시변 페이딩 채널에서 상관관계가 있는 안테나를 사용하는 DS-CDMA 통신 시스템을 위한 개선된 최대가능도 코드 타이밍 추정기

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Improved Code Timing Estimator for DS-CDMA Systems Using Correlated Antennas in Time-Varying Fading Channels

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이 논문은 2005년도 동아대학교 교내학술연구비(신진과제 전기)에 의해 연구되었음.

요 약

본 논문에서는 평탄 페이딩과 근거리/원거리 현상이 존재할 때 안테나 어레이를 사용하는 DS-CDMA 신호의 코드 타이밍을 추정하는 문제를 다루고자 한다. 페이딩 과정의 시변 페이딩 특성을 보다 잘 활용할 수 있도록 DS-CDMA 시스템에 대한 희망하는 사용자의 코드 타이밍을 추정하는 최대가능성 알고리즘을 유도한다. 코드 타이밍 추정기의 개발에 있어서 주어진 관찰 비트들을 포함하는 윈도우를 각각의 부-윈도우가 충분히 큰 길이를 갖도록 나눈다. 제안된 방법은 채널 페이딩의 일관성 시간과 연관된 똑같은 크기를 가지는 부-윈도우들을 이용한다. 또한 페이딩 속도를 추정하지 않고 충분히 긴 관찰 비트들을 두 개의 연속적인 부-윈도우로 나누는 방법을 대안으로 제시한다. 알고리즘의 유도과정은 공간적으로 상관성이 있는 다중안테나를 바탕으로 한다. 공간적인 페이딩 상관성이 제안된 알고리즘의 동기 획득 및 평균 동기 획득 성능에 미치는 영향을 분석한다.

ABSTRACT

We consider the problem of estimating a code-timing of DS-CDMA signal in antenna array systems in the presence of flat fading channels and near-far environments. We derive an approximate maximum likelihood algorithm of estimating the code-timing of a desired user for DS-CDMA systems to better utilize the time-varying characteristics of the fading process. In the development of code timing estimator, the given observation bits are divided into many sets of sub-windows with each sufficiently large. The proposed method uses sub-windows with equal size, associated with the coherence time of channel fading. The alternative approach is that without the estimation of the fading rate, the sufficiently given observation bits are simply separated into two consecutive sets of sub-windows. The derivation of the proposed algorithms is based on multiple antennas partially correlated in space. The impacts of spatial fading correlation on acquisition and mean acquisition time performance of the proposed algorithms are also examined.

키워드

Code division multiple access, code timing estimation, maximum-likelihood estimation, synchronization, code acquisition, spatial fading correlation

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Ⅰ. Introduction

Direct sequence code division multiple access (DS-CDMA) systems require code synchronization which is an essential operation. Code acquisition in DS-CDMA communication systems is to initially acquire the timing phase of the desired signal, usually within a fractional part of one chip duration. The correlator has been used as a conventional method to code acquisition. However, it might be highly sub-optimal in the presence of multiple access interference (MAI). It has been shown that the timing acquisition problem limits the capacity of a DS-CDMA system using the matched filter under the presence of MAI [1]. It is necessary to efficiently acquire the accurate timing information of each user even in multiuser environment. In addition, time-varying fading channels make the process of time delay estimation more difficult and complicated. Synchronization problem of DS-CDMA signals for a single antenna system has been extensively examined [1-5].

Antenna arrays that exploit the spatial dimension of wireless systems have attracted enormous interests for quality, coverage, or capacity improvement [6-8]. Furthermore, multiple-input multiple-output (MIMO) wireless communication has recently received tremendous interests [9]. Multiple antennas offering spatial dimension [10-14] can significantly improve the code acquisition performance in the presence of MAI and time-varying fading channels. The effects of spatial correlation on code acquisition have been examined in [11]. In [14], an approximate maximum likelihood (ML) approach of estimating the code-timing of a desired user for DS-CDMA systems has been developed as the multiple-antenna sensors-based estimator (MASE) algorithm exploiting receiver diversity. When the MASE algorithm is derived, the fading process has been assumed to be constant during a sufficiently given observation interval. Hence, the time-varying quantities due to the fading process may not be accurately represented in the process of timing estimation over time-varying channels, especially for large observation windows. In the presence of time-varying Rayleigh fading scenarios, the MASE estimator may not work well due to time-varying quantities even if the large observation window is used.

In this paper, we propose an efficient algorithm for an approximate ML approach of estimating the code-timing of a desired user for DS-CDMA systems in the presence of time-varying fading channels. In order for MASE algorithm to better exploit the time- varying characteristics of the fading process, the given observation bits are divided into many sets of sub-windows with each sufficiently large. The algorithm is referred to as the improved MASE (I-MASE). The I-MASE algorithm is based on the sub-window size, corresponding to the coherence time of channel fading. It requires the knowledge of the channel fading rate. The alternative approach called the modified MASE (M-MASE) is that without the estimation of the fading rate, the sufficiently given observation bits are simply separated into two consecutive sets of sub-windows. In real environments, the assumption that the received signals at different antennas experience either uncorrelated fading or perfectly correlated fading, which has been used in [12-15], cannot hold out. There will exist some degree of spatial correlation, depending on the relative spacing between antenna elements and/or surrounding scattering conditions. Hence, we have developed the I-MASE algorithm with multiple antennas partially correlated in space. The acquisition and mean acquisition time performance of the proposed algorithms is presented in correlated fading channels between antennas. In this paper, $(\cdot)^H$ and $(\cdot)^T$ denote the complex conjugate transpose and transpose, respectively, and Re(X) represents the real part of X.

Ⅱ. System Model

We consider an asynchronous K-user DS- CDMA system with BPSK modulation in flat fading environments. By pulse amplitude modulating the mth data symbol of the kth user, $d_k(m) \in \{+1, -1\}$, with a period of the spreading waveform $b_k(m)$, the baseband signal is expressed as $s_k(t) = \sum_{m=0}^{M-1} d_k(m) b_k(t-mT)$ where M is the number of bits considered for synchronization and $T=NT_c$ denotes the bit duration, with T_c and N being the chip interval and processing gain, respectively. The kth user's spreading waveform with period T is $b_k(t)=\sum_{n=1}^{N} c_k(n)\Pi_{T_c}(t-(n-1)T_c)$ where $c_k(n)\in\{+1,-1\}$ and $\Pi_{T_c}(t)$ denotes a unit rectangular pulse over the chip period $[0,T_c)$. The transmitted signal is formed by multiplying $s_k(t)$ with the carrier $\sqrt{2P_k}cos(w_c t)$, where P_k is the kth user's transmitted power.

The correlated channel coefficients between multiple antennas are generated with a simple stochastic MIMO model [16-18]. Here, the transmission connection between the mobile station and base station described by the narrowband single-input-multiple-output radio channel at the base station with P antennas can be expressed by Dirac delta function as $\boldsymbol{h}(t)=\boldsymbol{f}(t)\delta(t-\tau)\in\mathbb{C}^{P\times 1}$ with τ being a propagation delay and $\boldsymbol{f}(t)=[\beta^{(1)}(t)\ \beta^{(2)}(t)\ \ldots\ \beta^{(P)}(t)]^T$. Here, the correlated channel coefficients $\beta^{(p)}(t)$, $p=1,2,\ldots,P$, with

$$E\left[|\beta^{(p)}(t)|^2\right]=1 \tag{1}$$

are obtained according to $\boldsymbol{f}(t)=\boldsymbol{U}\boldsymbol{\alpha}(t)$ where $\boldsymbol{\alpha}(t)=[\alpha^{(1)}(t)\ \alpha^{(2)}(t)\ \ldots\ \alpha^{(P)}(t)]^T$ with $\alpha^{(p)}(t)$ being a zero-mean complex independent identically distributed random variable shaped by the desired Doppler spectrum and hence the amplitude and phase of the fading process $\alpha^{(p)}(t)$ having Rayleigh-distribution and uniform distribution, respectively. The lower triangular matrix $\boldsymbol{U}$ results from the Cholesky decomposition of the matrix $\boldsymbol{R}_u=\boldsymbol{U}\boldsymbol{U}^T\in\mathbb{R}^{P\times P}$ with

$$\boldsymbol{R}_u=\begin{bmatrix}\rho_{11} & \rho_{12} & \cdots & \rho_{1P}\\ \rho_{21} & \rho_{22} & \cdots & \rho_{2P}\\ \vdots & \vdots & \ddots & \vdots\\ \rho_{P1} & \rho_{P2} & \cdots & \rho_{PP}\end{bmatrix} \tag{2}$$

where the correlation coefficient $\rho_{p_1p_2}$ between two antennas p_1 and p_2 is defined as $\rho_{p_1p_2}=\langle |\beta^{(p_1)}|^2, |\beta^{(p_2)}|^2\rangle$ which depends on the distance between antenna elements, the power azimuth spectrum (PAS) type, azimuth spread (AS) and mean direction of arrival (DOA). The envelope correlation coefficient for uniform PAS can be obtained as

$$\rho_{p_1p_2}(|p_1-p_2|D) \triangleq |R_{II}(|p_1-p_2|D)+jR_{IQ}(|p_1-p_2|D)|^2 \tag{3}$$

where $D=d/\lambda$ with the element spacing d and wavelength λ, and the cross-correlation functions between the real parts and between a real part and an imaginary part, respectively, of the received signal at two antennas are given by [16,17]

$$R_{II}(|p_1-p_2|D)=J_0(2\pi|p_1-p_2|D)+4B\sum_{u=0}^{\infty}\frac{J_{2u}(2\pi|p_1-p_2|D)}{2u}cos(2u\phi)\sin(2u\Delta\phi) \tag{4}$$

$$R_{IQ}(|p_1-p_2|D)=4B\times\sum_{u=0}^{\infty}\frac{J_{(2u+1)}(2\pi|p_1-p_2|D)}{2u+1}sin[(2u+1)\phi]\times\sin[(2u+1)\Delta\phi] \tag{5}$$

where $J_u(.)$ is the Bessel function of the first kind with the uth order. Here, B is a normalization constant to make the uniform PAS a probability distribution function and the AS of uniform PAS defined over $[\phi-\Delta\phi,\ \phi+\Delta\phi]$ with the mean DOA ϕ and parameter $\Delta\phi$ is determined as $S_\phi=\Delta\phi/\sqrt{3}$ [19].

The received signal at the pth antenna can be written as

$$r^{(p)}(t)=\sum_{k=1}^{K}\beta_k^{(p)}(t)s_k(t-\tau_k)\sqrt{2P_k}cos(w_c t)+n^{(p)}(t) \tag{6}$$

where the noise waveform $n^{(p)}(t)$ is a spatially uncorrelated white Gaussian noise waveform at the pth antenna with a two-sided power spectral density of $N_0/2$. We assume that the propagation delay is considered to be fixed for $t\in[0,\ MT)$. The transmitter and receiver for the kth user's signal are assumed to have aligned their clocks to roughly within a bit interval. Hence, we consider only the relative propagation delay, that is, $\tau_k\in[0,\ T)$. Furthermore, the fading processes are assumed to be wide-sense stationary and to be constant during a symbol time. The receiver front-end at each antenna consists of a standard IQ-mixing stage followed by an integrate-and-dump section. Ignoring double frequency terms, the equivalent complex received sequence at the pth antenna can be expressed as

$$r_m^{(p)}(n) = \sum_{k=1}^{K} \beta_k^{(p)}(m)\sqrt{P_k} \times \frac{1}{T_c}\int_{mT+(n-1)T_c}^{mT+nT_c} s_k(t-\tau_k)dt + n_m^{(p)}(n) \quad (7)$$

where $n_m^{(p)}(n)$ is a zero mean white complex Gaussian sequence uncorrelated in time with variance

$$\sigma^2 = E\left[\,|n_m^{(p)}(n)|^2\right] = N_o/T_c = N_oN/E_{b,k} \quad (8)$$

with the transmitted energy per bit, $E_{b,k}$, for the kth user.

Ⅲ. Improved Code Timing Estimator

The MASE estimator [14] has been derived for receiver diversity DS-CDMA systems on the basis of multiple antennas with spatial decorrelation. It has been assumed that the fading process is constant over the total observation time interval $t\in[0, MT)$. Thus, it is derived as an approximate ML estimator for a time-invariant system. Nevertheless, simulations have been conducted for time- varying Rayleigh fading scenarios to investigate sensitivity to fading rate [14]. It has been shown that increasing observation window size up to the coherence time of the fading channel makes its performance to get better, but the use of observation window size exceeding the channel coherence time does not improve the performance. Similar results are shown in [8], where the MASE for a single antenna is the same as large sample maximum likelihood (LSML) algorithm [2]. Hence, we want to improve its acquisition performance with large observation window on time-varying fading channels.

In order to develop the large sample approximate ML timing estimates exploiting time-varying properties of fading channels, we assume that the fading process is constant over each sub-window interval $t\in\ [(q-1)L_sT, qL_sT)$, $q=1,2,\cdots,Q$,, where the total observation interval $t\in[0, MT)$ is equally divided into $Q(=M/L_s)$ sub-windows. We notice that the assumption and resultant cost function used to develop an ad-hoc approach in [15] are different from those used in this paper. Here the fading process is assumed to be constant over only each sub-window duration whereas in [15] it does not change during the total observation window.

Let the output vector of an antenna array at the qth sub-window's lth bit, $\boldsymbol{r}((q-1)L_s+l)$, be defined as

$$\boldsymbol{r}((q-1)L_s+l) = \begin{bmatrix} \boldsymbol{r}^{(1)}((q-1)L_s+l) \\ \boldsymbol{r}^{(2)}((q-1)L_s+l) \\ \vdots \\ \boldsymbol{r}^{(P)}((q-1)L_s+l) \end{bmatrix} \in \mathbb{C}^{NP\times 1} \quad (9)$$

Then,

$$\vec{\boldsymbol{r}}(l) = \left[\,\boldsymbol{r}^T(l)\ \ \boldsymbol{r}^T(L_s+l)\ \cdots\ \boldsymbol{r}^T((Q-1)L_s+l)\,\right]^T$$

is defined as the output vector of an antenna array at the lth bit for $l=1,2,\cdots,L_s$. By modeling the contributions of the MAI and additive noise as an unknown colored Gaussian random process, the MAI and noise vector $\vec{\boldsymbol{e}}(l)\in\mathbb{C}^{NPQ\times 1}$ is assumed to be a complex Gaussian random vector with zero-mean and arbitrary covariance matrix $\overline{\Omega}$ that satisfies $E[\,\vec{\boldsymbol{e}}(l)\vec{\boldsymbol{e}}^H(l)\,] = I_{PQ}\otimes\Omega = \overline{\Omega}$ with $\otimes$ denoting the Kronecker product and

$$\Omega = E\left[\boldsymbol{e}^{(p)}((q-1)L_s+l)\,\boldsymbol{e}^{(p)^H}((q-1)L_s+l)\right].$$

Assuming the first user's signal is the desired one, the antenna array's output vector $\vec{\boldsymbol{r}}(l)$ can be written as

$$\vec{\boldsymbol{r}}(l) = \boldsymbol{D}_1\vec{\boldsymbol{z}_1}(l) + \vec{\boldsymbol{e}}(l) \quad (10)$$

where

$$\boldsymbol{D}_1 = diag\left\{\boldsymbol{D}_1^{(1)}, \boldsymbol{D}_1^{(2)}, \cdots, \boldsymbol{D}_1^{(Q)}\right\} \in \mathbb{C}^{NPQ\times 2Q} \quad (11)$$

$$\boldsymbol{D}_1^{(q)} = \left[\,\boldsymbol{D}_1^{(1,q)^T}\ \boldsymbol{D}_1^{(2,q)^T}\ \cdots\ \boldsymbol{D}_1^{(P,q)^T}\,\right]^T,\ q=1,2,\cdots,Q \quad (12)$$

$$\vec{\boldsymbol{z}_1}(l) = \left[\,\boldsymbol{z}_1^T(l)\ \ \boldsymbol{z}_1^T(L_s+l)\ \cdots\ \boldsymbol{z}_1^T((Q-1)L_s+l)\,\right]^T \quad (13)$$

$$\boldsymbol{z}((q-1)L_s+l) = \left[\,\boldsymbol{z}_{1,1}((q-1)L_s+l)\ \ \boldsymbol{z}_{1,-1}((q-1)L_s+l)\,\right]^T \quad (14)$$

$$\boldsymbol{z}_{1,i}((q-1)L_s+l) = \frac{d_1((q-1)L_s+l) + \boldsymbol{i}\,zd_1((q-1)L_s+l-1)}{2} \quad (15)$$

for $i=1,-1$ and $\boldsymbol{D}_1^{(p,q)} = g_1^{(p,q)}\boldsymbol{A}_1(\tau_1)$. Here, $g_1^{(p,q)}$ $\left(=\beta_1^{(p,q)}\sqrt{P_1}\right)$ includes a channel coefficient of the qth sub-window at the pth antenna, which is assumed to be constant during the corresponding sub-window and

$A_1(\tau_1) = [\ a_{1,1}(\tau_1)\ \ a_{1,-1}(\tau_1)\ ] \in \mathbb{R}^{N\times 2}$. The form of $a_{1,i}(\tau_1)$ for $i=1,-1$ can be given with τ_1 and $c_1(n)$ by

$$a_{1,i}(\tau_1) = \left[\frac{\delta_1}{T_c} P(n_1+1, i) + \left(1-\frac{\delta_1}{T_c}\right) P(n_1, i) \right] c_1 \tag{16}$$

where $\tau_1 = n_1 T_c + \delta_1$ such that n_1 is an integer and $\delta_1 \in [0, T_c)$, and $c_1 = [\,c_1(N)\ c_1(N-1)\ \cdots\ c_1(1)\,]^T$. The permutation matrix $P(\alpha, \nu) \in R^{N\times N}$ is given in block form by

$$P(\alpha,\nu) = \begin{pmatrix} 0 & I_{N-\alpha} \\ \nu I_\alpha & 0 \end{pmatrix}, \quad \nu = \pm 1 \tag{17}$$

where I_α is the $\alpha \times \alpha$ identity matrix. The problem of interest is to estimate the code-timing τ_1 from the received data $\{\vec{r}(l)\}_{l=1}^{L_s}$ with an assumption that $\{c_1(n)\}_{n=1}^{N}$ and $\{\dot{d}_1(m)\}_{m=1}^{M}$ are known.

The joint probability distribution is then given by

$$f\left[\{\vec{r}(l)\}_{l=1}^{L_s} \middle| \{g_1^{(p,q)}\}_{p=1,q=1}^{P,\ Q}, \mu, \Omega\right] = \frac{1}{\pi^{PQL_sN}|\Omega|^{PQL_s}} exp\left\{ \sum_{p=1}^{P}\sum_{q=1}^{Q}\sum_{l=1}^{L_s} v_{(p,q,l)}^H \Omega^{-1} v_{(p,q,l)} \right\} \tag{18}$$

with

$$v_{(p,q,l)} = r^{(p)}((q-1)L_s+l) - D_1^{(p,q)} z_1((q-1)L_s+l)$$

and $\mu \triangleq \delta_1/T_c$. Here, $|\cdot|$ denotes the determinant of a matrix. We can then take a procedure of timing estimation algorithm in a manner similar to what was done for the MASE [14]. It can be shown that the maximization of log-likelihood function of the received output vector $r^{(p)}((q-1)L_s+l)$ amounts to maximizing the cost function which is given by (19) at the bottom of the page. Here, the consistent estimate $\hat{D}_1^{(p,q)}$ of $D_1^{(p,q)}$ and estimate $\hat{\Omega}$ of Ω, respectively, are given by

$$\hat{D}_1^{(p,q)} = \hat{R}_{zr_p}^{(p,q)^H} \hat{R}_{zz}^{(q)^{-1}} \triangleq [\,\hat{d}_1^{(p,q)}\ \hat{d}_{-1}^{(p,q)}\,] \tag{20}$$

$$\hat{\Omega} = \frac{1}{PQ}\sum_{p=1}^{P}\sum_{q=1}^{Q}\left(\hat{R}_{r_pr_p}^{(p,q)} - \hat{R}_{zr_p}^{(p,q)^H}\hat{R}_{zz}^{(q)^{-1}}\hat{R}_{zr_p}^{(p,q)}\right) \tag{21}$$

where

$$\hat{R}_{zr_p}^{(p,q)} = \frac{1}{L_s}\sum_{l=1}^{L_s} z_1((q-1)L_s+l)\, r^{(p)^H}((q-1)L_s+l) \tag{22}$$

$$\hat{R}_{zz}^{(q)} = \frac{1}{L_s}\sum_{l=1}^{L_s} z_1((q-1)L_s+l)\, z_1^H((q-1)L_s+l) \tag{23}$$

$$\hat{R}_{r_pr_p}^{(p,q)} = \frac{1}{L_s}\sum_{l=1}^{L_s} r^{(p)}((q-1)L_s+l)\, r^{(p)^H}((q-1)L_s+l) \tag{24}$$

Even if the exact ML optimization problem is simplified to the large sample approximate ML estimation approach, the direct optimization of expression (19) still requires a computationally expensive search over the parameter space. It can be easily shown that the maximization problem of the cost function C_1 in (19) amounts to find the zeros of a second-order polynomial for each chip interval. The improved MASE (called I-MASE) algorithm can be summarized by the following steps:

Step 1: Compute $\hat{D}_1^{(p,q)}$ for $p=1,2,\cdots,P$ and $q=1,2,\cdots,Q$, and $\hat{\Omega}$ using (22), (23) and (24).

Step 2: For $n=1,2,\cdots,N-1$, repeat the following two sub-steps to compute the corresponding cost $C_1(nT_c)$ and the cost $C_1([n+\hat{\mu}]T_c)$ which can be rewritten as $C_1(\mu) = \mu^H G\mu / \mu^H H\mu$ with $\mu = [1-\mu, \mu]^T$, G given by (25) at the bottom of the page and

$$H = Re\left\{ a_{1,1}^H \hat{\Omega}^{-1} a_{1,1} + a_{1,-1}^H \hat{\Omega}^{-1} a_{1,-1} \right\} \tag{26}$$

a) Form $a_{1,i}(nT_c)$, for $i=1,-1$, with (16) and compute G and H.

b) Find the roots $\hat{\mu}_1$ by solving the second-order polynomial and compute the corresponding cost $C_1([n_1+\hat{\mu}_1]T_c)$ with an estimate $\hat{\mu}_1 \in [0,1)$ and the cost $C_1(n_1T_c)$.

Step 3: Select $\hat{\tau}_1 = [\hat{n}_1+\hat{\mu}_1]T_c$ corresponding to the largest cost among the set of costs corresponding to the candidate timing estimates to be the estimate of the τ_1.

When deriving the I-MASE algorithm, it is assumed that the fading process is constant during each sub-window. The length of total observation bits in I-MASE can be selected such that they are divided into sub-windows with equal size, accounting for the fading coherence time. The length of sub-windows is determined by the channel coherence time that can be estimated by Doppler frequency or

$$C_1 = \sum_{p=1}^{P}\sum_{q=1}^{Q} \frac{\left(a_{1,1}^H(\hat{\mu})\hat{\Omega}^{-1}\hat{d}_1^{(p,q)} + a_{1,-1}^H(\hat{\mu})\hat{\Omega}^{-1}\hat{d}_{-1}^{(p,q)}\right)^H\left(a_{1,1}^H(\hat{\mu})\hat{\Omega}^{-1}\hat{d}_1^{(p,q)} + a_{1,-1}^H(\hat{\mu})\hat{\Omega}^{-1}\hat{d}_{-1}^{(p,q)}\right)}{a_{1,1}^H(\hat{\mu})\hat{\Omega}^{-1}a_{1,1}(\hat{\mu}) + a_{1,-1}^H(\hat{\mu})\hat{\Omega}^{-1}a_{1,-1}(\hat{\mu})} \tag{19}$$

$$G = Re\left\{\sum_{p=1}^{P}\sum_{q=1}^{Q}\left(a_{1,1}^H\hat{\Omega}^{-1}\hat{d}_1^{(p,q)} + a_{1,-1}^H\hat{\Omega}^{-1}\hat{d}_{-1}^{(p,q)}\right)^H\left(a_{1,1}^H\hat{\Omega}^{-1}\hat{d}_1^{(p,q)} + a_{1,-1}^H\hat{\Omega}^{-1}\hat{d}_{-1}^{(p,q)}\right)\right\} \tag{20}$$

velocity estimators [20,21]. With the knowledge of the fading rate, the sub-window size L_s can be obtained by using the popular rule of thumb definition of the channel coherence time [22], $L_s = \lfloor 0.423R_s/f_d \rfloor$ where f_d and R_s denote the fading and data rate, respectively, and $\lfloor x \rfloor$ is the integer part of x. Instead of having the total observation window size consisting of Q times L_s bits, we consider arbitrary observation window size M. If the given observation bits (M) are smaller than the obtained sub-window size L_s, it is used as one set of an observation window L_s. Otherwise, the number of sub-window sets for the total observation bits is given by $N_s = \lceil M/L_s \rceil$ where $\lceil x \rceil$ rounds x to the nearest integer towards infinity. Thus, if the observation bits (M) is larger than sub-window size L_s, the length of overlapped bits between two consecutive sub-windows is determined by $N_l = \lceil (L_sN_s - M)/(N_s - 1) \rceil$. It is expected that the I-MASE can provide approximately N_s-fold diversity gain, relative to MASE. Since the I-MASE needs to know the fading rate of the channel, the other simple approach referred to the modified MASE (M-MASE) is that without the estimation of the fading rate, the sufficiently given observation bits are partitioned into two consecutive sets of the received signals. With the observation bits more than coherence time, there may exist two times diversity effects. Note that without partitioning the observation bits, the M-MASE and I-MASE algorithms are the same as the MASE. The MAI and noise covariance matrix used in MASE, M-MASE and I-MASE algorithms is estimated as the average of those estimated from each antenna output, as shown in (21). Hence, the MASE, M-MASE and I-MASE require $MP \geq N$, $MP/2 \geq N$ and $L_sP \geq N$, respectively, so that $\hat{\Omega}$ is nonsingular with probability one. On the other hand, their computational complexities are approximately $O(N^3 + P(M+1)N^2)$, $O(N^3 + 2P(M/2+1)N^2)$ and $O(N^3 + PQ(L_s+1)N^2)$, respectively.

An interesting future subject is to expand the proposed I-MASE algorithm into multipath fading MIMO channels. Extension of the I-MASE restricted to one-path in this paper to the frequency selective multipath fading channel could involve deriving a timing estimator based on multipath channel model and solving a multidimensional search for multipath's delays, which would require highly expensive computations. Furthermore, it would be simplified to the successive optimization problem of the cost function for increasing numbers of multipath delays.

Ⅳ. Simulation Results

The performance of the timing estimators in each simulation was obtained from 2,000 Monte-Carlo runs. The simulated system uses Gold sequences with N=31 chips per bit operating over single path time-varying fading channels with arbitrary delays randomly generated for each run. The near-far ratio (NFR) is defined as P_k/P_1 for $k = 2, 3, \cdots, K$. The timing offset τ_k and data bits of all users are independent of each other. It is assumed that the PAS is uniformly distributed and the mean DOA is uniformly distributed over $[-\pi/3, \pi/3]$. White Gaussian noise passes through a third-order low pass filter in order to generate the fading processes that are independent standard Rayleigh fading processes [23]. The Doppler rate of the fading channel is denoted by f_d, which is defined as $f_d = \nu f_c / c$ where ν, f_c and c, respectively, are mobile speed, carrier frequency and speed of light. In these simulations, the data rate is assumed to be 9,600 bps. The fading process $\alpha^{(p)}(m)$ is independently generated for each antenna. When estimate $\hat{\tau}_1$ produced by the timing estimators over single path Rayleigh fading channels is less than a half chip duration away from the true delay of the corresponding path, there is a correct acquisition. Mean acquisition time (MAT) performance is also considered to measure the average time needed for acquisition process, which can be given by $T_{ACQ} = M / P_d$ [3] where P_d is the acquisition probability and the time to request a new preamble in case of misacquisition is set to zero. Using the knowledge of the channel fading rate as in I-MASE algorithm, the improved equal-gain combining correlator (IEC-Correlator) based on multiple antennas, which is computationally simple, is employed for comparison, as the following:

$$\hat{\tau}_1 = \arg\max_{\tau_1} \left\{ \frac{|\hat{y}\, a_{1,1}^H(\tau_1)|^2}{\| a_{1,1}(\tau_1) \|^2} \right\} \quad (27)$$

$$\hat{y} = \frac{1}{PQL_s} \sum_{p=1}^{P} \sum_{q=1}^{Q} \sum_{l=1}^{L_s} r^{(p)}((q-1)L_s + l) \quad (28)$$

If there is no overlap in sub-windows, an equal -gain combining correlator (EC-Correlator) is obtained. Note that the fading rate of the channel is assumed to be known for IEC-Correlator and I-MASE. EC-Correlator and IEC-Correlator require that the training sequence consists of all ones.

We first investigate the impacts of different observation and training window sizes on the acquisition and MAT performance. Fig. 1 and 2 plot the acquisition probability and MAT, respectively, of the desired signal as the length of observation bits is varying for P=2 antennas, $E_{b,1}/N_0 = 0$ dB, K=20 users, NFR=10 dB, a normalized Doppler rate $f_d T = 0.0104$ (i.e., carrier frequency = 900 MHz, symbol rate = 9.6 kHz, and mobile speed = 100 km/h), $S_\phi = 2^\circ$ and antenna spacing $D(= d/\lambda) = 4$. The use of increasing observation window size in an excess of around the fading coherence time does not improve the acquisition performance

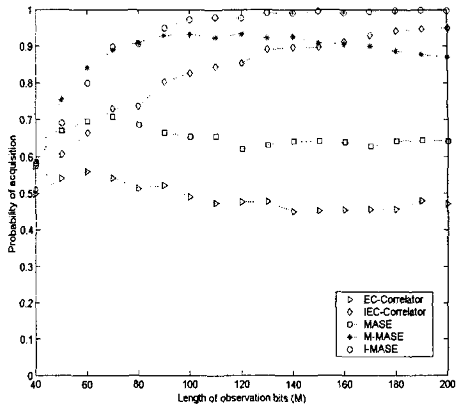

Fig. 1. Probability of correct acquisition for P=2 antennas, $E_{b,1}/N_0 = 0$ dB, K=20 users, NFR=10 dB, $f_d T = 0.0104$, $S_\phi = 2^\circ$ and antenna spacing D=4
그림 1. P=2 안테나, $E_{b,1}/N_0 = 0$ dB, K=20 사용자, NFR=10 dB, $f_d T = 0.0104$, $S_\phi = 2^\circ$, 그리고 안테나 간격 D=4일 때 올바른 동기 획득 확률

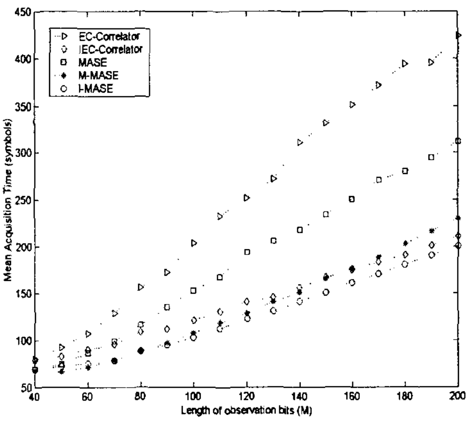

Fig. 2. Mean acquisition time for P=2 antennas, $E_{b,1}/N_0 = 0$ dB, K=20 users, NFR=10 dB, $f_d T = 0.0104$, $S_\phi = 2^\circ$ and antenna spacing D=4
그림 2. P=2 안테나, $E_{b,1}/N_0 = 0$ dB, K=20 사용자, NFR=10 dB, $f_d T = 0.0104$; $S_\phi = 2^\circ$, 그리고 안테나 간격 D=4일 때 평균 동기 획득 시간

of the MASE as well as EC-Correlator. Because in fact the fading process is time varying, the estimate $\hat{D}_1^{(p,q)}$ in MASE is not the exact value of the fading process, but the average one of it and hence could not be better ones even for much larger observation bits. That's why the acquisition performance of the MASE is not better even when the observation window length is larger than the coherence time. In contrast, the IEC-Correlator and I-MASE get better acquisition as the number of observation bits increases. It is shown that since the M-MASE takes into account the time diversity due to time-varying fading effect which can be used for better estimates of the and , it gives much better MAT as well as acquisition performance than MASE for the same length of observation bits. It is pointed out that the MASE has the largest acquisition performance with observation window of around 60 bits which is approximately corresponding to the channel coherence time for $f_d T = 0.0104$, whereas the M-MASE has the maximum performance for about 120 observation bits approximated twice the fading coherence time.

In the next experiment, we examine the effects of antenna spacing on the acquisition and MAT performance. In Fig. 3 and Fig. 4, it is seen that the timing estimators for antennas with larger separation perform better than the corresponding algorithms for closer antennas. It is observed that as the antenna spacing *D* increases, the acquisition performance slightly increases and the MAT gets better. It is due to the

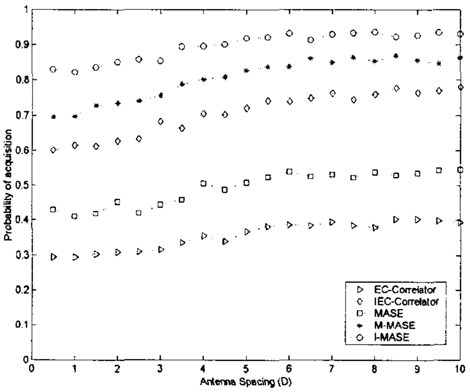

Fig. 3. Probability of correct acquisition for P=2 antennas, $E_{b,1}/N_0 = 0$ dB, K=30 users, NFR=10 dB, M=120 observation bits, $f_d T = 0.0104$ and $S_\phi = 2^\circ$
그림 3. P=2 안테나, $E_{b,1}/N_0 = 0$ dB, K=30 사용자, NFR=10 dB, M=120 관찰 비트, $f_d T = 0.0104$, 그리고 $S_\phi = 2^\circ$ 일 때 올바른 동기 획득 확률

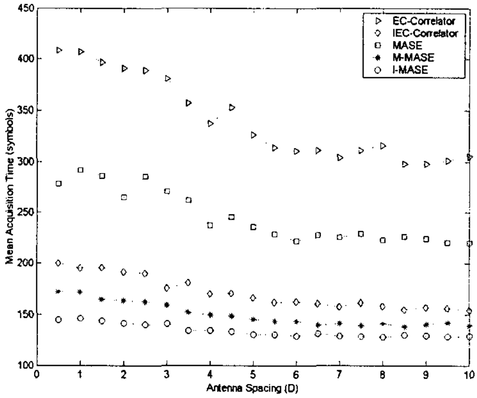

Fig. 4. Mean acquisition time for P=2 antennas, $E_{b,1}/N_0 = 0$ dB, K=30 users, NFR = 10 dB, M=120 observation bits, $f_d T = 0.0104$ and $S_\phi = 2^\circ$
그림 4. P=2 안테나, $E_{b,1}/N_0 = 0$ dB, K=30 사용자, NFR=10 dB, M=120 관찰 비트, $f_d T = 0.0104$, 그리고 $S_\phi = 2^\circ$ 일 때 평균 동기 획득 시간

gain of spatial diversity since the wide antenna spacing makes the fading correlation among antenna elements small. It is shown that the presence of spatial decorrelation is beneficial to acquisition and MAT performance of all code-timing estimators.

Now, we evaluate the sensitivity of timing estimators to the time-variation of the channel. Fig. 5 shows the effects of channel fading rate f_d on the acquisition performance of all algorithms. As the fading rate increases, the acquisition performance of all algorithms is getting worse. However, even though the fading rate is as large as 200 Hz, the M-MASE is still much better than MASE. It is shown that the EC-Correalor, MASE and M-MASE are highly sensitive to the fading rate whereas the IEC-Correalor and I-MASE are more robust. Note that the MASE has a similar performance to M-MASE and I-MASE for very slow time-varying fading. This result could be also obtained when the data rate is high and the vehicle velocity is low, and hence the normalized Doppler rate is small. Therefore, the improvement through I-MASE in such a scenario is slight. However, if the vehicle velocity goes higher even for high data rate, I-MASE and M-MASE could be applied to improve the acquisition

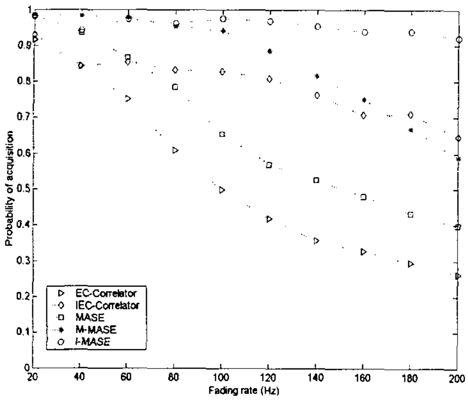

Fig. 5. Probability of correct acquisition for P=2 antennas, $E_{b,1}/N_0 = 0$ dB, K=20 users, NFR=10 dB, M=100 observation bits, $S_\phi = 2°$ and D=4

그림 5. P=2 안테나, $E_{b,1}/N_0 = 0$ dB, K=20 사용자, NFR=10 dB, M=120 관찰 비트, $S_\phi = 2°$, 그리고 D=4일 때 올바른 동기 획득 확률

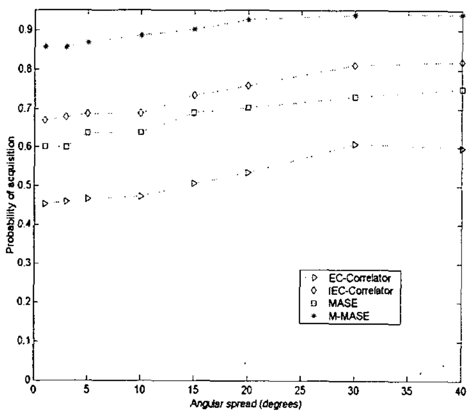

Fig. 6. Probability of correct acquisition for P=2 antennas, $E_{b,1}/N_0 = 0$ dB, K=20 users, NFR=10 dB, M=80 observation bits, $f_d T = 0.0104$ and D=0.5

그림 6. P=2 안테나, $E_{b,1}/N_0 = 0$ dB, K=20 사용자, NFR=10 dB, M=120 관찰 비트, $S_\phi = 2°$, 그리고 D=4일 때 평균 동기 획득 시간

performance. If the maximum Doppler frequency is known, the I-MASE algorithm utilizing sub-windows which can be obtained by the estimated coherence time may be more effective when using observation bits with much more than the coherence time. Doppler frequency estimator could be used to get the estimate of fading rate for I-MASE algorithm and the investigation of robustness to the estimation errors could be extended, but we do not consider it in this work. Fig. 6 shows the acquisition performance as a function of AS when P=2 antennas, $E_{b,1}/N_0 = 0$ dB, K=20 users, NFR=10 dB, M=80 bits, $f_d T = 0.0104$ and D=0.5. It is found that as the angular spread increases, or as the spatial correlation decreases, the acquisition performance is slightly improved.

V. Conclusions

An efficient algorithm for an approximate ML approach of estimating the code-timing of a desired user for DS-CDMA systems is developed to exploit both space and time domains in flat fading channels. In order to better exploit the time-varying property of the fading process, the I-MASE and M-MASE algorithms are presented. The code-timing estimate of desired signal is obtained by finding the zeros of second-order polynomials, which is computationally efficient. We have considered multiple antennas partially correlated in space, which could experience the case between two extreme cases of spatial fading correlation, the case of uncorrelated fading and the case of fully correlated one. The acquisition and MAT performance of the proposed algorithms is presented in the presence of correlated Rayleigh fading channels. The M-MASE algorithm much more improves the acquisition and MAT performance of the MASE in the time-varying fading channels. It is observed that the M-MASE algorithm significantly improves the acquisition performance over the correlator based on multiple antennas in the multiuser environments with near-far situation on the time-varying Rayleigh fading channels

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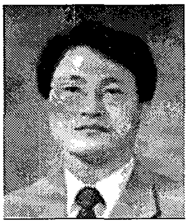

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