DS-CDMA Noncoherent Receiver with PN Code Tracking

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ABSTRACT

In this paper, a noncoherent receiver with PN code tracking for direct sequence code division multiple access (DS-CDMA) communication systems in multipath channels is proposed and analyzed. The decision-feedback differential detection method is employed to detect PSK signals. In this method an "error signal" is used to update the tap weights and the estimated code delay. By increasing the number of feedback symbols we can improve the performance of the proposed noncoherent receiver. With the large number of feedback symbols, the optimum weight can be derived analytically, and the performance of the proposed noncoherent receiver matches to that of the conventional coherent receiver. Simulations are shown in support.

General Terms

DS-CDMA, Code Tracking, Differential Detection

1 INTRODUCTION

In telecommunications, direct-sequence spread spectrum (DSSS) is a modulation technique, in which the signal takes up more bandwidth than the information signal that is being modulated. The name 'spread spectrum' comes from the fact that the carrier signals occur over the full bandwidth (spectrum) of a device's transmitting frequency. Direct sequence spread-spectrum (DS/SS) systems have recently received much interest for use in mobile radio systems for commercial applications [1]. This has motivated considerable research in different aspects of SS systems including multiple access, multiuser receiver design, code synchronization, power control, etc. Code synchronization is one of the most critical issues to be resolved by the system designer since lack of synchronization results in a complete loss of useful communication. Code

synchronization is usually obtained in two stages: acquisition (coarse alignment) and tracking (fine alignment). Acquisition is initially used to bring the delay offset between the incoming signal and the locally generated code to within the pull-in range of the tracking loop(usually one code symbol duration), then tracking is initiated to minimize the delay offset error and to compensate for the changes that may be caused by channel variations, code Doppler, and clock instabilities. Most of the previously proposed code tracking techniques use a closed loop form such as the delay-locked loop (DLL)and the tau–dither loop (TDL) [2]. The incoming signal is correlated either simultaneously (in DLL) or alternately (in TDL) with delayed and advanced versions of the local pseudo noise (PN) code (usually one chip or less apart), and the correlators' outputs are subtracted to generate an error signal. This signal is used to control a voltage-controlled clock (VCC) that derives the local PN code generator to close the loop that minimizes the error signal.

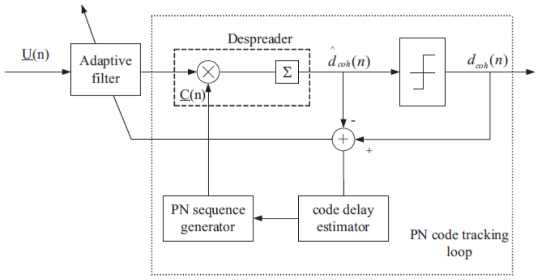


Fig.1. Conventional coherent receiver

The conventional coherent receiver deal with data detection and PN code timing recovery jointly is described in Fig. 1. PN code tracking can be categorized into coherent and noncoherent loops. When the demodulation is coherent, a coherent carrier reference must be generated prior to demodulation. The generation of coherent reference at low signal-to-noise ratio is difficult. This difficulty is from the fact that any communication system must convey information from transmitter to the receiver. This implies that the carrier is in some way modulated with this information. In contrast to coherent case, the noncoherent detection is less complex and more robust against carrier phase variations.

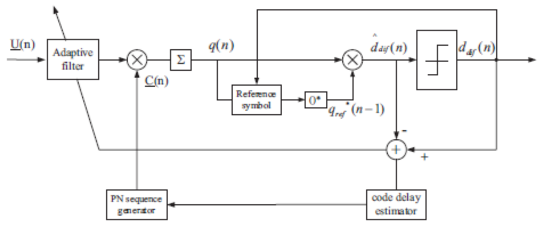


Fig. 2 Noncoherent receiver with PN code tracking

Fig. 2. Shows the noncoherent receiver this noncoherent detection requires the differential

Encoding in the transmitter and the reference symbol is needed for decoding in the receiver

2. System Model

In the CDMA system, the base station transmits *K* data vectors of active users. Each user, *k*, transmits a data symbol sequence *{bk*(*n*)*}*, *k* = 1, ...,*K*, which consists of

Independent and identically distributed (i.i.d.) *M*-ary PSK symbols at symbol interval *Ts*, i.e.,

$b_k(n) \in \{e^{j2\pi\nu/M} \mid \nu \in$

(0, 1, ...,*M* − 1)*}*. The data symbol sequences of different users are independent. The actual data rate may be varied due to the service provided or the actual transmission conditions. Each data symbol *bk*(*n*) of user *k* is first oversampled by a spreading factor *Lc* and the resulting vector is multiplied element-by-element by the elements of PN spreading code sequence *{pk*(*l*)*}*, *k* = 1, ...,*K*, which consists of *Lc* in general

complex chips at chip interval *Tc* and satisfies the relation *Tc* = *Ts/Lc*. The transmitted baseband spread spectrum
Signal is defined as

$$S(t)=\sum_{k=0}^{K-1}\sum_{n=-\infty}^{\infty} b_k(n)PN_k(t-nT_s) \qquad (1)$$

Where $bk(n)$ is the information-bearing symbol of the k-th user, and $PNk(t)$ is a wideband PN sequence defined by $PN_k(t) = \sum_{l=0}^{L_c-1} p_k(l)\Psi(t-lT_c)$ where $p_k(l) \in \{\pm 1\}$

is the l-th element of the PN code of user k, and $\Psi(t)$is the chip waveform. The CDMA downlink signal is then transmitted through a multipath channel. In the digital DSCDMA receiver, the incoming waveform is down converted in quadrature and sampled at the Nyquist rate. The bandwidth of PN code is approximately $1/Tc$. The Nyquist sampling interval is Td, $Td = Tc/D$, where D is the number of samples during one chip duration in the receiver. Therefore, the number of samples in one symbol duration is DLc. In general, the
Discrete-time channel is modeled relative to the bandwidth of the DS waveform. A multipath channel can be represented by a tapped-delay-line with spacing equal to $1/DB$, where B is the signal bandwidth. Thus, the channel can be modeled as an FIR filter of length Ln whose impulse response is $\{h_i\}_0^{L_n-1}$. ISI arises for Ln greater than one, and MAI arises due to channel distortion or non-orthogonal spreading code or both. We assume that the channel order $Ln << DLc$ since the maximum delay spread of the channel is usually insignificant in relative to the symbol period Ts [14]. The received sample
Sequence $\{ri\}$ can be expressed as [2][3][4]

$$r_i = e^{j\Theta}\sum_{l=0}^{L_n-1} h_l S_{i-l}+v_i = e^{j\Theta}\sum_{l=0}^{L_n-1} h_l S\left((i-l)T_d-\tau_0\right)+v_i \qquad (2)$$

Where $S(t-\tau 0)$ means that $S(t)$ suffers from timing offset Θ denotes a constant phase shift introduced by channel, and$\tau 0$ denotes the true code delay. The noise term $\{vi\}$ is an i.i.d. Gaussian random sequence with the variance $\sigma 2n$. The PN code samples generated in the receiver are represented by $\{ci\}$ which is obtained by sampling $PNk(t)$ and $ci = PNk(iTd-\tau)$ where τ is the code delay which can be adapted to track the true code delay $\tau 0$. Our problem is to design an
Adaptive filter $\{wi\}$ that estimates (or predicts) the desired signal. In the same time, we estimate the code delay $\tau 0$ to achieve code synchronization

3THECONVENTIONAL COHERENT RECEIVER

The conventional coherent receiver with PN code tracking is shown in Fig. 1. The constant phase shift Θ is assumed to be known perfectly. The tap weight vector is

$$\underline{W}(n) = [w_0(n) \;\; w_1(n) \;\; \cdots \;\; w_{L_w-1}(n)]^H, \qquad (3)$$

Where L_w is the number of taps of the transversal filter used in the receiver. The local PN code vector is.

$$\underline{C}(n) = [c_{DL_c(n-1)+1} \;\; c_{DL_c(n-1)+2} \;\; \cdots \;\; c_{DL_cn} \;\;]^T, \qquad (4)$$

Where $[\cdot]^T$ denotes Transposition and $[\cdot]^H$ denotes Hermitian operation. Define the sample matrix

$$\underline{U}(n) = \qquad (5)$$

$$\begin{bmatrix} r_{DL_c(n-1)+1} & r_{DL_c(n-1)+2} & \cdots & r_{DL_cn} \\ r_{DL_c(n-1)} & r_{DL_c(n-1)+1} & \cdots & r_{DL_cn-1} \\ \vdots & \vdots & \cdots & \vdots \\ r_{DL_c(n-1)-(L_w-2)} & r_{DL_c(n-1)-(L_w-3)} & \cdots & r_{DL_cn-(L_w-1)} \end{bmatrix}$$

The estimated signal $Y\wedge(n)$ is denoted by

$$\underline{\hat{Y}}(n) \stackrel{def}{=} [\hat{y}_{DL_c(n-1)+1} \;\; \hat{y}_{DL_c(n-1)+2} \;\; \cdots \;\; \hat{y}_{DL_cn}]^T \qquad (6)$$
$$= \underline{U}^H(n)\underline{W}(n).$$

Note that $C(n)$ has the property $C(n)^TC(n) = \beta$ where β is a constant that can be determined. That is, with sampling time Tc/D and without filtering, the maximum correlation of PN sequence is DLc. After filtering, β is no longer an integer, but Still selected by maximum correlation value in general. The value β depends on the sampling of the chip waveform $\Psi(t)$, that is

$$\beta = L_c \times \left\{\max_{\delta} \sum_{l=0}^{D-1} \Psi^2(\delta + lT_d)\right\},$$

Where $0 \leq \delta \leq Td$ if $\Psi(t)$ is time-limited. It is desired that the output of the adaptive filter is the estimation of the desired signal, so
That the normalized despreader output is

$$\hat{d}_{coh}(n) = \underline{W}^H(n)\underline{U}(n)\underline{C}(n)/\beta \qquad (7)$$

and the output behaves like an MPSK signal. In other words, we want that the signal at the despreader output to have the same statistic as that of an ideal MPSK demodulator output. To achieve this goal we can use the cost function $J_{COH=}$

$$E\left[\left|d_{coh}(n) - \hat{d}_{coh}(n)\right|^2\right]$$

Where $dcoh\ (n)$ is the hard decision
Result of $\hat{d}_{coh}(n)$

Let $ecoh\ (n) = dcoh\ (n) - dcoh\wedge(n)$ be the error signal, *the cost function Jcoh can be written as*

$$\begin{aligned} J_{coh} &= E[d_{coh}(n)d^*_{coh}(n) - d_{coh}(n)\underline{C}^T(n)\underline{U}^H(n)\underline{W}(n)/\beta \\ &\quad -d^*_{coh}(n)\underline{W}^H(n)\underline{U}(n)\underline{C}(n)/\beta \\ &\quad +\underline{W}^H(n)\underline{U}(n)\underline{C}(n)\underline{C}^T(n)\underline{U}^H(n)\underline{W}(n)/\beta^2] \end{aligned} \qquad (8)$$

$$\begin{aligned} \frac{\partial J_{coh}}{\partial \tau} &= -d_{coh}(n)\frac{\partial \underline{C}^T(n)}{\partial \tau}\underline{U}^H(n)\underline{W}(n)/\beta \\ &\quad -d^*_{coh}(n)\underline{W}^H(n)\underline{U}(n)\frac{\partial \underline{C}(n)}{\partial \tau}/\beta \\ &\quad + \underline{W}^H(n)\underline{U}(n)\left(\frac{\partial \underline{C}(n)}{\partial \tau}\underline{C}^T(n) + \underline{C}(n)\frac{\partial \underline{C}^T(n)}{\partial \tau}\right) \\ &\quad \times \underline{U}^H(n)\underline{W}(n)/\beta^2 \end{aligned} \qquad (9)$$

where

$$\frac{\partial \underline{C}(n)}{\partial \tau} = \left[\frac{\partial}{\partial \tau}c_{DL_c(n-1)+1} \;\; \frac{\partial}{\partial \tau}c_{DL_c(n-1)+2} \;\; \cdots \;\; \frac{\partial}{\partial \tau}c_{DL_cn}\right]^T \qquad (10)$$

$$\frac{\partial c_i}{\partial \tau} = \frac{\partial PN_k(iT_d - \tau)}{\partial \tau} = \sum_{l=0}^{L_c-1} p_k(l)\frac{\partial}{\partial \tau}\Psi(iT_d - \tau - lT_c) \qquad (11)$$

The tap weight of the adaptive filter is thus updated by

$$\begin{aligned} \underline{W}(n+1) &= \underline{W}(n) - \mu\frac{\partial J_{coh}}{\partial \underline{W}} \\ &= \underline{W}(n) + \mu\left[\frac{2}{\beta}e^*_{coh}(n)\underline{U}(n)\underline{C}(n)\right] \end{aligned} \qquad (12)$$

Where μ *is* the step-size. The code delay τ is updated through $(n+1) = \tau(n) - \lambda\partial Jcoh/\partial\tau$ where λ is the step-size. The quantity $-\frac{\partial J_{coh}}{\partial \tau}$ can be regarded as an" error signal", estimating the chip-timing error of a code tracking loop.

4. THE PROPOSED NONCOHERENT RECEIVER

Fig. 2 shows the block diagram of the proposed noncoherent receiver that combines the differential detection with PN code tracking for DS-CDMA systems. The i.i.d. information sequence *{ak(n)}* is first differentially encoded. The resulting MDPSK symbols *bk*(*n*) are given

$$b_k(n) = a_k(n)b_k(n-1) \qquad (13)$$

by

and the transmitted signal model is the same as (1). The received sample sequence *{ri}* is also expressed as (2). Here, the constant phase shift Θ is unknown. At the receiver, the sampled signals are first passed through the transversal filter and then despreaded by the local PN sequence. The tap weight vector *W*(*n*), local PN code vector *C*(*n*), and the received sample matrix *U*(*n*) can be described as equations(3), (4), and (5), respectively. The normalized despreader output *q*(*n*) is the same as (7), and can be represented as

q(*n*) = *WH*(*n*)*U*(*n*)*C*(*n*)*/β.* In the next stage, the differential detection is necessary to recover the MDPSK information sequence. The decision variable *ddif* (*n*) is obtained by noncoherent processing of the despreader output *q*(*n*),

ddif (*n*) =*q*(*n*)*q∗ref* (*n* − 1) where the reference symbol *qref* (*n* − 1) is generated [4] as follows

$$q_{ref}(n-1) = \frac{1}{N-1}\sum_{l=1}^{N-1} q(n-l)\prod_{m=1}^{l-1} d_{dif}(n-m) \quad (14)$$

Where *N*, *N* ≥ 2, is the number of despreader output symbols used to calculate d_{dif} (*n*). *ddif* (*n*) is the hard decision result of $\widehat{d}_{dif}(n)$. Note that for *N* = 2,

qref (*n* − 1) = *q*(*n* − 1), $\widehat{d}_{dif}(n)$ is the decision variable of a conventional differential detection. However, for *N* > 2, a significant performance improvement can be obtained. We can use the cost function

$$J_{dif} = E\left[\left|d_{dif}(n) - \hat{d}_{dif}(n)\right|^2\right]$$

. The error signal can be defined as

$$e_{dif}(n) \overset{def}{=} d_{dif}(n) - \hat{d}_{dif}(n).$$

Here, *edif* (*n*) at the *n*-th symbol time also depends on past tap weight vectors *W*(*n*−*v*), *v* ≥ 1. For the derivation of the adaptive algorithm, these past tap weight vectors are treated as constants since

|*edif* (*n*)|^2 is differentiated only with respect to *W*(*n*). The cost function of differential detection *Jdif* can be written as

$$\begin{aligned}
J_{dif} &= E[d_{dif}(n)d^*_{dif}(n) \qquad (15)\\
&-d_{dif}(n)\frac{q_{ref}(n-1)}{\beta}\underline{C}^T(n)\underline{U}^H(n)\underline{W}(n)\\
&-d^*_{dif}(n)\frac{q^*_{ref}(n-1)}{\beta}\underline{W}^H(n)\underline{U}(n)\underline{C}(n)\\
&+\frac{|q_{ref}(n-1)|^2}{\beta^2}\underline{W}^H(n)\underline{U}(n)\underline{C}(n)\underline{C}^T(n)\underline{U}^H(n)\underline{W}(n)]
\end{aligned}$$

The gradient of the cost function with respect to the tap weight vector is

$$\begin{aligned}
\frac{\partial J_{dif}}{\partial \underline{W}} &= -\frac{2}{\beta}q^*_{ref}(n-1)d^*_{dif}(n)\underline{U}(n)\underline{C}(n)\\
&+ \frac{2}{\beta^2}|q_{ref}(n-1)|^2\underline{U}(n)\underline{C}(n)\underline{C}^T(n)\underline{U}^H(n)\underline{W}(n)\\
&= -\frac{2}{\beta}q^*_{ref}(n-1)e^*_{dif}(n)\underline{U}(n)\underline{C}(n) \qquad (16)
\end{aligned}$$

and the gradient of the cost function with respect to the code delay

$$\begin{aligned}
\frac{\partial J_{dif}}{\partial \tau} &= -d_{dif}(n)\frac{q_{ref}(n-1)}{\beta}\frac{\partial \underline{C}^T(n)}{\partial \tau}\underline{U}^H(n)\underline{W}(n)\\
&\quad -d^*_{dif}(n)\frac{q^*_{ref}(n-1)}{\beta}\underline{W}^H(n)\underline{U}(n)\frac{\partial \underline{C}(n)}{\partial \tau}\\
&+\frac{|q_{ref}(n-1)|^2}{\beta^2}\underline{W}^H(n)\underline{U}(n)\\
&\times\left(\frac{\partial \underline{C}(n)}{\partial \tau}\underline{C}^T(n) + \underline{C}(n)\frac{\partial \underline{C}^T(n)}{\partial \tau}\right)\underline{U}^H(n)\underline{W}(n) \quad (17)
\end{aligned}$$

is

The gradient vector ∂*C*(*n*)/∂*τ* is the same as (10). The tap weight of the noncoherent adaptive filter is updated by

$$\underline{W}(n+1) = \underline{W}(n) + \mu\left[\frac{2}{\beta}q^*_{ref}(n-1)e^*_{dif}(n)\underline{U}(n)\underline{C}(n)\right] \quad (18)$$

And the code delay is updated by

$$\tau(n+1) = \tau(n) - \lambda\frac{\partial J_{dif}}{\partial \tau}.$$

5. PERFORMANCE ANALYSIS

Problem 1

In this problem, We implemented the corresponding programs to simulate the performance of a DS-SS/BPSK communication system in the following scenarios:

1. In the AWGN channel
2. In the presence of pulsed noise jamming and AWGN

The average BER obtained by simulation is attached.

Discussion

1. From the simulation, we find that the performance curves are very closed to the theoretical results.
2. We also find that the performance curves are similar for different code lengths.
3. Averaging over more experiments and using a larger symbol size will produce results closer to the theoretical results. The symbol size used in my simulation is10,000.

Problem 2

1. Besides implementing all the required simulations, we also tried to use error correcting codes to reduce the average BER. The code used in our experiment is binary BCH code with code word length and message length equal to 15 and 5, respectively. The experimental result is attached.
2.We also simulated the performance of the system in barrage noise jamming and AWGN. The experimental result is attached.

Discussion

1. From the experiment, we find that using error-correcting codes reduces the average BER obviously, especially when bit energy to noise ratio is relatively large.
2. The performance of the system in barrage noise jamming and AWGN environment is very close to the theoretical results

Problem 3

In this problem, We implemented the corresponding programs to simulate the average BER in:

1. Rayleigh fading and AWGN
2. Frequency selective Rayleigh fading and AWGN

The average BER obtained by simulation is attached.

Discussion

1. From the simulation, we find that the performance curves are very close to the theoretical results.
2. We also find that the performance curves are similar for different code lengths.
3. Averaging over more experiments and using a larger symbol size will produce results closer to the theoretical results. The symbol size used in our simulation is 10,000.

Besides implementing all the required simulations, we also implemented a pre-detection selective combining diversity receiver to reduce the average BER. In our experiment, we used two branches (channels) to transmit signals. The experimental result is attached.

From the experiment, we find that using pre-detection selective combining diversity receiver reduces the average BER obviously, especially when bit energy to noise ratio is relatively large

Problem 4

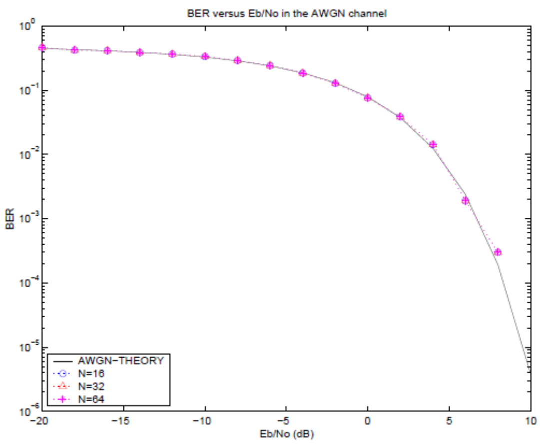
In this problem, we implemented the corresponding programs to simulate the average BER versus Eb/N0 of K users transmitting BPSK symbols at an equal power lever using DSCDMA in the following two scenarios:

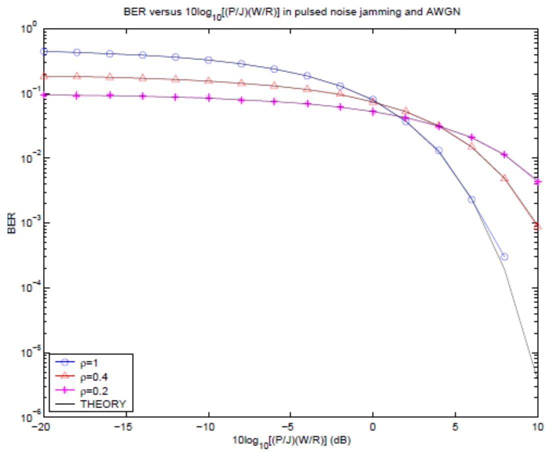
1. Perfect synchronism and orthogonal codes in AWGN
2. A random a synchronism between the users, uniformly distributed between 0 and 5 samples, **Discussion**

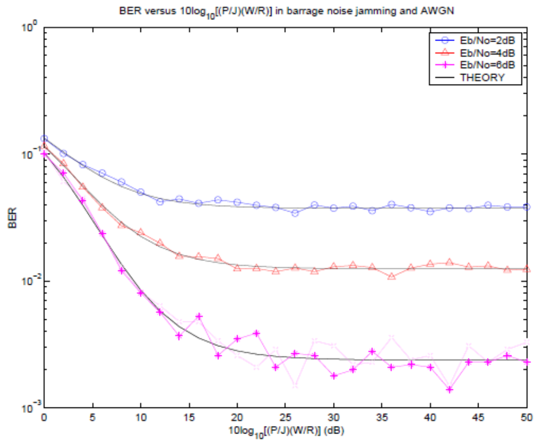
1. From the simulation, we find that the performance curves are very close to the theoretical results.
2. Averaging over more experiments and using a larger symbol size will produce results closer to the theoretical results. The symbol size used in my simulation is 10,000.
3. For the second part of the problem, when the number of users is equal to 4 and 8, we should use a very large symbol size in the simulation in order to get accurate results for large bit energy to noise ratio.
4. Although it takes me lots of time to do the simulation, I learn a lot and understand the theories better by doing it.
5. At last, I want to say that I really like this course. It has lots of fun and makes me like communications more.

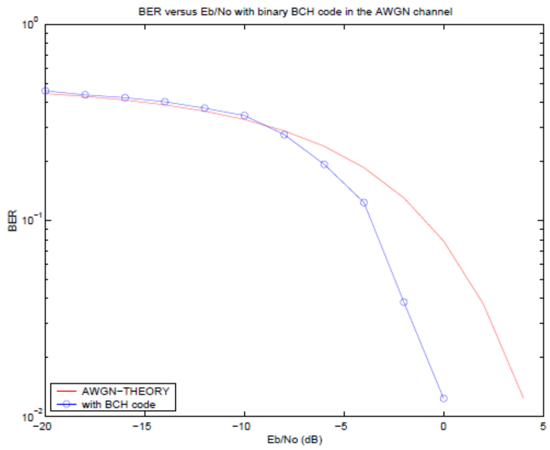
AWGN

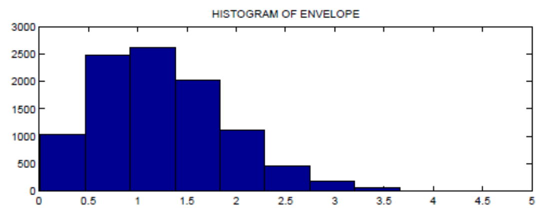
6. SIMULATION RESULTS

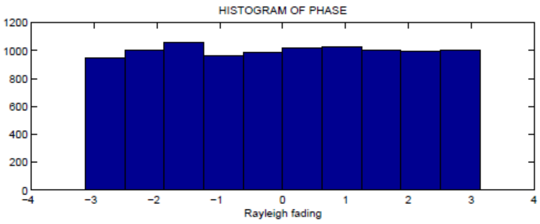


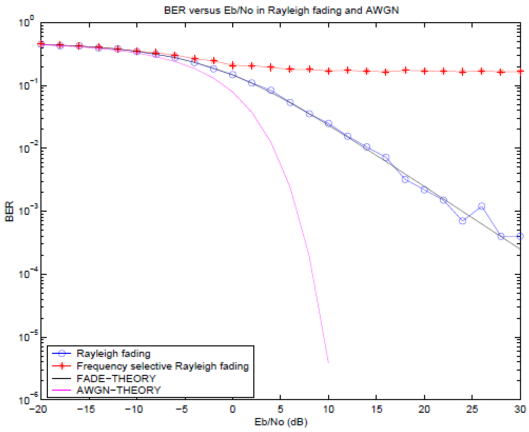


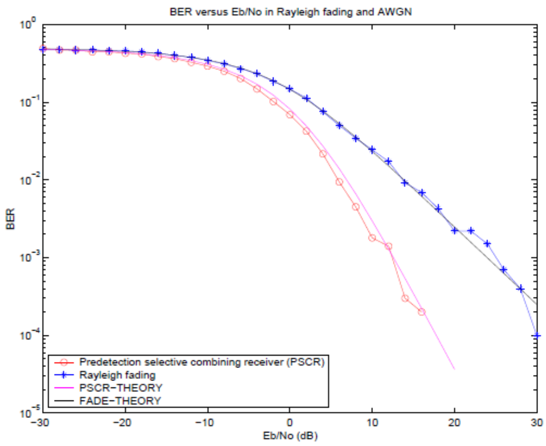


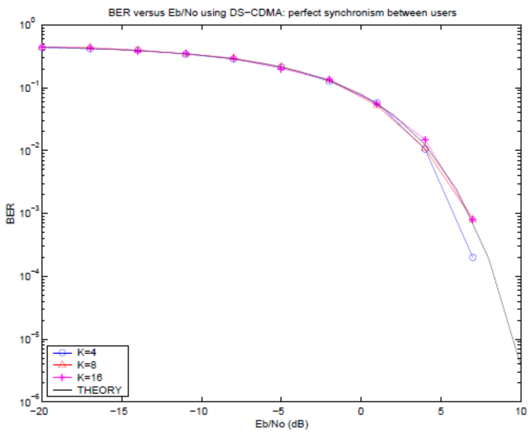


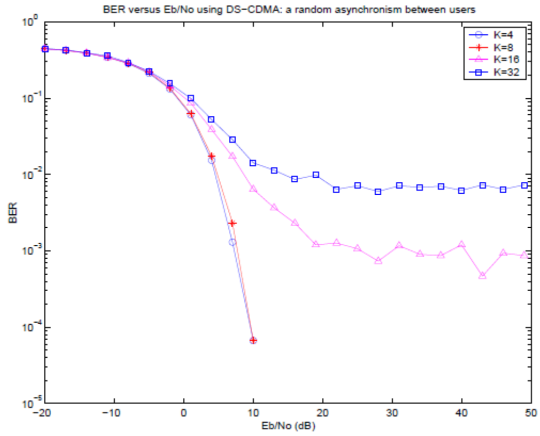












7. Conclusion

A novel noncoherent receiver for joint timing recovery and data detection in DS-CDMA systems is proposed in this work. It estimates the desired signal and code delay by LMS algorithm at the same time. It is shown that the timing offset can be rapidly tracked even if the mismatch is up to half chip time interval. The loss of noncoherent detection compared with conventional coherent detection is limited and can be adjusted via the generation of the reference symbol for the decision-feedback Differential detection. The performance of the noncoherent receiver can approach the performance of the conventional coherent receiver if an infinite number of feedback symbols are used. Furthermore, simulations show that the proposed receiver in an asynchronous situation approaches the performance as that of the receivers with perfect synchronization.

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