BLIND SINGLE-USER ARRAY RECEIVER FOR MAI CANCELLATION IN MULTIPATH FADING DS-CDMA CHANNELS

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ABSTRACT
In this paper, a blind subspace-type single-user array direct-sequence code division multiple access (DS-CDMA) receiver is proposed based on a new joint space-time channel estimation technique for frequency-selective channels. In the proposed approach, the spatio-temporal multipath channel parameters associated with the desired user are jointly estimated with the number of identifiable paths not limited by the number of antennas. Then these estimated parameters are employed to efficiently combine the desired multipath rays while, at the same time, (asymptotically) complete MAI cancellation is achieved.

1 INTRODUCTION
With DS-CDMA chosen as the air-interface for the next generation of wireless communication systems/networks, the intention is to support as many users as possible operating at very high data rates in the presence of severe channel distortion effects. In these systems the multiple users transmit asynchronously, overlapping in both time and frequency and, therefore, resulting in multiple-access interference (MAI) which, eventually, limits the system performance [1,2]. Furthermore, the presence of multipath propagation effects not only increases the MAI by introducing extra interfering paths, but also distorts the desired signal component (frequency-selective fading). However, a well designed DS-CDMA system may exploit:

- the multipath channel by recombining the multipath rays in a positive way, thereby combating the multipath fading effects;
- the highly structured signal format of the MAI to provide MAI cancellation;

thus significantly improving the overall system’s performance. However, for this exploitation to be effective, channel parameters (such as multipath delays, path coefficients, etc.) must be estimated prior to the subsequent data symbol detection. One of the main concerns of this paper is channel estimation using an array of antennas. The use of an antenna-array makes it possible to mitigate the above channel effects and provides the potential benefit of exploiting the spatial structure of the channel leading to joint spatio-temporal reception – space-time communications. It has been argued that joint spatio-temporal reception/processing is a breakthrough approach for future generation of wireless communications [3,4,5] and the estimation of the spatio-temporal channel parameters is central to the research work presented in this paper.

2 SIGNAL MODELLING
In an $M$-user asynchronous DS-CDMA system, the $i^{th}$ user’s baseband transmitted signal may be written as

$$m_i(t) = \sum_{n=-\infty}^{\infty} a_i[n] c_{PN_i}(t-nT_{\text{se}}), \quad nT_{\text{se}} \leq t \leq (n+1)T_{\text{se}}$$

where $\{a_i[n] : \forall n \in N\}$ is the $i^{th}$ user’s data symbol sequence of period $\pm 1$. $T_{\text{se}}$ is the data symbol period and, $c_{PN_i}(t)$ is one period of the pseudo random spreading waveform associated with the $i^{th}$ user, which is modelled as follows:

$$c_{PN_i}(t) = \sum_{k=0}^{N-1} a_i[k] c_i(t-kT_c), \quad kT_c \leq t \leq (k+1)T_c$$

In Equation (2) $\{a_i[k] \in \pm 1\}$ represents the $i^{th}$ user’s PN-sequence of period $N = T_{\text{se}}/T_c$ and $c_i(t)$ denotes the chip pulse shaping waveform of duration $T_c$.

Let us assume that the base station has an array of $N$ antennas and that the transmitted signal of the $i^{th}$ user arrives at the base station via $K_i$ paths, $\forall i = 1, ..., M$. Consider the $j^{th}$ path of the $i^{th}$ user arrives at the array from direction $\theta_i$ with channel propagation parameters $\beta_j$ and $\gamma_j$ representing the complex path gain and path delay, respectively. Let $\hat{S}_j = \hat{S}(\theta_i)$ denote the array manifold vector (array response) associated with the $j^{th}$ path of the $i^{th}$ user, which has the form:

$$\hat{S}(\theta_i) = \exp \left( - j \mathbf{r}^T \hat{b}_i \right)$$

where $\mathbf{r} = [\xi_1, \xi_2, ..., \xi_N]$ is a $3 \times N$ real matrix with columns the positions of antenna elements (in half wavelengths) and $\hat{b}_i$ is the wavenumber vector pointing towards the direction $\theta_i$. Then, based on the above description, the received complex baseband signal-vector at the antenna array (see Figure-1, point-A) can be represented as follows:

$$\mathbf{z}(t) = \sum_{i=1}^{M} \mathbf{S}_i \text{diag}(\hat{\beta}) \mathbf{m}_i(t) + \mathbf{u}(t)$$

where $\mathbf{u}(t)$ is a complex white Gaussian noise vector with covariance matrix $\sigma^2 \mathbf{I}_N$ and

$$\mathbf{S}_i = \left[ \mathbf{S}_1, \mathbf{S}_2, ..., \mathbf{S}_{K_i} \right] \in \mathbb{C}^{N \times K_i}$$

$$\hat{\beta} = [\hat{\beta}_1, \hat{\beta}_2, ..., \hat{\beta}_{K_i}]^T \in \mathbb{C}^{K_i \times 1}$$

$$\mathbf{m}_i(t) = [m_i(t-\tau_1), m_i(t-\tau_2), ..., m_i(t-\tau_{K_i})]^T \in \mathbb{R}^{K_i \times 1}$$

By discretising the signal-vector $\mathbf{z}(t)$ (using chip matched filters followed by chip rate samplers) and then passing the samples
through a tapped-delay line of length equal to $2N_c$, as shown in Figure-1, the data matrix $X[n] \in \mathcal{C}^{K \times 2N_c}$ is formed, i.e.,
\[ X[n] = [x_0[n], x_2[n], \ldots, x_{2N_c}[n]]^T \] (4)
representing the contents of all $N$ tapped-delay lines associated with the $n^{th}$ data symbol period. Note that due to the asynchronous operation of the system, the tapped-delay lines contain contributions from the previous and next symbols as well as the current symbol. To further model the received signal we need to define the reference temporal manifold vector $z_i$ of the $i^{th}$ user and a shift operator matrix $J$. The vector $z_i$ is defined as below
\[ z_i = [\alpha_i[0], \alpha_i[1], \ldots, \alpha_i[N_c - 1], \Omega_{\bar{u}_i}]^T \in \mathcal{C}^{2N_c \times 1} \] (5)
and represents one period of the PN-sequence of the $i^{th}$ user extended with zeros to have a length of $2N_c$. This corresponds to a zero path delay situation. On the other hand, $J$ is a $2N_c \times 2N_c$ matrix defined as follows
\[ J = \begin{bmatrix} 1 & 0 & \ldots & 0 \\ 0 & 1 & \ldots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \ldots & 1 \end{bmatrix} = \begin{bmatrix} \Omega_{\bar{u}_{N_c - 1}} & 0 \\ 0 & \Omega_{\bar{u}_{N_c - 1}} \end{bmatrix} \] (6)
having the following property: every time the matrix $J$ (or $J^T$) operates on a column vector it down-shifts (or up-shifts) the elements of the vector by one. For instance, $J^n z_i$ is a version of $z_i$ down-shifted by $n$ elements, while $(J^T)^n z_i$ is a version of $z_i$ up-shifted by $n$ elements. Based on the definitions of Equations (5) and (6), the $k^{th}$ column of $X[n]$, i.e., $x_k[n]$, (see Equation (4)) can be modelled in terms of the up and down-shifted versions of $z_i$ to represent the contributions of the previous, current, and next data symbols as follows:
\[ x_k[n] = \sum_{i=1}^{K} \sum_{j=1}^{K} \left( J^T \right)^k \Omega_{\bar{u}_j} \alpha_i[n - j] + J^k \Omega_{\bar{u}_i} \alpha_i[n] \] (7)
where $l_i$ is the discrete version of the path delay $\tau_{ii}$, i.e.,
\[ l_i = [\tau_{ii}/T_c] \mod N_c. \]

Based on Equation (7), we can formulate the spatio-temporal received signal-vector
\[ z[n] = [z_0[n] \, z_2[n] \, \cdots \, z_{2N_c}[n]]^T \] (8)
\[ = \sum_{i=1}^{K} \sum_{j=1}^{K} (\Theta_i)_{\bar{u}_j} \left( J^T \right)^N \Omega_{\bar{u}_j} \alpha_i[n - j] + J^N \Omega_{\bar{u}_i} \alpha_i[n] \]

with $\otimes$ denoting the Kronecker product. In Equation (8), the term $\Theta_i \otimes (J^T \Omega_{\bar{u}_j})$ is defined as the spatio-temporal manifold vector for the $i^{th}$ path of the $j^{th}$ user. Using this spatio-temporal manifold vector, $z[n]$ can be rewritten in a compact form as
\[ z[n] = \sum_{i=1}^{K} \left( H_{i,\text{prev}} \Psi_{i} \right) \tilde{z}[n] + z[n] \] (9)
where, for the $i^{th}$ user, the matrix $H_{i} \in \mathcal{C}^{2N_c \times K}$ has as columns the spatio-temporal manifold vectors $\Psi_{i} \forall j = 1, \ldots, K_i$, while $H_{i,\text{prev}}$ and $H_{i,\text{accr}}$ can be expressed as a function of $H_{i}$ as follows:
\[ H_{i,\text{prev}} = \left( I_N \otimes \left( J^T \right)^N \right) H_{i} \]
\[ H_{i,\text{accr}} = \left( I_N \otimes \left( J^T \right)^N \right) H_{i} \]
The received signal vector in Equation (9) can be decoupled into four terms representing the desired, ISI (due to the length of the tapped-delay line), MAI and noise terms, as follows (assuming the first user is the desired user):
\[ z[n] = H_{i,\text{prev}} \Psi_{i} \tilde{z}[n] + L_{i,i} [n] + L_{i,\text{MAI}} [n] + z[n] \] (10)
where
\[ L_{i,i} [n] = H_{i,\text{prev}} \Psi_{i} \left( \alpha_i[n-1] \right) \]
\[ L_{i,\text{MAI}} [n] = \sum_{i'=2}^{M} H_{i,\text{prev}} \Psi_{i} \left( \alpha_i[n-1] \right) \]

In the following section, by exploiting the above modelling, a novel receiver is proposed which provides (asymptotically) complete cancellation of both ISI and MAI effects and, consequently, detects the data sequence $\{a_i[n] \forall n\}$, by estimating the spatio-temporal manifold vectors associated with the desired user (i.e. the matrix $H_{i}$).

\[ H_{i,\text{prev}} = \left( I_N \otimes \left( J^T \right)^N \right) H_{i} \]

The 3 BDIN SINGLE-USER CDMA ARRAY RECEIVER

Before we proceed let us define the $(2N_c \times 2N_c)$ Fourier Transform matrix $\mathcal{F}$
\[ \mathcal{F} = \left[ \phi^0, \phi^1, \phi^2, \ldots, \phi^{2N_c-1} \right] \] (11)
where $\phi = [1, \phi^1, \phi^2, \ldots, \phi^{2N_c-1}]^T$ with $\phi = \exp(-j\pi N_c)$. And let us apply this to the $(2N_c \times 1)$ spatio-temporal manifold vector $\Psi_{i}$, associated with the $i^{th}$ path of the $i^{th}$ user, by using the operation $\mathcal{F} \otimes \Psi_{i}$. Then, it can be easily proven that the manifold vector $\mathcal{F} \Psi_{i}$ is transformed to another $(2N_c \times 1)$ vector, as follows:
\[ (I_N \otimes \mathcal{F}) \Psi_{i} = (I_N \otimes \mathcal{F}) (\Theta_i \otimes (J^T \Omega_{\bar{u}_j})) \]
\[ \Rightarrow (I_N \otimes \mathcal{F}) \Psi_{i} = \Theta_i \otimes (\mathcal{F} \otimes \Psi_{i}) \] (12)
Next let us apply the Fourier Transform matrix to the received signal vector $z[n]$, using the same operation, i.e.
\[ (I_N \otimes \mathcal{F}) z[n] \]

It is clear from Equation (9) that this operation directly affects the matrices $H_{i,\text{prev}}$, $\Psi_{i}$, and $H_{i,\text{accr}}$. In particular, the columns of $\Psi_{i}$ are transformed according to Equation (12). In addition, if the operation $\mathcal{F}^{-1}$ related to the desired user is applied, the overall transformation of the columns of $H_{i}$ can be represented as
\[ H_{i,\text{prev}} = \left( I_N \otimes \left( J^T \right)^N \right) H_{i} \]

The front-end of the proposed array receiver is shown in Figure-1.
\[ S(\theta_j) \otimes \left( \text{diag}(\mathbf{F}_g) \right)^{-1} \text{diag}(\mathbf{F}_g) \phi^q \] for \( i = 1, ..., M \) \hspace{1cm} (13)

However, it is obvious that, if and only if \( i = 1 \) (i.e. only for the desired user), Equation (13) simplifies to \( S(\theta_j) \otimes \phi^q \). We call the vector \( S(\theta_j) \otimes \phi^q \) the transformed spatio-temporal manifold vector associated with the \( j \)th path of the desired user.

The proposed receiver exploits the above property associated with the desired signal in a number of different ways. Its first task is to preprocess the input signal \( \mathbf{z}[n] \) by the known transformation matrix \( \mathbf{Z}_w = \mathcal{I}_{N} \otimes \left( \text{diag}(\mathbf{F}_g) \right)^{-1} \mathcal{P} \) which, from now on, will be referred to as the spatial-temporal preprocessing associated with the desired user. Thus,

\[ \mathbf{z}_q[n] = \mathbf{Z}_w \mathbf{x}[n] = \mathcal{H}_w \mathbf{z}_1[n] + \mathcal{H}_w \mathbf{z}_2[n] + \mathcal{H}_w \mathbf{z}_3[n] + \mathcal{H}_w \mathbf{z}_4[n] \] \hspace{1cm} (14)

where \( \mathcal{H}_w \) is a matrix with columns the transformed spatio-temporal manifold vectors of the desired user.

Note that the preprocessor \( \mathbf{Z}_w \) is a function of the PN-sequence of the desired user, known to the receiver, with \( \mathcal{H}_w \mathbf{z}_1[n] \) being a diagonal matrix.

Having transformed the signal \( \mathbf{z}[n] \) to \( \mathbf{z}_q[n] \) the next task of the proposed receiver is to estimate the matrix \( \mathcal{H}_w \), or equivalently, to estimate the parameters \( \left( \theta_j, l_j \right) \) for \( j = 1, ..., K_1 \). This is done as follows. If a signal subspace technique was employed based on the second order statistics of the vector \( \mathbf{z}_q[n] \), the contribution of the desired (first) term of Equation (14) would be a rank one matrix representing a one-dimensional space spanned by a linear combination of the columns of \( \mathcal{H}_w \). To avoid this problem and restore the dimensionality of this subspace to \( K_1 \) dimensions, the Vandermonde structure of the submatrices of \( \mathcal{H}_w \) is exploited by the proposed receiver employing a temporal smoothing approach similar to the spatial smoothing approach presented in [6].

Thus, initially, the smoothing matrix \( \mathbf{R}_{\text{smooth}N} \) is formed

\[ \mathbf{R}_{\text{smooth}N} = \frac{1}{Q} \sum_{i=1}^{Q} \mathbf{R}_{\text{sn}} \in \mathbb{C}^{dN \times dN} \] \hspace{1cm} (16)

with \( \mathbf{R}_{\text{sn}} \) representing the covariance matrix of a \( dN \) element subvector \( \mathbf{z}_q[n] \) of \( \mathbf{z}_1[n] \) (formed according to Figure-2) with \( d < 2N_k \). Note that \( Q \) overlapping subvectors are needed in this procedure satisfying the following conditions:

\[ Q > K_1 \] and \[ 3(M-1) + 2K_1 < dN-1. \] \hspace{1cm} (17)

Then, a 2D MUSIC spectrum is constructed based on the following cost function

\[ \xi_{\text{MUSIC}}(\theta, l) = \frac{1}{\mathbf{E}_n \mathbf{E}_n^H \left( S(\theta) \otimes \phi^q L_{\text{aut}}^H \right)} \] \hspace{1cm} (18)

where \( \phi^q \) is a subvector of \( \phi \) of length \( d \) and \( \mathbf{E}_n \) is the matrix with columns the noise eigenvectors of \( \mathbf{R}_{\text{smooth}N} \).

The peaks of \( \xi_{\text{MUSIC}}(\theta, l) \) provide the values of \( \left( \theta_j, l_j \right) \) for \( j = 1, ..., K_1 \) associated with the spatio-temporal manifold vector of the desired user. Consequently, the matrix \( \mathcal{H}_w \); and, therefore, the matrix \( \mathcal{H}_f \) is determined. Then, the parameter \( \mathbf{p} \) can be estimated in a similar fashion to that presented in [5].

The final step of the proposed receiver is concerned with the estimation of the weight vector to despread and, at the same time, isolate the desired signal by removing the unwanted interference effects associated with the second and third terms of the RHS of Equation (10). Thus, initially, the proposed receiver (based on the parameters estimated so far) removes the desired signal effects from the covariance matrix of \( \mathbf{z}_q[n] \) and forms the noise-plus-interference covariance matrix \( \mathbf{R}_{\text{noise} \text{+ inter}} \), associated with the last three terms of the RHS of Equation (10). That is,

\[ \mathbf{R}_{\text{noise} \text{+ inter}} = \mathbf{R}_x - \mathcal{H}_w \mathbf{E}_n \mathbf{E}_n^H \] \hspace{1cm} (19)

where \( \mathbf{E}_n = \mathcal{L}[\mathbf{z}_q[n]]^H \). If \( \mathcal{P}_n \) represents the projection operator onto the subspace spanned by the noise eigenvectors of \( \mathbf{R}_{\text{noise} \text{+ inter}} \), then the following weight vector

\[ \mathbf{w}_n = \mathcal{P}_n \mathcal{H}_w \left( \mathcal{H}_w^H \mathcal{P}_n \mathcal{H}_w \right)^{-1} \mathbf{a}_j \] \hspace{1cm} (19)

where \( \mathbf{a}_j \) is a normalising factor such that \( \| \mathbf{w}_n \| = 1 \). Orthogonal to the interference subspace to which the vectors \( L_{\text{aut}}[n] \) and \( L_{\text{aut}}[n] \) belong. Using the weight-vector of Equation (19) the decision variable for the \( n^{th} \) data symbol can be estimated as

\[ d_n[n] = \mathbf{w}_n^H \mathbf{z}_q[n] \] \hspace{1cm} (20)

where it can be easily proven that the power of the noise term \( n_1[n] = \mathbf{w}_n^H \mathbf{E}_n[n] \) is equal to \( \sigma^2 \).

Figure 2: Forming the Q overlapping subvectors for temporal smoothing.
4 SIMULATIONS

In this section, some illustrative simulation results are presented in order to demonstrate the main features of the proposed receiver. The first set of results is related to the joint direction and time of arrival (DOA, TOA) estimation part of the proposed receiver and is shown in Figure-3. The scenario used is a uniform linear array of five antennas operating in the presence of three co-channel users. Each user is assigned a Gold sequence of length 31 and has 10 multipath rays. The parameters of the 10 paths associated with the desired user are given in Table-1.

<table>
<thead>
<tr>
<th>Table-1: Desired user's parameters</th>
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<tbody>
<tr>
<td>$j = 1$</td>
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<tr>
<td>$\theta_1$</td>
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<td>$\theta_2$</td>
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It is clear from Figure-3 that all paths of the desired user (User 1) are successfully identified/estimated.

The second set of results is related to the decision variable $d_i[n]$ of the proposed receiver when the receiver operates in the presence of finite averaging effects. This implies that there are estimation errors due to the finite observation interval which allow some interference components (terms MAI and ISI of Equation (10)) to pass to the decision variable. Two observation intervals of length equivalent to 100 bits and 1000 bits are considered. The desired user transmits these bits over a multipath channel with three resolvable multipaths in the presence of 10 single path MAI-users with each user assigned a Gold sequence of length 31. The power of the noise is assumed 20dB below, and the power of the MAI-users 20dB above, the desired signal power and a 5-element uniform linear array of antennas is used. The results are shown in Figure-4 demonstrating that the larger the observation interval the smaller the MAI components passing to the proposed receiver's output. In addition, for the 1000 bits case, the RAKE receiver, as well as the 'decorrelating' receiver, have been simulated for the same array environment and the results are shown in Figures 5a and 5b respectively. It is clear from these figures that the proposed receiver provides a significant improvement over the RAKE receiver. It is important to highlight that both the RAKE and proposed receiver use only the PN code of the desired user (single user). In contrast the 'decorrelating' receiver uses the PN codes of the whole set of users (multi-user receiver) but, despite this, its performance is comparable to the single user receiver proposed in this paper.

Other simulation results, not presented in this paper, also show that the proposed structure outperforms the decorrelating receiver in the case of imperfect channel estimates and demonstrate its robustness against unresolved multipath rays.

5 CONCLUSIONS

This paper presents a blind single-user space-time receiver for frequency-selective DS-CDMA channels, based on an innovative joint space-time multipath channel estimation procedure. The proposed receiver has a subspace-type structure, and is able to identify jointly the coherent multipath delays and DOA in the presence of MAI's.

REFERENCES