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Joint Parameter Estimation of Multiuser Asynchronous DS CDMA Signals in Unknown Fading Channel in DS CDMA System with Multiple Antennas at the Base-Station and Single Antenna at the Mobile

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Abstract: We present MUSIC type algorithm for joint parameter estimation in asynchronous DS CDMA system with multiple antenna system at base station and single antenna system at the mobile users. The proposed algorithm is applied in downlink processing and it enables the joint estimation of time delay, frequency shift as well as spatial parameters (direction of departure - DOD of transmitted signal) of asynchronous DS CDMA signals using single antenna system at the mobile. The proposed algorithm is well suited to unknown fading channel, frequency selective or dispersive, and it can be primarily used as a method of multipath channel identification. It seems to the authors that it can be used as a basis for the further improvement of process of detection at the mobile receiver. The proposed algorithm is near-far resistant, requires no preamble, and can be applied when the user sequences are not perfectly orthogonal.

Keywords: Signal processing, MUSIC algorithm, CDMA system, Fading channel, antenna.

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1 Introduction

The application of antenna arrays in mobile communication systems is of great theoretical interest in recent years. A number of theoretical papers prove that antenna arrays can significantly improve the system performances such as increasing channel capacity and spectrum efficiency and reducing the effects of multipath fading, etc., [1]. But, there are some limitations in practical application of antenna arrays. First of all, the application of antenna arrays is better suited to the base stations. At the portable mobile the application of antenna arrays is limited and it is not practical.

In recent years many papers have been published that are related to the problem of transmit antenna diversity, [2]. The authors have noticed that in many papers related to that, the parameters of antenna array geometry (transmit antenna array steering vector) is not incorporated in mathematical model of the transmitted signal. The problem of transmit antenna geometry is almost the same as the geometry problem of the receive antenna arrays. In whatever way the antenna elements in transmit antenna array are located, the effect of antenna geometry can be resulted in spatial modulation of the transmitted signal, with different effects from one antenna geometry to the other. Then, an idea appeared to the authors: How can this fact be exploited on the mobile side with single antenna receiver to improve the global performances of detection? We supposed a kind of multiple antenna system at the base station which the same data stream is SS modulated in each transmit antenna channel with different sequences. So the set of orthogonal sequences is assigned to each user. In such system, the structure of transmitted signal is spatially dependent and in each point in the space carries the information about the space parameters (azimuth and elevation of the signal departure from the base station). If the set of own user code sequence and transmit antenna array manifold of the base station is known to the mobile receiver, it is possible to formulate an algorithm for joint estimation of spatial parameters (direction of departure - DOD), time delay and frequency shift. From the mobile receiver point of view, spatial dependency of transmitted signal have as a consequence that all paths in discrete multipath fading channel with different direction of arrival can be detected as a separate user. At the mobile single antenna receiver the processing is basically in time domain but the effects are almost the same as the antenna array is applied to the mobile.

Starting from the previous published subspace based algorithms for joint parameter estimation of asynchronous DS CDMA signals [3] -[12], a MUSIC based algorithm for joint direction of departure- DOD, frequency shift and time delay estimation in the downlink of the DS CDMA system with multiple antennas at the base station and single antenna at the mobile users is proposed. For the estimated direction of departure DOD of direct path (which usually have the shortest time delay) the direction of arrival can be directly calculated. The proposed algorithm is well suited to unknown fading channel and it can be used as a method for channel identification. The proposed algorithm is near-far resistant, requires no preamble, and can be applied when the user code sequences are not perfectly orthogonal. Algorithm can be used for both single user and multi-user estimation.

2 System Model

A block diagram of the model of the supposed asynchronous DS CDMA system is presented on Fig. 1. Signal of the kth user transmitted from the

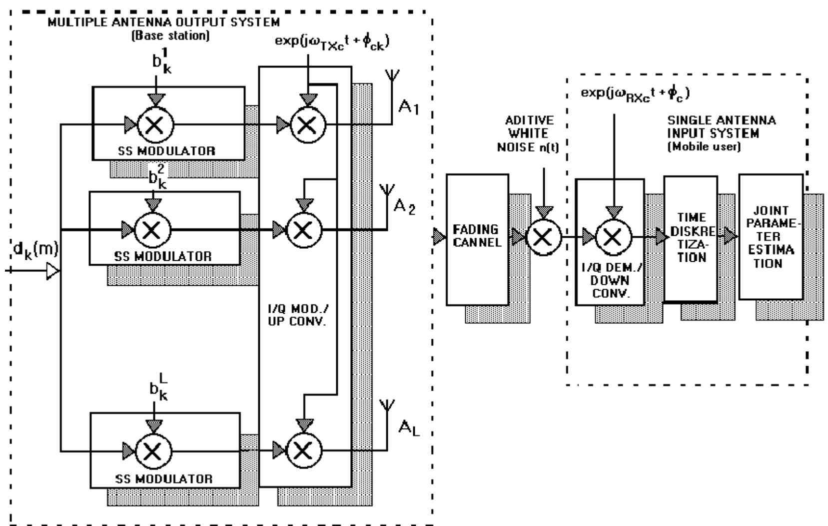


Fig. 1. A block diagram of the supposed system.

antenna array composed of antennas which are nonuniformly distributed and located at points $\{\boldsymbol{z}_l\} \in \mathcal{R}^{3X1}$, $l = 1, \ldots, L$ in real three space can be modeled as

$$\begin{aligned} y_k(t) =&\Re\{\sqrt{2P_{TXk}}e^{j\omega_{TXc}t} \\ &\times \sum_{i=1}^{L} s_k^l[t-\tau_k-\tau_{TX}^l(\theta_{TX},\varphi_{TX})]e^{j\phi_{TXck}}g_{TX}^l(\theta_{TX},\varphi_{TX}) \\ &\times e^{j\omega_{TXc}\tau_{TX}^l(\theta_{TX},\varphi_{TX})}\} \end{aligned} \quad (1)$$

where the subscripts k denotes kth user and superscript l denotes the lth antenna; P_{TXk} is summary transmitted power of the kth user which is equal to $P_{TXk} = \sum_{l=1}^{L} P_{TXk}^{l}$, where P_{TXk}^{l} is transmitted power of the kth user emitted from lth antenna; ω_{TXc} is the transmitter carrier frequency; ϕ_{TXck} is the carrier phase of the transmitter up converter which have to be the same on the all L antennas, but it can be different from the user to the user; $s_k^l[t - \tau_k - \tau_{TX}^l(\theta_{TX}, \varphi_{TX})]$ is the baseband signal of the kth user on the lth transmit antenna formed by the data stream $d_k(m) \in A_k$, A_k being the kth user alphabet, i.e. $s_{lk}(t) = \sum_{m=-\infty}^{\infty} d_k(m) b_k^l(t - mT)$; $b_k^l(t)$ is the complex valued user code sequence of the kth user on lth antenna, where $b_k^l(t)$ for $t \in [0, T)$; T and $T_c = T/N$ are bit and the chip duration respectively; τ_k is the time delay (time offset) of the kth user related to time reference on the transmitter, which is due to asynchronous nature of the users; $\tau_{TX}^l(\theta_{TX}, \varphi_{TX}) = \boldsymbol{v}^T(\theta_{TX}, \varphi_{TX})\boldsymbol{z}_l/c$ is time delay of the signal on the lth antenna relative to the reference point (origin of the spherical coordinate system of transmit antenna array), transmitted to direction $\{\theta_{TX}, \varphi_{TX}\}$, where θ_{TX} denotes azimuth and φ_{TX} denotes elevation which are defined in coordinate system of the transmit antenna array; $\boldsymbol{z}_l$ is vector of location of the lth antenna in real three space, c is propagation velocity; $\boldsymbol{v}(\theta_{TX}, \varphi_{TX})$ is the unit vector in direction $\{\theta_{TX}, \varphi_{TX}\}$; $g_{TX}^l(\theta_{TX}, \varphi_{TX})$ is the lth antenna gain in direction of$\{\theta_{TX}, \varphi_{TX}\}$. The value $a_{TX}^l(\theta_{TX}, \varphi_{TX}) = g_{TX}^l(\theta_{TX}, \varphi_{TX}) \exp[j\omega_{TXc}\tau_{TX}^l(\theta_{TX}, \varphi_{TX})]$ is the lth element of steering vector of the transmit antenna array.

2.1 Channel model

Suppose that the antenna aperture of the transmit antenna array is small $(L * \lambda_c/2)$ and that there are no any obstacles near the transmit antenna array. Then the same channel model can be supposed for all transmit-receive antenna pairs. We presume the time varying channel with discrete multipath, [13]. The channel impulse response can be modeled as

$$h(t, \tau) = \sum_{p=1}^{R_p(t)} \alpha_p(t) e^{j\phi_p(t)} e^{j\Delta\omega_{D_p} t} \delta[t - \tau_p(t)] \tag{2}$$

where $R_p(t)$ denotes the number of discrete paths, which is time varying. $\alpha_p(p)$ is random attenuation of the pth path due to propagation channel; $\phi_p(t)$ is the random carrier phase of the pth path due to channel propagation which is distributed uniformly in $[0, 2\pi)$; $\Delta\omega_{D_p}$ is Doppler frequency-shift of the pth path due to the channel characteristic; $\delta(t)$ is Dirac function; $\tau_p(t)$ is

pth path delay due to multipath structure of the channel. Multi-user signal at the point of the antenna of the receiver of the kth mobile user can be modeled as

$$r(t) = \sum_{k=1}^{K} y_k(t) * h(t, \tau) \tag{3}$$

where K denotes the number of users and symbol $*$ denotes convolution. It is clear that due to multipath structure of the propagation channel many paths of the transmitted signals are received. Multi-user signal at the input of the receiver antenna can be modeled as

$$\begin{aligned} r(t) =& \Re\{n(t) + \sum_{k=1}^{K} \sqrt{2P_{TXk}} e^{j\omega_{TXc}t} \\ & \times \sum_{p=1}^{R_p(t)} g_{RX}(\theta_{RXp}, \varphi_{RXp}) \alpha_p(t) e^{j\phi_p(t)} e^{j(\Delta\omega_{Dp} + \Delta\omega_{DRXp})t} \\ & \times \sum_{l=1}^{L} s_k^l[t - \tau_k - \tau_p(t) - \tau_{TX}^l(\theta_p, \varphi_p)] e^{j\phi_{TXck}} a_{TX}^l(\theta_p, \varphi_p)\} \end{aligned} \tag{4}$$

where $q_{RX}(\theta_{RXp}, \varphi_{RXp})$ is receiver antenna gain where $\{\theta_{RXp}, \varphi_{RXp}\}$ denotes azimuth and elevation of arrival to the receiver antenna of the pth path; $\Delta\omega_{DRXp}$ is Doppler frequency-shift of the pth path due to the receiver motion. $n_1(t)$ is the additive white Gaussian noise with two-sided power spectral density $N_0/2$.

2.2 System model restrictions

Let suppose, for simplicity, that omnidirectional antenna elements are used in transmit antenna array at base station and omnidirectional antenna in mobile receiver so $g_{TX}^l(\theta_{TX}, \varphi_{TX}) = 1$, $l = 1, \ldots, L$, and $g_{RX}(\theta_{RXp}, \varphi_{RXp}) = 1$. Let's also suppose that incident CDMA signals can be approximated as narrow-band in the array processing sense, so the approximation $s_{lk}[t - \tau_k(t) - \tau_p(t) - \tau_{TX}^l(\theta_p, \varphi_p)] \approx S_{lk}[t - \tau_k(t) - \tau_p(t)]$ for $l = 1, 2, \ldots, L$; $k = 1, 2, \ldots, K$ is suitably accurate for all values of time t. Suppose that time delay $\tau_p(t)$ of the pth path and the number of the paths $R_p(t)$ and azimuth and elevations $\{\theta_{RXp}, \varphi_{RXp}\}$; $p = 1, \ldots, R_p$ are time invariant (constant) during the M symbol intervals so $\tau_p(t) \approx \tau_p$ and $R_p(t) = R_p$. Starting from the previous discussions and assumptions, equation (4) can be simplified to

$$\begin{aligned} r(t) =&\Re\{n(t) + \sum_{k=1}^{K} e^{j\omega_{TXc}t} \sum_{p=1}^{R_p} \rho_{kp}(t) e^{j(\Delta\omega_{Dp}+\Delta\omega_{DRXp})t} \\ &\times \sum_{t=1}^{L} s_{lk}(t-\tau_{kp}) e^{j\phi_{TXck}} a^l(\theta_{TXkp}, \varphi_{TXkp})\} \end{aligned} \tag{5}$$

where $\tau_{kp} = \tau_k + \tau_p$ denotes the summary time delay of the pth path of the kth user related to time reference of the transmitter, and $\rho_{kp}(t) = \sqrt{2P_{TXk}}\alpha_p(t)\exp[j\phi_p(t)]$is a complex value. Assume that the fading processes are slowly time varying relative to the duration of each data symbol, so that $\rho_{kp}(t)$ can be taken as a constant for the duration of the mth symbol interval.

3 Formulation a Problem of Joint Parameter Estimation

At the receiver front-end received at the mobile side the continuous-time signal $r(t)$is down converted (by IQ mixing to baseband) and then time discretized by sampling the outputs of an integrators which integrates receiving signals over subinterval $T_i = T_c/Q$ where Q is the over sampling factor. The obtained equivalent complex sequence $r(q)$ can be expressed as

$$r(q) = n(q) + \sum_{k=1}^{K}\sum_{p=1}^{R_p} \int_{(q-1)T_i}^{qTi} \rho_{kp}(t) e^{j\Delta\omega_p t} \sum_{l=1}^{L} s_{lk}(t-\Delta\tau_{kp}) e^{j\phi_{ck}} a^l(\theta_{kp}, \varphi_{kp}) \tag{6}$$

were $\Delta\omega_p = \omega_{TXc} - \omega_{RXc} + \Delta\omega_{Dp} + \Delta\omega_{DRXp}$ is the summary frequency shift which is due to the mismatch of the transmitter/receiver oscillators, Doppler frequency shift due to channel and Doppler frequency shift due to receiver motion; $\Delta\tau kp$ is time delay of the pth path of the kth user related to the receiver time reference; $n(q)$ is the zero-mean white complex Gaussian sequence with the variance $\sigma^2 = E[\mid n(l) \mid^2] = N_0/T_i = P_l N_0 QN/E_{b,l}$ and $E_{b,l} = P_l T$ is the energy per bit for the first user. From the resulting discrete-time signal samples, a sequence of observation vectors $\boldsymbol{r}(m) = [r(mQN + QN), \ldots, r(mQN+1)]^T \in C^{QN\times l}$ which correspond to one symbol interval are formed [4]. Time delay $\Delta\tau_{kp} \in [0,T)$is defined in relation to the start of the observation vector. Similarly, the noise vector $\boldsymbol{n}(m) \in C^{QN}$ can be defined as $\boldsymbol{n}(m) = [n(mQN+QN), \ldots, n(mQN+1)]^T$. Since the system is asynchronous, the bit-timing is unknown to the receiver and the observation vector will contain the end of the previous symbol and the beginning of the current symbol for each user. Now, let us define the contribution $\boldsymbol{r}_{kp}(m)$ of the pth path of the kth user to $\boldsymbol{r}(m)$. It can be modeled as:

$$\begin{aligned}\boldsymbol{r}_{kp}(m) =& [\boldsymbol{u}_{kp}^{R}(\Delta\tau_{kp},\omega_{Dp},\theta_{TX_p},\varphi_{TX_p})\ \boldsymbol{u}_{kp}^{L}(\Delta\tau_{kp},\omega_{Dp},\theta_{TX_p},\varphi_{TX_p})] \\ &\times\begin{bmatrix}\gamma_{kp}(m-1)\\ \gamma_{kp}(m)\end{bmatrix}\end{aligned} \tag{7}$$

The vector pair: $\boldsymbol{u}_{kp}^{R}(\Delta\tau_{kp},\omega_{Dp},\theta_{TXp},\varphi_{TXp})\ \boldsymbol{u}_{kp}^{L}(\Delta\tau_{kp},\omega_{dp},\theta_{TXp},\varphi_{TXp})\}$ are signal vectors, which are defined similarly to [7] as

$$\begin{aligned}\boldsymbol{u}_{kp}^{R}(\Delta\tau_{kp},\omega_{Dp},\theta_{TXp},\varphi_{TXp}) =& \Big[\frac{\delta_k}{T_i}D(p_k+1,1,0) \\ &+(1-\frac{\delta_k}{T_i})D(p_k,1,0)\Big]\boldsymbol{g}_{kp}\end{aligned} \tag{8}$$

$$\begin{aligned}\boldsymbol{u}_{kp}^{R}(\Delta\tau_{kp},\omega_{Dp},\theta_{TXp},\varphi_{TXp}) =& \Big[\frac{\delta_k}{T_i}D(p_k+1,0,1) \\ &+(1-\frac{\delta_k}{T_i})D(p_k,0,1)\Big]\boldsymbol{g}_{kp}\end{aligned} \tag{9}$$

where $\tau_{kp} = p_{kp}T_i + \delta_{kp}$, p_{kp} is an integer, $\delta_{kp} \in [0, T_i)$, the vector $\boldsymbol{g}_{kp} = \boldsymbol{C}^{QN\times l}$ is defined as

$$\boldsymbol{g}_{kp} = [g_{kp}(QN), g_{kp}(QN-1), \ldots, g_{kp}(1)]^T \tag{10}$$

where

$$g_k(l) = \frac{1}{T_i}\int_{(l-1)T_i}^{lT_i} e^{j\Delta\omega_p t}\sum_{l=1}^{L} b_k^l(t)a_{TX}(\theta_{TXp},\varphi_{TXp})\mathrm{d}t \tag{11}$$

The permutation matrix $\boldsymbol{D}(r,\alpha,v) \in R^{QN\times QN}$ is a matrix defined in block form identically as in [13]. Let's notice that signal vectors are known function of unknown parameters. The scalar quantities $\{\gamma_{kp}(m)\gamma_{kp}(m-1)\}$ are complex constants which depend on the power, the phase and the transmitted symbols in mth and mth-1 bit interval of the kth user. The observation vector $\boldsymbol{r}(m)$can be expressed as:

$$\boldsymbol{r}(m) = \sum_{k=1}^{K}\sum_{p=1}^{Rp}\boldsymbol{r}_{kp}(m) + \boldsymbol{n}(m) = \boldsymbol{A}\boldsymbol{c}(m) + \boldsymbol{n}(m) \tag{12}$$

The columns of the matrix $\boldsymbol{A} \in C^{QN\times 2KR_p}$ are the KR_p pairs of the signal vectors of the general form $\{\boldsymbol{u}_{kp}^R(\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp})\ \boldsymbol{u}_{kp}^L(\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp})\}$, $k = 1,\dots,K$, $p = 1,\dots,R_p$. The elements of the vector $\boldsymbol{c}(m) \in C^{2KR_p}$ are the KR_p pairs of the complex quantities $\{\gamma_{kp}(m)\ \gamma_{kp}(m-1)\}$, $k = 1,\dots,K\ p = 1,\dots,R_p$.

3.1 Formulation of MUSIC based algorithm for joint DOD, time delay and frequency shift estimation

The joint parameter estimation problem can be formulated as follows: The M observation vectors $\boldsymbol{r}(m)$ are given, it is necessary to estimate deterministic unknown set of parameters $\{\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi TXp\}$ for $k = 1,\dots,K$, $p = 1,\dots,R_p$. It was supposed that those parameters remain approximately constant during the period of M bits and that $\Delta\tau_{kp} \in [0,T)$. Suppose that receiver of the kth user knows: set of own user sequences $\{b_k^l\}$, $l = 1,\dots,L$ and transmit antenna array manifold (number of antenna elements L, gain of antenna elements $\{g_{TX}^l(\theta_{TX}, \varphi_{TX})\}$, $l = 1,\dots,L$, array geometry contained in vectors $\boldsymbol{z}_l$, $l = 1\dots,L$). The first step in formulation of MUSIC base algorithm for joint parameter estimation is the estimation of covariance matrix of the observation vectors. Since it is generally unknown, it can be M estimated from the available observation vectors as

$$\boldsymbol{R} = \frac{1}{M}\sum_{m=1}^{M}\boldsymbol{r}(m)\boldsymbol{r}^*(m) = \boldsymbol{A}\boldsymbol{C}\boldsymbol{A}^* + \boldsymbol{N} \tag{13}$$

We assume that a) columns of the matrix A are linearly independent for all possible values $\{\Delta\tau, \Delta\omega, \theta_{TX}, \varphi_{TX}\}$, which can be provided by appropriate sequence and antenna array geometry design, and b) correlation matrix $\boldsymbol{C} = \boldsymbol{E}\{\boldsymbol{c}(m)\boldsymbol{c}(m)^*\}$ has to have full (maximal) rank which is, especially in fading channel, almost always fulfilled. Both conditions are (or can be) fulfilled in practice. The signal vectors (columns of matrix $\boldsymbol{A}$) are the known function of unknown parameters $\{\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp}\}$ that we would like to estimate and they are contained in signal subspace $\boldsymbol{E}_s \in C^{QN\times 2KR_p}$ and for true values $\{\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp}\}$ those signal vectors are orthogonal to the noise subspace $\boldsymbol{E}_n \in C^{QN\times(QN-2KR_p)}$ of covariance matrix $\boldsymbol{R}$. **MUSIC** type algorithm for the joint estimation of unknown parameters $\{\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp}\}$ can be formulated as

$$\{\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp}\} = \text{arg.max}\{P_{MUS}(\Delta\tau, \Delta\omega, \theta_{TX}, \varphi_{TX})\} \tag{14}$$

where

$$P_{MUS}(\Delta\tau, \Delta\omega_{Dp}, \theta_{TX}, \varphi_{TX}) = \frac{\| \boldsymbol{u}_k^R(\Delta\tau, \Delta\omega, \theta_{TX}, \varphi_{TX}) \|}{\| \boldsymbol{E}_n \boldsymbol{u}_k^R(\Delta\tau, \Delta\omega, \theta_{TX}, \varphi_{TX})} + \frac{\| \boldsymbol{u}_k^L(\Delta\tau, \Delta\omega, \theta_{TX}, \varphi_{TX}) \|}{\| \boldsymbol{E}_n \boldsymbol{u}_k^L(\Delta\tau, \Delta\omega, \theta_{TX}, \varphi_{TX}) \|} \tag{15}$$

The unknown set of parameters $\{\Delta\tau_{kp}, \omega_{Dp}, \theta_{TXp}, \varphi_{TXp}\}$ of the kth user are determined as arguments of maximums of $P_{MUS}(\Delta\tau, \Delta\omega, \theta_{TXp}, \varphi_{TX})$. The signal vectors for the $\tau \in [pT_i, (p+1)T_i]$ can be modelled identically as in [7].

4 Ambiguity Characterization of the Space-Time-Frequency Manifold

Continuum of all possible signal vectors $\{\boldsymbol{u}_{kp}^R(\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp})$ $\boldsymbol{u}_{kp}^L(\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TX})\}$ of the kth user lie in four dimensional space-time-frequency manifold. It is supposed that the manifold is known to the receiver of the kth user (theoretically evaluated or measured by calibration procedure). In fact, the manifold represents priory knowledge of the signal (system) model. Incident signal vectors $\{\boldsymbol{u}_{kp}^R(\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp})$ $\boldsymbol{u}_{kp}^L(\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp})\}$ theoretically are perfect orthogonal to the noise subspace of covariance matrix. The MUSIC algorithm generally exploit that fact. If there are many signal vectors in the array manifold which do not correspond to actual set of parameters $\{\Delta\tau_{kp}, \Delta\omega_{Dp}, \theta_{TXp}, \varphi_{TXp}\}$ but which are collinear (or near to collinear) to the incident signal vectors, then false pick(s) or side lobes in joint parameter estimation occur(s) due to ambiguity. This is ambiguity of the type I. In [13] the type I ambiguity function is defined as a normalized measure of collinearity the signal vectors in array manifold. This function can be generalized and defined for space-time-frequency manifold of the kth user in following way:

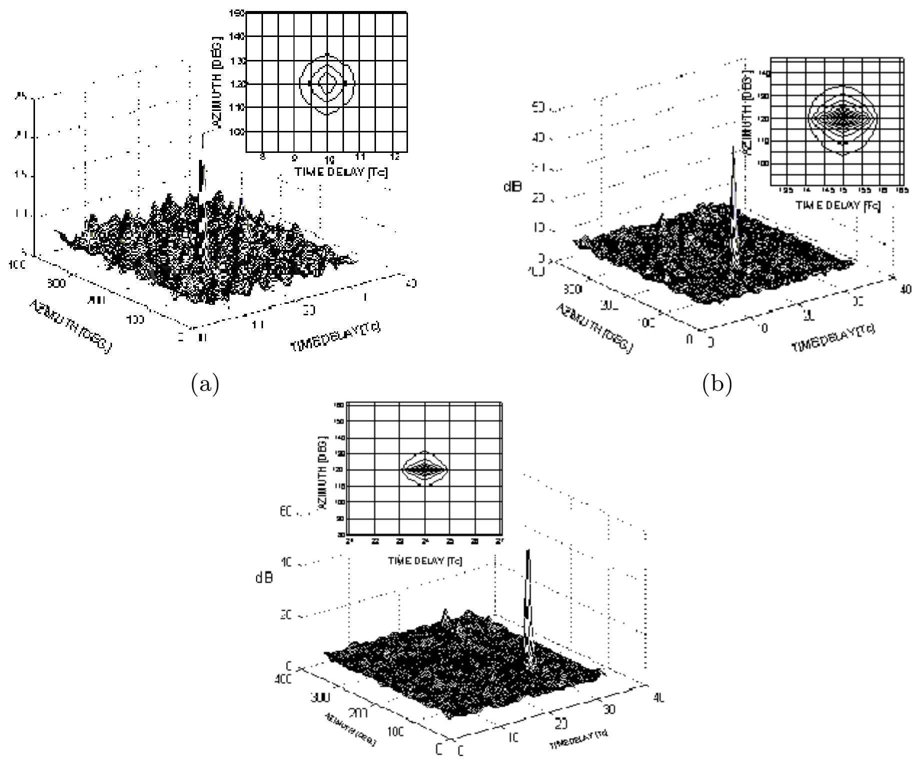
$$\chi_{kij}^I(\Delta\tau, \Delta\omega_{Di}, \theta_{TXi}, \varphi_{TXi}, \Delta\tau_j, \Delta\omega_{Dj}, \theta_{TXj}, \varphi_{TXj}) = \frac{\boldsymbol{u}_{ki}^R(*)^* \boldsymbol{u}_{kj}^R(*) + \boldsymbol{u}_{ki}^L(*)^* \boldsymbol{u}_{kj}^L(*)}{\| \boldsymbol{u}_{ki}^R(*) \| \| \boldsymbol{u}_{kj}^R(*) \| + \| \boldsymbol{u}_{ki}^L(*) \| \| \boldsymbol{u}_{kj}^L(*) \|} \tag{16}$$

where $\boldsymbol{u}_{ki}^R(*) = \boldsymbol{u}_{ki}^R(\Delta\tau_i, \Delta\omega_{Di}, \theta_{TXi}, \varphi_{TXi})$ and $\boldsymbol{u}_{kj}^R(*) = \boldsymbol{u}_{kj}^R(\Delta\tau_j, \Delta\omega_{Dj}, \theta_{TXj}, \varphi_{TXj})$ The symbols $*$ and $\| \cdot \|$ stands for conjugate transposition and norm of vector. The ambiguity properties of the space-time-frequency manifold determine in deterministic way the characteristics of any algorithm for joint parameter estimation. In other words, using the ambiguity function, false picks and side lobes in joint parameter estimation can be perfectly predicted.

5 Results of Simulation and Discussion

Two examples are presented. In both examples the circular transmit antenna array with $L = 10$ antenna elements was assumed. The fading channel is modeled with the commonly used wide-sense stationary uncorrelated scatterer Rayleigh fading channel model [15], so the are the $\rho_{kp}(t)$ zero-mean, complex Gaussian slow time varying fading processes (the value $\rho_k(t)$ is constant during the symbol interval) , uncorrelated from different paths of the same user. But for the same paths of different users at the same location with the same direction of arrival it differs just in complex constant. In the first example, a system with tree asynchronous DS CDMA users is simulated. The flat, single path Rayleigh fading channel is supposed. The modulation is BPSK. User code sequences are binary random sequence generated using MATLAB rand function with $N = T/T_c = 32$. Signal to noise ratio for the first user defined as E_{b1}/N_0 is 0 dB. Near far ratio defined as P_k/P_1 are 0 dB for $k = 1$, 10 dB for $k = 2$ and 30 dB for $k = 3$. The number of symbols is $M = 200$. The normalized time delays are: $\Delta\tau_1/T_c = 10$, $\Delta\tau_2/T_c = 15$, $\Delta\tau_3/T_c = 24$. Normalized frequency shifts $\omega_{Dk}/(2\pi/T_c)$ are 0, 0.0001 and –0.0001 respectively. The azimuths and elevations of signal departure from the base station are $\theta_{TX1} = \theta_{TX2} = \theta_{TX3} = 120°$, $\varphi_{TX1} = \varphi_{TX2} = \varphi_{TX3} = 0°$. It was supposed that the direct path is received so the azimuths and elevations of arrival are $\theta_{RX1} = \theta_{RX2} = \theta_{RX3} = 300°$, $\varphi_{RX1} = \varphi_{RX2} = \varphi_{RX3} = 0°$. The results of joint time delay and direction of departure (arrival) for both tree DS CDMA users are presented comparatively on the Figure 2(a), 2(b) and 2(c). From the Figure 2 it can be seen that the proposed algorithm provide correct parameter estimation of both three users practically without user interference and that the proposed algorithm is near-far robust (due to MUSIC). The frequency shifts are fixed to zero when MUSIC cost function is calculated. Simulations show the proposed algorithm is relatively robust to frequency shifts.

In the second example, a system with one DS CDMA user is simulated. The antenna array was the same as in the previous example. A four discrete path Reyleigh fading channel is supposed. User code sequence is binary random sequence with $N = T/T_c = 32$; Oversampling is used with $Q =$



(a) (b) (c)

Fig. 2. Joint time-delay and azimuth estimation for the (a) first, (b) second and (c) third user

2, The normalized time delays of the path are: $\Delta\tau_1/T_c = 15$, $\Delta\tau_2/T_c = 15.55$, $\Delta\tau_3/T_c = 15.5$, $\Delta\tau_4/T_c = 24.5$; The azimuths and elevations of paths departure are: $\theta_{TX1} = 200°$, $\theta_{TX2} = 250°$, $\theta_{TX3} = 180°$, $\theta_{TX4} = 220°$, $\varphi_{TX1} = \varphi_{TX2} = \varphi_{TX3} = \varphi_{TX4} = 0^0$. The modulation is 4-PSK. The signal to noise ratio, defined as E_{b1p/N_0} are: 10 dB for the first, 5 dB for the second, 0 dB for the third and fourth path. The ambiguity function $\chi^I_{1ij}(\Delta\tau_i, \Delta\tau_j)$ for the $\Delta\omega_{Di} = \Delta\omega_{Dj} = 0$, $\theta_{TXj} = \theta_{TXi} = 200°$ and $\varphi_{TXi} = \varphi_{TXj} = 0°$ is shown on the Figure 3(a). The form of that function points to the time correlation properties of user code sequence. As it can be seen, this function depends on $(\Delta\tau_i - \Delta\tau_j)$ but not on $\Delta\tau_i, \Delta\tau_j$. The ambiguity function $\chi^I_{kij}(\theta_{TXi}, \theta_{TXj})$ for $\Delta\tau_i = \Delta\tau_j = 15T_c$ $\Delta\omega_{Di} = \Delta\omega_{Dj} = 0$, and $\varphi_{TXi} = \varphi_{TXj} = 0°$ is presented on the Figure 3(b). As it can be seen, the ambiguity function depends on θ_i, θ_j but not on $(\theta_i - \theta_j)$. The main lobe of the function is primarily determined by antenna aperture. Generally its form is not the

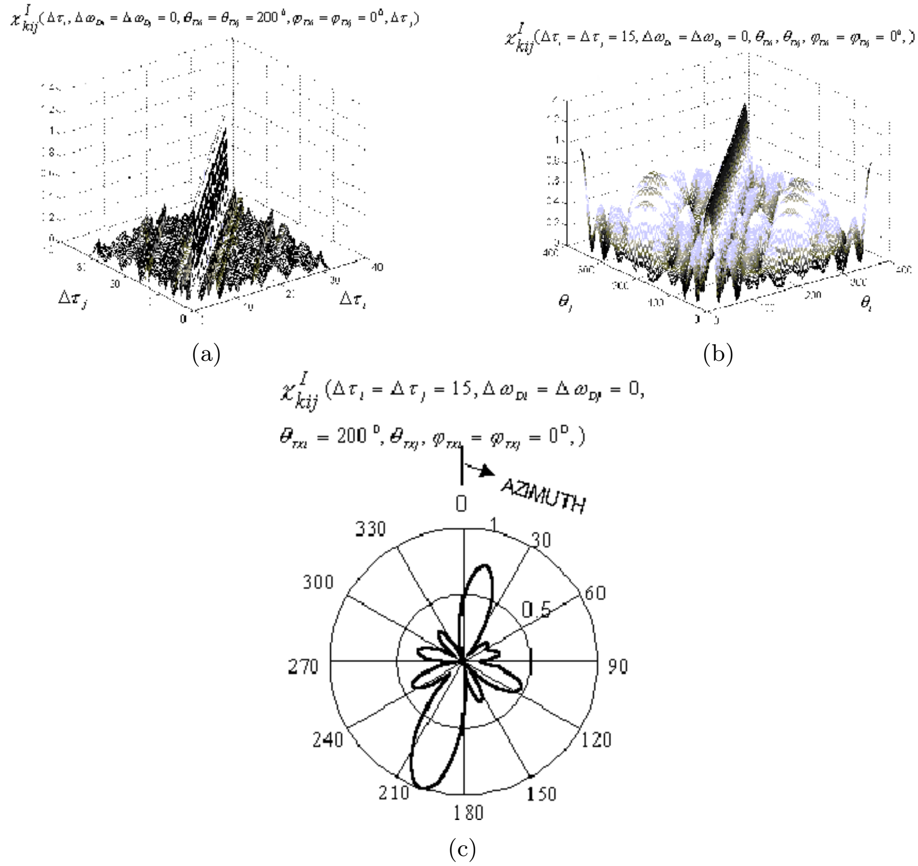
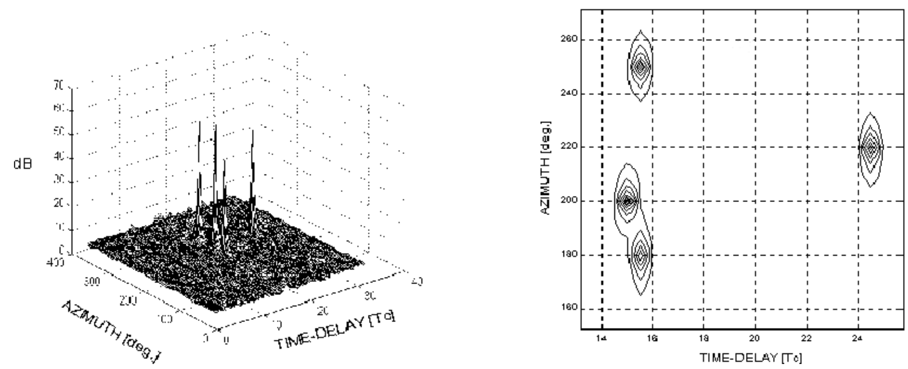


Fig. 3. Ambiguity function: (a) $\chi^I_{lij}(\Delta\tau_i, \Delta\tau_j)$ (b) $\chi^{(}_{lij}\theta_{TXi}, \theta_{TXj}$, c) Polar plot of $\chi^I_{lij}(\theta_{TXi} = 200^\circ, \theta_{TXj})$.

same along the azimuth (it depends on antenna geometry). For the same antenna array, the user code sequence influences side lobes. The polar plot of the ambiguity function $\chi^I_{kij}(\theta_{TXi} = 200^\circ, \theta_{TXj})$ is presented on Figure 3(c). The main lobe of this function directly determines the possibility of spatial resolvability of the different paths of the same user. From this picture it can be seen that the effects are the same, as if the antenna array was used at the mobile. The mesh plot of he result of the joint time delay and azimuth estimation is presented on the figure 4a , but the contour plot of the same is presented on Figure 4(b). As it can be seen, the proposed algorithm enables the correct parameter estimation of all three paths. So it is evident that it can be used for multipath channel identification.



(a) (b)

Fig. 4. Joint time-delay and azimuth estimation for the in fading multipath channel. (a) Mesh plot. (b) Contour plot.

6 Conclusion

The presented results of simulation shows that the proposed MUSIC type algorithm, applied in downlink processing, is well suited to joint parameter estimation of asynchronous DS CDMA signals in unknown slow varying multipath fading channel. It enables the joint estimation of time delay, frequency shift as well as direction of departure (DOD) of transmitted asynchronous DS CDMA signals in the system with multiple antennas at the base station and single antenna at the mobile. It can be primarily used as a method for multipath channel identification. It seems to the authors that it can be used as a basis for further improvement of process of detection at the mobile receiver. The results of simulation prove that the proposed algorithm is near-far resistant, requires no preamble, and can be applied when the user sequences are not perfectly orthogonal. The algorithm can be used for both single-user and multi-user estimation and there are no constraints on the code sequences (sequences need not be binary) and antenna array geometry (antenna array can be non-uniform volume).

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