Code Timing Acquisition Using an Antenna Array for Asynchronous DS-CDMA Systems in a Near-Far Environment

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Abstract

In this paper, an adaptive near-far resistant detector using a multiple antennas for acquiring synchronization of an asynchronous DS-CDMA system is presented. The only requirement is the prior knowledge of the user-spreading sequence of interest. Also, there is no need of a training sequence and a training period. Synchronization acquisition of asynchronous DS-CDMA signals is treated as a binary-hypothesis test. Under the binary hypotheses, an adaptive generalized maximum likelihood ratio test (GLRT) is developed to acquire the multipath code timings of a single desired user in both a fading and a near-far interference environment. The synchronization is obtained by sliding a search window in a sample-by-sample manner to perform the adaptive GLRT with the non-constrained maximum likelihood estimator (MLE) of Direction-Of-Arrival (DOA) during the symbol interval. The acquisition performance of the proposed detector is illustrated by computer simulations of an asynchronous BPSK DS-CDMA system and is shown to be potential to against multipath fading and the problems of near-far interference.

I. Introduction

In recent years, direct-sequence CDMA has become one of the most popular multiple-access techniques for cellular mobile radio. In DS-CDMA communications, users are multiplexed by distinct code waveforms, rather than by orthogonal frequency bands in frequency-division multiple access (FDMA), or by orthogonal time slots in time-division multiple access (TDMA). All system users are allowed to transmit information simultaneously and occupy the entire broad-band frequency spectrum. However, DS-CDMA systems often suffer from multiple-access interference (MAI) and near-far interference problems. In addition, there is an important issue of code synchronization of asynchronous DS-CDMA signals using an antenna array has been receiving much attention.

It is well known that the standard method of code acquisition, i.e., code cross-correlator and its modifications [3-4], performs well under situations of a limited number of users and strict transmit power control, but degrades drastically in the presence of strong MAI or the near-far interference. In [5-7], the subspace-based approaches with only a single antenna are employed to estimate the different users' propagation delays over fading channels. Recently, there has been considerable effort to take advantage of the spatial diversity of a "smart" antenna (a receiver antenna array) to reduce the effects of channel fading due to multipath propagation. Hence, better acquisition performance of asynchronous DS-CDMA signals can be expected in comparison to the single antenna case. In [8-9], multiple antennas are based on the large sample maximum likelihood (LSML) and the subspace-based multiple signal classification (MUSIC) timing estimators, to perform the code timing acquisition over a time-varying fading channel.

In this paper, acquiring synchronization by the use of an antenna array is formulated as a binary-hypothesis test. An adaptive GLRT is developed to acquire the multipath code timings of a desired user in fading and near-far environment. The non-constrained MLE of DOA is obtained without assumption on the array geometry. Due to its simplicity, the adaptive GLRT with the non-constrained MLE of DOA can be utilized to achieve code synchronization. By sliding a search window on a sample-by-sample basis and performing the GLRT on the data in each search window, the multipath arrival times of the wanted user-spreading code pattern can be found.

II. Asynchronous DS-CDMA Signal Format

Consider an asynchronous DS-CDMA mobile radio network with M users that employ the spreading waveforms $s_1(t), s_2(t), \ldots, s_M(t)$ and transmitting sequences of binary phase-shift keying (BPSK) symbols. The transmitted baseband signal of the l^{th} user can be written in the form,

$$r_l(t) = \sum_{i=-\infty}^{\infty} A_l d_l[i] s_l(t - iT_b), \;\; l = 1, \ldots, M.$$

Here, $s_l(t)$ is the spreading waveform of the l^{th} user, given by

$$s_l(t) = \sum_{m=0}^{N-1} c_{l,m} p_{T_c}(t - mT_c), \;\; 0 \leq \; t \; \leq T_b$$

where

$$p_{T_c}(t) = \begin{cases} \frac{1}{\sqrt{T_c}} & , \; t \in [0, T_c], \\ 0 & , \; \text{otherwise}. \end{cases}$$

The parameters and other quantities are defined as follows:

A_l	Amplitude of the l^{th} user.
$d_l[i] \in \{\pm 1\}$	The i^{th} symbol of the l^{th} user.
T_b	Information symbol interval.
$\{c_{l,m}\}_{m=0}^{N-1}$	The spreading code of ± 1 of the l^{th} user.
T_c	The chip interval.
N	The spreading gain ($= T_b/T_c$).

On the Assumption that each transmitter is equipped with a single antenna, then the baseband multipath channel between each user's transmitter and the base station receiver can be modeled as a single-input multiple-output channel with the following vector impulse response [10]:

$$\mathbf{h}_l(t) = \sum_{j=1}^{K} \mathbf{b}_{l,j} a_{l,j} \delta(t - \tau_{l,j}),$$

where $\delta(\cdot)$ denotes the continuous-time unit-impulse (or Dirac delta) function and K is the number of paths in each user's channel. $a_{l,j}$, $\tau_{l,j}$, and $\mathbf{b}_{l,j} = [b^1_{l,j}, \ldots, b^J_{l,j}]^T$ are, respectively, the complex gain, the propagation delay, and the array response vector of the j^{th} path of the l^{th} user's signal. "$_T$" denotes the transpose operator.

Assuming that there are M active system users, each with K propagation paths, impinge on the receiving antenna array with J sensors in an additive white Gaussian noise (AWGN) channel. The total received signal at the receiver can be written in the form,

$$\begin{aligned}\mathbf{x}(t) &= \sum_{l=1}^{M} \mathbf{h}_l(t) \star r_l(t) + \mathbf{n}(t), \\ &= \sum_{l=1}^{M} A_l \sum_{j=1}^{K} a_{l,j}\mathbf{b}_{l,j} \sum_{i=-\infty}^{\infty} d_l[i] s_l(t - iT_b - \tau_{l,j}) + \mathbf{n}(t).\end{aligned}$$

where "$\star$" indicates the convolution operator and $\mathbf{n}(t) = [n_1(t), \ldots, n_J(t)]^T$ represents a column J-vector of complex white Gaussian noise processes.

III. Formulation of Problem

Here the detection of a single desired user's spreading code sequence that is embedded in MAI is treated as a binary-hypothesis test. In order to distinguish the binary hypotheses, a GLRT, that is described by a probability density function (pdf), $\mathbf{P}$, on the sample space, is derived next.

Consider a receiving array with J elements, the received data $\mathbf{x}(t)$ is first downconverted to baseband and then passed through a chip matched filter (MF). The output of the chip MF is sampled every T_s seconds, where $S_a (= T_c/T_s)$ is an integer and ≥ 1. The discrete-time output is then used as the input of the adaptive N-element tapped delay line (TDL) to form N-element data vector. Assuming that the output signal of the chip MF is sampled at time $t'T_s$ during the i^{th} symbol interval, $iT_b \leq t \leq (i+1)T_b$, so that the output of the adaptive TDL for the p^{th} antenna element can be expressed as the column N-vector given by

$$\begin{aligned}\mathbf{x}^{(i)}_p(t') = [x_p(t'T_s + iT_b), x_p(t'T_s - T_c + iT_b), \\ \ldots, x_p(t'T_s - (N-1)T_c + iT_b)]^T, \\ 0 \leq t' \leq NS_a - 1, 1 \leq p \leq J.\end{aligned}$$

For simplicity of notation, the superscript "$_{(i)}$" is suppressed for the remainder of the paper. Then the output of the antenna array can be represented by a $J \times N$ matrix given by

$$\mathbf{X}_{t'} = [\mathbf{x}_1(t'), \mathbf{x}_2(t'), \ldots, \mathbf{x}_J(t')]^T.$$

To search for the unknown locations of the desired spreading code pattern, the detector is designed by sliding a "search window" with the size N, sample-by-sample through all of the received data . A binary hypotheses are defined as follows: the null hypothesis H_0 is composed of an interference-plus-noise only process, and the signal-interference-plus-noise hypothesis H_l are the processes given by

$$\begin{aligned}H_0 &: \mathbf{X}_{t'} = \mathbf{V}_{t'}, \\ H_l &: \mathbf{X}_{t'} = \mathbf{g}_{l,j}\mathbf{s}_l + \mathbf{V}_{t'},\end{aligned}$$

where $\mathbf{g}_{l,j} = A_l a_{l,j} d_l \mathbf{b}_{l,j} = \alpha_{l,j}\mathbf{b}_{l,j}$ and $\alpha_{l,j} = A_l a_{l,j} d_l$, $1 \leq j \leq K$. Let $\mathbf{s}_l = [c_{l,0}, \ldots, c_{l,N-1}]$ denote the l^{th} user-spreading code sequence of length N. $\mathbf{V}_{t'}(k) = [V_{t',1}(k), \ldots, V_{t',J}(k)]^T$ represents the noise environment which is composed of the directional interference sources and the spatially white noise. The noise process can be assumed to be Gaussian zero-mean and independent from sample-to-sample [11]. Hence, the mean value and the covariance matrix of $\mathbf{X}_{t'}$ under hypotheses H_0 and H_l are given by

$$\begin{aligned}E\{\mathbf{X}_{t'}|H_0\} = \mathbf{0}, \ E\{\mathbf{X}_{t'}|H_l\} = \mathbf{g}_{l,j}\mathbf{s}_l, \ and \\ \mathbf{R}_{t'} \triangleq E[(\mathbf{X}_{t'} - E[\mathbf{X}_{t'}])(\mathbf{X}_{t'} - E[\mathbf{X}_{t'}])^\dagger].\end{aligned}$$

$E[\cdot]$ denotes the expectation operator and the notation "$_\dagger$" is the conjugate transpose operation.

The generalized likelihood ratio function, $\Lambda_l(\mathbf{X}_{t'})$, of the l^{th} spreading code sequence, depending on the maximum likelihood estimate $\hat{\mathbf{\Theta}}_l$ of the parameter vector $\mathbf{\Theta}_l$ which is expressed as the pair $[\mathbf{g}_{l,j}, \mathbf{R}_{t',l}]$ under the binary hypotheses, takes the following form [1, Appendix A], [2, Page 1762]:

$$\begin{aligned}\Lambda_l(\mathbf{X}_{t'}) &= \frac{max_{\mathbf{R}_{t',l}, \mathbf{g}_{l,j}} \mathbf{P}(\mathbf{X}_{t'}|H_l, \ \mathbf{R}_{t',l}, \ \mathbf{g}_{l,j})}{max_{\mathbf{R}_{t',0}} \mathbf{P}(\mathbf{X}_{t'}|H_0, \ \mathbf{R}_{t',0})} \\ &= \frac{\mathbf{P}(\mathbf{X}_{t'}; \hat{\mathbf{\Theta}}_l(\mathbf{X}_{t'}))}{\mathbf{P}(\mathbf{X}_{t'}; \hat{\mathbf{\Theta}}_0(\mathbf{X}_{t'}))} \\ &= \frac{|\hat{\mathbf{R}}_{t',0}|^N}{\min_{\mathbf{g}_{l,j}} |\hat{\mathbf{R}}_{t',l}|^N}, \qquad (1)\end{aligned}$$

where

$$\begin{aligned}\mathbf{P}(\mathbf{X}_{t'}|H_l, \mathbf{R}_{t',l}, \mathbf{g}_{l,j}) &= \frac{1}{\pi^{NJ}|\mathbf{R}_{t'}|^N} e^{-\frac{N}{2}Tr(\mathbf{R}^{-1}_{t'}\hat{\mathbf{R}}_{t',l})}, \\ \mathbf{P}(\mathbf{X}_{t'}|H_0, \mathbf{R}_{t',0}) &= \frac{1}{\pi^{NJ}|\mathbf{R}_{t'}|^N} e^{-\frac{N}{2}Tr(\mathbf{R}^{-1}_{t'}\hat{\mathbf{R}}_{t',0})}, \\ \hat{\mathbf{R}}_{t',0} &= \frac{1}{N}\mathbf{X}_{t'}\mathbf{X}^\dagger_{t'}, \\ \hat{\mathbf{R}}_{t',l} &= \frac{1}{N}(\mathbf{X}_{t'} - \mathbf{g}_{l,j}\mathbf{s}_l)(\mathbf{X}_{t'} - \mathbf{g}_{l,j}\mathbf{s}_l)^\dagger,\end{aligned}$$

where $|\mathbf{R}_{t'}|$ denotes the determinant of $\mathbf{R}_{t'}$. $\mathbf{P}(\mathbf{X}_{t'}|H_0, \mathbf{R}_{t',0})$ and $\mathbf{P}(\mathbf{X}_{t'}|H_l, \mathbf{R}_{t',l}, \mathbf{g}_{l,j})$ represent the joint probability density functions of $\mathbf{X}_{t'}$ under hypotheses H_0 and H_l which are in terms of the trace function, $_{Tr}$, respectively. $\hat{\mathbf{R}}_{t',0}$ and $\hat{\mathbf{R}}_{t',l}$ are the well-known MLEs of the unknown parameters $\mathbf{g}_{l,j}$ and $\mathbf{R}_{t'}$ under hypotheses H_0 and H_l, respectively, see [12, p.430, Theorem:10.1.1].

By an algebraic manipulation, Eq.(1) can be simplified as

$$\lambda_l(\mathbf{X}_{t'}) = \frac{|\mathbf{X}_{t'}\mathbf{X}^\dagger_{t'}|}{\min_{\mathbf{g}_{l,j}} |\mathbf{F}_{\mathbf{g}_{l,j}}|}, \qquad (2)$$

where

$$\mathbf{F}_{\mathbf{g}_{l,j}} \triangleq (\mathbf{X}_{t'} - \mathbf{g}_{l,j}\mathbf{s}_l)(\mathbf{X}_{t'} - \mathbf{g}_{l,j}\mathbf{s}_l)^\dagger. \qquad (3)$$

To find $\min_{\mathbf{g}_{l,j}} |\mathbf{F}_{\mathbf{g}_{l,j}}|$ in Eq.(2), one may expand and make some variables substitution for $\mathbf{F}_{\mathbf{g}_{l,j}}$ in Eq.(3).

$$\begin{aligned}\mathbf{F}_{\mathbf{g}_{l,j}} &= \mathbf{X}_{t'}\mathbf{X}_{t'}^{\dagger} - \tilde{\mathbf{g}}_{l,j}\tilde{\mathbf{s}}_l\mathbf{X}_{t'}^{\dagger} - \mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger}\tilde{\mathbf{g}}_{l,j}^{\dagger} + \tilde{\mathbf{g}}_{l,j}\tilde{\mathbf{g}}_{l,j}^{\dagger} \\ &= (\tilde{\mathbf{g}}_{l,j} - \mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})(\tilde{\mathbf{g}}_{l,j} - \mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})^{\dagger} + \mathbf{G}_l, \end{aligned} \tag{4}$$

where

$$\begin{aligned}\tilde{\mathbf{g}}_{l,j} &= \alpha_{l,j}\tilde{\mathbf{b}}_{l,j}, \\ \mathbf{G}_l &= \mathbf{X}_{t'}\mathbf{X}_{t'}^{\dagger} - (\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})(\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})^{\dagger},\end{aligned}$$

and the sampled spreading code sequence is normalized by letting,

$$\tilde{\mathbf{s}}_l = (\mathbf{s}_l\mathbf{s}_l^{\dagger})^{-1/2}\mathbf{s}_l,$$

which yields $\tilde{\mathbf{s}}_l\tilde{\mathbf{s}}_l^{\dagger} = 1$, and

$$\tilde{\mathbf{b}}_{l,j}\tilde{\mathbf{s}}_l = \mathbf{b}_{l,j}\mathbf{s}_l,$$

where $\tilde{\mathbf{b}}_{l,j}$ is the modified complex gain vector of the J-element antenna array, given by

$$\tilde{\mathbf{b}}_{l,j} = (\mathbf{s}_l\mathbf{s}_l^{\dagger})^{1/2}\mathbf{b}_{l,j}.$$

Evidently, the matrix $\mathbf{G}_l$ has a quadratic form and can be proved to be positive definite with probability one, see [1, Appendix A], [2, Appendix A]. Hence, after factoring $\mathbf{G}_l$ out of Eq.(4), the determinant of Eq.(4) can be expressed as follows:

$$|\mathbf{F}_{\mathbf{g}_{l,j}}| = |\mathbf{G}_l|[1 + (\tilde{\mathbf{g}}_{l,j} - \mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})^{\dagger}\mathbf{G}_l^{-1}(\tilde{\mathbf{g}}_{l,j} - \mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})]. \tag{5}$$

If no default form is assumed for the antenna pattern $\mathbf{b}_{l,j}$, the maximum solution of Eq.(2) can be achieved when the second term in Eq.(5) vanishes, i.e.,

$$\hat{\tilde{\mathbf{g}}}_{l,j} = \mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger} \;\Rightarrow\; \hat{\tilde{\mathbf{b}}}_{l,j} = \alpha_{l,j}^{-1}\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger}$$

so that

$$\begin{aligned}\lambda_l(\mathbf{X}_{t'}) &= \frac{|\mathbf{X}_{t'}\mathbf{X}_{t'}^{\dagger}|}{|\mathbf{X}_{t'}\mathbf{X}_{t'}^{\dagger} - (\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})(\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})^{\dagger}|} \\ &= \frac{1}{1 - (\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})^{\dagger}(\mathbf{X}_{t'}\mathbf{X}_{t'}^{\dagger})^{-1}(\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})}.\end{aligned} \tag{6}$$

Apparently, using Eq.(6) the test in Eq.(2) is equivalent to the test statistic given by

$$\begin{aligned}q_l(\mathbf{X}_{t'}) &= (\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger})^{\dagger}(\mathbf{X}_{t'}\mathbf{X}_{t'}^{\dagger})^{-1}(\mathbf{X}_{t'}\tilde{\mathbf{s}}_l^{\dagger}) \\ &= \frac{(\mathbf{X}_{t'}\mathbf{s}_l^{\dagger})^{\dagger}(\mathbf{X}_{t'}\mathbf{X}_{t'}^{\dagger})^{-1}(\mathbf{X}_{t'}\mathbf{s}_l^{\dagger})}{(\mathbf{s}_l\mathbf{s}_l^{\dagger})}.\end{aligned}$$

The test statistic q_l is used to test at each time phase within time period NT_c for the existence of the signal of the l^{th} user. The decision on which timing phase is most likely to occur in a symbol interval is attained by finding the maximum over the filter bank of tests in $Q \triangleq \{q_l(\mathbf{X}_{t'}) |\, 0 \le t' \le NS_a - 1\}$.

IV. Computer Simulations

In this section, two simulation examples are conducted to demonstrate the performance of the proposed detector for timing acquisition of DS-CDMA signals. In both examples, an asynchronous BPSK DS-CDMA system with the number of users $M = 8$ is considered. User 1, 2, and 3 are assumed to generate the multipath signals due to multipath reflections. The spreading code sequence of each user in the active cell is a Gold sequence of length $N = 31$ while the spreading codes for intercell interferers are random codes. The detector employs a uniformly linear antenna array with twelve sensors and half-wavelength spacing. A single user of interest with 3 multipath duplicates, say user 1, is acquired by the proposed detector in the presence of strong MAI and near-far environment. The level of MAI is designated in terms of the ratio of the power of any interfering user to the 1^{th} path of the desired user. In Scenario I, all interfering users are assumed to have the same power ratio with respect to the desired user while it is selected randomly in Scenario II. This power advantage is denoted by a quantity called the *near-far ratio* (NFR), i.e.,

$$\text{NFR} = \frac{|\alpha_{l,j}|^2}{|\alpha_{1,1}|^2},$$

where the subscript '${}_{l,j}$' denotes the j^{th} path of l^{th} user and '${}_{1,1}$' is assumed to be the 1^{th} path of the desired user. The multipath delays, the DOAs and the NFRs of all users in both simulated examples are tabulated in Fig.1. For convenience, it is assumed in the experiments that the propagation delays are set to multiples of T_c. All the simulation curves are derived by performing 1000 *Monte-Carlo* trials.

# User	Delay (in Tc)	Signature Sequence	DOA (in degree)	NFR (Scenario I)	NFR (Scenario II)
1	3 7 14	Gold Code Gold Code Gold Code	60 0 15	1 1 1	1 1 1
2	0 11	Gold Code Gold Code	10 -15	1 1	2 4
3	5 9	Gold Code Gold Code	25 -45	1 1	1 2
4	16	Gold Code	-75	1	8
5	13	Gold Code	-60	1	1
6	21	Gold Code	40	1	4
7	19	Random Code	-20	1	2
8	25	Random Code	80	1	2

Fig. 1 Simulated multipath CDMA channel parameters for Scenario I and II.

A. Scenario I

In Fig.2, the probability of correct synchronization is presented as a function of the generalized signal-to-noise ratio ($GSNR$) (i.e., $GSNR \triangleq \mathbf{s}_1\mathbf{s}_1^{\dagger}\mathbf{b}_{1,1}^{\dagger}\mathbf{R}_{\mathbf{n}}^{-1}\mathbf{b}_{1,1}$) under the perfect power control scenario, i.e., all users have the same signal power (NFR = 0 dB). The "correct" in the phrase of correct synchronization means that all the three versions generated by user 1 must be identified correctly. Synchronization results in Fig.2 show that the performance of the proposed detector is substantially improved by a larger antenna array. When the antenna number is slightly over 7, the synchronization performance approaches to that of 12 antenna sensors case. The conventional detector using the sample cross-correlation of the received signal and the user-spreading code sequence of interest (a standard matched filter) fails to identify the multipath timings as the results shown.

B. Scenario II

In order to consider near-far situation, power ratios with respect to the desired user are assumed random, i.e., no power control is employed, in Scenario II. The other parameters are

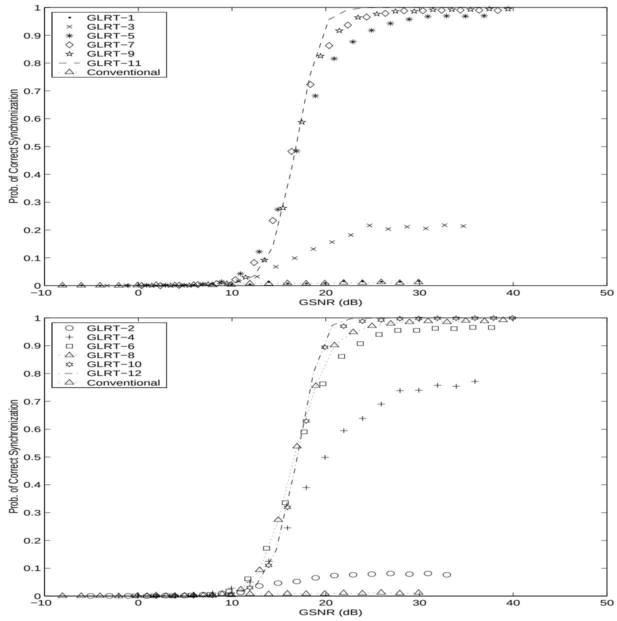

Fig. 2 Synchronization results for CDMA Scenario I, (Upper): odd antenna sensors, (Lower): even antenna sensors.

set identically to Scenario I. Observe Fig.3, the performance of acquiring synchronization is drastically degraded under near-far situation except the larger antenna array case. In the absence of power control, the antenna number must be increased to achieve the same performance as the good power control case. In Fig.4, the proposed algorithm compares and identifies the code timings of the desired spreading code sequence accurately, while it is hard for the sample cross-correlation method to make the right recognition.

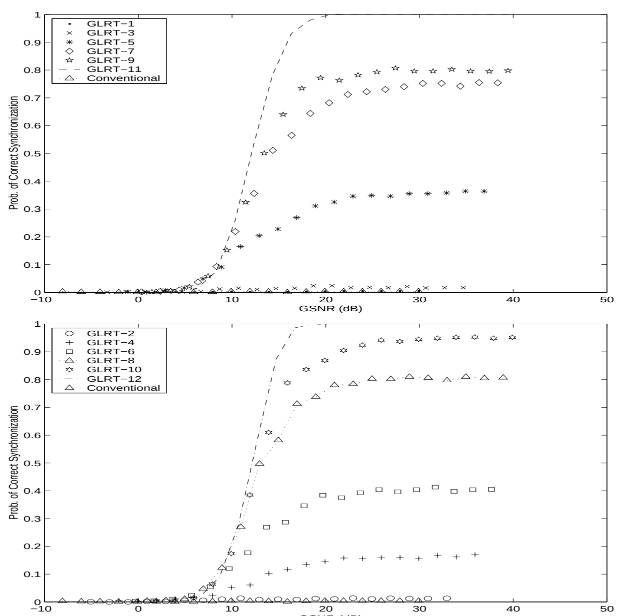

Fig. 3 Synchronization results for CDMA Scenario II, (Upper): odd antenna sensors, (Lower): even antenna sensors.

V. **Conclusion**

Code timing acquisition algorithm of an asynchronous DS-CDMA system exploiting spatial-temporal diversity is proposed for detecting the multipath timings of a single desired

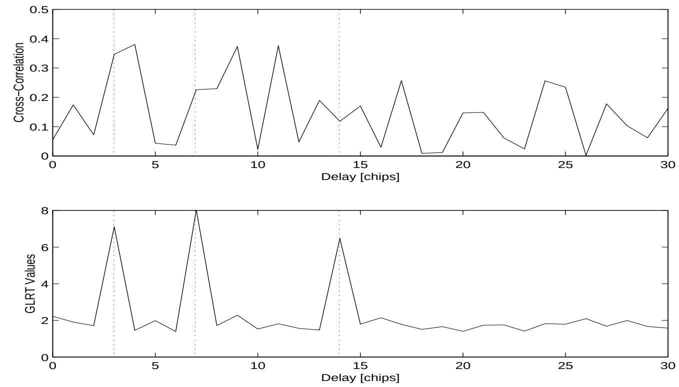

Fig. 4 Synchronization results of the sample cross-correlation algorithm (Upper) and the proposed algorithm (Lower). (M=8, SNR=6dB, and MAI≈14dB in a Near-Far Interference Environment)

user under multipath fading and near-far environment. It should be noted that the only requirement for the proposed detector is the prior knowledge of the desired user's spreading sequence. No side information of system users is necessary while focusing on a given user. Also the proposed algorithm can be easily extended to multiuser detection by forming all the single-user detectors in parallel if all users' spreading code sequences are available. In other words, the proposed algorithm can be used to identify the multipath timings for either multiuser or a single-user case. The acquisition performance is also shown to be significantly improved as the number of antenna array increases and to be near-far resistant detector.

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