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BLIND PARALIND MULTIUSER DETECTION FOR SMART ANTENNA CDMA SYSTEM OVER MULTIPATH FADING CHANNEL

X. Zhang, X. Gao, and Z. Wang

Electronic Engineering Department
Nanjing University of Aeronautics & Astronautics
Nanjing 210016, China

Abstract—This paper links the multipath smart antenna CDMA system signal detection problem to the PARAllel profiles with LINear Dependencies (PARALIND), and derives a deterministic blind PARALIND algorithm whose performance is very close to nonblind space-time minimum mean square error (ST-MMSE) method. The blind PARALIND algorithm has the better performance than space-time matched filter. The proposed PARALIND algorithm also works well in array error condition. Most notably, it does not require knowledge of the DOA (Direction Of Arrival) and channel fading information.

1. INTRODUCTION

Recently mobile communication is developing toward the beyond third Generation, which can provide higher speed data service and spectrum efficiency, and larger system capacity [1, 2]. So we should take effective measures to suppress interferences. Smart antenna [39–49]not only combats multipath fading, but also suppresses interference signals. When spatial signal processing achieved through smart antenna is combined with temporal signal processing, space-time 2D RAKE receiver can repair signal impairments to result in a higher network capacity, coverage, and quality. Smart antenna technique has been applied to mobile station, which becomes an important technique to meet WCDMA system demands [3]. However, 2D-RAKE is still single-user detection essentially [4]. Space-time multiuser detection combines space-time processing with multiuser detection to jointly suppress MAI

Corresponding author: X. Zhang (fei_zxf@163.com).

(multi-access interference) and ISI (inter-symbol interference) with the knowledge of multi-path and space-time signature. Miller [5] made researches on optimal space-time multiuser detection in the case of synchronization and asynchronization. Nagatsuka [6] studied on optimal space-time multiuser detection with multipath channel. However, its complexity in [5] or [6] increases exponentially with the number of users. It is not practical for a large number of users. So we look for other ways for space-time multiuser detection, like space-time MMSE multiuser detection [7], RLS based space-time multiuser detection [8]. The algorithms in [5–7] require the knowledge of space time channel. The algorithm in [8] needs the train sequence. Blind space-time multiuser detection doesn't require the knowledge of space time channel information and the train sequence. Blind space-time multiuser detections use special property of the sources, such as finite-alphabet (FA), constant-modulus (CM) [9] and cyclostationarity. Blind space-time multiuser detections also use space-time minimum output energy [10] or space-time constrained minimum output energy [11]. A novel blind space-time multiuser detection algorithm for smart antenna CDMA system over multipath fading channel is investigated in this paper.

Parallel factor (PARAFAC) analysis has been first introduced as a data analysis tool in psychometrics, most of the research in the area is conducted in the context of chemometrics [12], spectrophotometric, chromatographic and flow injection analyses. Harshman [13] developed the PARAFAC model. At the same time, Caroll and Chang [14] introduced the canonical decomposition model, which is essentially identical to PARAFAC. In signal processing field, PARAFAC can be thought of as a generalization of ESPRIT and joint approximate diagonalization ideas [15, 16]. PARAFAC is thus naturally related to linear algebra for multi-way arrays [17]. PARAFAC has been used widely in blind receiver detection for Direct-sequence code-division multiple access (CDMA) system [18], array signal processing [19–26], blind estimation of Multi-Input-Multi-Output (MIMO) system [27], blind speech separation [28], downlink receiver for space-time block-boded CDMA System [29] and multiuser detection for Single-Input-Multi-Output (SIMO) CDMA system [30].

When data have PARAFAC structure, if some of the factors are linearly dependent in one mode, the PARAFAC solution will usually be nonunique for those factors involved in the dependency. In the PARAFAC model there is no provision for explicitly specifying 'interactions' or linear dependencies between factors. PARAllel profiles with LINear Dependencies (PARALIND) model was developed to handle linearly dependent factor variation [32, 33]. PARALIND both displays the dependency structure and minimizes the loss of uniqueness. In fact, it usually holds that the interpretation is related to stating the PARALIND model as a constrained PARAFAC model. PARALIND was used in blind multiuser detection in antenna array WCDMA systems with large delay spread [34], but delay spread is integer-symbol, which isn't applicable for the practical system.

The guard interval insertion is used to blind multiuser detection for multitpath CDMA channel [18]. Guard interval can deal with inter-symbol interference caused by channel delay spreading, but it leads to spectrum efficiency decrease. Our work is to investigate blind PARALIND multiuser detection for smart antenna CDMA system over multipath fading channel. Our proposed algorithm does not require guard interval.

Our work links the multipath smart antenna CDMA system signal detection problem to PARALIND model and derives a deterministic blind PARALIND algorithm whose performance is very close to nonblind space-time minimum mean square error (ST-MMSE). The blind PARALIND algorithm has much better performance than space-time matched filter. The proposed PARALIND algorithm works well in array error condition. Most notably, it does not require knowledge of the DOA (direction of arrival) and channel fading information.

This paper is structured as follows. Section 2 develops data model. Section 3 proposes algorithm. Section 4 presents simulation results, and Section 5 summarizes our conclusions.

2. DATA MODEL

Assume there are K users that all employ BPSK spread spectrum modulation in DS-CDMA system. The kth user signal has r_k independent paths to base station. The received signal at base station, where uniform circular array antennas with M elements is employed, is shown as follows.

$$\mathbf{x}(t) = \sum_{k=1}^{K} \sum_{l=1}^{r_k} \alpha_{kl} b_k(t) \mathbf{s}_k(t - \tau_{kl}) \beta_{kl} + \mathbf{n}(t) \tag{1}$$

where $\mathbf{n}(t)$ is space-time channel white noise. β_{kl} is the channel fading of the lth path of the kth user. α_{kl} is the direction vector of the lth path of the kth user. τ_{kl} is the delay of the lth path of the kth user. $\mathbf{s}_k$ is the spread code of the kth user. $b_k(t)$ is the transmit signal of the kth user. r_k is the multipath number of the kth user. Define $\mathbf{h}_{kl} = \alpha_{kl}\beta_{kl}$ is space-time channel vector of the lth path of the kth user. Chip rate sampling is used for the received signal.

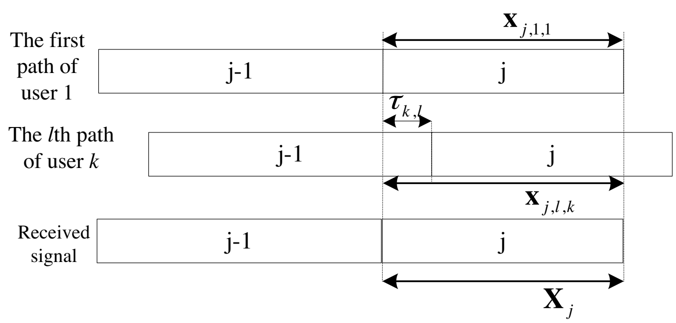

Figure 1. Asynchronous problem.

Assume the delay of the lth path of the kth user is τ_{kl} chips, and $\tau_{k,l} < P$ (P is spreading gain), which is shown in Fig. 1. The lth path signal component of the kth user's in the jth symbol of the received signal is shown as

$$\mathbf{x}_{j,l,k} = \mathbf{h}_{k,l}\mathbf{s}_{k,l}^{(1)}b_k(j-1) + \mathbf{h}_{k,l}\mathbf{s}_{k,l}^{(2)}b_k(j) \tag{2}$$

where $b_k(j)$ is the jth transmitted symbol of the kth user.

$$\mathbf{s}_{k,l}^{(1)} = [\ s_k(P-\tau_{k,l}+1) \quad \dots \quad s_k(P) \quad 0 \quad \dots \quad 0\ ] \in \mathbb{R}^{1\times P} \tag{3}$$

$$\mathbf{s}_{k,l}^{(2)} = [\ 0 \quad \dots \quad 0 \quad s_k(1) \quad \dots \quad s_k(P-\tau_{k,l})\ ] \in \mathbb{R}^{1\times P} \tag{4}$$

$[s_k(1), s_k(2), \dots, s_k(P)]$ is spread code of the kth user. The first part in (2) can be regarded as interference between the $(j-1)$th symbol and the jth symbol.

Define $r = \sum_{k=1}^{K} r_k$. The jth symbol of the received noiseless signal can be shown

$$\mathbf{X}_j = \sum_{k=1}^{K}\sum_{l=1}^{r_k}\mathbf{x}_{j,l,k} = \sum_{k=1}^{K}\sum_{l=1}^{r_k}\mathbf{h}_{k,l}\mathbf{s}_{k,l}^{(1)}b_k(j-1) + \sum_{k=1}^{K}\sum_{l=1}^{r_k}\mathbf{h}_{k,l}\mathbf{s}_{k,l}^{(2)}b_k(j) \tag{5}$$

Define

$$\mathbf{X}_j^{(2)} = \sum_{k=1}^{K}\sum_{l=1}^{r_k}\mathbf{h}_{k,l}\mathbf{s}_{k,l}^{(2)}b_k(j) = \mathbf{H}diag\left(\mathbf{b}_j^{(2)}\right)\mathbf{S}^{(2)} \tag{6}$$

where $diag(.)$ is to construct diagonal matrix.

$$\mathbf{H} = [\mathbf{h}_{1,1}, \dots, \mathbf{h}_{1,r_1}, \mathbf{h}_{2,1}, \dots, \mathbf{h}_{2,r_2}, \dots, \mathbf{h}_{K,1}, \dots, \mathbf{h}_{K,r_K}] \in \mathbb{C}^{M\times r} \tag{7}$$

$$\mathbf{S}^{(2)} = \left[\mathbf{s}_{1,1}^{(2)T}, \ldots, \mathbf{s}_{1,r_1}^{(2)T}, \mathbf{s}_{2,1}^{(2)T}, \ldots, \mathbf{s}_{2,r_2}^{(2)T}, \ldots, \mathbf{s}_{K,1}^{(2)T}, \ldots, \mathbf{s}_{K,r_K}^{(2)T}\right]^T \in \mathbb{R}^{r\times P} \quad (8)$$

$$\mathbf{b}_j^{(2)} = \left[\underbrace{b_1(j), \ldots, b_1(j)}_{r_1}, \ldots, \underbrace{b_K(j), \ldots, b_K(j)}_{r_K}\right] \in \mathbb{R}^{1\times r} \quad (9)$$

It is assumed that space-time channel state information is constant for J symbols. Eq. (6) can be denoted as

$$\mathbf{X}_j^{(2)} = \mathbf{H}D_j\left(\mathbf{B}^{(2)}\right)\mathbf{S}^{(2)}, \quad j = 1, 2, \ldots, J \quad (10)$$

where $D_j(.)$ is to extract the jth row of its matrix and construct a diagonal matrix out of it. $\mathbf{B}^{(2)}$ is shown as follows

$$\mathbf{B}^{(2)} = \left[\mathbf{b}_1^{(2)T}, \mathbf{b}_2^{(2)^T}, \ldots, \mathbf{b}_J^{(2)T}\right]^T \in \mathbb{R}^{J\times r} \quad (11)$$

Similarly define

$$\mathbf{X}_j^{(1)} = \sum_{q=1}^{K} \mathbf{h}_q \mathbf{s}_q^{(1)} b_q(j-1) = \mathbf{H}D_j\left(\mathbf{B}^{(1)}\right)\mathbf{S}^{(1)}, \quad j = 1, 2, \ldots, J \quad (12)$$

where

$$\mathbf{S}^{(1)} = \left[\mathbf{s}_{1,1}^{(1)T}, \ldots, \mathbf{s}_{1,r_1}^{(1)T}, \mathbf{s}_{2,1}^{(1)T}, \ldots, \mathbf{s}_{2,r_2}^{(1)T}, \ldots, \mathbf{s}_{K,1}^{(1)T}, \ldots, \mathbf{s}_{K,r_K}^{(1)T}\right]^T \in \mathbb{R}^{r\times P} \quad (13)$$

$$\mathbf{B}^{(1)} = \left[\mathbf{b}_1^{(1)T}, \mathbf{b}_2^{(1)T}, \ldots, \mathbf{b}_J^{(1)T}\right]^T \in \mathbb{R}^{J\times r} \quad (14)$$

where

$$\mathbf{b}_j^{(1)} = \left[\underbrace{b_1(j-1), \ldots, b_1(j-1)}_{r_1}, \ldots, \underbrace{b_K(j-1), \ldots, b_K(j-1)}_{r_K}\right] \in \mathbb{R}^{1\times r} \quad (15)$$

The jth symbol of the received noiseless signal can be shown

$$\mathbf{X}_j = \mathbf{X}_j^{(1)} + \mathbf{X}_j^{(2)} = \mathbf{H}D_j\left(\mathbf{B}^{(1)}\right)\mathbf{S}^{(1)} + \mathbf{H}D_j\left(\mathbf{B}^{(2)}\right)\mathbf{S}^{(2)}, \quad j = 1, 2, \ldots, J \quad (16)$$

Eq. (16) is also denoted as

$$\mathbf{X}_j = \left[\begin{array}{cc} \mathbf{H} & \mathbf{H} \end{array}\right] D_j(\mathbf{B}_E) \left[\begin{array}{c} \mathbf{S}^{(1)} \\ \mathbf{S}^{(2)} \end{array}\right] = \mathbf{H}\Gamma D_j(\mathbf{B}_E)\mathbf{S}_E^T, \quad j = 1, 2, \ldots, J \quad (17)$$

where

$$\mathbf{S}_E = \begin{bmatrix} \mathbf{S}^{(1)} \\ \mathbf{S}^{(2)} \end{bmatrix}^T \in \mathbb{R}^{P\times 2r} \tag{18}$$

$$\Gamma = \begin{bmatrix} 1 & & & & 1 & & & \\ & 1 & & & & 1 & & \\ & & \ddots & & & & \ddots & \\ & & & 1 & & & & 1 \end{bmatrix} \in \mathbb{R}^{r\times 2r} \tag{19}$$

$$\mathbf{B}_E = \begin{bmatrix} \mathbf{B}^{(1)} & \mathbf{B}^{(2)} \end{bmatrix} = [\mathbf{B}_1, \mathbf{B}_2]\mathbf{T} = \mathbf{BT} \in \mathbb{R}^{J\times 2r} \tag{20}$$

where $\mathbf{T} = \begin{bmatrix} \bar{\mathbf{T}} \\ \bar{\mathbf{T}} \end{bmatrix} \in \mathbb{R}^{2K\times 2r}$, and $\bar{\mathbf{T}} = \begin{bmatrix} 1_{r_1} & & 0 \\ & \ddots & \\ 0 & & 1_{r_K} \end{bmatrix} \in \mathbb{R}^{K\times r}$, 1_{r_i} is a $1 \times r_i$ vector.

$$\mathbf{B}_1 = \begin{bmatrix} b_1(0) & b_2(0) & \dots & b_K(0) \\ b_1(1) & b_2(1) & \dots & b_K(1) \\ \vdots & \vdots & \ddots & \vdots \\ b_1(J-1) & b_2(J-1) & \dots & b_K(J-1) \end{bmatrix} \in \mathbb{R}^{J\times K} \tag{21}$$

$$\mathbf{B}_2 = \begin{bmatrix} b_1(1) & b_2(1) & \dots & b_K(1) \\ b_1(2) & b_2(2) & \dots & b_K(2) \\ \vdots & \vdots & \ddots & \vdots \\ b_1(J) & b_2(J) & \dots & b_K(J) \end{bmatrix} \in \mathbb{R}^{J\times K} \tag{22}$$

$$\mathbf{B} = [\mathbf{B}_1, \mathbf{B}_2] \in \mathbb{R}^{J\times 2K} \tag{23}$$

Eq. (17) is also denoted as

$$\mathbf{X}_j = \mathbf{H}\Gamma D_j(\mathbf{BT})\mathbf{S}_E^T, \quad j = 1, 2, \dots, J \tag{24}$$

Γ and $\mathbf{T}$ are dependency matrices. The received signal is modeled as PARAllel profiles with PARALIND model or trilinear model with collinear. $\mathbf{X}_j$ can be regarded as the jth slice in the temporal direction. According to trilinear model in Eq. (24), Eq. (24) can be denoted with the following form through rearrangements, $\mathbf{Y}'_p = \mathbf{BT}D_p(\mathbf{S}_E)(\mathbf{H}\Gamma)^T$, $p = 1, 2, \dots, P$; $\mathbf{Z}'_m = \mathbf{S}_E D_m(\mathbf{H}\Gamma)(\mathbf{BT})^T, m = 1, 2, \dots, M$.

$$\mathbf{Y}_p = \left[\mathbf{Y}'_p\right]^T = \mathbf{H}\Gamma D_p\left(\mathbf{S}_E\right)\mathbf{T}^T\mathbf{B}^T, \quad p = 1, 2, \dots, P \tag{25}$$

According to Eq. (25), the received noiseless signal is shown

$$\begin{aligned} \mathbf{Y} &= \begin{bmatrix} \mathbf{Y}_1 & \mathbf{Y}_2 & \dots & \mathbf{Y}_P \end{bmatrix} \\ &= \mathbf{H}\Gamma \begin{bmatrix} D_1\left(\mathbf{S}_E\right)\mathbf{T}^T\mathbf{B}^T & D_2\left(\mathbf{S}_E\right)\mathbf{T}^T\mathbf{B}^T & \dots & D_P\left(\mathbf{S}_E\right)\mathbf{T}^T\mathbf{B}^T \end{bmatrix} \\ &= \mathbf{H}\Gamma(\mathbf{S}_E \odot \mathbf{BT})^T \end{aligned} \tag{26}$$

where $\odot$ is the Khatri-Rao product.

3. BLIND PARALIND MULTIUSER DETECTION

When linear dependencies exist, the appropriate PARAFAC solution can be called rank deficient. PARAFAC-ALS algorithm [35] can not converge to global minimum for collinear in the trilinear model. PARAFAC-ALS fails to work well in the condition of collinear [33]. PARALIND is used to resolve this problem. The signal in (26) is modeled as PARALIND. The cost function of the PARALIND model is

$$\min \left\| \tilde{\mathbf{Y}} - \mathbf{H}\Gamma(\mathbf{S}_E \odot \mathbf{BT})^T \right\|_F^2 \tag{27}$$

where $\tilde{\mathbf{Y}}$ is the received noisy signal. $\|\|_F$ stands for the Frobenius norm.

According to (25),

$$vec(\mathbf{Y}_p) = vec\left(\mathbf{H}\Gamma D_p\left(\mathbf{S}_E\right)\mathbf{T}^T\mathbf{B}^T\right) = \left[\mathbf{BT}D_p\left(\mathbf{S}_E\right) \otimes \mathbf{H}\right] vec(\Gamma) \tag{28}$$

where $vec(\mathbf{X})$ is the vector obtained by stringing out $\mathbf{X}$ column-wise to a column vector. $\otimes$ stands for Kronecker-product. Considering all slices, this leads to

$$\begin{bmatrix} vec(\mathbf{Y}_1) \\ vec(\mathbf{Y}_2) \\ \vdots \\ vec(\mathbf{Y}_P) \end{bmatrix} = \begin{bmatrix} \mathbf{BT}D_1\left(\mathbf{S}_E\right) \otimes \mathbf{H} \\ \mathbf{BT}D_2\left(\mathbf{S}_E\right) \otimes \mathbf{H} \\ \vdots \\ \mathbf{BT}D_P\left(\mathbf{S}_E\right) \otimes \mathbf{H} \end{bmatrix} vec(\Gamma) \tag{29}$$

Eq. (29) can be denoted as

$$vec(\mathbf{Y}) = [(\mathbf{S}_E \odot \mathbf{BT}) \otimes \mathbf{H}]vec(\Gamma) \tag{30}$$

Update of dependency matrix Γ is

$$\begin{aligned} vec(\Gamma) \;=\; & \left[\left(\mathbf{T}^T\mathbf{B}^T\mathbf{BT}\right) * \left(\mathbf{S}_E^T\mathbf{S}_E\right) \otimes \left(\mathbf{H}^T\mathbf{H}\right)\right]^{-1} \\ & vec\left(\sum_{p=1}^{P} \mathbf{H}^T\mathbf{Y}_p\mathbf{BT}D_p\left(\mathbf{S}_E\right)\right) \end{aligned} \tag{31}$$

where $*$ stands for Hadamard-product.

Update of Γ follows taking into account that

$$\left(\mathbf{S}_E \odot (\mathbf{BT})\right)^T \left(\mathbf{S}_E \odot (\mathbf{BT})\right) = \left(\mathbf{T}^T\mathbf{B}^T\mathbf{BT}\right) * \left(\mathbf{S}_E^T\mathbf{S}_E\right)$$

Similarly, updates of the matrices $\mathbf{B}$, $\mathbf{H}$ and $\mathbf{S}_E$ are

$$\mathbf{B} = \left(\sum_{p=1}^{P} \mathbf{Y}_p^T \mathbf{H} \Gamma D_p \left(\mathbf{S}_E \right) \mathbf{T}^T \right) \left\{ \mathbf{T} \left[\left(\Gamma^T \mathbf{H}^T \mathbf{H} \Gamma \right) * \left(\mathbf{S}_E^T \mathbf{S}_E \right) \right] \mathbf{T}^T \right\}^{-1} \tag{32}$$

$$\mathbf{H} = \left(\sum_{p=1}^{P} \mathbf{Y}_p \mathbf{B} \mathbf{T} D_p \left(\mathbf{S}_E \right) \Gamma^T \right) \left\{ \Gamma \left[\left(\mathbf{T}^T \mathbf{B}^T \mathbf{B} \mathbf{T} \right) * \left(\mathbf{S}_E^T \mathbf{S}_E \right) \right] \Gamma^T \right\}^{-1} \tag{33}$$

$$\begin{aligned} Diag(D_p \left(\mathbf{S}_E \right)) \; = \; & \left\{ \left(\mathbf{T}^T \mathbf{B}^T \mathbf{B} \mathbf{T} \right) * \left(\Gamma^T \mathbf{H}^T \mathbf{H} \Gamma \right) \right\}^{-1} \\ & Diag \left(\Gamma^T \mathbf{H}^T \mathbf{Y}_p \mathbf{B} \mathbf{T} \right), \quad p = 1, \ldots, P \end{aligned} \tag{34}$$

where the operator $Diag(\mathbf{X})$ means a column vector with the diagonal elements of $\mathbf{X}$.

Blind PARALIND multiuser detection algorithm for smart antenna CDMA system over multipath fading channel is proposed in this paper. This algorithm can be called 'PARALIND'. The detail steps are shown as follows:

Step1: Initialization for the matrices $\mathbf{B}$, $\mathbf{H}$ and $\mathbf{S}_E$. The dependency matrices Γ and $\mathbf{T}$ are known.

Step2. update for $\mathbf{B}$ according to (32)

Step3. update for $\mathbf{H}$ according to (33)

Step4. update for $\mathbf{S}_E$ according to (34)

Step5. repeat Step2 to Step4 until convergence

Step6. make decision for the estimated matrix $\mathbf{B}$ to implement the blind multiuser detection.

4. SIMULATION RESULTS

Let $\tilde{\mathbf{Y}}_p = \mathbf{H} \Gamma D_p \left(\mathbf{S}_E \right) \mathbf{T}^T \mathbf{B}^T + \mathbf{N}_p$, $p = 1, 2, \ldots, P$, be the received noisy data, where $\mathbf{N}_p$ are the AWGN matrices. We define SNR

$$\mathrm{SNR} = 10 \log_{10} \frac{\sum\limits_{p=1}^{P} \left\| \mathbf{H} \Gamma D_p \left(\mathbf{S}_E \right) \mathbf{T}^T \mathbf{B}^T \right\|_F^2}{\sum\limits_{p=1}^{P} \left\| \mathbf{N}_p \right\|_F^2} \, \mathrm{dB} \tag{35}$$

We present Monte Carlo simulations that are to assess the bit error rate (BER) performance of the proposed blind PARALIND algorithm. The number of Monte Carlo trials is 1000. We compare our PARALIND algorithms with nonblind space-time minimum mean square error (ST-MMSE) [36, 37] and space time matched filter (ST-MF). ST-MMSE offers a performance bound against which blind algorithms are measured [11, 38]. Compared with our blind PARALIND algorithm, the nonblind ST-MMSE receiver assumes the perfect knowledge of DOA, SNR and channel fading information. The PARALIND algorithm doesn't require DOA (direction of arrival) information and channel fading information.

8-element circular array is used at base stations. The spread gain is 64. We assume Binary Phase Shift Keying (BPSK) modulated signal and multipath fading channel in the simulations.

Note that J is the number of symbol-snapshots and K is the number of users.

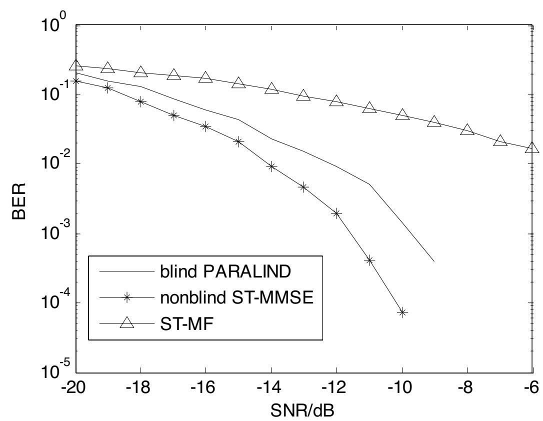

Figure 2. Algorithm performance comparison with $K = 4$ and $J = 200$.

Simulation 1. The performances of these algorithms under different J are shown in Fig. 2–Fig. 4. There are 4 users in this simulation, and each user has two paths to base station.

Figure 2 presents large sample results for $J = 200$. From Fig. 2, we find that the BER performance of blind PARALIND algorithm is close to nonblind ST-MMSE, and PARALIND algorithm has much better performance than ST-MF receiver.

Figure 3–Fig. 4 depict results for $J = 100$ and $J = 50$, respectively. From Fig. 2–Fig. 4, we find that the gap between blind PARALIND and (nonblind) ST-MMSE receiver increases as J decreases.

Simulation 2. The performance of the PARALIND algorithm under different user number K is investigated in this simulation. Fig. 5 depicts result for $J = 100$ and different K. It is clear that PARALIND algorithm performance degrades with the number of users K increasing. When the number of users increases, if satisfying the condition of (the definition of k-rank is shown in [17]. r is total path number of all user), our proposed algorithm can works well."

Simulation 3. The actual array parameters may differ from the nominal array in several ways-gain, phase and sensor location errors.

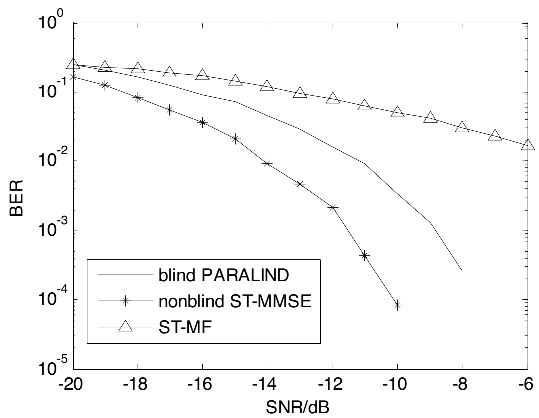

Figure 3. Algorithm performance comparison with $K = 4$ and $J = 100$.

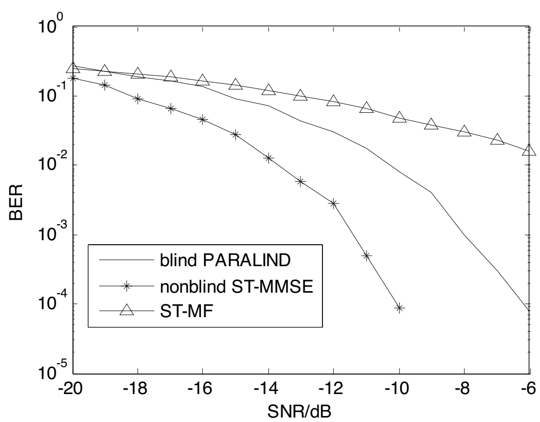

Figure 4. Algorithm performance comparison with $K = 4$ and $J = 50$.

Gain and phase errors occur when the response of each antenna to a known signal has the different amplitude and/or phase response than expected. Blind PARALIND multiuser detection algorithm performance in the array error condition is also investigated. In this simulation, array error vector is the array gain and phase error vector. The array error vector $\mathbf{g} = [1,\ 0.6071-0.6953i,\ 1.0083+0.9059i,\ 0.3497-0.7167i,\ 0.9693+1.2916i,\ 0.5343+0.3883i,\ 0.7330+1.5894i,\ 0.8878-1.5133i]$. Assume that array response vector is α, and then the array response vector with array error is $diag(\mathbf{g})\alpha$. The sample number J is 100 and the user number K is 4 in this simulation. The

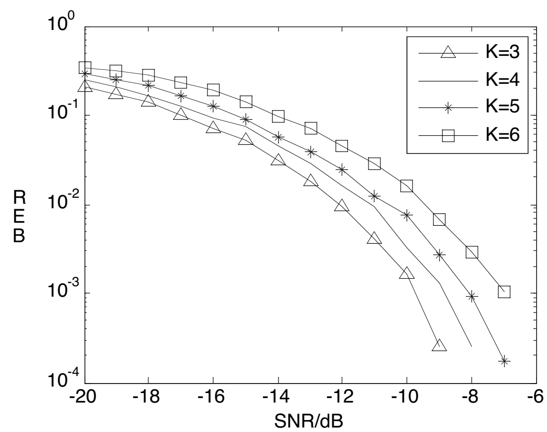

Figure 5. Algorithm performance under different K.

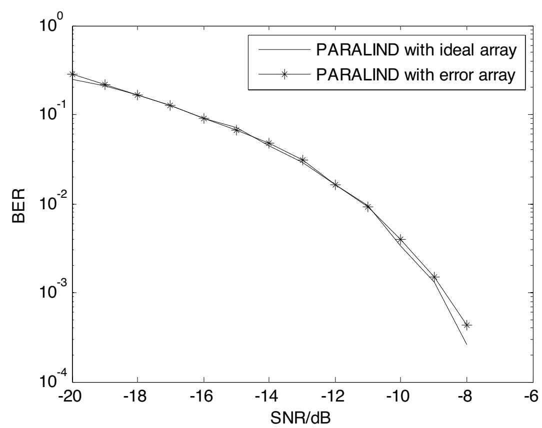

Figure 6. Algorithm performance with array error.

performance of blind PARALIND algorithm in array error condition is shown in Fig. 6. Fig. 6 shows that blind PARALIND algorithm has the better performance in the array error condition. Blind PARALIND algorithm has robust characteristics to array error.

5. CONCLUSIONS

Our work links the multipath smart antenna CDMA system signal detection problem to the PARAllel profiles with LINear Dependencies (PARALIND) and derives a deterministic blind PARALIND algorithm whose performance is very close to nonblind space-time minimum mean square error (ST-MMSE). The blind PARALIND algorithm has the much better performance than space-time matched filter. The proposed PARALIND algorithm also works well in array error condition. Most notably, it does not require knowledge of the DOA and channel fading information. Compared with the multiuser detection algorithm in [34], our proposed algorithm can work wider than the algorithm in [34]. Multiuser detection algorithm in [34] only works well in integer-symbol delay condition. Compared with Ref. [18], our proposed algorithm does not require guard interval, and increases spectrum efficiency.

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