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A Novel DS-CDMA Direction of Arrival Estimator for Frequency- Selective Fading Channel with Correlated Multipath by Using Beamforming Filter

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Abstract—**In this paper, a new method for estimating the direction of arrival of Asynchronous DS-CDMA signals in multipath fading channel is proposed. In the proposed method, first, for removing the effects of undesired paths, we project the coherent signals of the undesired paths perpendicularly to the under process signal, next, the signal is passed through a filter bank.Byusing beam forming filters, the effectof other users on the desired users signal is decreased and hence, the search area is decreased almost to one tenth. In this algorithm, searching all angles is not required,to estimate the number of sources does not require any information criteria, andalso, the number of users can exceed the number of antenna arrays contrary to many of the conventional methods. Simulation results are illustrated to confirm the efficiency of the method.**

Keywords; ***Direction of arrival estimation, orthogonal projection, beam forming, frequency-selective fading channel.***

I. Introduction

The system capacity can be increased by using temporal or spatial signal processing techniques to transmit and receive signals. The smart antenna techniques, including beamforming and diversity ones, are especially useful in mobile direct-sequence code division multiple access (DS-CDMA) communication systems [1].

Here, we consider the state that a beamforming array is adopted at the base station while each mobile user transmits the signal by using a single antenna. In such a system, direction of arrival (DOA) of the received signals is a parameter for the base station to be estimated, in order to set up reliable connection for data transmission among mobile users. The DOAs of users are not available in practice and should be estimated for implementation of optimum receivers.

Furthermore in recent years, estimating the direction of arrival has been an attractive area of research because of its important application in radar and wireless location finding. Among the proposed methods, the signal subspace algorithms have attracted a lot of interest due to their high resolution. However, in a highly correlated or coherent environment due to multipath propagation, the direction of coherent signals cannot be detected via conventional subspace methods like the MUltipleSIgnal Classification (MUSIC) algorithmandEstimation of Signal Parameters via Rotational Invariance Techniques(ESPRIT),since the spatial signatures cannot be resolved in the signal subspace [2-4].

Furthermore, in order to estimate DOA by conventional methods, the number of array elements must be greater than the number of users; this is impractical in CDMA systems with a large number of active users. A DOA estimator employing code matched filters and parallel MUSIC is proposed in [5]. In this method other users except of active users are considered Gaussian noise and as aninterference. In [6] a subspace method for estimating DOA for multiple coherent narrowband signals is proposed.

In this paper, we propose a method for DS-CDMA signal DOA estimation with an array of antennas at the receiver. First, we decorrelate the received signal of each path by orthogonal projection and then apply a bank of $2M$ (M is the number of arrays) beam forming filters to the received two-dimensional (2-D) signal which results in $2M$ time domain sequences. The total energy of interfering signals is reduced after each beam forming filter. The energy of desired signalis also reduced, since the DOA is still unknown at this stage. However, the proposed algorithm, together with this receiver structure is able to control the loss of energy while maintaining the SNR level.

The paper is organized as follows. In Section II, we describe the mathematical model of the system, set the underlying assumptions and define the problem objectives. In Section III, the proposed method is developed for synchronous single path and asynchronous multipath cases. In Section IV, performance of the proposed method is evaluated by simulation, and Section V concludes the paper.

II. MATHEMATICAL MODEL

A. Notation

The notations employed in this paper are standard. Signals are discrete-time and complex in general. Upper and lower-case bold letters denote matrices and vectors, respectively. The operators $(\cdot)^T$ and $(\cdot)^H$ denote transpose and hermitianoperationsrespectively. Finally $\otimes$ isKronecker matrix product.

B. System Model

Consider a DS-CDMA system with K active users transmitting binary information sequences of $b_1, b_2, \ldots, b_K$ with normalized spreading waveforms $s_1, s_2, \ldots, s_K$ that are randomly distributed in space. A Q bit transmitted baseband signal from the *k-th* user is:

$$x_k(t) = A_k \sum_{i=0}^{Q-1} b_k(i) s_k(t - iT_b) \qquad (1)$$
$$k = 1, 2, \ldots, K$$

Where T_b is the bit interval, $b_k(i) \in \{-1, +1\}$ is the *i-th* bit of a sequence of independent and identically distributed(i.i.d) random variables transmitted by the *k-th* user and A_k denotes the amplitude of the *k-th* user. $s_k(t)$ is defined as follows and its energy is limited to $[0, T_b]$:

$$s_k(t) = \sum_{j=0}^{N_c-1} c_k(j)\psi(t - jT_c) \quad 0 \le t \le T_b \qquad (2)$$

where $N_c = T_b / T_c$ is the processing gain; and $\psi(t)$ is a chip waveform of duration T_c and $\{c_k(n)\}_{n=1}^{n=N_c-1}$ is a signature code sequence of ± 1s assigned to the *k-th* user that can be represented as $\underline{c}_k = [c_k(0) \;\; c_k(1) \;\; \ldots \;\; c_k(N_c - 1)]^T$.

At the receiver, an antenna array of M elements is employed and the baseband multipath channel of the *k-th* user can be modeled as a single-input multiple-output channel with $M \times 1$ vector impulse response $\underline{h}_k(t)$ given as [5]

$$\underline{h}_k(t) = \sum_{l=1}^{L} g_{kl} \delta(t - \tau_{kl}) \underline{a}_{\theta_{kl}} \qquad (3)$$

where L is the number of paths in each user's channel, g_{kl} and τ_{kl} are gain and delay of the *l-th* path of the *k-th* user's signal respectively, $\delta(.)$ is the Dirac delta function and $\underline{a}_{\theta_{kl}} = [a_1^{\theta_{kl}}, \ldots, a_M^{\theta_{kl}}]^T$ is the array response vector corresponding to the *l-th* path of the *k-th* user's signal with DOA of θ_{kl}.

Thetotal received baseband signal at the *i-th* antenna denoted by $r_i(t)$ is the superposition of the signals from all users plus the additive ambient noise and the $M \times 1$ vector $\underline{r}(t) = [r_1(t), \ldots, r_M(t)]^T$ can be expressed as:

$$\begin{aligned} \underline{r}(t) &= \sum_{k=1}^{K} x_k(t) * \underline{h}_k(t) + \sigma \underline{n}(t) \\ &= \sum_{i=0}^{Q-1} \sum_{k=1}^{K} A_k b_k(i) \sum_{l=1}^{L} \underline{a}(\theta_{kl}) g_{kl} \\ &\times s_k(t - iT_b - \tau_{kl}) + \sigma \underline{n}(t) \end{aligned} \qquad (4)$$

where $*$ denotes convolution, σ^2 is the variance of the ambient noise at each antenna element and $\underline{n}(t) = [n_1(t), \ldots, n_M(t)]^T$ is a vector of independent zero mean complex white Gaussian noise processes with unit variance, i.e.

$$E\{\underline{n}(t)\underline{n}(t')^{H}\} = I_{M}\delta(t-t') \quad (5)$$

Where E is the expectation operator and I_M is the $M \times M$ identity matrix. We also assume that the noise processes and transmitted sequences of users are statistically independent.

To find the directions of the received signals from the *l-th* path of the *k-th* user, the receiver's chip matched filter is synchronized with the delay of τ_{kl}. The receiver works at the discretechip rate.The sample of *m-th* antenna is:

$$y_{kl,m}(i,j) = \int_{iT_b+\tau_{kl}+jT_c}^{iT_b+\tau_{kl}+(j+1)T_c} r_m(t)\psi^*(t-iT_b-\tau_{kl}-jT_c)dt \quad (6)$$

M samples of antenna output can be represented by a $M \times 1$ vector as:

$$\underline{y}_{kl}(i,j) = \begin{bmatrix} y_{kl,1}(i,j) \\ y_{kl,2}(i,j) \\ \vdots \\ y_{kl,M}(i,j) \end{bmatrix} \quad (7)$$

Or:

$$\underline{y}_{kl}(i,j) = \begin{bmatrix} y_{kl,1}(i,j) \\ y_{kl,2}(i,j) \\ \vdots \\ y_{kl,M}(i,j) \end{bmatrix} = \int_{iT_b+\tau_{kl}+jT_c}^{iT_b+\tau_{kl}+(j+1)T_c} \underline{r}(t)\psi^*(t-iT_b-\tau_{kl}-jT_c)dt \quad (8)$$

By combining thevectors $\underline{y}_{kl}$, we define the $M \times N_c$ matrix $BL_{kl}(i)$ as follows:

$$BL_{kl}(i) = [\underline{y}_{kl}(i,0), \underline{y}_{kl}(i,1), \ldots, \underline{y}_{kl}(i,N_c-1)] \quad (9)$$

The contribution of *k-th* user at *m-th* antenna ($r_m(t)$) is:

$$r_m^k(t) = A_k \sum_{q=0}^{Q-1} b_k(i) \sum_{l=1}^{L} g_{kl} a_m(\theta_{kl}) s_k(t-qT_b-\tau_{kl}) \quad (10)$$

Therefore the contribution of *k-th* user at $y_{kl,m}(i,j)$ is:

$$BL_{kl,m}^k(i) = y_{kl,m}^k(i,j) \quad 0 \le j \le N_c-1$$
$$= A_k \sum_{q=0}^{Q-1} b_k(q) \sum_{p=1}^{L} g_{kp} a_m(\theta_{kp})$$
$$\times \int_{iT_b+jT_c+\tau_{kl}}^{iT_b+(j+1)T_c+\tau_{kl}} s_k(t-qT_b-\tau_{kp})\psi^*(t-iT_b-jT_c-\tau_{kl})dt \quad (11)$$

Due to $\psi(t)$ waveform at $[0,T_c]$ and $s_k(t)$ at $[0,T_b]$, product of $s_k(t-qT_b-\tau_{kp})$ and $\psi(t-iT_b-jT_c-\tau_{kl})$ is always zero except in two cases:

$$iT_b + jT_c + \tau_{kl} + T_c \ge qT_b + \tau_{kp} \quad (12)$$

$$iT_b + jT_c + \tau_{kl} < qT_b + \tau_{kp} + T_b \quad (13)$$

Since the effect of i-1, i, i+1-th bits in integral, $y_{kl,m}^k(i,j)$ simplifies to:

$$y_{kl,m}^k(i,j) = A_k b_k(i-1) \sum_{p=1}^{L} g_{kp} a_m(\theta_{kp}) \underline{c}_{kp}^{l,-1} + A_k b_k(i) \sum_{p=1}^{L} g_{kp} a_m(\theta_{kp}) \underline{c}_{kp}^{l,0} + A_k b_k(i+1) \sum_{p=1}^{L} g_{kp} a_m(\theta_{kp}) \underline{c}_{kp}^{l,+1} \quad (14)$$

Where $\underline{c}_{kp}^{l,s}$ is a $1 \times N_c$.

$$\underline{c}_{kp}^{l,s} = [c_{kp}^{l,s}(0), c_{kp}^{l,s}(1), \ldots, c_{kp}^{l,s}(N_c-1)] \quad (15$$

$$c_{kp}^{l,-1} = \int_{jT_c+\tau_{kl}}^{(j+1)T_c+\tau_{kl}} s_k(t+T_b-\tau_{kp})\psi^*(t-jT_c-\tau_{kl})\,dt \quad (16)$$

$$c_{kp}^{l,0} = \int_{jT_c+\tau_{kl}}^{(j+1)T_c+\tau_{kl}} s_k(t-\tau_{kp})\psi^*(t-jT_c-\tau_{kl})\,dt \quad (17)$$

$$c_{kp}^{l,+1} = \int_{jT_c+\tau_{kl}}^{(j+1)T_c+\tau_{kl}} s_k(t-T_b-\tau_{kp})\psi^*(t-jT_c-\tau_{kl})\,dt \quad (18)$$

As a result, the contribution of *n-th* user at $BL(i)$ matrix is:

$$BL_{kl}^k(i) = A_k b(i-1) \sum_{p=1}^{L} g\underline{a}(\theta_{kp}) \otimes \underline{c}_{kp}^{l,-1} + A_k b(i) \sum_{p=1}^{L} g\underline{a}(\theta_{kp}) \otimes \underline{c}_{kp}^{l,0} + A_k b(i+1) \sum_{p=1}^{L} g\underline{a}(\theta_{kp}) \otimes \underline{c}_{kp}^{l,+1} \quad (19)$$

III. THE PROPOSED METHOD

A. Proposed method for synchronous single path case

First we consider the case where all users are transmitting synchronously in a single path fading system, i.e. $L=1$. Without loss of generality, assume $\tau_1=\tau_2=\ldots=\tau_K$. In this case BL matrix is independent of indices k and l and simplifies to:

$$BL_{kl}^{k}(i)=A_k b_k(i) g_k \underline{a}(\theta_k)\otimes \underline{c}_k \tag{20}$$

A bank of $2M$ beamforming filters W_s is applied to BL. The beamforming filterbank W_s is in the formof $[\underline{w}_{s,1},\underline{w}_{s,2},\ldots,\underline{w}_{s,2M}]$,where it has $M\times 2M$ dimension. Each beamforming filter steers at a different direction. The normalized *m-th* beamforming is set up as:

$$\underline{w}_{s,m}=\frac{1}{M}[1,e^{-j\pi\frac{(m-1)}{M}},\ldots,e^{-j\pi\frac{(M-1)(m-1)}{M}}]^T \tag{21}$$

which steers at $\sin\theta=m-1/M$ for $1\le m\le M+1$ or $\sin\theta=\frac{m-1-2M}{M}$ as $m>M+1$. The example of M=2 is shown in Fig. 2.

The filter output has $2M\times M$ dimension :

$$X=W_s^{\,H}BL(i)=\begin{bmatrix}\underline{w}_{s,1}^{H}BL(i)\\ \vdots\\ \underline{w}_{s,2M}^{H}BL(i)\end{bmatrix}= \tag{22}$$

$$\begin{bmatrix}x_1 & \cdots & x_{2M}\end{bmatrix}^T$$

The filter response for direction of *k-th* user is:

$$\underline{\Gamma}_k=W_s^H\underline{a}(\theta_k) \tag{23}$$

And if we use uniform linear array (ULA) antenna, then:

$$\underline{\Gamma}_k=\frac{1}{M^2}\begin{bmatrix}\sum_{v=0}^{M-1}e^{-jv\pi\sin\theta}\\ \vdots\\ \sum_{v=0}^{M-1}e^{-jv\pi(\sin\theta-\frac{2M-1}{M})}\end{bmatrix} \tag{24}$$

The signal in *i-th* block is:

$$SW=\sum_{k=1}^{K}A_k g_k\underline{\Gamma}_k d_k(i)\underline{c}_k^T \tag{25}$$

If the direction of *k-th* user is desired, by assuming:

$$\begin{cases}E[d_k(j_1)d_k(j_2)]=0 & j_1\ne j_2\\ E[d_k(j_1)d_m(j_2)]=0 & k\ne m\end{cases} \tag{26}$$

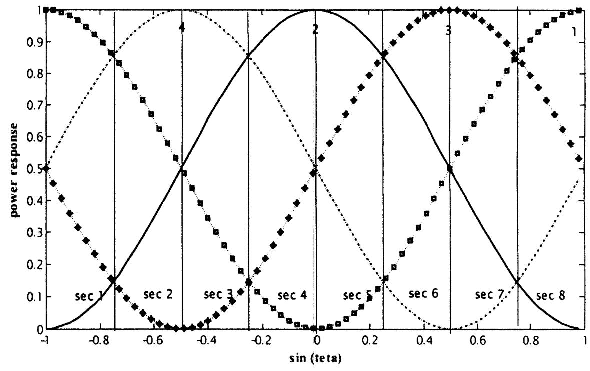

Figure 1. The beam shape at space-doamin for the 2-antenna case: the power response from all beamforming filters

The correlation matrix for the *m-th* row (i.e. the *m-th* beam forming filter output) is:

$$R_m=E\{\underline{x}\underline{x}^H\}=E\{\underline{x}_m^T\underline{x}_m^*\} \tag{27}$$

Where:

$$\underline{x}_m=\sum_{k=1}^{K}A_k g_k d_k(i)\sum_{v=0}^{M-1}e^{-j\pi v(\sin\theta_k-\frac{m-1}{M})}\underline{c}_k^T \tag{28}$$

Thus:

$$R_m=(\sum_{k=1}^{K}A_k g_k d_k(i)\Gamma_{k,m}\underline{c}_k)(\sum_{k=1}^{K}A_k g_k d_k(i)\Gamma_{k,m}^*\underline{c}_k^T) \tag{29}$$

Where:

$$\Gamma_{k,m}=\sum_{v=0}^{M-1}e^{-j\pi v(\sin\theta_k-\frac{m-1}{M})} \tag{30}$$

(e,f)-th element of this matrix is:

$$\begin{aligned}&\{A_1g_1d_1(i)\Gamma_{1,m}c_1(e)+A_2g_2d_2(i)\Gamma_{2,m}c_2(e)+\ldots+\\ &A_Kg_Kd_K(i)\Gamma_{K,m}c_K(e)\}\times\\ &\{A_1g_1d_1(i)\Gamma_{1,m}^*c_1(f)+A_2g_2d_2(i)\Gamma_{2,m}^*c_2(f)+\ldots+\\ &A_Kg_Kd_K(i)\Gamma_{K,m}^*c_K(f)\}\end{aligned} \tag{31}$$

R_m stands for the MAI plus noise covariance matrix:

$$R_m=R_{\underline{N},m}+\left|A_k g_k\underline{\Gamma}_{k,m}\right|^2\underline{c}_k\underline{c}_k^T \tag{32}$$

The direction finding algorithm first identifies a section that the desired signal may fall in. To determine the desired section, we use the largest responses belonging to the beam forming filter. Table I and fig. 1 show the mapping for the 2-antenna system. To achieve this goal, we use the following function that is constructed for each row [7]:

$$\varphi_m = \frac{1}{\underline{c}_k^T R_m^{-1} \underline{c}_k} \quad (33)$$

Now we can compute equation 33 for each rowthen select two largest values of them named$M1$ and $M2$. For a given $\sin\theta$ value in the desired section, we can have two response values into a vector representation as:

$$\underline{b}_s(\sin(\theta)) = \frac{1}{M^2}\begin{bmatrix} \left|\sum_{v=0}^{M-1} e^{-j\pi v\sin(\theta)}\right|^2 \\ \left|\sum_{v=0}^{M-1} e^{-j\pi v(\sin(\theta)-\frac{1}{M})}\right|^2 \end{bmatrix} \quad (34)$$

Where $\sin\theta$ ranges from 0 to $1/2M$. After identification of target section, a fine search is performed. $\underline{b}_s(\sin(\theta))$ can be treated as a steering vector to search in the range $\sin\theta = 0 \sim 1/2M$. By collecting the two largest soft values from (33) to form a vector and constructing an orthogonal projector, we obtain:

$$\underline{\hat{\varphi}} = \begin{bmatrix} \underline{\hat{\varphi}}_{M1} \\ \underline{\hat{\varphi}}_{M2} \end{bmatrix} \quad (35)$$

$$P_{\underline{\hat{\varphi}}} = \frac{\underline{\hat{\varphi}}\,\underline{\hat{\varphi}}^T}{\left\|\underline{\hat{\varphi}}\right\|^2} \quad (36)$$

$$\sin(\hat{\theta}) = \arg\max_{\sin(\theta)} \frac{\left\|\underline{b}_s(\sin\theta)\right\|^2}{\underline{b}_s(\sin\theta)^T (I - P_{\underline{\hat{\varphi}}})\underline{b}_s(\sin\theta)} \quad (37)$$

Let sec stands for the estimated section, then the angle θ_k can be obtained by :

$$\hat{\theta}_k = \begin{cases} \sin^{-1}(\sin(\hat{\theta}) + (\sec - 2M - 1)\frac{1}{2M}) & \sec \textit{ is odd} \\ \sin^{-1}(\frac{1}{2M} - \sin(\hat{\theta}) + (\sec - 2M - 1)\frac{1}{2M}) & \sec \textit{ is even} \end{cases}$$

(38)

B. *Proposed method for asynchronous multipath case*

The algorithm for asynchronous multipath fading is anextension of the method for the single-path synchronouscase. In this case, we collect the signals of the desired userfrom all L different paths. The method of DOA estimation forthe single-path case (relation (38)) is not applicable directlysince the correlation between signals from different paths ofthe *k-th* user will hinder its applicability. We should try todecorrelate the received signal of each path from otherpaths of the *k-th* user. Without loss of generality, we assumethat the different paths of users are numbered in increasing order of path delays, i.e. $\tau_{k1} \le \tau_{k2} \le \ldots \le \tau_{kL}$ for $k \in \{1,\ldots,K\}$. The received vector of N_c samples of chip matched filter synchronised with τ_{kl} (the delay of the *l-th* path of the *k-th* user) at any antenna will contain interference of the samesymbol from the *m-th* path of the *k-th* user, which can berepresented by a $N_c \times 1$ vector as:

$$\underline{t}_m = [t_m(0), \ldots, t_m(N_c - 1)]^T \quad (39)$$

$$t_m(j) = \int_{jT_c+\tau_{kl}}^{(j+1)T_c+\tau_{kl}} s_k(t-\tau_{km})\psi^*(t - jT_c - \tau_{kl})dt \quad (40)$$

In fact $\underline{t}_m$ is the interference of *m-th* path of the *k-th* user on the *l-th* path of the *k-th* one. We define the $N_c \times (L-1)$ matrix C_{kl} with columns $\underline{t}_m$ for $\{m = 1,\ldots,L \; m \ne l\}$

$$C_{kl} = [\underline{t}_1 \vdots \underline{t}_2 \vdots \ldots \vdots \underline{t}_{l-1} \vdots \underline{t}_{l+1} \vdots \ldots \vdots \underline{t}_L] \quad (41)$$

The column space of matrix C_{kl} (space spanned bycolumns of C_{kl}) is in fact the interference space caused bythe different paths of the *k-th* user on the *l-th* path of the *k-th*user. To decorrelate the received signal from the *l-th* path ofthe *k-th*user and the other paths of the *k-th* user,we just need to project the received vector of N_c samples of the chip matched filter into a space orthogonal to the column space of matrix C_{kl} . The projection operator into the orthogonal space of C_{kl} is [8]:

$$P_{kl}^{\perp} = I_N - C_{kl}(C_{kl}^H C_{kl})^{-1} C_{kl}^H \quad (42)$$

In other words, the vector $BL_{kl}^{k\,\perp}(i) = P_{kl}^{\perp} BL_{kl}^k(i)$ is not affectedby other paths and is totally decorrelated fromthe usersof other path signals, then we can apply the previous method to $BL_{kl}^{k\,\perp}$ and estimate the direction of the arrival.

IV. SIMULATION

In this section, the performance of the proposed method is evaluated in a multipath DS-CDMA system using gold code sequences of length 31. In all simulations, the receiver has a uniform linear array with half wavelength spacing between adjacent antennas and the antenna elements are assumed to be omnidirectional. There are eight active users (K=8) and three paths for each user. The real and the imaginary parts of the complex channel gains have been generated randomly by a zero mean Gaussian distribution with unit variance.

The root mean square error (RMSE) and bias versus number of data samples, SNR and number of array elements are evaluated for the first, second and third paths of 1-*th* user and plotted in Fig.2-7.To compare the performance of the algorithm, we compare this method with theproposed algorithms in [5], [8]and [9] in Fig. 8 and 9.In Fig. 10 and 11, the bias and RMSE of proposed algorithm for DOA estimation are illustrated for different value of SNR respectively. As we expected, with increase of SNR, bias and RMSE of proposed algorithm for DOA estimation has been decreased. In the following, to show the consistency of our algorithm the RMSE of proposed algorithm for DOA estimation is illustrated for different value of the data samples in Fig. 13. As we see in simulation result with increase of the number of data samples RMSE tend to zero. This clearly shows the consistency of the proposed algorithm for DOA estimation.

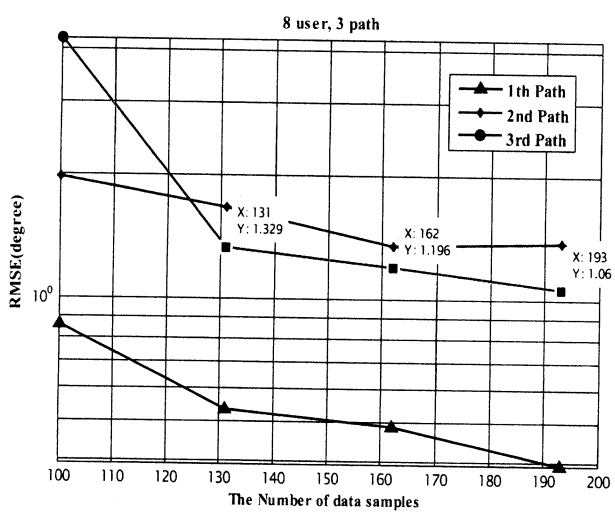

Figure 2. RMSE versus the Number of data samples (M=3)

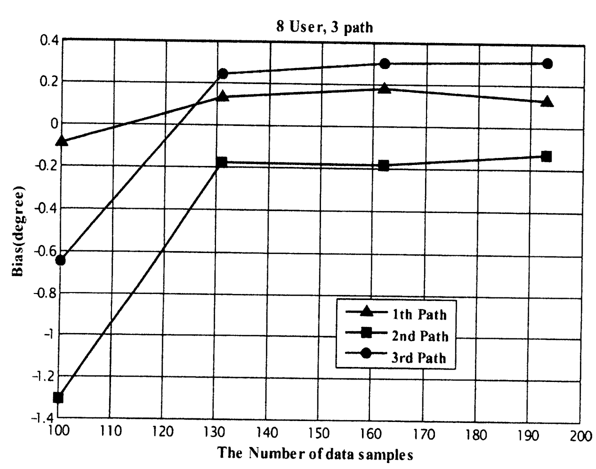

Figure 3. Bias versus the Number of data samples (M=3)

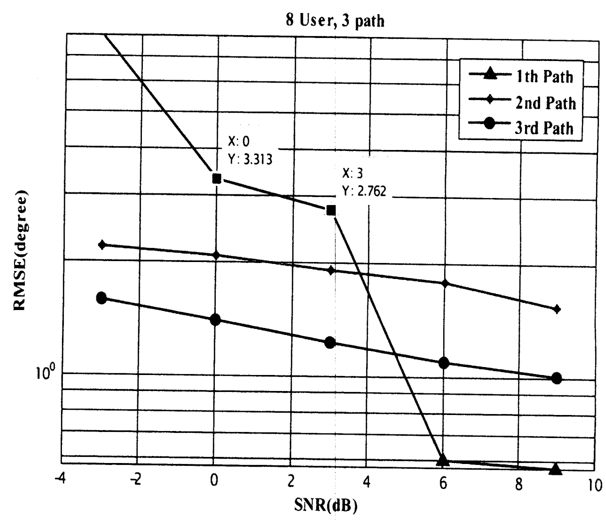

Figure 4. RMSE versus SNR (M=3)

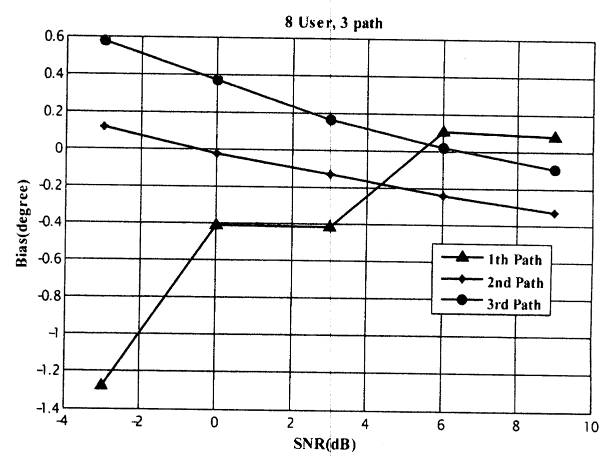

Figure 5. Bias versus SNR (M=3)

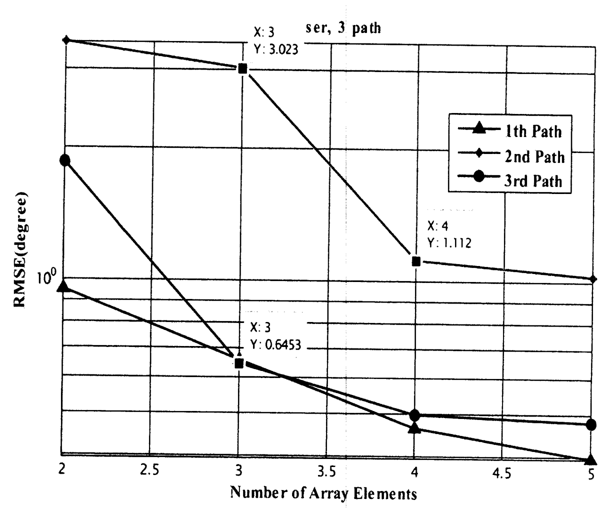

Figure 6. RMSE versus Number of Array Elements

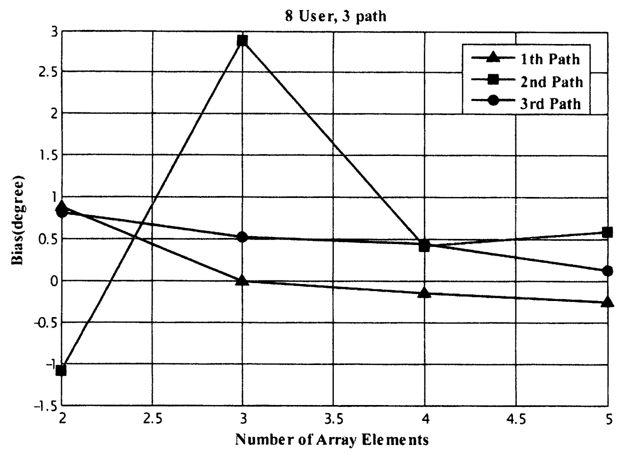

Figure 7. Bias versus Number of Array Elements

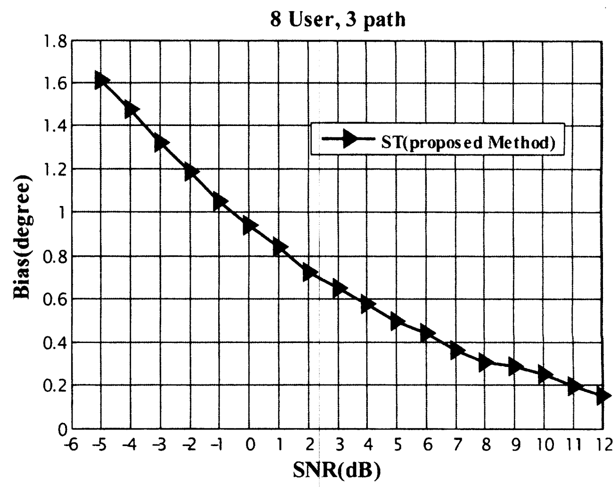

Figure 10. Bias versus SNR for proposed method

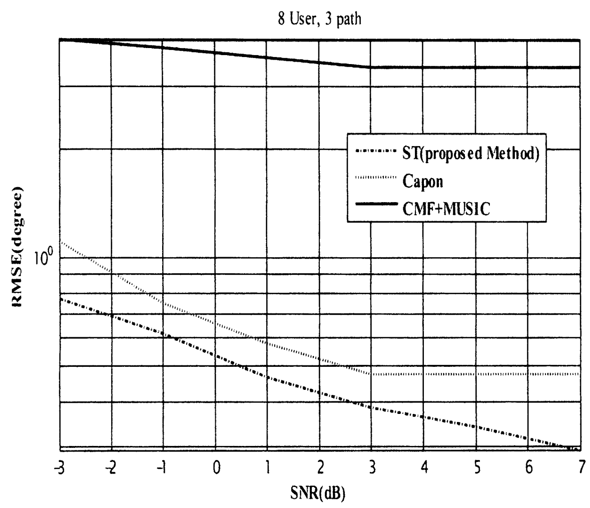

Figure 8. Comparison of RMSE against SNR for three methods

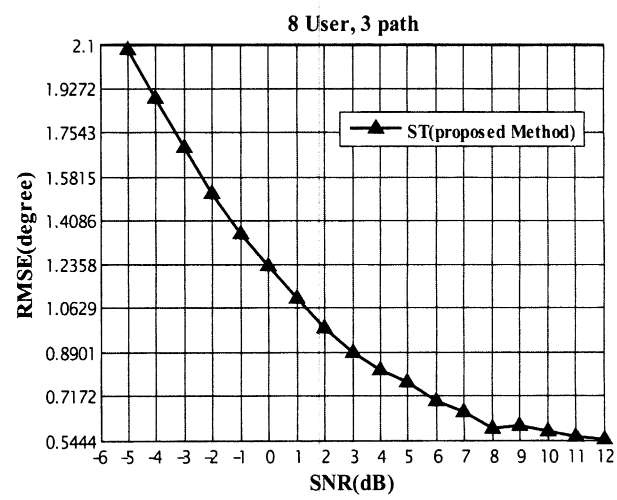

Figure 11. RMSE versus SNR for propsed method

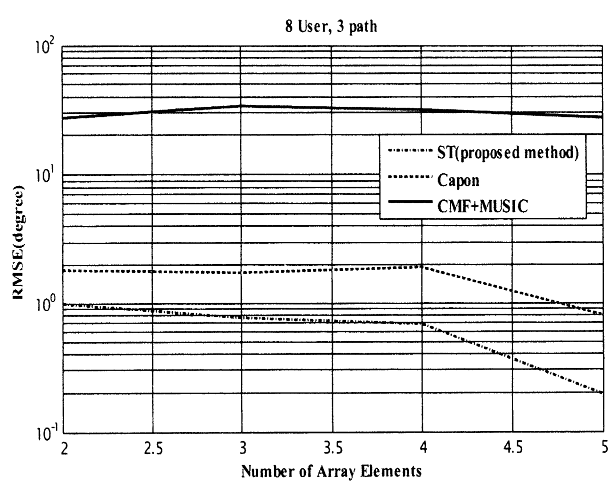

Figure 9. Comparison of RMSE against Number of Array Elements for three methods

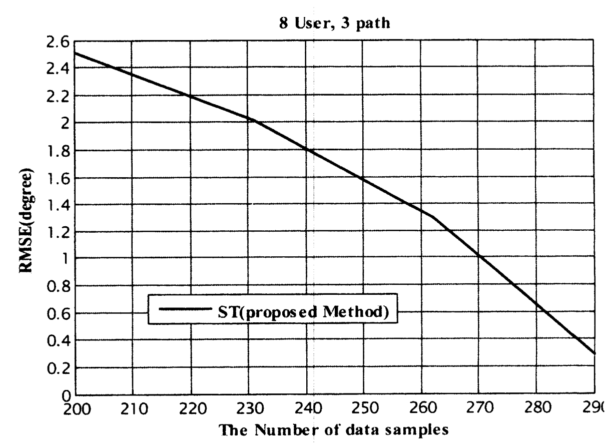

Figure 12. RMSE versus the number of data samples for proposed method

TABLE I. THE MAPPING OF SECTIONS AND BEAMFORMING FILTERS.

section	range for 2-antenna	2-antenna , m=
1	-1,-3/4	3,4
2	-3/4,-2/4	4,3
3	-2/4,-1/4	4,1
4	-1/4,0	1,4
5	0,1/4	1,2
6	1/4,2/4	2,1
7	2/4,3/4	2,3
8	3/4,1	3,2

V. CONCLUSION

In this paper a new method for estimating the direction of arrival of signals in frequency-selective fading channel with correlated multipaths is proposed. At first, we decorrelated the received signal of each path from other paths, and then passed it from a beam forming filter. In this approach in contrast to other DOA estimation approaches, it was not required to search all angles, and it did not require EVD. Due to the equivalency of thescenario to a DOA estimation of single source in the noise environment, AIC and MDL criteria werenot required for estimating the number of sources. Since beam forming filters wereused, the destroyingeffect ofother users on the desired users signal wasalso decreased and the efficiency of the algorithm was increasedand the search area wasdecreased almost to one tenth. Some of the other advantageous of the proposed method are, forming of correlation matrix including vectors with smaller dimension which results in lower complexity of calculation. Also, Simulations have shown that the suggested approach works properly in the case that the number of sources exceeds the number of array elements.The simulation results show that the proposed method has good stability in the multipath fading channel and the estimates of the proposed method are consistent. Intensification of the multipath fading reduces the performance of estimator. This efficiency decrease can be compensated by increasing.

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