}[n] \tag{13}
\]

where the first term on the right-hand side of (13) denotes the discretised signal vector associated with the desired user, the second term contains the interferences due to \( (U_1 - 1) \) of its co-code users (arising from either code-reuse interferers or intercepting co-code jammers), the third term represents the remaining interferences from the rest of the users together with their corresponding co-code partners, and the last term is the sampled noise vector. The \( j \)-th user's data vector constitutes contributions from not only the current but also the previous and next symbols given as \( \mathbf{a}_j[n] = [ a_j[n-1], a_j[n], a_j[n+1] ]^T \), the matrix \( \mathbf{G}_j = I_3 \otimes \mathbf{\kappa}_j \), and \( \mathbf{H}_i = \{ I_{2qN_c} \otimes (\mathbf{J}^T)^{V_i} \} \mathbf{H}_i^{V_i} \), \( (\mathbf{I}_{2qN_c} \otimes (\mathbf{J}^T)^{V_i}) \mathbf{H}_i^{V_i} \), with \( \mathbf{H}_i^{V_i} \) having columns the STAR manifold.

**Figure 1:** Scattering propagation channel.
vectors corresponding to the \(i\)th user, i.e. \(\mathbb{H}_i = [\mathbf{h}_{i1}, \mathbf{h}_{i2}, \ldots , \mathbf{h}_{iK_i}]\).

\[\] III. CHANNEL ESTIMATION AND RECEPTION

To perform an estimation of the channel parameters pertaining to the desired user and all of its co-code interferers, we proposed that the discretised signal vector \(\mathbf{z}[n]\) in (9) be preprocessed by the desired user's preprocessor (i.e. user 1) defined as \(\mathbf{Z}_1 = \mathbf{L}_v \otimes (\text{diag}(\xi_1))^{-1} \mathbf{F}\), where \(\tilde{\xi}_1\) is the Fourier transformed version of \(\xi_1\), that is \(\tilde{\xi}_1 = \mathcal{F} \xi_1\) with \(\mathcal{F}\) being the Fourier transformation matrix given by

\[
\mathcal{F} = [\Phi^0, \Phi^1, \Phi^2, \ldots, \Phi^{(2qN_s-1)}]
\]

where \(\Phi = [1, \Phi^1, \Phi^2, \ldots, \Phi^{(2qN_s-1)}]^T\) and

\[
\Phi = \exp(-j \frac{2\pi}{2qN_c})
\]

To get a clearer picture of its operation on the discretised signal vector \(\mathbf{z}[n]\), let's apply the preprocessor to the STAR manifold vector in (5) as follows

\[
\mathbf{\hat{h}}_{ij} = \mathbf{Z}_1 \mathbf{h}_{ij} = \mathbf{A}_{ij} \otimes \{\text{diag}(\xi)^{-1} \mathbf{F}(\mathbf{z}^i_j)\}
\]

or

\[
\mathbf{\hat{h}}_{ij} = \mathbf{A}_{ij} \otimes \{\text{diag}(\xi)^{-1} \text{diag}(\xi) \mathbf{F}^i_j\}
\]

(11)

It is clear that the above expression in (11) can be reduced to simply \(\mathbf{\hat{h}}_{ij} = \mathbf{A}_{ij} \otimes \Phi^i_j\) if and only if \(i = 1\), which is corresponding to the desired user's PN-code sequence. Hence, by applying the same operation to (9), the discretised signal vector is therefore transformed to

\[
z[i][n] = \mathbf{Z}_i \mathbf{x}[n]
\]

(12)

\[
= \sum_{i=1}^{U_1} \mathbf{\hat{h}}_{i} \mathbf{z}_{i}[n] + \mathbf{Z}_1 \mathbf{L}_{\text{ISI}}[n] + \mathbf{Z}_1 \mathbf{L}_{\text{MAI}}[n] + \mathbf{Z}_1 \mathbf{n}[n]
\]

where \(\mathbf{\hat{h}}_{i} = [\mathbf{\hat{h}}_{i1}, \mathbf{\hat{h}}_{i2}, \ldots, \mathbf{\hat{h}}_{iK_i}]\). Notice that the signal vector \(\mathbf{z}_{i}[n]\) is effectively decoupled into four terms, namely the desired signal constituent \(\mathbf{\hat{h}}_{i} \mathbf{z}_{i}[n]\) as well as that of its \((U_i - 1)\) co-code partners, the ISI, MAI and noise components respectively.

But its second order statistics \(\mathbb{R}_{\mathbf{z}_i} \in \mathbb{C}^{2qN_s \times 2qN_s}\), instead of providing a basis for the desired signal subspace, would result in a rank deficiency with the desired signal subspace dimension being reduced to one. To restore the dimensionality of this subspace back to \(\sum_{i=1}^{U_1} \mathbf{K}_i\), we will make use of the Vandermonde structure of the submatrices of \(\mathbf{\hat{h}}_{i}\) provided by the preprocessing operation. This can be achieved by performing a technique referred to as temporal smoothing. First let's partition the correlation matrix \(\mathbb{R}_{\mathbf{z}_i}\) into \(N^2\) segmented matrices of size \(2qN_c \times 2qN_c\). By averaging a set of \(Q\) (where \(Q = 2qN_c - d + 1\) and \(d < 2qN_c\)) overlapping \(d \times d\) submatrices along the main diagonal of each segmented matrices simultaneously, a \(Nd \times Nd\) temporal-smoothed covariance matrix \(\mathbb{R}_{T\text{smooth}}\) is therefore constructed. However its dimensionality can only be successfully restored provided that \(Q \geq \sum_{i=1}^{U_1} \mathbf{K}_i\). Note that such averaging operation is similar to the well known spatial smoothing (averaging) technique as described in [6]. This technique can also be applied for clusters arriving from the same direction (co-directional). But for clusters of a particular user arriving at the same time (co-delay), singularity in \(\mathbb{R}_{T\text{smooth}}\) will occur. This special case cannot be resolved for a general array geometry but for a uniform linear array where spatial smoothing can be overlaid on top of \(\mathbb{R}_{T\text{smooth}}\) to form the spatial-temporal-smoothed covariance matrix \(\mathbb{R}_{T\text{smooth}}\).

Having obtained the covariance matrix, the clusters' channel parameters can be found by minimising the following MUSIC-type cost function, which is derived using the transformed STAR manifold vector \(\mathbf{\hat{h}}\), given as

\[
\xi(\theta, l, \varphi) = \mathbf{\hat{h}}^H \mathbb{E}_n \mathbb{E}_n^H \mathbf{\hat{h}}
\]

(13)

where \(\mathbf{\hat{h}} = \mathbf{A} \otimes \Phi^i_j\) in which \(\Phi^i_j\) is a subvector of \(\Phi\) with length \(d\), and \(\mathbb{E}_n\) is a matrix whose columns are the generalised noise eigenvectors of \((\mathbb{R}_{T\text{smooth}}, \mathbb{D})\) due to the transformed noise in (12), with \(\mathbb{D}\) representing the spatial-temporal-smoothed diagonal matrix \(\mathbf{Z}_1 \mathbf{Z}_1^H\).

However, the minimisation process involves a multidimensional search over \(\theta\), \(l\) and \(\varphi\) for its minima. To simplify the process, it is noted that the manifold vector \(\mathbf{A}\) within the transformed STAR manifold vector \(\mathbf{\hat{h}}\) can be rewritten as \(\mathbf{A}(\theta, \varphi) = \mathbb{M}(\theta) \cdot \mathbb{P}(\varphi)\) where

\[
\mathbb{M}(\theta) = \begin{bmatrix} \mathbf{S}(\theta) & \mathbf{S}(\theta) \\ \mathbf{D}(\theta) \end{bmatrix} ; \quad \mathbb{P}(\varphi) = [1 \quad \varphi]^T
\]

(14)

Note that for a linear array geometry, for instance an array lying along the \(x\)-axis, the term \(\mathbf{S}(\theta)\) will be nullified for \(\theta = 0\) and \(\pi\) due to its derivative coefficient being proportional to an azimuth-dependent function \(\sin(\theta)\). In this case, the true \(\mathbf{\hat{S}}(\theta)\) can be replaced by lumping its azimuth-dependent function together with \(\varphi\), i.e. \(\mathbb{M}(\theta) = [\mathbf{S}(\theta), \mathbf{D}(\theta)]\) and \(\mathbb{P}(\varphi, \theta) = [1, \varphi \sin(\theta)]^T\) where \(\mathbf{D}(\theta)\) is the derivative term \(\mathbf{\hat{S}}(\theta)\) after discarding its azimuth-dependent function.

The re-expression in (14) thus yields a cost function similar to that commonly-used for diversely polarised arrays [1] given as

\[
\xi(\theta, l, \varphi) = \mathbf{\hat{h}}^H \mathbb{M}(\theta) \cdot \mathbb{E}_n \mathbb{E}_n^H \mathbb{M}(\theta) \cdot \mathbb{P}(\varphi) \quad \forall \mathbb{V} = \begin{bmatrix} \mathbf{c}_1 \mathbf{v}_1 \mathbf{v}_2 \mathbf{v}_2 \\ \mathbf{v}_1 \mathbf{v}_2 \mathbf{v}_2 \mathbf{v}_2 \end{bmatrix}
\]

As in the diversely polarised arrays scenario, the minimisation over the \(\mathbb{P}\) search space is equivalent to finding the eigenvectors corresponding to the minimum eigenvalues of \(\mathbb{V}\). Hence by dropping the two-dimensional vector \(\mathbb{P}\) and applying the quadratic formula, the cost function in (15) can be simplified to

\[
\xi(\theta, l) = \text{trace}(\mathbb{V}) - \sqrt{\text{trace}(\mathbb{V})^2 - 4 \det(\mathbb{V})}
\]

(16)
where \( \text{trace}(\bullet) \) denotes the trace operation and \( \det(\bullet) \) represents the determinant of the \( 2 \times 2 \) matrix. Now let \( \xi_{\text{min}} \) be the minima obtained from the spectrum constructed using the cost function \( \xi(\theta, l) \) in (16). The location of the minima, as such, provides the joint estimate of its direction of arrival (DOA) and time of arrival (TOA). Its corresponding weighted perturbation factor, on the other hand, is estimated as follows

\[
\varphi = \begin{cases} 
\frac{(\xi_{\text{min}} - 2v_{11})/2v_{12}}{2v_{21}/(\xi_{\text{min}} - 2v_{22})} 
\end{cases} 
\]  (17)

However in the case of a linear array, the estimated weighted perturbation factor \( \varphi \) has to be divided by its azimuth-dependent term \( \sin(\theta) \). It is also worthwhile to note that the above expression in (17) can be further simplified in most circumstances since the value \( \xi_{\text{min}} \) is usually close to zero.

Now for a set of \( B \) unique PN-code signatures, the estimated composite channel parameters can be obtained as

\[
\hat{H}_{\text{comp}} = \left[ \hat{H}_{PN,1}, \hat{H}_{PN,2}, \ldots, \hat{H}_{PN,B} \right] 
\]  (18)

where \( \hat{H}_{PN,i} \) has columns the estimated STAR manifold vectors associated with the \( i^{th} \) user and its corresponding co-code partners. In order for the blind space-time receiver to suppress the contributions from the MAI and ISI interferences, it is proposed that the received discretised signal vector \( \bar{x}[n] \) in (9) be passed through a novel multi-cluster filter bank to yield

\[
y[n] = L^H \cdot \bar{x}[n] 
\]  (19)

where \( L \) is the multi-cluster filter bank based on the orthogonal projection of the interference subspace, i.e.

\[
L = \left[ \hat{H}_{\text{interf}}^+ \hat{H}_{PN,1} \left( \hat{H}_{PN,1}^H \hat{H}_{\text{interf}} \right)^{-1} \right] 
\]  (20)

where

\[
\hat{H}_{\text{interf}} = \left[ \left( I_N \otimes \mathcal{F} \right) \hat{H}_{\text{comp}}, \left( \hat{H}_{\text{comp}}^H \right)^{-1} \hat{H}_{PN,1} \right] 
\]

is the estimated composite channel parameters \( \hat{H}_{\text{comp}} \) with the exclusion of the matrix \( \hat{H}_{PN,1} \) = \( \hat{H}_{1, \ldots, K} \) corresponding to the desired user's PN-code sequence.

In a non co-code environment, the outputs of the filter bank can then be simply combined to realise the decision statistic for the \( n^{th} \) symbol of the desired user

\[
b[n] = w^H \cdot y[n] 
\]  (21)

where \( w \) is the combining weight vector obtained using the principal eigenvector of the autocorrelation matrix of (19). But such methodology will result in an absolute phase ambiguity which can however be easily resolved by differential encoding, with the transmitted data symbol \( e[n] = s_k[n-1], k[n] \) and the decoding according to the following criterion \( \tilde{e}[n] = \text{sgn}(Re\{b[n-1]^* \cdot b[n]\}) \).

On the other hand, in the presence of co-code users, the filter bank output, comprising of \( \sum_{i=1}^{K} K_i \) branches, need to be partitioned to differentiate those branches belonging to the desired user. This can be done by assigning a short random header sequence common to the designated non co-code user group. The filter bank output, following the decision device, are then passed to the Branch Identification Process, whereby a correlation with the header is performed to identify the dominant branch exhibiting the strongest presence of the header sequence. The dominant branch is subsequently cross-correlated with each of the \( \sum_{i=1}^{K} K_i \) branches and compared with a prespecified threshold value. When the correlated output exceeds the threshold value, it is assigned to the desired user. Having segregated the branches, the desired user's filter bank outputs can then be combined using (21). Note that it is not necessary to assign all the \( K_1 \) branches to the desired user. If the channel parameters of any particular branch is erroneous (incorrect channel estimation) or unidentified (incomplete channel estimation), the Branch Identification Process will leave that branch unassigned, thus inducing robustness to the receiver. In a similar manner, the above technique can also be applied to decode the remaining users in the designated group as well as their corresponding co-code partners.

### IV. Simulations

To demonstrate the key features of the proposed blind space-time receiver, let's consider a uniform \( N = 5 \) element circular array (with half-wavelength spacing) operating in the presence of \( M = 7 \) co-channel DS-CDMA users, each being assigned a Gold sequence of length \( N_f = 31 \) (of rectangular chip pulseshaping). The number of unique code signatures is \( B = 3 \), with the designated user group consisting of \( i = 1, 3 \) and 6, as listed in Table I. The angular spread \( \sigma \) (SPD) of each cluster is generated by rendering 40 spatially uniform-distributed local scatterers. The array is assumed to collect 150 data symbols for processing with a chip-rate sampler of \( q = 1 \).

<table>
<thead>
<tr>
<th>Code signature vector 1 (( a_1 ))</th>
<th>Code signature vector 2 (( a_2 ))</th>
<th>Code signature vector 3 (( a_3 ))</th>
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Assume that user 1 is the desired user having an input SNR of 20dB; while its corresponding co-code partner (i.e. user 2) constitutes an interference ratio of 0dB, together with the rest of the remaining interferers having a SIR = -20dB (i.e. near-far problem). By setting \( d = 55 \) for temporal smoothing, it is seen from Fig. 2 that all the clusters, be it diffused or line-of-sight occupations, associated with the desired user as well as its co-code partners can be identified/estimated successfully using the proposed algorithm. Notice that the algorithm can still operate even in situations whereby (i) the 2\(^{nd}\) cluster of user 1 is co-located in both the code and space domains due to the 2\(^{nd}\) cluster of user 2, (ii) the 3\(^{rd}\) cluster of user 1 is co-located in both the
code and time domains due to the 1st cluster of user 2, and (iii) the 3rd cluster of user 1 is co-located in both the space and time domains due to the 2nd cluster of user 3. In addition to that, the number of clusters that can be resolved by the algorithm is also not constrained by the number of sensors available in the array. The clusters belonging to the desired user are then singled out, as illustrated in Table II, by applying the Branch Identification Process with an header sequence of length 7. Fig. 3 depicts the signal constellation of the proposed blind receiver as compared with a conventional 2D RAKE receiver, with the latter assuming full knowledge of the channel information (including knowledge of all the desired user's co-code interferences). It is clear that the proposed receiver offers a significant performance enhancement than that of the RAKE receiver.

Next, let's take a look at the performance of the receiver with framework based on (i) the proposed model \( \hat{A}(\theta, \varphi) = \hat{S}(\theta) + \varphi \hat{S}(\theta) \) as compared with that based on (ii) the conventional model \( \hat{A}(\theta) = \hat{S}(\theta) \). For comparison clarity, let's assume that the all the clusters under consideration have the same angular spread \( \sigma \). Using the same scenario as above, the output SNIR of the receivers are averaged over 1000 Monte Carlo runs and plotted as shown in Fig. 4. It is therefore apparent that by incorporating the diffusion aspect into the framework, the proposed receiver provides a substantial performance gain than that based on the conventional model which is observed to be deteriorating with the angular spread \( \sigma \).

V. CONCLUSIONS

A blind space-time receiver based on a computationally efficient near-far resistant channel estimation technique is presented for multiple point or diffused signal paths in an asynchronous DS-CDMA system. Due to its underlying architecture, the receiver can be easily extended to cope in a non-unique code signature environment by simply adding a novel Branch Identification Process in the detection procedure. The receiver is robust to erroneous or incomplete channel estimation since its operation requires only the existence of an estimable cluster due to the desired user separable/identifiable in at least one of the following domains: space, time and code. The number of resolvable clusters is also no longer limited by the number of sensors available in the array.

REFERENCES