CHAOTIC SEQUENCES AS SPREADING CODES FOR DS/CDMA SYSTEM

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Thesis Title	Chaotic Sequences as Spreading Codes for DS/CDMA System
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ABSTRACT

Direct sequence code division multiple access (DS/CDMA) technique has been used extensively in communication system since it can accommodate the limited availability of the frequency bandwidth. In addition, a DS/CDMA system is also able to well-maintain the quality of the system in the multiple access interference (MAI) and inter symbol interference (ISI) environments due to multipath nature of channels in presence of additive white Gaussian noise (AWGN). In the DS/CDMA system, spreading codes play an important role in multiple access capacity. M-sequence, gold sequence, as namely a few, has been traditionally used as spreading codes. These sequences are generated by shift registers and are periodic in nature causing less number of sequences and limiting the security.

This thesis presents an investigation of using chaotic sequences as spreading code for a DS/CDMA system. The chaotic sequences are generated by a chaotic map generator. The superior beneficial of chaotic spreading code than conventional one is that the chaotic sequences are easy to generate for a very long sequence. In addition, an enormous number of different sequences can simply be generated by changing its initial condition. Chaotic sequences are deterministic, reproducible, uncorrelated and random-like, which can be very helpful in enhancing the security of transmission in communication system. The performance of chaotic sequences in a DS/CDMA communication system using adaptive iterative receiver is investigated in this thesis. The simulation results prove that the chaotic sequences provide better performance compared with the conventional spreading code. Using chaotic sequences as spreading codes can accommodate a large number of users and spreading sequences without degrading the BER performance and the security of the system.

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Bangkok, Thailand
May, 2011

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Chapter 1

Introduction

1.1 Background

The new technology in wireless communication is noteworthy to study and much more challenging since the demand of high performance in telecommunication system increase faster over in the past decades. High capacity, transmission velocity, transmission quality, and security are the main performance needed by users. Many researches were proposed to attain the superior performance in wireless communication systems [1-4]. Some issues concerning with wireless communication become the main aspect for many researchers to gain the users' need regarding to their necessity, such as how to communicate in easier and faster way and also how to get information in real-time processing.

The explosive number of users in recent years has allowed the wireless communication system to provide the simultaneous access with the same frequency bands, especially when radio links are required for a large number of terminals in a density population area. Hence, the high frequency, low power and low-voltage circuitry are needed to fulfill the demands of the overall system. But then the other problems are emerged, such as the interference by adjacent users due to the limited availability of frequency bands.

Direct Sequence Code Division Multiple Access (DS/CDMA) was proposed as a technique to combat the interference between adjacent users and to improve the efficiency bandwidth of the transmitted signal using pseudorandom sequence as spreading sequences [5, 6]. In this technique, the transmitted signal appears as noise-like and the power spectral density is low so that the signal will be difficult to jam. The orthogonal spreading sequences allow the users to communicate simultaneously on the same channels by their unique spreading sequence on each user.

The conventional spreading sequences which commonly used in DS/CDMA system are maximal length shift register sequences (M-sequences) and Gold sequences [7, 8]. The M-sequences are the longest codes that can be generated by a given shift register or a delay element of the fixed length. Therefore, the number of sequences is usually too small and is not suitable for spread spectrum systems. Gold sequences have been proposed to cope with the drawback of M-sequences in 1967 by Gold [9]. These sequences are constructed by the EXOR of two M-sequences with the same length. The simple

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circuitry generator to generate a large number of spreading sequences makes Gold sequences more popular to be implemented in DS/CDMA system than M-sequences.

As the number of users increase dramatically, DS/CDMA system must provide more spreading sequences to handle more users. But, since the spreading sequences in the DS/CDMA need more memory to generate a large number of spreading sequences, then a chaotic system was proposed to replace spreading sequences in DS/CDMA system [10-15]. Although a chaotic sequence is still relatively new to be applied in communication system, but the performance of this system can boost the superiority of a DS/CDMA system compare with the conventional spreading sequences concerning with the security and unlimited sequences to generate. The benefits of using chaotic sequences compared to conventional spreading sequences are described below:

- With given code length, a larger number of sequences is available compared to the number of conventional spreading sequences
- Chaotic sequences can be easily generated and their statistical characteristics can be adjusted to their user's needs.

In this thesis, layered space time coding (LSTC) DS/CDMA structure of transmitter and receiver is presented to simulate chaotic spreading code.

This chapter begins with some background issues in DS/CDMA communication system and the motivations of the work are presented in Section 1.2. The objectives of the thesis are well described in Section 1.3 and finally the outline of this thesis will be given in Section 1.4 as the last section in this chapter.

1.2 Motivations

The growing demand of the performance in wireless communication has created an impending need for development of the new technique to improve the speed, reliability and capacity of communication system or for decreasing the size, complexity and cost. The DS/CDMA with spread spectrum technique was investigated to suppress the drawback of wireless communication. Since the conventional spreading sequences in DS/CDMA have some limitation in order to achieve and maintain carrier and symbol synchronization due to the different propagation times for different users, so the perfect synchronization can never be achieved in the real system. The other issues concerning with the spread spectrum system are the carrier regeneration and the code clock regeneration detectors, because of the binary nature spreading sequences used in binary waveforms. This emerges possibility to investigate newly chaotic sequences in DS/CDMA communication system in performance comparison with the conventional one.

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A chaotic system as a deterministically dynamical system with their special characteristics, predictable in the short-time and exhibiting long-term unpredictability, has a potential superiority referred to the conventional spreading sequences [16]. Details of the chaotic behavior are hypersensitive to change in initial conditions (even minor changes in the starting values of the variables). Equivalently, chaotic signals rapidly de-correlate with themselves [17, 18]. The auto-correlation functions of the chaotic signal have a large peak at zero and decays rapidly. It becomes good benefit in synchronization for each of the spreading sequences. While the low cross-correlation function are useful for increasing the user's capacity in DS/CDMA communication systems.

The implementation and investigation to enhance the performance of chaotic sequences in DS/CDMA system were discussed in [19-21]. Synchronization is still the important issued for applying a chaotic signal in DS/CDMA due to its characteristics.

1.3 Objectives

In this research, chaotic sequences with one-dimensional map and three-dimensional map are proposed to investigate the enhancement of the DS/CDMA system as pseudorandom sequences in terms of BER performance and complexity analysis for the whole systems. Implementation of the new structure of the receivers is also discussed. The recursive least square (RLS) as an adaptive iterative receivers which can elevate the interference between adjacent users and channels are the matched ones to collaborate with chaotic sequences, so that the superior performance of DS/CDMA can be achieved.

1.4 Thesis Outline

The thesis is organized in five chapters and is briefly summarized as follows.

Chapter 1 introduces some background information and previous works about the research in the chaotic system and the application in communication systems. The key idea in this chapter is to show the potential benefit of using chaos for communication especially in the DS/CDMA system.

Chapter 2 describes the literature reviews of DS/CDMA and the chaotic system. The concept of DS/CDMA with the basic structure of the transmitter and the receiver using a conventional spreading code such as M-sequences and Gold sequences is discussed. Several dimensional-chaotic maps are also explained in this chapter as spreading sequences such as Logistic map and Chebyshev map as a one-dimensional map and Chen's attractor as a three-dimensional map.

Chapter 3 presents methodology of the research. The proposed method of chaotic signals as spreading sequences and the structure of the adaptive RLS receivers are described in detail.

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Chapter 4 gives the simulation results and analysis of chaotic performances in the DS/CDMA as pseudorandom sequences compared with the conventional spreading sequences using adaptive iterative RLS receivers.

Chapter 5 concludes the thesis in reviewing and highlighting the contributions that have been presented in this thesis. The suggestions for future works related to this research are given as well.

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Chapter 2

Literature Reviews

The main interest in this thesis is the analysis and design of transceivers in a DS/CDMA system using chaotic spreading sequence. Some preliminaries necessary are presented in this chapter, which will be described in the following chapter. Section 2.1 describes the overviews of a DS/CDMA system as spread spectrum communication system. Section 2.2 introduces the basic description of a chaotic system, while Section 2.3 reviews about channel coding principles, and Section 2.4 presents the multiuser detection techniques.

2.1 DS/CDMA systems

The basic concept of direct sequence spread spectrum communication systems as the core of the CDMA system is the multiple access using spread spectrum modulation with their unique spreading sequence to identify each user. To mitigate the interference among the users, the unique spreading sequence must have the good properties of autocorrelation and cross correlation. An asynchronous CDMA system was proposed the spreading sequence using pseudorandom (PN) sequence to assign each user. The kind of the PN sequence can be gold sequence, Kasami sequence, maximum length sequence (m-sequence) to name a few.

In this section, the briefly introduction of a DS/CDMA system will be presented. The definition of the DS/CDMA system will be described by the model of functional block DS/CDMA communication system. The generation of the PN sequence is also discussed in this section as well, while the application for the conventional spreading sequence are given at the last section.

2.1.1 Concept and Modeling

For a DS/CDMA system, a direct sequence spread spectrum technique allows users to occupy the same time and frequency. In this system, each user is identified by the unique codes [22, 23]. These unique codes are set binary sequences which are orthogonal or almost orthogonal to one another. Figure 2.1 demonstrates the functional block of a transceiver in the DS/CDMA system.

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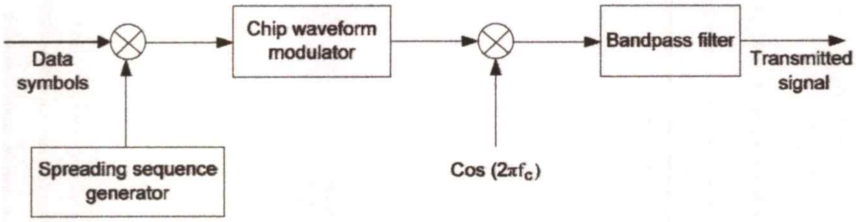
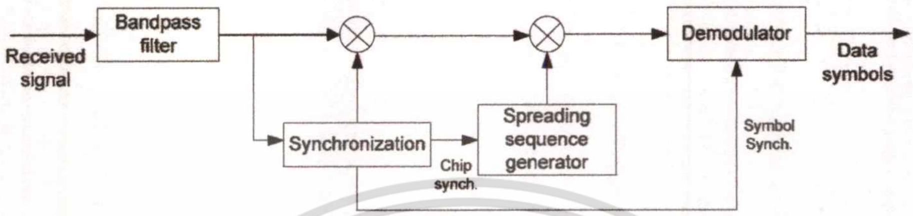


(a) Transmitter



(b) Receiver

Fig. 2.1 The functional block diagram of a DS/CDMA transceiver

At the transmitter, the binary message signal is multiplied by the spreading sequence, generated by a spreading sequence generator. While, the receiver correlates with the received signal with a replica of the spreading sequence. The threshold element estimates the transmitted signal from the result of the correlation. Spreading sequences are required to have a broadband spectrum; i.e., they should be uncorrelated. In order to reduce crosstalk, spreading sequence from different transmitter-receiver pairs must not correlate to each others.

2.1.2 PN sequence generator

A DS/CDMA system allows a PN sequence to modulate the information of the data with noise like waveform [24, 25]. A PN sequence is generated by such a shift register that has autocorrelation properties similar to those of white noise. There are several PN sequences that are used to generate the binary spreading sequence in DS/CDMA system. In this section the detail descriptions of generation the PN sequence are presented below. These are a maximal-length sequence (m-sequence) generator and a gold sequence generator.

A PN-sequence, $\ldots \alpha_{-2}, \alpha_{-1}, \alpha_0, \alpha_1, \alpha_2 \ldots$, where $\alpha_n \in \{0,1\}$, generated by shift registers is a periodic sequence of ones and zeros with period N_p. Because the sequence is periodic, $a_n = a_{n+N_p}$ for any n. The spreading signal a (t) derived from this pseudorandom sequence is also a periodic pseudorandom function of time with period $\mathcal{T} = N_p T_c$ and is defined by

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$$a(t) = \sum_{n=-\infty}^{\infty} a_n \, h_{T_c}(t - nT_c) \tag{2.1}$$

where $a_n = (-1)^{\alpha_n}$, and $h_{T_c}(t)$ is a pulse or chip period T_c. The signal $a(t)$ is deterministic, so that its autocorrelation function is defined by

$$R_a(\tau) = \frac{1}{\mathcal{T}} \int_0^{\mathcal{T}} a(t)a(t+\tau)\, dt \tag{2.2}$$

Because $a(t)$ is periodic with period $\mathcal{T}$, it follows that $R_a(\tau)$ is also periodic with period $\mathcal{T}$, the power spectrum A (f) is found by taking the Fourier transform of (2.2). This power spectrum consists of discrete spectral lines at all harmonics of $^1/_{N_p T_c}$

$$A(f) = \sum_{k=-\infty}^{\infty} A_k \delta\,(f - \frac{k}{N_p T_c}) \tag{2.3}$$

Where δ (.) is the delta function. Consider two different signals $a(t)$ and $'(t)$. The cross correlation function of these two deterministic signals is

$$R_{aa'}(\tau) = \frac{1}{\mathcal{T}} \int_0^{\mathcal{T}} a(t)a'(t+\tau)\, dt \tag{2.4}$$

Where it has been assumed that $a(t)$ and $a'(t)$ have the same period $\mathcal{T}$, then the cross-correlation function is also periodic with period $\mathcal{T}$.

2.1.2.1 Maximum-length sequence (m-sequence)

An M-sequence can be generated using linear feedback shift registers (LSFR) and the combination logic that will produce the sequence which likes a high frequency noise. This pseudo-noise sequence is not completely random, but it is generated by a well defined logic. The same logic is used at the transmitter and the receiver. Figure 2.2 shows the illustration to generate *m*-sequences using *m*-registers, where the coefficient g_1, g_2, ….., g_m are elements of the binary Galois Field, GF(2), with the value 0 or 1.

The symbols of the field element over GF(2^m) in polynomial form are shown below :

$$\beta(D) = \beta_0 + \beta_1 D + \cdots \beta_{m-1} D^{m-1} \tag{2.5}$$

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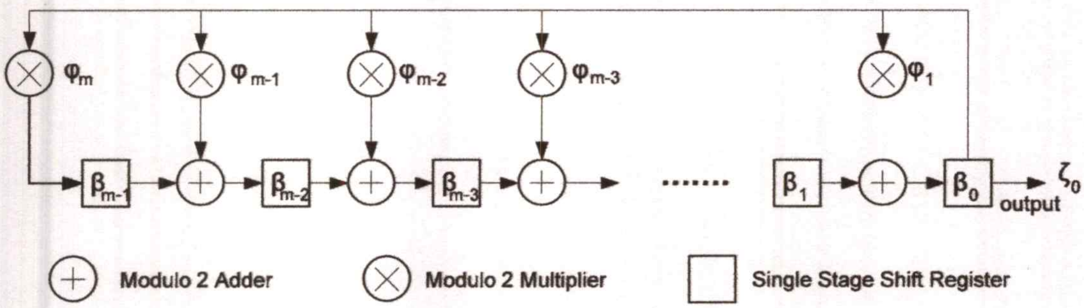

Fig. 2.2 An M-stage linear shift register (Galois feedback generator)

These polynomial forms are initially stored in the stages of a shift register as can be seen at Figure 2.2. The first operation is to add (in GF(2)) $\beta_{0\varphi_i}$ to β_i for $1 \le i \le m-1$ and then to shift the register one position to the right, replacing β_{m-1} in the leftmost position by $\beta_{0\varphi_m}$. It should be verified that the content of the shift register at this point in polynomial form is

$$\beta^{(1)}(D) = \sum_{i=1}^{m-1} \beta_i D^{i-1} + \beta_0 \sum_{i-1}^{m} \varphi_i D^{i-1} \tag{2.6}$$

From Equation (2.6), it follows that

$$\beta^{(1)}(D)D = \beta(D) + \beta_{0\varphi}(D) \tag{2.7}$$

where

$$\varphi(D) = 1 + \sum_{i=1}^{m} \varphi_i D^i \tag{2.8}$$

$\varphi(D)$ is the characteristic polynomial of the shift register.

Suppose now that $\varphi(D)$ is a primitive polynomial over GF(2) and this polynomial generates GF(2^m). Because a primitive elements of the field is a root of $\varphi(D)$ we get from Equation (2.8) that

$$\beta^{(1)}(\alpha)\alpha = \beta(\alpha). \tag{2.9}$$

If $\beta(\alpha) = \alpha^j$, then $\beta^{(1)}(\alpha) = \alpha^{j-1}$.

The circuit in Figure 2.2 generates the power of α in inverse order. Correspondingly, the output of the shift register, ζ (*D*), is the sequence of zero coefficients of the elements of this sequence. The period N_p of this sequence is $(2^m - 1)$, because the number of the nonzero elements in GF(2^m) is equal to $(2^m - 1)$. This is the maximal possible period for shift registers of length *m*. Such a shift register is

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called a maximum-length (linear) shift register (MLSR). The sequence can be generated by an MLSR sequence, maximum-length sequence, maximal-length, or *m*-sequence.

Equation (2.7) can be written in the form

$$\beta^{(t+1)}(D)D^{t+1} = \beta^{(t)}(D)D^{t} + \zeta_t D^t \varphi(D), \tag{2.10}$$

where $\beta^{(t)}$ is the register content at the t-th moment and ζ_t is the register output at the t-moment, t = 0, 1, 2, ….

The summation of Equation (2.10) over *t* from *t* = 0 to *t* = ∞ gives

$$\sum_{t=o}^{\infty} \beta^{(t)}(D)D^{t} + \beta^{(0)}(D) = \sum_{t=0}^{\infty} \beta^{(t)}(D)D^{t} + \zeta(D)\varphi(D), \tag{2.11}$$

where

$$\zeta(D) = \sum_{t=0}^{\infty} \zeta_t D^t \tag{2.12}$$

is the output sequence of the shift register. From (2.12), we have

$$\zeta(D) = \frac{\beta^{(0)}(D)}{\varphi(D)}, \tag{2.13}$$

where $\beta^{(0)}(D)$ is the initial condition polynomial.

Randomness property of the PN-sequence is the main point to support the CDMA system. M-sequences almost satisfy the randomness properties of Bernoulli sequences. The parameters of the generation process (the coefficient φ_i of the generator polynomial (D)) are deterministic but the *m* terms of the initial vector, or equivalently, the time shift of the MLSR sequence can be chosen randomly. As pseudorandom sequence, that eligible to use in the predecessor of CDMA system, *m*-sequences have the best periodic autocorrelation in terms of minimizing the maximum value of the out-of-phase autocorrelation, while for the cross correlation, some of m-sequence pairs have large cross correlation values that make the m-sequence no longer exist to implement in multiple users environment. The limited number of the sequences that can be generated by the m-sequence generator is also the problem that lead out to the new generation of spreading sequences.

Table 2.1 shows the list of the shift register connections for generating the m-sequences. An *m*-sequence is a maximum-length nonlinear shift register sequences, which is said to be a primitive *m*-sequence if it is obtained by adding all zero *n*-taps in ones of a period of an *m*-sequence.

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Table 2.1 Shift-register connection for generating *m*-sequences

m	Stage Connected	*m*	Stage Connected	*m*	Stage Connected
2	1, 2	13	1, 10, 11, 13	24	1, 18, 23, 24
3	1, 3	14	1, 5, 9, 14	25	1, 23
4	1, 4	15	1, 15	26	1, 21, 25, 26
5	2, 5	16	1, 5, 14, 16	27	1, 23, 26, 27
6	1, 6	17	1, 15	28	1, 26
7	1, 7	18	1, 12	29	1, 28
8	1, 5, 6, 7	19	1, 15, 18, 19	30	1, 8, 29, 30
9	1, 6	20	1, 18	31	1, 29
10	1, 8	21	1, 20	32	1, 11, 31, 32
11	1, 10	22	1, 22	33	1, 21
12	1, 7, 9, 12	23	1, 19	34	1, 8, 33, 34

Delay and add property

Consider an output sequence ζ_0 (D) of the shift register in Figure 2.3. According to Equation (2.13), it can be represented as

$$\zeta_0(\mathrm{D}) = \frac{\beta^{(0)}(D)}{\varphi(D)}, \tag{2.14}$$

where $\beta^{(0)}$ is the initial condition polynomial and $\varphi(D)$ is the characteristic polynomial of the shift register. If the sequence ζ_0 (D) are shifted by an arbitrary number of cycles k, $k < 2^m - 1$, the output sequence of the shift register ζ_k (D), for different initial vector $\beta^{(k)}(D)$ is below

$$\zeta_k(\mathrm{D}) = \frac{\beta^{(k)}(D)}{\varphi(D)}. \tag{2.15}$$

If the original sequence ζ_0 (D) and the shifted sequence ζ_k (D) is added, term by term modulo 2, we obtain a new sequence that itself is an output sequence of the shift register with initial condition vector $\beta^{(0)}(D) + \beta^k(D)$, that is

$$\zeta_0(D) + \zeta_k(D) = \frac{\beta^{(0)}(D)+\beta^{(k)}(D)}{\varphi(D)}. \tag{2.16}$$

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Hence, the generated sequence is itself a time shift of the same *m*-sequence. This is the *delay and add property*. The sequence $\zeta_0(D) + \zeta_k(D)$ satisfies the balance property. It follows that two sequences $\zeta_0(D)$ and $\zeta_k(D)$, each of the length 2^m-1, differ in 2^{m-1} positions and agree in $2^{m-1} - 1$ positions.

We can formulate the first and third properties in terms of the real number of alphabet {1, -1} instead of the binary logical alphabet {0, 1}. The first property is reformulated as equality

$$\frac{1}{N_p} \sum_{n=1}^{N} a_n = -\frac{1}{N_p}, \qquad (2.17)$$

where a_n is the equivalent (in the real number alphabet) of the *n*-th term of the MLSR sequence, ζ_n, that is, $a_n = (-1)^{\zeta_n}$. Equation (2.12) follows from the balance property in Equation (2.11)

The third property is

$$\theta_a(k) = \frac{1}{N_p} \sum_{n=1}^{N} a_n a_{n+k} = \begin{cases} 1, & k = 0 \\ -1/N_p, & k \neq 0 \end{cases} \qquad (2.18)$$

This is the discrete autocorrelation function of the *m* sequence takes only two values. The autocorrelation of the m-sequence can be defined by averaging over a complete cycle of the sequence. In practice, synchronization of signals spread by the pseudorandom sequence often require that an estimate of the correlation between the received signal and the receiver dispreading signal is made in less than a full period.

If we consider the sequence $\{a_n\}$ as a discrete time periodical random process, Equation (2.12) implies that the time average of the process is equal to $-1/N_p$; Equation (2.13) implies that the time correlation of the process is equal to 1 or $-1/N_p$. If we treat the initial condition vector as a random vector with II D equiprobable binary components, the *m*-sequence becomes a stationary ergodic process.

In some CDMA applications, it is useful to be able to generate *m*-sequence in different phase. For example in a DS/CDMA system, different phase shift of a very long spreading sequence may be used to distinguish different users. The DS/CDMA receiver selects the desired signal by choosing the phase of the reference dispreading signal. The phase difference may be equivalent to thousand chips, thus making the use of straightforward phase shift impractical. A practical method is simply to calculate the shift register initial conditions required to generate a sequence delayed by *k* chips from the sequence generated from another specific initial condition.

From the consideration of the Galois feedback generator, it follows that these initial contents of the shift register are two elements of GF (2^m) such that $\beta^{(k)} = \beta^{(0)}\alpha^{-k}$.

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2.1.2.2 Gold sequence

The gold spreading sequences were introduced by Gold in 1967. Gold sequence that was invented by Gold has a small value of cross correlation within a set, that lead out to the good impact in multiple user environments. This class of pseudorandom sequences with their good properties can decrease the multiple access interference that can be found in multiuser DS-CDMA communication systems [26]. Furthermore, a gold sequence generator is also can generate more pseudorandom sequences, so that the multiple users in DS/CDMA can be handled.

Gold sequence can be generated by a pair of *m*-sequences, called preferred *m*-sequence and forming the modulo-2 of two sequences. Consider *a* maximum-length sequence *a* of period $N_p = 2^m - 1$ and a second sequence is said to be the decimation of the sequence *a* . the notation *a'* =a[q] is used for sequence *a'*. Sequence *a'* has period N_p if and only if gcd(N_p,q) = 1, where "gcd" means the greatest common advisor. Any pairs of maximum-length sequences *a* and *a'* having the same period N_p are related as *a'* = *a*[*q*] for some *q*.

The cross-correlation spectrum of pairs of maximum-length sequences can be three-valued, four-valued, or many-valued. All pairs of maximum-length sequences whose cross-correlation is three-valued are called *preferred pairs* of *m* sequences. Those three-values are $-\frac{1(m)}{N_p}, -\frac{1}{N_p}, \frac{[t(m)-2]}{N_p}$, where $= log_2(N_P + 1), N_P$, N_P is the sequence's period,

$$\begin{cases} 1 + 2^{\frac{m+1}{2}}, & for\ m\ odd, \\ 1 + 2^{\frac{m+2}{2}}, & for\ m\ even \end{cases} \tag{2.19}$$

The following conditions are sufficient to define a preferred pair *a* and *a'* of maximum-length sequences with period $N_p = 2^m - 1$:

- $m \neq 0$ (mod 4); that is, *m* is odd or *m* = 2 (mod 4)
- *a'* = *a* [*q*], where either $q = 2^k + 1$ or $q = 2^{2k} - 2^{2k} + 1$
- gcd (m,k) = $\begin{cases} 1, for\ m\ odd \\ 2, for\ m = 2(mod\ 4) \end{cases}$

where gcd means greatest common devisor.

In defining sets of Gold sequences, it is necessary to find *preferred pairs* of maximum-length sequences. Let *a* and *a'* represent a preferred pair of maximum-length sequences having period $N_p = 2^m - 1$. Consider the set of $N_p + 2 = 2^m + 1$ sequences {*a*, *a'*, *a*+*a'*, *a*+*Da'*, … , $a+D^{N-1}a'$}, where $D_{a'}^{j}$ means phase shift of *a'* by *j* units. This set is called the set of Gold sequences. It can be proved that any pair of sequences in the set has three-valued cross-correlation spectrum. With the

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exception of sequences *a* and *a'*, the set of Gold sequences does not contain maximum-length sequences. Hence, their autocorrelation functions are not two-values but four-valued and they take the same three-values as the cross-correlation plus the value 1.

Gold sequences form a family of spreading sequences with good cross-correlation properties, while the auto-correlation properties is not as well as cross-correlation properties [27].

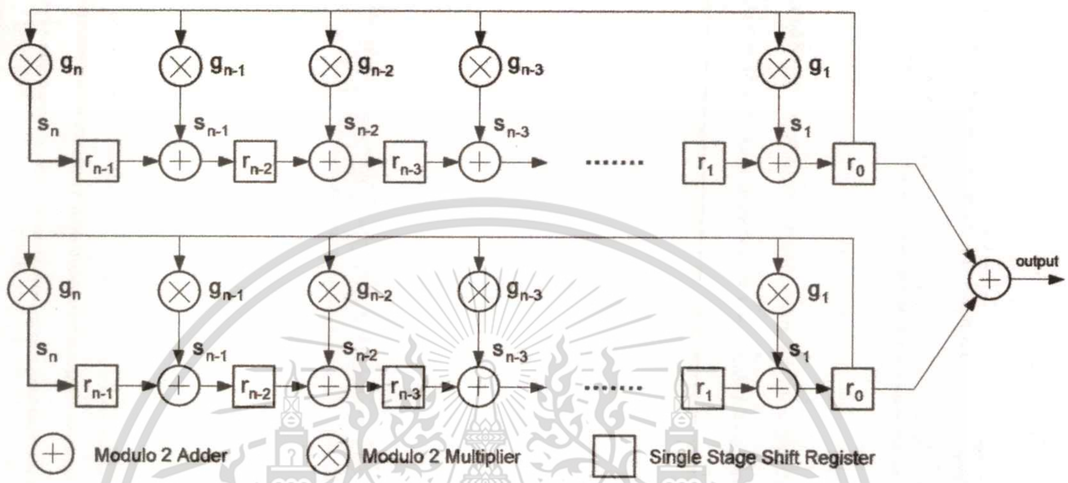

Fig. 2.3 The Gold code generator with *m*-shift registers

2.2 Chaotic System

The superiority performance of chaotic in nonlinear dynamics attracted many researchers to employ in communication system, since the natural behavior of chaos can turn out to be the beneficial factor to gain security and capacity in terms of data communication [28-33]. In this section, the concept and modeling of the chaos will be discussed in details and the generation of chaotic sequences using several map of a chaotic generator will be presented. To support the compatibility in communication system, some synchronization techniques of coupled chaotic system [34] are also given in this section.

2.2.1 Concept and modeling

The idea of using chaos in communications systems sparked in the early 1990s. A brief summarize of the properties of chaotic signals are described in Table 2.2. These applications result from three core aspects of chaos, broadband, complexity, and orthogonality [35-38].

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Table 2.2 Properties of chaotic sequences

Property	Description	Application
Deterministic	Generated by deterministic systems (algorithms); can be reproduced in principle	Information coding, measurements
Noise-like	Seemingly random; irregular to an observer not knowing the generation algorithm	Pseudorandom signal generators, information encryption
Broadband	Continuous power density spectrum over an extended frequency range	Broadband stimulation, measurement, and communication
Decorrelated	Signals from : - The same generator with different initial conditions - Structurally identical systems with slight parameter differences that are not correlated	CDMA, communication and broadband measurements

2.2.2 Chaotic sequence generator

The major difference of chaotic sequence than a PN sequence is the characteristic of chaotic sequence that are no longer binary. Therefore, the chaotic sequence must be transferred into binary sequences. Figure 2.4 is shown the process of binary converter in chaotic generation. The chaotic sequences are generated by chaotic map generators. This continuous chaotic sequence then feeds forward to the quantization and encoding process.

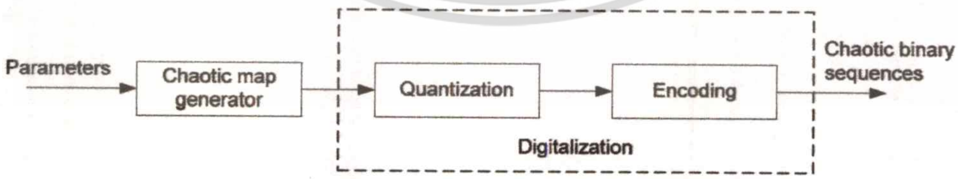

Fig. 2.4 A chaotic sequence generator

In this research, the T. Kohda binary quantification algorithm is used to assure the security of binary sequences as applied in cryptography [39]. Chaotic real-valued sequence is transformed into the

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binary by using the algorithm. The procedure of T. Kohda binary qualification algorithm is described as follows.

Firstly, the threshold function of $\sigma_v(x)$ is defined as

$$\sigma_v(x) = \begin{cases} 0 & (x < v) \\ 1 & (x \geq v) \end{cases} \tag{2.20}$$

Secondly, the absolute value of x in a floating-point number with m bits is written by

$$|x| = 0.c^1(x)c^2(x) \ldots c^m(x) \qquad c^i(x) \in \{0,1\} \tag{2.21}$$

The i-th bit can be expressed as shown at Equation below. So that, the binary sequences can be obtained from the seed generator of chaotic system,

$$c^i(x) = \sigma_{\frac{1}{2}}\left(2^{i-1}|x| - \textstyle\sum_{p=1}^{i-1} 2^{i-p-1}c^p(x)\right) \tag{2.22}$$

Figure 2.5 shows the basic processing scheme of a CDMA system using chaotic code sequences. The two chaos generators in the transmitter and receiver have to generate the same sequence $x(k)$, so they have to be synchronized. If a reliable synchronization channel is not available, this can be done by generating the sequences in advance and storing them in the receiver. This automatically means truncation and periodic repetition, which is an important condition for synchronization. Synchronization is then achieved in the usual way, as it is done with PN sequences, or alternatively, a corresponding matched filter with a chaotically generated sequence of coefficients is used.

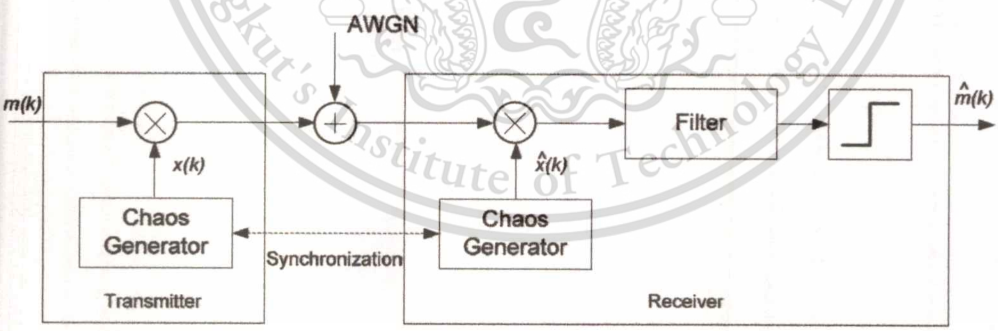

Fig. 2.5 Baseband model of a CDMA system using chaotic sequences

The performance of coherent chaos-based CDMA systems can be characterized as follows: as long as a correlation receiver is used for the calculation of the bit error rate, only the second-order characteristics are taken into account. So the receiver cannot distinguish chaotic and random signals,

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and the performance of chaos-based systems should compare with that of conventional systems. The benefit of using chaotic sequences consists of two points:

1. With given code length, a larger number of sequences can be generated by a chaotic sequence generator compared with the conventional spreading sequences.
2. Chaotic sequences can be easily generated and their statistical characteristics can be adjusted with the users' needs.

The latter point is especially interesting; hence, it has been found recently that slightly correlated sequences show a better performance than purely uncorrelated ones [40]. From the desired correlation or spectral properties of the sequence a return map g can be constructed that, from different initial values, generates a variety of code sequences. Another aspect to take into account is the structure of chaotic sequences for the receiver design. This leads to the nonlinear receivers, which might perform better than the classical ones [41-44].

2.2.2.1 Logistic map

One of the simplest and most widely studied nonlinear dynamical systems capable of exhibiting chaos is the logistic map [46]. Logistic map is one-dimensional map and nonlinear. One important feature of thre logistic map is the passage to chaos through a sequence of period doublings; the bifurcation where this doubling occurs is called a *pitchfork* bifurcation, because the local shape of the bifurcation diagram resembles a pitchfork [47],

$$F(x,r) = rx(1-x) \tag{2.23}$$

Or written in its recursive form, we have

$$x_{n+1} = rx_n(1-x_n), \quad 0 \le x_n \le 1, \quad 0 \le r \le 4 \tag{2.24}$$

Here, F is the transformation mapping function, and r is called the bifurcation parameter, that is shown in Figure 2.6 with $2,8 \le r \le 4$. Depending on the value of r, the dynamics of this system can be changed attractively, exhibiting periodicity or chaos. The bifurcation occurs at $r = 3$, leading to a stable period-2 cycle which eventually lose stability, as $r \approx 3,45$, giving rise to a stable period-4 cycle. As r increases further, the scenario repeats itself over and over again : each time a period-2^k cycle of the map F loses stability through a bifurcation of the map F^{k+1} , which gives rise to initially stable period-2^{k+1} cycle, where F is often-mentioned logistic map with a periodic point of prime period k. for $0 \le r \le r_c = 3.57 \dots$, the sequence $\{x_n\}$ of values of r at which cycles of period 2^k has a finite accumulation

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point $r \approx 3.57$. For $r_c \leq r \leq 4$, the sequence is chaotic, for all practical purposes, non-periodic and non-converging [48]. Figure 2.6 shows the bifurcation diagram of Logistic map.

Simultaneously, it is mathematically proven that excepts for negligibly short intervals where the sequence has odd periodicities, this particular range of values of r causes the logistic map to be chaotic over $\{0\ 1\}$. However, further investigation provides that map has indeed, a period-2 cycle for r slightly greater than three, equivalently, or a fixed point.

Furthermore, this map has a very sensitive dependence upon its initial value x_0, for those values of r. This sensitive dependence can be illustrated by giving a large initial points range to the iterative map. After a few iterations, the two resulting sequences will look completely uncorrelated.

Figure 2.7 is given as one dimension and phase trajectories of Logistic map using the value of initial condition is 0.5.

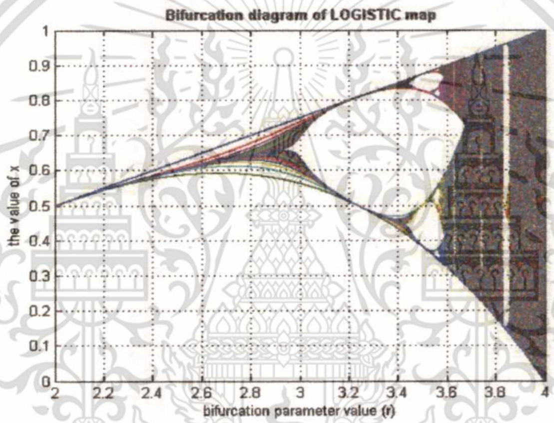
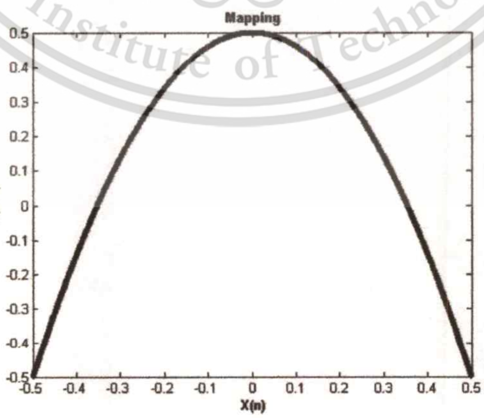


Fig. 2.6 Bifurcation diagram of Logistic map with initial value x= 0.5



a. One dimension of logistic map

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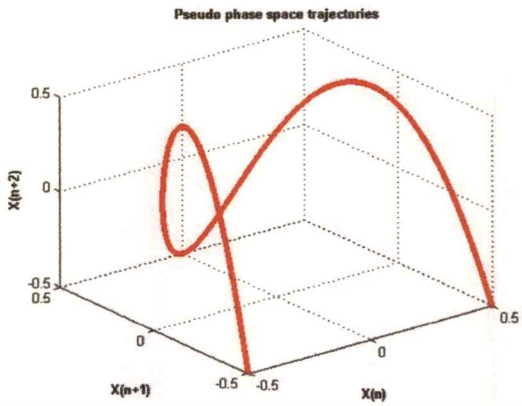

b. Phase space trajectories of logistic map

Fig. 2.7 One dimension and phase trajectories of Logistic map with initial value 0.5

2.2.2.2 Chebyshev map

The principle of Chebyshev map of degree *k*, is defined by Adler (1964) and Rivlin (1990) as shown at the Equation (2.32).

$$x_{n+1} = \cos(k\, arcos(x_n)), \quad -1 \leq x_n \leq 1, \quad n = 0, 1, 2, ... \tag{2.25}$$

Here, the map is chaotic when the value of k is $k \geq 2$. As chaotic maps, the Chebyshev equation has the properties that are sensitive to the initial value. If the value of ≥ 2 , they have positive Lyapunov exponents for any initial value, the Chebyshev is also have good properties with mixture and ergodicity. The real-valued sequences generated by the Chebyshev maps are orthogonal. The auto-correlation function of real-valued sequences generated by the Chebyshev maps is δ function [Niansheng Liu-86]

As presented in [49] that the examples of Chebyshev polynomials are given by

$$T_0(x) = 1, \qquad T_1(x) = 1,$$
$$T_2(x) = 2x^2 - 1, \qquad T_3(x) = 4x^3 - 3x, ...$$

Adler and Rivin [50] have shown Chebyshev polynomials of degree $k \geq 2$ are mixing and thus ergodic, and their invariant measure is given by $(x)dx = \frac{dx}{\pi\sqrt{1-x^2}}$. Furthermore, by investigating the asymptotical stability of the Frobenius-Perron operator corresponding to Chebyshev transformations, Chebyshev polynomials are shown to be exact and thus mixing and ergodic transformation [51].

The Chebyshev polynomials have the orthogonality as below

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$$\int_{-1}^{1} T_i(x)T_j(x)\rho(x)dx = \delta_{i,j}\frac{1+\delta_{i,j}}{2} \tag{2.27}$$

The auto-correlation functions for sequences generated by the Chebyshev polynomial are given

$$\begin{aligned}\langle C(l)\rangle &\equiv \left\langle \frac{1}{N}\sum_{j=1}^{n} T_p(x_j)T_p(x_j+1)\right\rangle \\ &= \int_{-1}^{1} T_p(x)T_{p_{t+1}}(x)\,\rho(x)dx \\ &= \frac{1}{2}\delta(l).\end{aligned} \tag{2.28}$$

2.2.2.3 Chen's attractor

In this section, the generalization of Chen's attractor is presented. The research of Chen's attractor in secure communication was proposed by Chen [52]. The definition of the Chen's attractor is described in Equation as below:

$$\begin{cases} \dot{x} = a(y-x) \\ \dot{y} = (c-a)x + cy - xz, \\ \dot{z} = xy - bz \end{cases} \tag{2.29}$$

where $a > 0$, $b > 0$, and c are the parameters ($2c > a$). With the typical parameter values of $a = 35$, $b = 3$ and $c = 28$, the corresponding numerical Chen's attractor can be found and is depicted in Fig. 2.8.

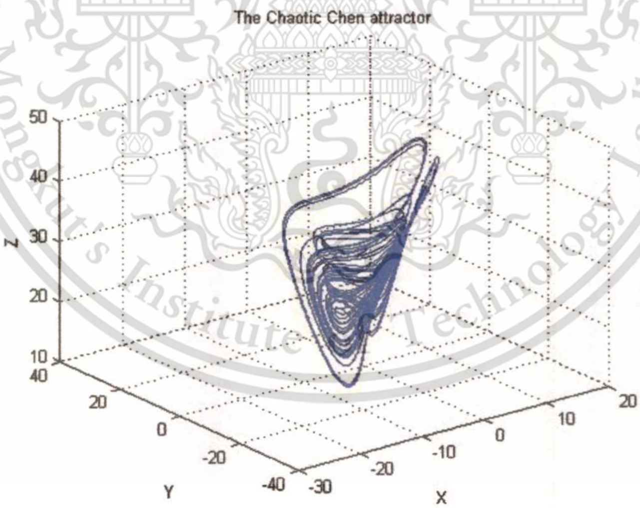

Fig. 2.8 The chaotic Chen's attractor

The previous works of using chaotic in communication system have been investigated in [74-80]. V.Nagarajan and P.Dananjayan [74] presented the implementation of logistic map in multi-carrier (MC) DS-CDMA system. In their proposed model, the chaotic sequence is using in time domain and

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frequency domain as given in Figures 2.9 and 2.10 as transmitter and receiver model respectively with *k* number of users.

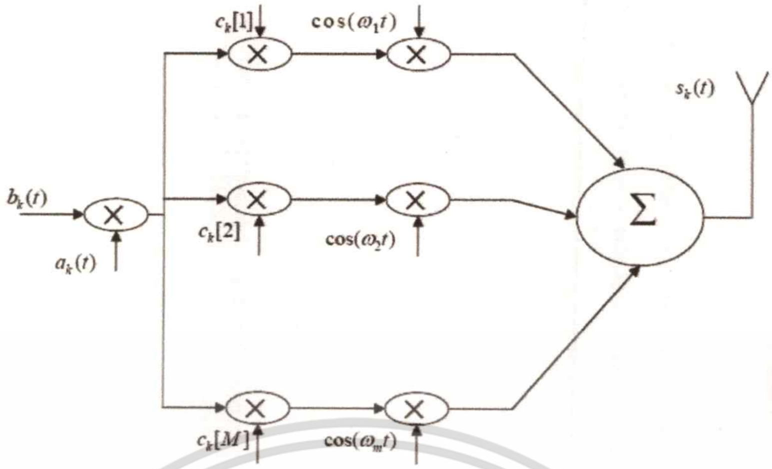

Fig. 2.9 The transmitter model of MC DS-CDMA in time domain and frequency-domain

In this technique, the transmitted signal can be expressed as

$$S_k(t) = \sqrt{\frac{2P_K}{M}} \sum_{m=1}^{M} b_k(t)a_k(t)c_k[m]cps(w_m t), \quad k = 1,2,3 \dots . K, \tag{2.30}$$

where P represents the transmitted power of each user and $\{w_m\}$, m=1,...,M represents the subcarrier frequency set. The binary data stream's $b_k(t)$ waveform consists of sequence of mutually independent rectangular pulses P_{Tb} of duration T_b and of amplitude +1 or -1 with equal probability. The spreading sequence $a_k(t)$ denotes the T-domain spreading sequence waveform of the user. By assuming the T-domain spreading factor $N=T_b/T_c$ to represents the number of the chips per bit-duration, it is can be assumed that the subcarrier signals are orthogonal and the spectral main-lobes of the sub-carrier signals are not overlapping with each other. The received signal can be shown by

$$r(t) = \sum_{k=1}^{M} \sqrt{\frac{2P_K}{M}} b_k(t)a_k(t-t_k)c_k[m]g_{m,k}cos(w_m t + \emptyset_{m,k}) + n(t), \tag{2.31}$$

Where n(t) represents the AWGN with zero mean and double-sided power spectral density of $N_0/2$. As shown in Figure 2.10 and Equation (2.29), each time and frequency (TF) domain spread MC DS/CDMA signal that is identified by the aid of two spreading sequences, one applied in the context of the T-domain.

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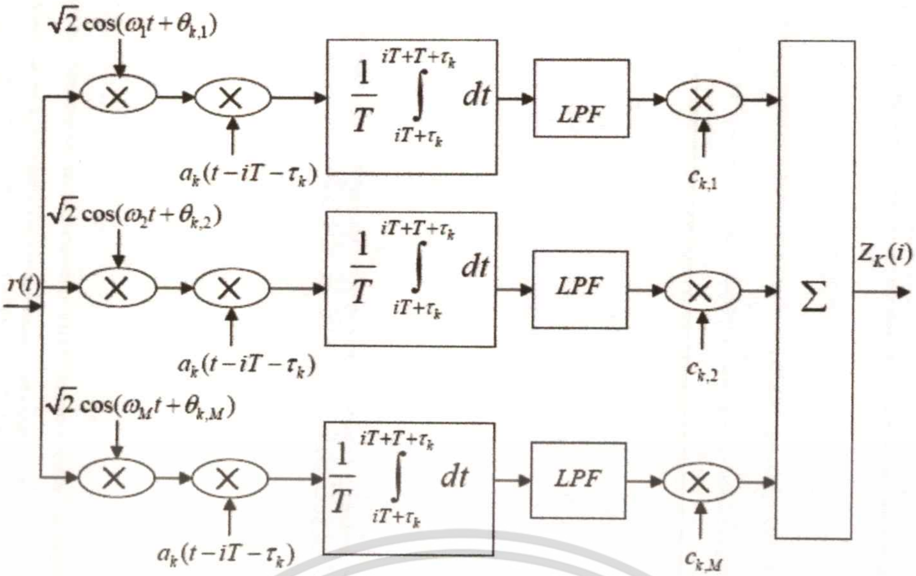

Fig. 2.10 Receiver model of MC DS-CDMA using both time-domain and frequency domain

2.3 Channel Coding Principles

During the last few decades, there has been an increasing demand for efficient and reliable digital data transmission and storage systems. To reliably reproduce the data, a major concern of the system design is the control of errors. With his 1948 paper, "A Mathematical Theory of Communication", Shannon [58] demonstrated that, by properly encoding of the information, errors induced by a noisy channel or storage medium can be reduced to any desired level without sacrificing the rate of information transmission or storages, as long as the information rate is less than the capacity of the channel. The fundamental philosophical contribution inspired the research in the error control coding areas.

2.3.1 Interleaving

Interleaving plays an important role in many digital communication systems for manifold reasons. In the context of wireless communications, fading channels often lead to bursty errors. Then, several successive symbols may be corrupted by deep fades. This has a crucial impact on the decoding performance, for example, the performance of convolutional codes because of its sensitivity to bursty. In order to overcome this difficulty, interleaving is applied. At the transmitter, an interleaver simply permutes the data stream in a specified manner, so that the symbols are transmitted in a different order. Consequently, a de-interleaver has to be employed at the receiver in order to reorder the symbols back into the original succession.

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2.3.2 Convolutional Codes

Convolutional codes are used extensively in numerous applications in order to achieve reliable data transfer, including digital video, radio, mobile communication and satellite communication. A convolutional code is described by three integers, which are the number of input symbols, *k*, the total number of output symbols, *n*, and memory order, *m* [59]. The *n*-tuple emitted by the convolutional encoding procedure is not only a function of an input *k*-tuple, but is also a function of the previous *mk*-input tuples. Figure 2.13 shows a simple convolutional code encoder, which is a linear feed forward shift register. The connection between the shift register elements and the modulo 2 adders can be conveniently described by the following two generator sequences:

$$g^{(1)} = (g_0^{(1)}\, g_1^{(1)}\, g_2^{(1)}) = (101)$$
$$g^{(2)} = (g_0^{(2)}\, g_1^{(2)}\, g_2^{(2)}) = (111) \tag{2.32}$$

If $u = (\ldots, u_{-1}, u_0, u_1, \ldots, u_l, \ldots)$ is the input data stream, then the two output sequences, denoted by $v^{(1)} = (\ldots, v_{-1}^{(1)}, v_0^{(1)}, v_1^{(1)}, \ldots, v_l^{(1)}, \ldots)$ and $v^{(2)} = (\ldots, v_{-1}^{(2)}, v_0^{(2)}, v_1^{(2)}, \ldots, v_l^{(2)}, \ldots)$ can be obtained as

$$v^{(i)} = u * g^{(i)}, \quad i = 1, 2 \tag{2.33}$$

where * denotes the convolutional operator.

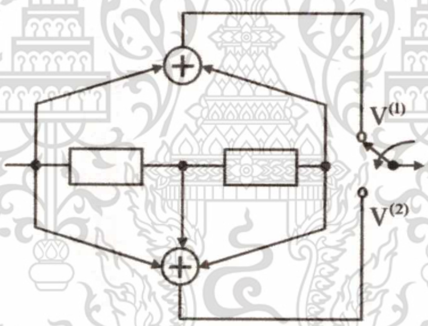

Fig. 2.13 A (2,1,2) convolutional encoder.

A convenient and common way of describing encoding and decoding operations is using trellis diagrams. A trellis stage for an input at time *t* for binary (2,1,2) code is shown in Figure. 2.14. A trellis diagram consists of *N* stages, where *N* is the number of input words, each consisting of *k* input data bits. The stage of the encoder is defined as the content of its shift register. For the encoder with total memory *K*, the stage number is 2*K*. Each new block of *k* inputs causes the transition to a new stage. That is, there are 2*K* branches leaving each stage. Each branch is labeled the *k* input causing the transition at time unit *t*, denoted by $u_t = [u_{t,1}, u_{t,2}, \ldots, u_{t,k}]$ and n corresponding to outputs, denoted by $v_t = [v_{t,0}, v_{t,1}, \ldots, v_{t,n-1}]$.

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The performance of a convolutional code depends on the decoding algorithm and distance property. If a hard-decision decoding algorithm is used, the code performance is measured by Hamming distance. The minimum free distance of a convolutional code is defined as the minimum Hamming distance between any two code sequences, which is the minimum weight of all non-zero code sequences of any length. If a soft-decision decoding algorithm is used, the code performance is measured by Euclidean distance. The minimum free Euclidean distance is defined as the minimum Euclidean distance between any two code sequences. However, the using of decision decoding algorithms is depends on both the convolutional code trellis and modulation type [60].

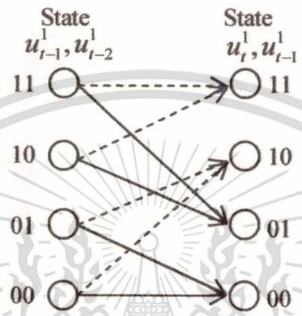

Fig. 2.14 One stage in the trellis diagram for binary (2,1,2) convolutional code.

2.3.3 MAP Decoding Algorithms

The decoder can apply a number of the soft output decoding algorithms. As the overall receiver complexity is mainly dominated by the decoder complexity, the choice of the decoding algorithm depends on the available processing power at the receiver. The maximum a posteriori (MAP) approach [61] is optimum in the sense that it minimizes the bit error probability at the decoder output and is usually implemented by using a forward and backward recursion algorithms. The MAP decoding algorithm operates on a trellis representation of the convolutional code as shown in Figure 2.14.

The conditional error probability of the decoder is defined as:

$$P(E \mid \mathbf{r}) = P(\hat{\mathbf{u}} \neq \mathbf{u}) = P(\hat{\mathbf{v}} \neq \mathbf{v}) \tag{2.34}$$

where – P(E) is the probability of error in the decoder

- **r** is the received where the term P(r) is independent of the decoding algorithms.

P(E) is given by:

$$P(E) = \sum_{\mathbf{r}} P(E \mid \mathbf{r}) P(\mathbf{r}) \tag{2.35}$$

So, the minimum probability of error in Equation (2.34) is achieved by minimizing $P(\hat{\mathrm{v}} = \mathrm{v} \mid \mathbf{r})$ for all **r** which is represented by:

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$$P(\hat{v} = v \mid \mathbf{r}) = \frac{P(\mathbf{r} \mid v)P(v)}{P(\mathbf{r})} \tag{2.36}$$

If the coded sequences are all likely equal, the probability of optimal decoding for a discrete memories channel is denoted as:

$$P(\mathbf{r} \mid v) = \prod_i P(\mathbf{r}_i \mid v_i) \tag{2.37}$$

The Equation (2.37) is rewrite in monotone increasing function as:

$$\log P(\mathbf{r} \mid v) = \sum_i \log P(\mathbf{r}_i \mid v_i) \tag{2.38}$$

where log P(r|v) is known as log-likelihood function.

The soft-output MAP decoder calculates the a posteriori log-likelihood ratio for data bit u_t as:

$$\Lambda(u_t) = \log \frac{p\{u_t = 1 \mid \mathbf{r}\}}{p\{u_t = 0 \mid \mathbf{r}\}} \tag{2.39}$$

where $P\{u_t=i|r\}$ is the a posteriori probability (APP) of the data bit u_t.

Then, the decoder makes the hard decision by:

$$\Lambda(u_t) = \begin{cases} 1, & \text{if } \Lambda(u_t) > 0 \\ 0, & \text{otherwise} \end{cases} \tag{2.40}$$

The APPs in Equation (2.39) can be computed from the trellis diagram as:

$$p\{u_t = 0 \mid \mathbf{r}\} = \sum_{(m',m)\in B_t^0} P\{S_{t-1} = m', S_t = m \mid \mathbf{r}\} \tag{2.41}$$

$$p\{u_t = 1 \mid \mathbf{r}\} = \sum_{(m',m)\in B_t^1} P\{S_{t-1} = m', S_t = m \mid \mathbf{r}\} \tag{2.42}$$

where S_{t-1} and S_t are the encoder stages at time (t-1) and t, respectively, and B_t^0 and B_t^1 are set of transitions from state m' to stage m caused by u_t=0 and u_t=1, respectively. (2.40) and (2.41) can be written as

$$p\{u_t = 0 \mid \mathbf{r}\} = \sum_{(m',m)\in B_t^0} \frac{P\{S_{t-1} = m', S_t = m \mid \mathbf{r}\}}{P(\mathbf{r})} \tag{2.43}$$

$$p\{u_t = 1 \mid \mathbf{r}\} = \sum_{(m',m)\in B_t^1} \frac{P\{S_{t-1} = m', S_t = m \mid \mathbf{r}\}}{P(\mathbf{r})} \tag{2.44}$$

where P(r) is a constant.

In order to efficiently calculate the information bits APPs, the following probability functions are defined [59]

$$\alpha_t(m) = P\{S_t = m, \mathbf{r}_1^t\} \tag{2.45}$$

$$\beta_t(m) = P\{\mathbf{r}_{t-1}^N \mid S_t = m\} \tag{2.46}$$

$$\gamma_t^i(m', m) = P\{u_t = i, S_t = m, \mathbf{r}_t \mid S_{t-1} = m'\} \tag{2.47}$$

where

$$\mathbf{r}_t = (r_{t,0}, \ldots, r_{t,i}, \ldots, r_{t,n-1}) \tag{2.48}$$

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and $$\mathbf{r}_t^k = (\mathbf{r}_t, ..., \mathbf{r}_{t+1}, ..., \mathbf{r}_k) \tag{2.49}$$

The joint transition probability, $P\{S_{t-1} = m', S_1 = m, \mathbf{r}\}$, can be expressed as

$$P\{S_{t-1} = m', S_1 = m, \mathbf{r}\} = \alpha_{t-1}(m) \sum_{i\in 0,1} \gamma_t^i(m', m)\beta_t(m) \tag{2.50}$$

where $\alpha_t(m)$ and $\beta_t(m)$ are obtained recursively as

$$\alpha_t(m) = \sum_{m'} \alpha_{t-1}(m) \sum_{i\in 0,1} \gamma_t^i(m', m) \tag{2.51}$$

$$\beta_t(m) = \sum_{m'} \beta_{t+1}(m) \sum_{i\in 0,1} \gamma_t^i(m', m) \tag{2.52}$$

and $\gamma_t^i(m', m)$ is a channel transition probability weighted by the information bit a priori probability $p_t(u_t = i), i = 0,1$ where u_t is the information symbol associated with transition $S_{t-1} = m' \rightarrow S_1 = m$. Coefficient $\gamma_t^i(m', m)$ can be written as

$$\gamma_t^i(m', m) = p_t(u_t = i) \prod_{j=0}^{j=n-1} P\{r_{t,j} \mid x_{t,j}\} \tag{2.53}$$

$$P\{r_{t,j} \mid x_{t,j}\} = \frac{1}{\sqrt{2\pi\sigma}} e^{\frac{(r_{t,j}-x_{t,j})^2}{2\sigma^2}} \tag{2.54}$$

where $x_{t,j}$, $j = 0, ..., n-1$, is a binary phase shift keying (BPSK) modulated symbol in the codeword associated with transition $S_{t-1} = m' \rightarrow S_1 = m$.

If we assume that the encoder starts and ends in a zero stage, the boundary conditions are

$$\alpha_0(0) = 1,\ \alpha_0(m) = 0,\ \text{for } m \neq 0 \tag{2.55}$$

$$\beta_N(0) = 1,\ \beta_N(m) = 0,\ \text{for } m \neq 0 \tag{2.56}$$

The log-likelihood ration now can be written as

$$\Lambda(u_t) = \log \frac{\sum_{(m',m)\in B^1} \alpha_{t-1}(m')\gamma_t^1(m', m)\beta_t(m)}{\sum_{(m',m)\in B^0} \alpha_{t-1}(m')\gamma_t^0(m', m)\beta_t(m)} \tag{2.57}$$

The above algorithm is usually referred to as a forward/backward algorithm, since the coefficients $\alpha_t(m)$ are calculated recursively starting from the beginning of the trellis, and the $\beta_t(m)$ coefficients are calculated recursively starting from the end of the trellis.

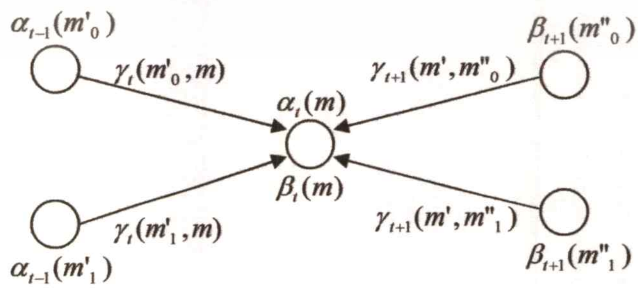

Fig. 2.15 Graphical representation of the forward and backwards recursions.

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Figure 2.15 shows the graphical representation of the forward and backward recursion. In this figure, $\alpha_{t-1}(m'_t)$ represents α coefficients for state m'_t in the (*t*-1)th stage which is connected with stage m in *t*-th trellis stage and where the transition $s_{t-1}=m' \rightarrow S_1 = m$ is caused by the information bit $u_t = i, i = 0,1$. Similarly, the $\beta_{t+1}(m")$ denotes β coefficient for stage $m"$ in the (*t*-1)th trellis stage which is connected with stage m in *t*-th trellis stage and where the transition $s_{t-1}=m' \rightarrow S_1 = m$ is caused by the information bit $u_t = i, i = 0,1$. The a posteriori probabilities of the information bits can be calculates as

$$P\{u_t = 1 | \mathbf{r}\} = \frac{e^{\Lambda(u_t)}}{1+e^{\Lambda(u_t)}} \tag{2.58}$$

$$P\{u_t = 0 | \mathbf{r}\} = \frac{1}{1+e^{\Lambda(u_t)}} \tag{2.59}$$

The a posteriori probabilities of the transmitted bits can be calculated by adding the probabilities of the codeword that contain a particular transmitted, are denoted as

$$P\{x_{t,j} = 1 | \mathbf{r}\} = \sum_{u_t=i, x_{t,j}=1} P\{u_t = i | \mathbf{r}\} \tag{2.60}$$

$$P\{x_{t,j} = -1 | \mathbf{r}\} = \sum_{u_t=i, x_{t,j}=-1} P\{u_t = i | \mathbf{r}\}$$

2.4 Multiuser Detection Techniques

Over the past twenty years, a number of sophisticated receiver designs, for example, multiuser detection, have been proposed for interference suppression under various settings. Generally, multiuser detection techniques are applied to joint detection of different signals transmitted over MIMO channel. Multiuser detection [62-65] was first proposed for CDMA systems rather than TDMA systems. The simples approach to demodulate CDMA signals is the single-user matched filter (MF), which was adopted in CDMA receiver. However, conventional single-user receivers are designed to operate in the fading channel without CCI and MAI. Hence, they do not work well in a multiuser system. The studies of multiuser detection techniques for CDMA system was started with the work of verdu [66]. For the AWGN channel, Verdu presented an optimum multiuser receiver based on maximum likelihood (ML) or maximum a posteriori probability (MAP) algorithm. This optimum technique however requires a prior knowledge of the signal amplitudes and phases and involves a high degree of computational complexity. Therefore, some suboptimum multiuser detection, named as linear detection which included zero-forcing (ZF) and minimum mean square error (MMSE) detection have been investigated. They can suppress the interference by means of linear processing in much lower complexity. Other sub-optimal

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schemes were proposed in hopes of reducing complexity while maintaining good performance, are referred as decision-feedback detection such as successive interference cancellation (SIC) and parallel interference cancellation (PIC). However, in designing a multiuser detector for MIMO channels, the knowledge of MAI characteristics such as the spreading sequences, the availability of the CSI at the receiver, must be jointly take into account in order to mitigate ISI, CCI, and MAI. Another successful technique, named adaptive detector, was introduced, based on an adaptive signal processing scheme such as recursive solutions. The advantages of using adaptive signal processing techniques are its capacity to track the channel variation without prior knowledge of CSI and its simplicity and robustness to the signal.

2.4.1 Asynchronous CDMA Systems

In a *K*-user synchronous CDMA system, the received signal consists of the sum of all users' spread signals embedded in additive white Gaussian noise (AWGN) as shown in Figure 2.16.

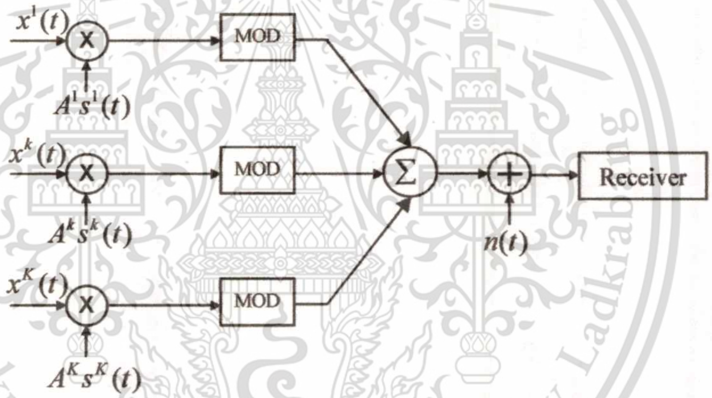

Fig. 2.16 DS-CDMA model.

The received signal is represented as:

$$r(t) = \sum_{k=1}^{K} s^k(t) A^k x^k + \sigma n(t) \ , \qquad t = [0, T] \tag{2.61}$$

where *T* is the data symbol period. $s^k(t)$ is the signature waveform assigned to the *k*-th user as:

$$s^k(t) = \sum_{n=1}^{N} s^k(n) q(t-(n-1)T_c) \tag{2.62}$$

where: $s^k(n)$ denotes the *n*-th chip value of s^k

$q(t)$ is the chip waveform

T_c is the chip interval.

Then, $s^k(t)$ is normalized to have unit energy as given by:

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$$\|s^k\|^2 = \int_0^T s^{k^2}(t)dt = 1 \tag{2.63}$$

The signature waveforms are assumed to be zero outside the interval [0,T], and hence there is no inter-symbol interference (ISI). A^k is the received amplitude of the k-th user's signal. A^{k^2} is referred to as the energy of the k-th user. $x^k \in [-1,+1]$ is the symbol transmitted by the k-th user and n(t) is additive white Gaussian noise with zero mean and variance σ^2.

Equation (2.61) can be written in a discrete-time form as the following:

$$\mathbf{r} = \mathbf{SAx} + \sigma\mathbf{n} \tag{2.64}$$

where

$$\mathbf{r} = [r(1), r(2), ..., r(N)]^T \tag{2.65}$$

and

$$S = \begin{bmatrix} s^1(1) & s^2(1) & \dots & s^K(1) \\ s^1(2) & s^2(2) & \dots & s^K(2) \\ \vdots & \vdots & \ddots & \vdots \\ s^1(N) & s^2(N) & \dots & s^K(N) \end{bmatrix} \tag{2.66}$$

$$\mathbf{A} = diag[A^1, A^2, ..., A^K] \tag{2.67}$$

$$\mathbf{b} = [b^1, b^2, ..., b^K]^T \tag{2.68}$$

$$\mathbf{n} = [n(1), n(2), ..., n(N)]^T \tag{2.69}$$

where N is number of data frame length.

2.4.3.2 Maximum A Posteriori Probability Detection

The Maximum a Posteriori (MAP) detector selects:

$$\hat{\mathbf{x}} = \arg\max_{\mathbf{b}} \Pr(\mathbf{x} \text{ transmitted} \mid \mathbf{r} \text{ received}) \tag{2.70}$$

which minimizes the probability of error. This is the same as the ML estimate if the symbols are likely equal. However, when combined with error control coding and iterative soft decoding, the decoder can pass reliability information to the multiuser detector in the form of the a priori distribution or likelihood ratio for each transmitted symbol. Hence, in that scenario the MAP estimate generally differs from the ML estimate. Furthermore, the MAP detector itself computes soft estimates of each symbol, although the final (hard) estimates are obtained from the soft estimates by thresholding. If the receiver detects a subset of the vector **x**, then the MAP estimate maximizes the corresponding marginal distribution.

The MAP detector suffers from the same drawbacks as the ML detector, namely, the complexity grows exponentially with the size of **x**, and it requires knowledge of **H**. However, in some applications where the system size is relatively small, the complexity may be manageable.

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2.4.4 Interference Cancellation Detections

Besides the linear detection schemes, researchers have also proposed non-linear detectors that use the interferers' data to detect that of the desired user. Interference cancellation (IC) [62] detectors employ temporary data estimates to reconstruct the interference, and then subtract it from the received signal. The IC detector includes two classes: successive interference cancellation (SIC) [63-65] and parallel interference cancellation (PIC) [66-69].

2.4.4.1 Successive Interference Cancellation Detection

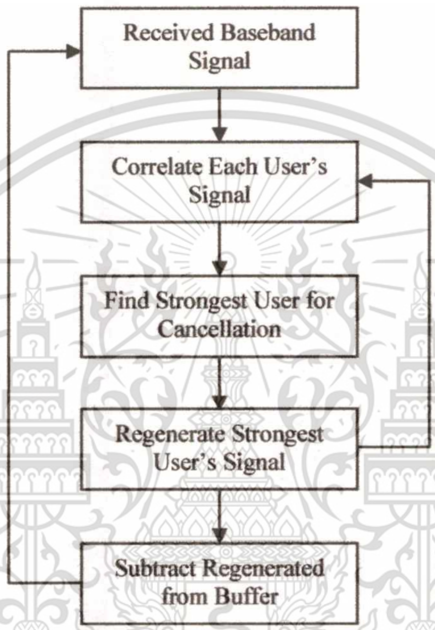

Fig. 2.19 Algorithm for successive interference cancellation.

The successive interference cancellation scheme uses the algorithm shown in Figure 2.19. At each iteration of the scheme, all the user's signals are estimated. The signal with the largest power is then regenerated and subtracted from the buffered received signal. The remaining signals are now re-estimated and a new largest user is selected. The process will continue until all the users' signals have been recovered or the maximum allowable number of cancellations is reached.

Successive interference cancellation has been shown to be very robust to imperfect power control in a DS-CDMA system. This comes from the strongest users (and thus best estimated) all having been cancelled from the received waveform. Successive interference cancellation is considered one of the simplest forms of interference cancellation because of the single stage of cancellation. However, the processor performing the cancellation must perform all the cancellations while maintaining the

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necessary data rate. As show in [70], this class of detectors has the almost paradoxical feature that interference cancellation can be performed most reliably when the interference is strong relative to the desired signal, i.e. when there is a significant power difference between each of the users signals. However, its performance is poor when power levels are approximately equal. At the same time, the signals need to be sorted by power correctly, and signal reordering is required whenever the power profile changes. This will be a particular risk in a high capacity system with widely variable power levels. Moreover, serial cancellation one by one will lead to a relatively long processing delay.

2.4.4.2 Parallel Interference Cancellation Detection

Parallel interference cancellation detection simultaneously removes from each user the interference produced by the remaining users. In this detector, each user in the system receives equal treatment in the attempt to cancel its MAI. Compared with the SIC, since the IC is performed in parallel for all users, the delay required to complete the process is dramatically reduced. Figure 2.24 illustrates the structure of the PIC detector in a two-user scenario. The PIC performs the estimation for all users' signals separately at first. Then the signal decisions are used to reconstruct the MAI, which is subtracted later. It's clear that the parallel cancellation process should be divided into at least two stages as shown in figure 2.20 and 2.21.

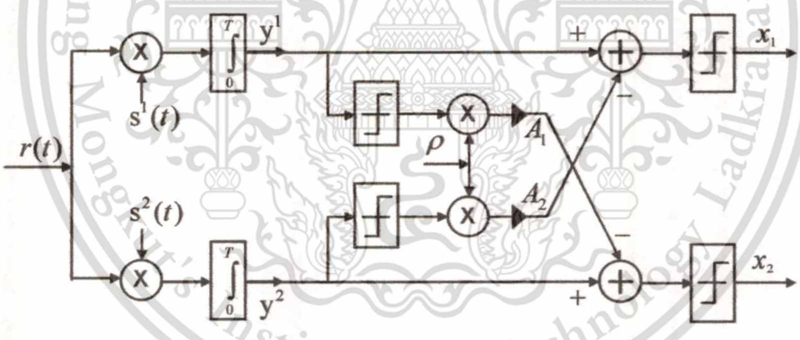

Fig. 2.20 Parallel Interference Cancellation for CDMA systems.

In the first stage which can be provided by the MF or linear detectors, the signal from each users are demodulated. Then, the data estimate of the kth user is given by:

$$\tilde{x}_k = \mathrm{sgn}(y^k_{MF}) \qquad (2.71)$$

where sgn(.) is the sign of a real number.

For the desired user k, remodulating the signal of user j with the signature wave, we get the MAI construction $s^j(t)s^k(t)A_j\tilde{x}_j$, which is subtracted from the received signal:

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$$y_{PIC}^{k}(t) = y_{MF}^{k}(t) - \sum_{j \neq k} \rho_{jk}(t) A_j \tilde{x}_j^{MF}$$
$$= A_k x_k - \sum_{j \neq k} \rho_{jk}(t) A_j (x_j - \tilde{x}_j) + n'(t) \quad (2.72)$$

Assuming that a hard decision is employed, the output of the two-stage PIC detector is:

$$\tilde{x}_{PIC}^{k} = \text{sgn}(y_{PIC}^{k})$$
$$= \text{sgn}(A_k x_k - \sum_{j \neq k} \rho_{jk}(t) A_j (x_j - \tilde{x}_j) + n'(t)) \quad (2.73)$$

The PIC approach can be further iterated by replacing the conventional MF outputs with the increasingly reliable temporary decisions of the interfering users. It is made by PIC detector in the last iteration as shown in Figure 2.21.

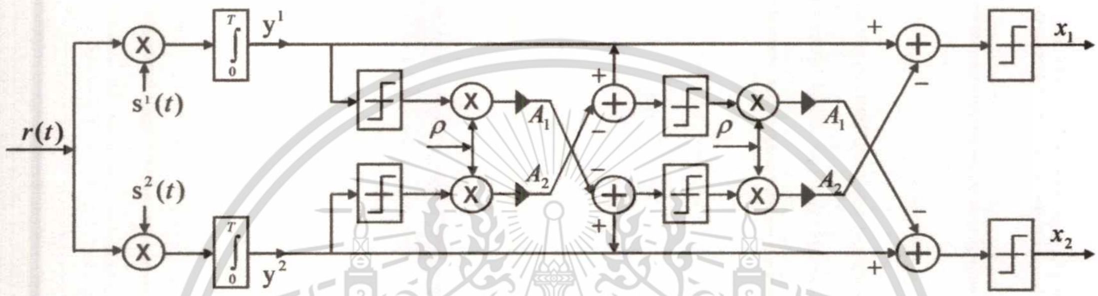

Fig. 2.21 Iterative PIC detector.

Then, the received signal is rewritten as:

$$y_{PIC}^{k,l}(t) = y_{MF}^{k}(t) - \sum_{j \neq k} \rho_{jk}(t) A_j \tilde{x}_{PIC}^{j,l-1} \quad (2.74)$$

where $y_{PIC}^{k,l}$ is the l-th iteration PIC detection result and $\tilde{x}_{PIC}^{j,l-1}$ is the data decision made by the PIC detection in the (l-1) th iteration. Moreover, the iteration can be performed as many times as needed. In general, the performance of the detector should continuously improve with increasing number of iterations. However, the disadvantage of PIC is that the risk of error propagation still exists. From Equation (2.80), the MAI can be fully removed if the temporary decision is correct. On the other hand, if the temporary decision is wrong, the error will be doubled after the PIC.

2.4.5 Adaptive Detections Recursive Least Squares Algorithm

Another alternative detection technique is an adaptive signal processing scheme, based on adaptive filters and adaptive algorithm. The advantages of using adaptive detection are its capacity to track the channel variations without prior knowledge of channel state information and its simplicity and robustness to the signal. The most common adaptive filters, which are used during the adaptation process, are the finite impulse response filters (FIR) types. These are preferable because they are stable,

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and no special adjustments are needed for their implementation. The adaptation algorithms, which will be introduced in this thesis is Recursive Least Squares (RLS) Algorithm.

RLS algorithm has higher computational requirement than LMS, but behaves much better in terms of steady state MSE and transient time [72]. The RLS algorithm is based on the Least Squares estimate of the filter coefficients **w**(n-1) at iteration (n-1), by computing its estimate at iteration n using the newly arrive data [72]. Figure 2.23 presents the general RLS block diagram.

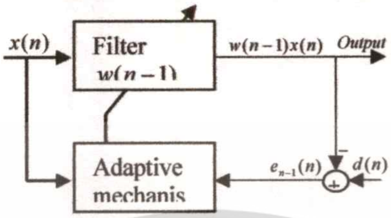

Fig. 2.23 RLS block diagram algorithm.

According to the exponentially weight least squares solution, the means square error function in Equation (2.95) is rewrite with the criterion of exponential forgetting factor as:

$$\zeta(n) = \sum_{i=i_1}^{n} \lambda^{n-i}[e(i)^2] = \sum_{i=i_1}^{n} \lambda^{n-i}[d(i) - \sum_{k=0}^{M-1} w_k(n) r(i-k)]^2 \quad (2.87)$$

Make the following variable change:

$$\left.\begin{aligned} r'(i) &= \sqrt{\lambda^{n-i}}\, r(i) \\ d'(i) &= \sqrt{\lambda^{n-i}}\, d(i) \end{aligned}\right\} \quad (2.88)$$

Then, (2.100) is rewritten as:

$$\zeta(n) = \sum_{i=i_1}^{n} [d'(i) - \sum_{k=0}^{M-1} w_k(n) r'(i-k)]^2 \quad (2.89)$$

So, the LS solution can be obtained as:

$$w(n) = [\Phi(n)]^{-1} z(n) \quad (2.90)$$

where

$$\Phi(n) = \sum_{i=1}^{n} \lambda^{i-1} r(i) r^T(i) \quad (2.91)$$

$$z(n) = \sum_{i=1}^{n} \lambda^{i-1} r(i) d(i) \quad (2.92)$$

Note that the data before (i=0) is zero. By using the information already available at time (n-1), (2.106) is defined as:

$$w(n-1) = [\Phi(n-1)]^{-1} z(n-1) \quad (2.93)$$

Rewrite the variable $\Phi(n)$ and $z(n)$ as the functions of $z(n-1)$ and $\Phi(n-1)$:

$$\Phi(n) = \lambda\Phi(n-1) + r(n) r^T(n) \quad (2.94)$$

$$z(n) = \lambda z(n-1) + r(n) d(n) \quad (2.95)$$

By using the matrix inversion formula [71], we have:

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$$A = B^{-1} + CD^{-1}C^T \tag{2.96}$$

then,
$$A^{-1} = B - BC(D + C^T BC)^{-1} C^T B \tag{2.97}$$

where,

$$\left.\begin{aligned} A &= \Phi(n) \\ B^{-1} &= \lambda\Phi(n-1) \\ C &= r(n) \\ D &= 1 \end{aligned}\right\} \tag{2.98}$$

By applying the matrix inversion formula to (2.107), we obtain:

$$\Phi^{-1}(n) = \lambda^{-1}\Phi^{-1}(n-1) - \frac{\lambda^{-2}\Phi^{-1}(n-1)r(n)r^T(n)\Phi^{-1}(n-1)}{1 + \lambda^{-1}r^T(n)\Phi^{-1}(n-1)r(n)} \tag{2.99}$$

Denoting:

$$U(n) = \Phi^{-1}(n) = \lambda^{-1}U(n-1) - \lambda^{-1}g(n)r^T(n)U(n-1) \tag{2.100}$$

$$g(n) = \frac{\lambda^{-1}U(n-1)r(n)}{1 + \lambda^{-1}r^T(n)U(n-1)r(n)} \tag{2.101}$$

$$e(n) = d(n) - r^T(n)w(n-1) \tag{2.102}$$

Therefore, the tap coefficient **w**(n) can be recursively calculated by:

$$w(n) = w(n-1) + g(n)e(n) \tag{2.103}$$

where $w(0) = 0$ and $U(0) = \delta I$.

Necessary basic knowledge to be used as tools and the simulation architecture has been presented in this chapter. Next chapter presents the research methodology linking knowledge in this chapter with the findings from computer simulations.

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Chapter 3

Research Methodology

In this chapter, the methodology of this research is presented. Layered space time coding (LSTC) of a Direct Sequence Code Division Multiple Access (DS/CDMA) system was proposed to simulate chaotic sequences as spreading code. The previous research of LSTC DS/CDMA using adaptive iterative receiver RLS presented in [73] is used to compare the performance of chaotic and gold as spreading codes. The adaptive iterative receivers have ability to reduce the co-channel interference (CCI) and multiple access interference (MAI) using the interference suppression and cancellation techniques. In this research, the performance of chaotic spreading code is simulated using LSTC DS/CDMA transmitters and receiver structures.

3.1 System Model

A downlink CDMA system as depicted in Figure 3.1 is considered in this study with no a priori knowledge of Channel State Information (CSI), spreading sequences and fading coefficients except the training sequence for each user. Data from different users is modulated by different signature waveforms before being transmitted asynchronously through a wireless channel [73], which is modeled as multipath Rayleigh fading. Then, the entire signals are combined with noises through Addictive White Gaussian Noise (AWGN) channels. At the receiver, the user data is retrieved adaptively by using the proposed adaptive iterative detector.

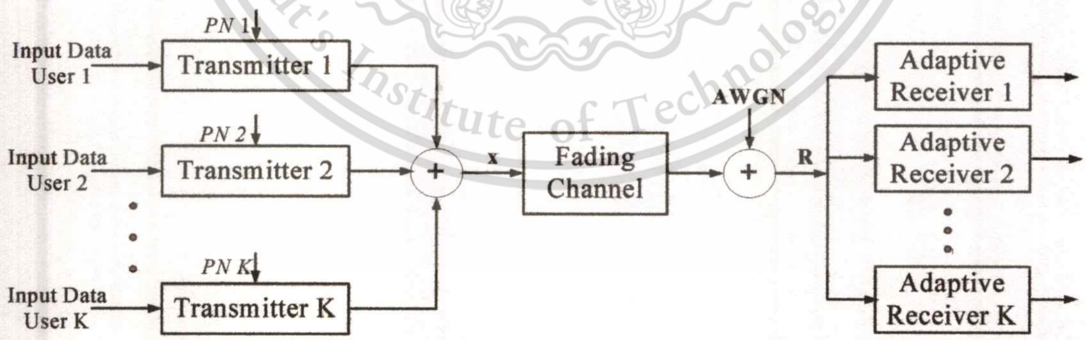

Fig. 3.1 An adaptive CDMA system model

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3.2 Transmitter Structure

The layered space-time (LST) architecture is used in the proposed transmitter structure. In LST coding algorithm, *N* data streams are transmitted simultaneously over *N* antennas over the same frequency band. The receiver uses *M* received antennas and interference canceling/suppression techniques to minimize interference caused by simultaneous transmission from *N* transmit antennas.

There are a number of various LST architectures, depending on whether error control coding is used or not and on the way the modulated symbols are assigned to the transmit antennas. The proposed LSTC-CDMA transmitter structure is described in detail as follows.

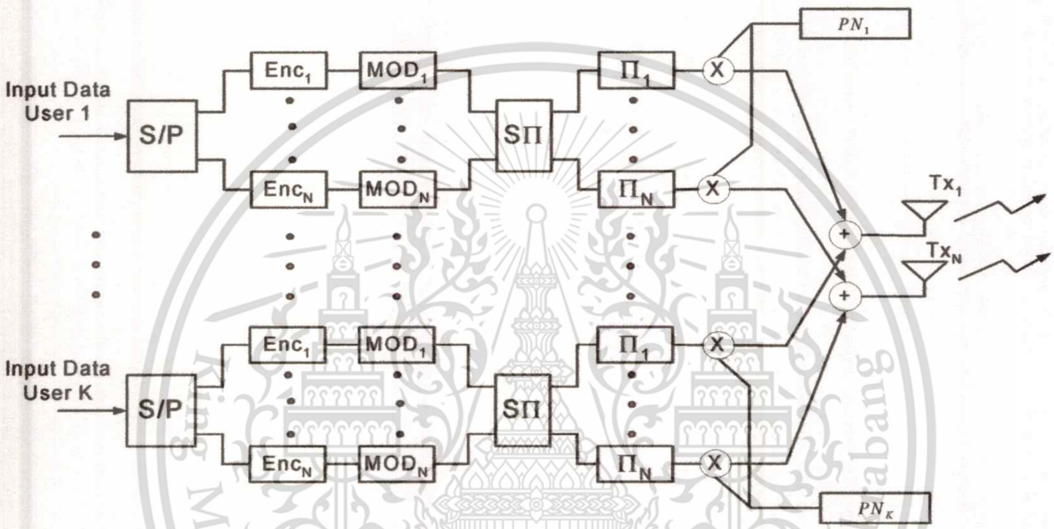

Fig. 3.2 Block diagram of LSTC-CDMA transmitter structure

In a downlink LSTC-CDMA system, all user signals are transmitted simultaneously with *N* transmitters and *M* receiver antennas. Figure 3.2 shows the LSTC-CDMA transmitter structure with *K* users. The binary information of each user is transmitted at a data rate of $r_b = 1/T_b$, where T_b is the bit interval. This data information is first converted into layered data information streams by a serial to parallel (S/P) converter. Before modulating, each information stream is encoded by a convolutional encoder to produce coded data stream for each layer. The layered coded data streams are then fed into a spatial time interleaver (SΠ) and time interleaver (Π). Next, the coded data streams are spread by its signature sequence using a spreading gain of *L* with $L = T/T_c$, where *T* is the symbol interval, and T_c is chip duration of spreading sequences. Finally, the spread symbols of all users are then combined together and simultaneously transmitted through *N* transmit antennas.

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Let's $\mathbf{b}$ is the coded signal vectors transmitted by K users through N transmit antennas and is defined by:

$$\mathbf{b} = [\mathbf{b}_1, \mathbf{b}_2, ..., \mathbf{b}_p, ..., \mathbf{b}_k]^{\mathrm{T}} \tag{3.1}$$

where
$$\mathbf{b}_p = [b_p^1, ..., b_p^n, ..., b_p^N], \tag{3.2}$$

and b_p^n is the information bit of the p-th user for n-th transmit antenna with n=1,..,N and p=1,..,K. Let $\mathbf{S}$ represents the L x KN spread transmitted sequences of K users for n transmit antennas, as given by:

$$\mathbf{S} = [\mathbf{s}_1^1, \ldots, \mathbf{s}_1^N, \ldots, \mathbf{s}_p^n \ldots \mathbf{s}_K^1, \ldots \mathbf{s}_K^N] \tag{3.3}$$

where
$$\mathbf{s}_p^n = [s_p^{n,1}, \ldots, s_p^{n,q}, \ldots, s_p^{n,L}]^T, \tag{3.4}$$

and $s_p^{n,q}$ is the q-th chip of a spreading sequence for the p-th user and n-th transmit antenna with n =1,…, N; p=1,..,K; and q=1,..,L.

Let's $\mathbf{r}_{t,j}^p$ is the received signal vector for the p-th user at the receiver antenna j , $j = 1, \ldots, M$, for symbol t, is represented by:

$$\mathbf{r}_{t,j}^p = \mathbf{S}\mathbf{H}_{t,j}^p \mathbf{b} + \mathbf{n}_{t,j}^p \tag{3.5}$$

where
$$\mathbf{r}_{t,j}^p = [r_{t,j}^{p,1}, \ldots r_{t,j}^{p,q}, \ldots, r_{t,j}^{p,L}]^T, \tag{3.6}$$

and $r_{t,j}^{p,q}$ is the received signal for the p-th user at the q-th chip of the t-th symbol for j-th antenna.

The received signals for all receive antennas of the p-th user are given by

$$\mathbf{R}_t^p = [\mathbf{r}_{t,1}^p, \ldots, \mathbf{r}_{t,j}^p, \ldots, \mathbf{r}_{t,M}^p]^T, \tag{3.7}$$

$\mathbf{H}_{t,j}^p$ is defined by:

$$\mathbf{H}_{t,j}^p = diag(\mathbf{h}_{t,j}^p, ..., \mathbf{h}_{t,j}^p)_{KNxKN} \tag{3.8}$$

where
$$\mathbf{h}_{t,j}^p = diag\,[h_{j,1}^p(t), ..., h_{j,n}^p(t), ..., h_{j,N}^p(t)]_{NxN}, \tag{3.9}$$

and $h_{j,n}^p(t)$ represents the fading coefficient from j-th receive antenna to n-th transmit antenna of the p-th user.

The $\mathbf{n}_{t,j}^p$ is defined as an $L\times1$ noise vector at the receive antenna j of the p-th user, given by:

$$\mathbf{n}_{t,j}^p = [n_{j,1}^p(t), \ldots, n_{j,q}^p(t), \ldots, n_{j,L}^p(t)]^T, \tag{3.10}$$

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where $n_{j,q}^{p}(t)$ is a Gaussian random variable with a zero mean and two sided power spectral density $N_o/2$ per dimension.

3.3 Receiver Structure

In downlink LSTC-CDMA system, there is an assumption that the system has no knowledge of Channel State Information (CSI), spreading sequences, and fading coefficients except the training sequence. A block diagram of the proposed adaptive iterative LSTC-CDMA multiuser receiver for the *p*-th user is shown in Figure 3.3. This structure consists of *K* adaptive iterative LSTC-CDMA single user receivers, each with an adaptive detector followed by *N* parallel soft-input soft-output channel decoders. The received signals are first input to the adaptive detector. The detector outputs are then fed to the time and spatial deinterleavers, denoted by Π^{-1} and $S\Pi^{-1}$, respectively. The deinterleavers outputs are then decoded by a Maximum A Posteriori (MAP) decoder. The estimated soft symbols of MAP decoders from all users are sent to the spatial and time interleavers, and then fed back to the adaptive detector under iterative technique to cancel the interference from adjacent antennas of the user, called CCI, and the interference from other users, called MAI.

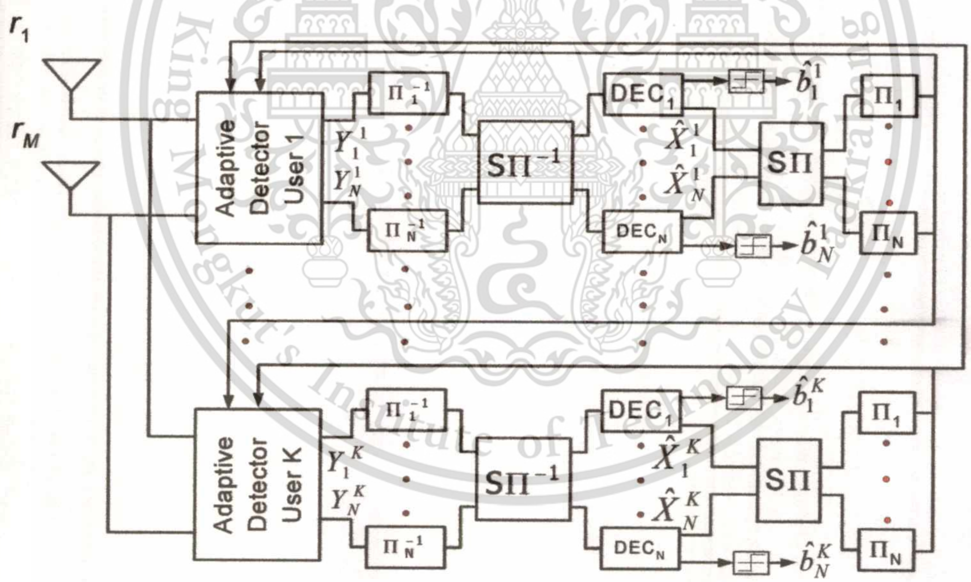

Fig. 3.3 Block diagram of LSTC- CDMA receiver structure

The challenge in detection of space-time signals is to design a low-complexity detector with improving system performance. Hence, detector structures in time domain system are investigated in this proposed receiver design. The recursive least square (RLS) as the main detection algorithms is

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studied. The MAP approaches [74], which minimizes the big error probability at the decoder output, is employed in this structure.

3.3.1 Time Domain Adaptive Iterative LSTC-CDMA receiver

The block diagram of the adaptive iterative LSTC-CDMA receiver for the p-th user in time domain system is shown in Figure 3.4. In the proposed adaptive receiver structure, N sets of adaptive detector consist of M equalizers for the feed-forward filter and an equalizer for the feedback filter modules. An equalizer, employed in the proposed adaptive detector, is based on the RLS algorithms. The detector output for each layer is obtained from combining a feed-forward and a feedback filter output. In the iterative process, the feed-forward filter is compensated for the channel estimation error, and the feedback filter is used to cancel the interference from adjacent antennas and other users. In the first iteration, there are no estimated symbols from the decoders and the feedback filter coefficients are zeros; thus, the feedback filter output is also zero. In the feed-forward filter, the M adaptive equalizers are used to estimate the channel coefficients and signature sequence for each layer of each user. The equalizer outputs from all receive antennas are added to obtain a feed-forward filter output signal for each transmit antenna as shown in Figure 3.4.

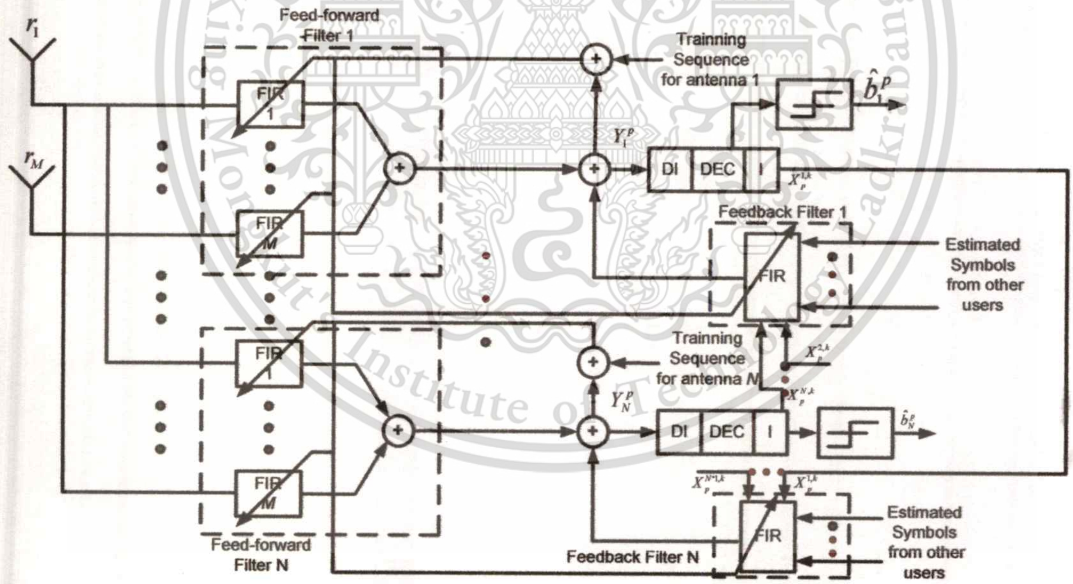

Fig. 3.4 Block diagram of the time domain adaptive iterative LSTC-CDMA receiver

Let $\mathbf{w}_j^{p,k}(t)$ be an Lx1 feed-forward tap coefficients vector for the j-th receive antenna of the p-th user during the k-th iteration at symbol interval t and be given by:

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$$\mathbf{w}_j^{p,k}(t)=[w_{t,j}^{p,k}(0),\dots,w_{t,j}^{p,k}(q),\dots,w_{t,j}^{p,k}(L\text{-}1)]^T \quad (3.11)$$

where $w_{t,j}^{p,k}(q)$ is the feed-forward tap coefficient corresponding to the q-th chip of the spreading sequence.

The $\underline{\hat{\mathbf{x}}}_{t,p}^{i,k}$ is $(KN-1)\times 1$ vector of the estimated soft symbols, at the k-th iteration, from MAP decoders of every antenna of all users, except the i-th antenna of p-th user at symbol interval t, given by:

$$\underline{\hat{\mathbf{x}}}_{t,p}^{i,k} = (\hat{\mathbf{x}}_{t,1}^{1,k},\dots,\hat{\mathbf{x}}_{t,1}^{N,k},\dots,\hat{\mathbf{x}}_{t,p}^{i,k},\dots,\hat{\mathbf{x}}_{t,K}^{1,k},\dots,\hat{\mathbf{x}}_{t,K}^{N,k})^{\mathrm{T}} \quad (3.12)$$

where

$$\hat{\mathbf{x}}_{t,p}^{i,k} = (\hat{\mathbf{x}}_{t,p}^{1,k},\hat{\mathbf{x}}_{t,p}^{2,k},\dots,\hat{\mathbf{x}}_{t,p}^{i-1,k},\hat{\mathbf{x}}_{t,p}^{i+1,k},\dots,\hat{\mathbf{x}}_{t,p}^{N,k}) \quad (3.13)$$

Let $\mathbf{w}_b^{i,k}(t)$ be the feedback filter coefficients of all users, except the i-th antenna of the p-th user, at symbol interval t in time domain, is expressed as:

$$\mathbf{w}_b^{i,k}(t)=[\mathbf{w}_{b,1}^{1,k}(t),\dots,\mathbf{w}_{b,1}^{N,k}(t),\dots,\mathbf{w}_{b,p}^{i,k}(t),\dots,\mathbf{w}_{b,K}^{1,k}(t),\dots,\mathbf{w}_{b,K}^{N,k}(t)]^T \quad (3.14)$$

where

$$\mathbf{w}_{b,p}^{i,k}(t)=[w_{b,p}^{1,k}(t),\dots,w_{b,p}^{i-1,k}(t),w_{b,p}^{i+1,k}(t),\dots,w_{b,p}^{N,k}(t)]\,. \quad (3.15)$$

The detected symbol of the p-th user obtained at the output of the adaptive detector for the i-th antenna during the k-th iteration at symbol interval t, denoted by $y_{t,p}^{i,k}$, is defined as:

$$y_{t,p}^{i,k} = \sum_{j=0}^{M} \mathbf{w}_j^{p,k}(t)^H\,\mathbf{r}_{t,j}^{p} + \mathbf{w}_b^{i,k}(t)^H\,\underline{\hat{\mathbf{x}}}_{t,p}^{i,k}\,. \quad (3.16)$$

The detector soft output $y_{t,p}^{i,k}$ in the time domain is then compared to training symbol, denoted by $\mathbf{x}_{t,p}^{i,k}$. The difference between them is calculated as the detection error. The detection error for the p-th user in the k-th iteration at symbol interval t, for i-th antenna, denoted by $e_{t,p}^{i,k}$, is represented by:

$$e_{t,p}^{i,k}=y_{t,p}^{i,k}\text{-}x_{t,p}^{i,k}\,. \quad (3.17)$$

The detection error is then used to adapt the feed-forward filter and feedback filter tap coefficients in the time domain. After the Mean Square Error (MSE) approaches a specified value, the training mode is switched to the decision directed mode, in which the training sequence is replaced by the hard decision output of each user detector. In the decision directed mode, the detection error is given by the difference between the detector output and the hard decision of the detector output.

The values tap coefficients, $\mathbf{w}_j^{p,k}(t)$ in Equation (3.11) and feedback filter tap coefficients, $\mathbf{w}_b^{i,k}(t)$ in Equation (3.14) are calculated by minimizing the mean square error, defined as ζ, and given by

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$$\zeta = mim\left(E\left|e_{t,p}^{i,k}\right|^2\right) = \min\left(E\left[\left|y_{t,p}^{i,k} - x_{t,p}^{i,k}\right|^2\right]\right). \tag{3.18}$$

Finally, the value of $\mathbf{w}_j^{p,k}(t)$ and $\mathbf{w}_b^{i,k}(t)$ can be calculated recursively by the adaptive detection algorithms RLS, as shown in Section 3.3.2.

3.3.2 Adaptive Detection Algorithms

In the adaptation process of both proposed receiver structure, the adaptive detection algorithms are used in order to compute the value of feed forward and feedback tap coefficients. It is obvious that the RLS adaptive filter has higher computational complexity than the predecessor of adaptive filter such as LMS or PFGLMS adaptive filters.

The adaptive filter Recursive Least Squares algorithm behaves much better in term of steady state Means Square Error (MSE) and faster transient time. So, tap coefficients of the feed forward and the feedback filters, can be determined recursively by Recursive Least Squares (RLS) algorithms as the following Equations (3.19)

Feed-Forward tap coefficients

$$\mathbf{w}_j^{p,k}(t+1) = \mathbf{w}_j^{p,k}(t) + \mathbf{g}_j^{p,k}(t+1)e(t) \tag{3.19}$$

where

$$\mathbf{g}_j^{p,k}(t+1) = \frac{\mathbf{U}_j^{p,k}(t)\mathbf{r}_{t,j}^{p}}{\lambda_f + \mathbf{r}_{t,j}^{p\,T}\mathbf{U}_j^{p,k}(t)\mathbf{r}_{t,j}^{p}}$$
$$\mathbf{U}_j^{p,k}(t) = \lambda_f^{-1}[\mathbf{U}_j^{p,k}(t-1) - \mathbf{g}_j^{p,k}(t)\mathbf{r}_{t,j}^{p\,T}\mathbf{U}_j^{p,k}(t-1)]; \tag{3.20}$$

Feedback tap coefficients

$$\mathbf{w}_b^{p,k}(t+1) = \mathbf{w}_b^{p,k}(t) + \mathbf{g}_b^{p,k}(t+1)e(t) \tag{3.21}$$

where

$$\mathbf{g}_b^{p,k}(t+1) = \frac{\mathbf{U}_b^{p,k}(t)\hat{\underline{\mathbf{x}}}_{t,p}^{i,k}}{\lambda_b + \hat{\underline{\mathbf{x}}}_{t,p}^{i,k\,T}\mathbf{U}_b^{p,k}(t)\hat{\underline{\mathbf{x}}}_{t,p}^{i,k}}$$
$$\mathbf{U}_b^{p,k}(t) = \lambda_b^{-1}[\mathbf{U}_b^{p,k}(t-1) - \mathbf{g}_b^{p,k}(t)\hat{\underline{\mathbf{x}}}_{t,p}^{i,k\,T}\mathbf{U}_b^{p,k}(t-1)] \tag{3.22}$$

where $\mathbf{r}_{t,j}^{p}$ and $\hat{\underline{\mathbf{x}}}_{t,p}^{i,k}$ are defined in Equations (3.6) and (3.12), (λ_f, λ_b) are the forgetting factor of feed-forward and feedback adaptation respectively. The initializations of RLS algorithm are:

$$\mathbf{w}(0) = 0,$$
$$\mathbf{U}(0) = \delta^{-1}I,$$

where δ is the inverse of the input signal power estimate and the recommended value is $\delta > 100\sigma^2$, I is the $(p+1)$ by $(p+1)$ identity matrix and p is the filter order.

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3.3.3 MAP Decoder

In the proposed receiver structure, detector output for the user p, denoted by $y_{t,p}^{i,k}$, is decoded by MAP decoder. The soft-output from the decoder is used to suppress the interference in the feedback filter in the next iteration. The process of this adaptive iterative detection/decoding is performed until the symbol estimation can converge to the optimal performance. The soft-output from the decoder in the last iteration is then fed into a decision device to produce a decision. For a Binary Phase Shift Keying (BPSK), the functions for the transmitted modulated symbols 1 and –1 can be written as:

$$P\left(y_{t,p}^{i,k} \middle| x_{t,p}^{i,k} = \pm 1\right) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp \frac{-\left(y_{t,p}^{i,k} \mp 1\right)^2}{2\sigma^2} . \tag{3.23}$$

The Log-Likelihood Ratios (LLR) is determined in the k-th iteration for the i-th transmit layer of p-th user, denoted by $\lambda_{t,p}^{i,k}$:

$$\lambda_{t,p}^{i,k} = \log\left(\frac{P\left(x_{t,p}^{i,k} = 1 \middle| y_{t,p}^{i,k}\right)}{P\left(x_{t,p}^{i,k} = -1 \middle| y_{t,p}^{i,k}\right)}\right) . \tag{3.24}$$

The symbol A Posteriori Probabilities (APP) $P(x_{t,p}^{i,k} = q| y_{t,p}^{i,k})$, with q equal 1 and -1, conditioned on the output variable which is defined as $\lambda_{t,p}^{i,k}$ for p-th user, can be obtained as:

$$P\left(x_{t,p}^{i,k} = 1 \middle| y_{t,p}^{i,k}\right) = \frac{e^{\lambda_{t,p}^{i,k}}}{e^{\lambda_{t,p}^{i,k}} + 1} \tag{3.25}$$

and

$$P\left(x_{t,p}^{i,k} = -1 \middle| y_{t,p}^{i,k}\right) = \frac{1}{e^{\lambda_{t,p}^{i,k}} + 1} . \tag{3.26}$$

The soft-output symbols are estimated in the i-th layer and k-th iteration for p-th user, calculated as:

$$\hat{x}_{t,p}^{i,k} = \frac{e^{\lambda_{t,p}^{i,k}} - 1}{e^{\lambda_{t,p}^{i,k}} + 1} . \tag{3.27}$$

We have discussed the structure for computer simulation in this chapter. Next chapter provides simulation parameters and data as well as the results.

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Chapter 4

Results and Analysis

This section describes the methodology used in this research. The performance evaluation is the BER performance of a direct sequence code division multiple access (DS/CDMA) system. Figure 3.1 shows a block diagram of the chaotic communication scheme based on the DS/CDMA system. The data information of individual user will be spread employed their own sequences, which generated the chaotic spreading code by designed method in which be emphasized discussion on the next section.

4.1 Simulation Systems and Parameters

The system models of conventional spreading code and chaotic spreading code in layered space-time coding (LSTC) DS/CDMA is described as shown in Table 4.1.

Table 4.1 Simulation systems

Components	Selected Components
Encoder	Convolutional Code
Decoder	MAP
Modulator	BPSK
Spreading Codes	*m*-sequence, Gold sequence, chaotic sequence
Interleaver	Random Interleaving
Channels Model	Rayleigh Fading Channels
Noise Channels	AWGN

In this simulation, the nonsystematic convolutional codes with code rate ½ and memory order of 3 is presented. The proposed system is simulated with 2 transmit and 2 receive antennas with 130 information bits in each frame per layer for each user. Each layer consists of 266 encoded symbols per frame. The data rate is 1 Mb/s at the carrier frequency, f_c, of 2 GH_z. The simulation results are represented in terms of average BER versus the ratio of averaged energy per bit, denoted by *Eb*, to the power spectral density of the AWGN, denoted by *No*.

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4.2 System Performances Results

4.2.1 Correlation properties of the spreading code

4.2.1.1 Auto-correlation

One of the most important characteristics of spreading sequence is its auto-correlation properties. At the receiver, the received signal is mixed with locally generated spreading sequence. The value of the auto-correlation at the receiver side must be maximum or signal strength at the point of synchronization for the best decision to be made on the incoming signal.

Figures 4.1 and 4.2 show the autocorrelation of the 31-bit *m*-sequence and the Gold sequence respectively. Figures 4.3 and 4.4 show the cross-correlation with the length of $L = 31$ chaotic sequence. M-sequences are said to have very good autocorrelation properties because of their balance property, which means that the sequence will have exactly one higher bit than lower bit. This is the closest a sequence consisting of an odd number of 1 s can get to a zero average. The autocorrelation is given as N when the time lag is 0 and given as $-1/N$ at all other times. Gold sequences can be unbalanced and thus causing some spikes in their autocorrelation properties. The table below shows the correlation bounds of the non-zero lags of the autocorrelation. We can notice that the auto-correlation bounds are comparable between the Gold sequence and the chaotic sequence given in Figure 4.3.

Table 4.2 shows the details of the value of auto correlation properties of *m*-sequence, Gold sequences and Chaotic sequences.

Table 4.2 Auto-correlation properties of spreading sequence

	Mean	Max	Bounds
m-sequence	5.22e-4	1	[0, - 0,09]
Gold sequence	0.0259	1	[0.23, - 0,21]
chaotic sequence	5.28e-4	1	[0.21, - 0.24]

The auto-correlation properties in DS/CDMA have the important impact to recover the signal at the receiver. In order to recover the received signal, the code which the signal was spread within the transmitter is reproduced in the receiver and mixed with spreading signal. If the incoming signal and the locally generated PN code are synchronized, the original signal after correlation can be recovered. This correlation to recover the incoming signal is named auto-correlation properties.

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m-sequence

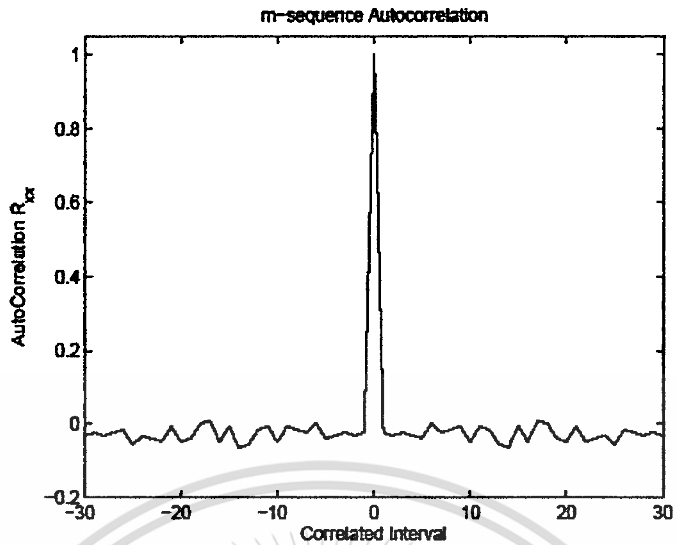

Fig. 4.1 Auto-correlation of *m*-sequences

Gold-sequence

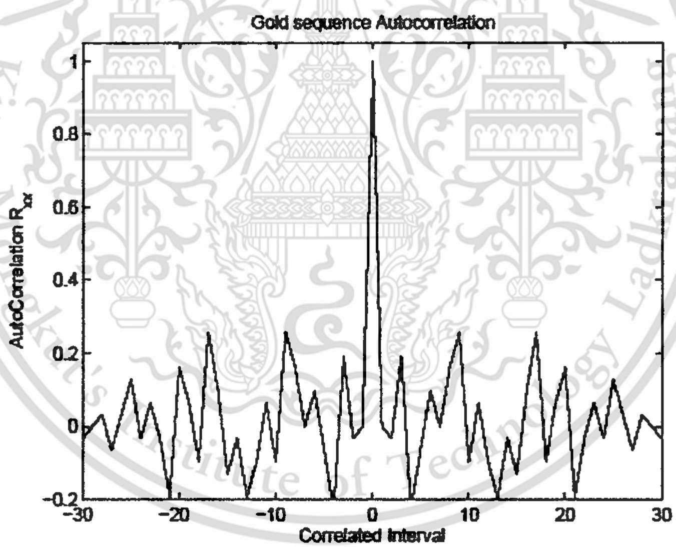

Fig. 4.2 Auto-correlation of Gold sequences

Figures 4.1 and 4.2 present the auto-correlation properties of the conventional spreading sequences. The *m*-sequences perform the better auto-correlation properties than the Gold sequences. Chaotic sequences also show the better performance than the Gold sequence, but compared to the *m*-sequence the correlation properties of chaotic sequences is not as good as the *m*-sequence. Figure 4.3 shows the auto-correlation of chaotic sequences.

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Chaotic sequence

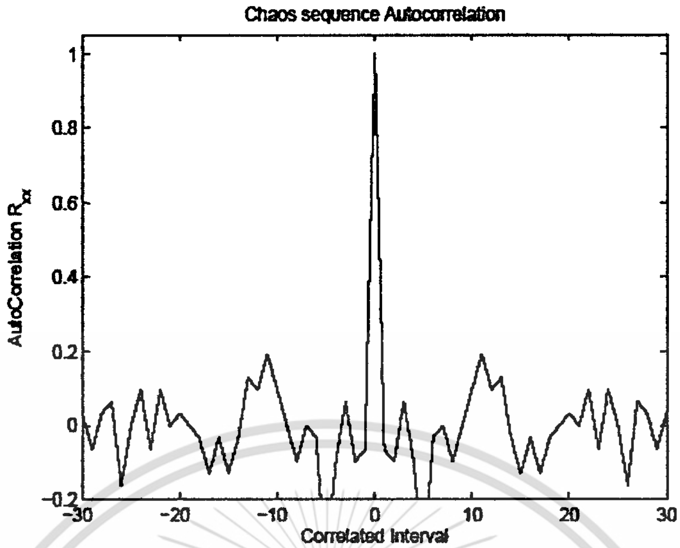

Fig. 4.3 Auto-correlation of Chaotic sequences

4.2.1.2 Cross-correlation

Another desired property of spreading sequence is the minimum cross-correlation. In a spread spectrum system with multiple users, when the received signal is mixed with the locally generated spreading sequence, it must result in minimum signal strength. This would ensure the receiver would be able to differentiate between the transmitted spreading sequence and the other spreading sequence from other users.

Figures 4.4, 4.5, and 4.6 show the cross correlation of the *m*-, Gold and chaotic sequences. We can see from the table below that the cross-correlation bounds are approximately the same for all three sequences. Studies show that m-sequences have very bad cross correlation properties. In this simulation the correlation bounds are about 0.21 and -0.24 for all three sequences. For much longer sequences, Gold codes have much lower cross-correlation than *m*-sequences, and chaotic sequences have been known to achieve much lower cross-correlation bounds than Gold codes.

Table 4.3 shows the details value of cross correlation properties of *m*-sequences, Gold sequences and chaotic sequences. This cross-correlation property is useful to mitigate the interference in multiple access (MAI). As performed in Figures 4.5, 4.6, and 4.7, chaotic sequences have the superior cross-correlation properties than their predecessor spreading codes.

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Table 4.3 Cross-correlation properties of spreading sequence

	Mean	Var	Max	Bounds
m-sequence	-1.22e-18	0.5328	0.2395	[0.21, - 0,24]
Gold sequence	-1.47e-18	0.6294	0.3316	[0.31, - 0,24]
Chaotic sequence	-6.82e-19	0.5907	0.3112	[0.29, - 0.21]

m-sequence

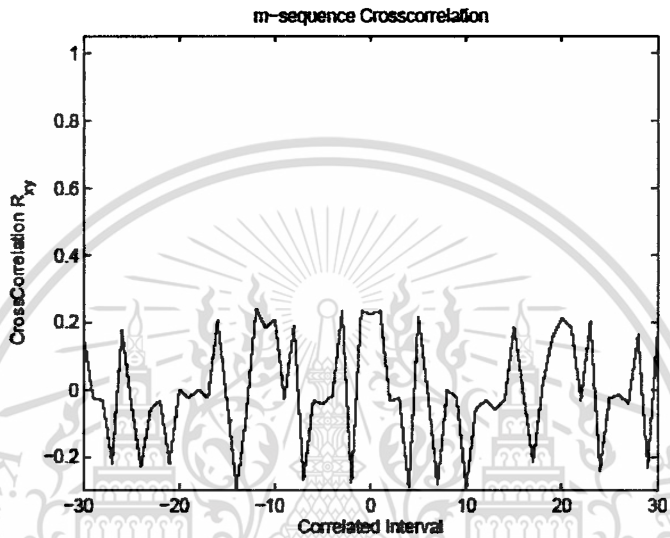

Fig. 4.4 Cross correlation of *m*-sequences

Gold-sequence

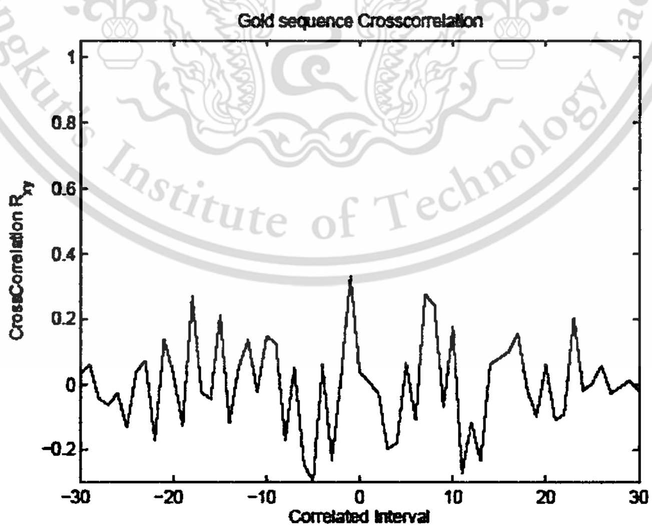

Fig. 4.5 Cross correlation of Gold sequences

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Chaotic sequence

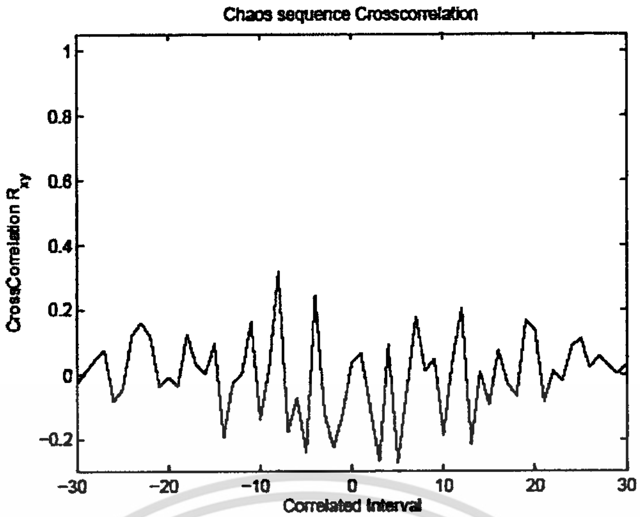

Fig. 4.6 Cross correlation of chaotic sequences

4.2.2 BER Results

The simulation results for the adaptive iterative LSTC DS/CDMA receiver with BPSK modulation in time domain systems are presented in this section. The system operates in the training mode until the mean square error (MSE) approaches the minimum mean square error (MMSE), then it switches to the decision directed mode.

The performance of chaotic and Gold sequences using adaptive iterative recursive least square (RLS) LSTC DS/CDMA receiver is presented. In this simulation, multiuser performance with the length of the spreading sequence is 31 with a given AWGN channel. Figures 4.7, 4.8, and 4.9 show the BER performance of the conventional spreading code and the chaotic spreading code for 2 users. The results show that the chaotic sequences seem to perform the best, followed by the Gold sequences and finally *m*-sequences. Chaotic sequences in one-dimensional map given by the logistic map shows the best performance than other two chaotic ones with higher dimensional map, i.e. the Chebyshev map and the Chen's attractor. As one-dimensional map, the logistic map is the easier and simpler sequences to generate and synchronize. As mentioned in Chapter 2, that the parameter of the initial condition in the chaotic map is the key value to be assigned for each user. In this simulation, the initial condition of each chaotic map is set by value $x_o = 0.5$. For the logistic map, the bifurcation parameter of r is $3.57 < r < 4$ to achieve the perfectly chaotic sequences as non-periodic sequences. Figure 4.7 shows the performance of the logistic map when the value of the bifurcation parameter $r = 3.89$.

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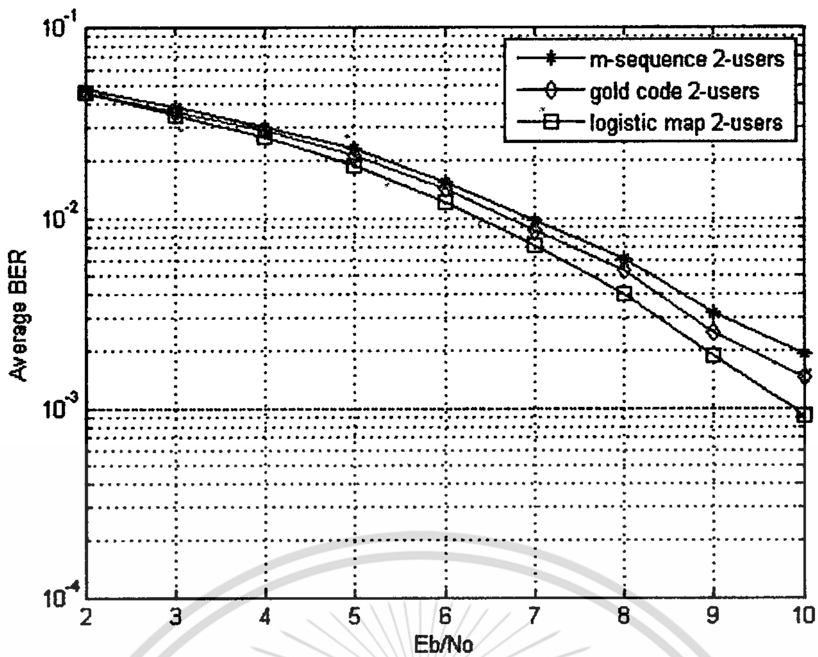

Fig. 4.7 The BER performance comparison using *m*-sequences, Gold sequences and the logistic map for 2 users

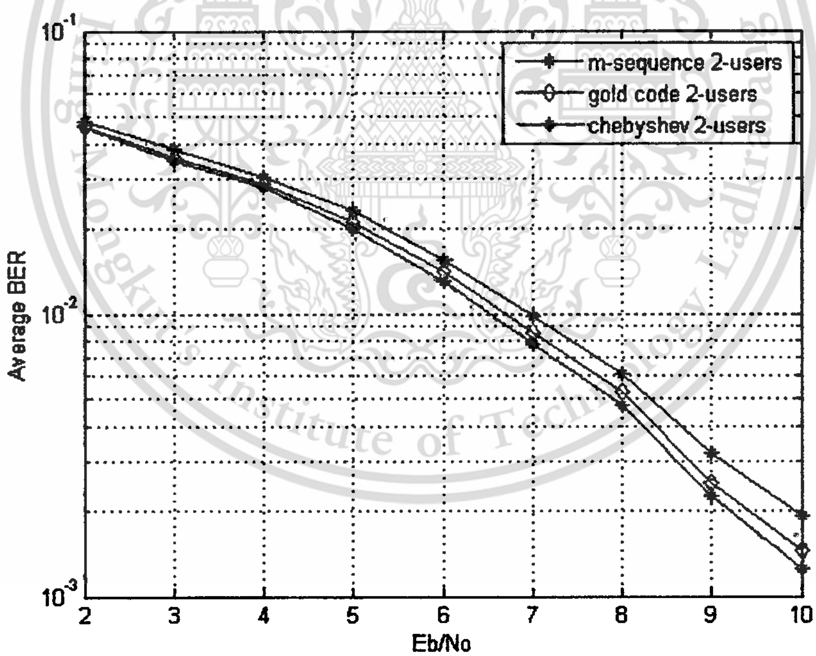

Fig. 4.8 The BER performance comparison for 2 users

There can be seen at the Figure 4.8 that the BER performance of the Chebyshev map is better than Gold sequences and *m*-sequences. For Chebyshev polynomial of degree-*k*, we use initial condition value x_0

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=0.5, and polynomial degree of $k = 2$. While for the Chen's attractor as three-dimensional chaotic map seems not so superior compared to Gold code, but better with *m*-sequence, it seems that Chen's attractor is much better as shown in Figure 4.9. Note that the chaotic parameter used herein for the Chen's attractor is the *x* parameter.

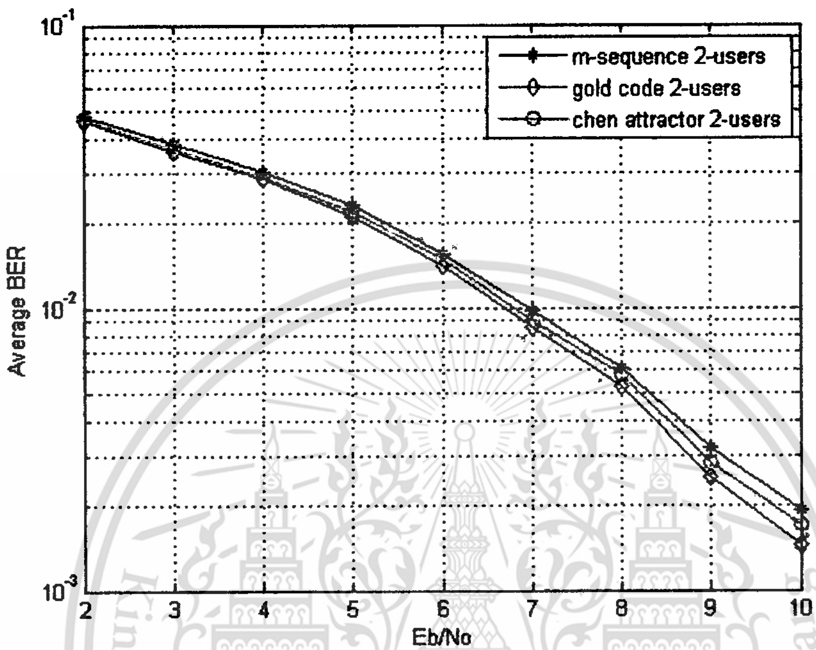

Fig. 4.9 The BER performance comparison for 2 users

Figure 4.10 presents the BER performance comparison of the chaotic sequence when the users are 2 users. Using the same initial condition for each of the chaotic map, i.e the logistic map, the Chebyshev map, and the Chen's attractor map, it is shown that the logistic map has the better performance in term of BER than the higher-dimensional chaotic map, i.e a Chebyshev map as a two-dimensional map and a Chen's attractor as a three-dimensional map.

The performance of chaotic spreading sequences and conventional spreading sequences for 4 users are given in Figures 4.11, 4:12, and 4.13. The results show that the performance of chaotic spreading code and conventional spreading code are degraded when the number of users are changed from 2 to 4 users.

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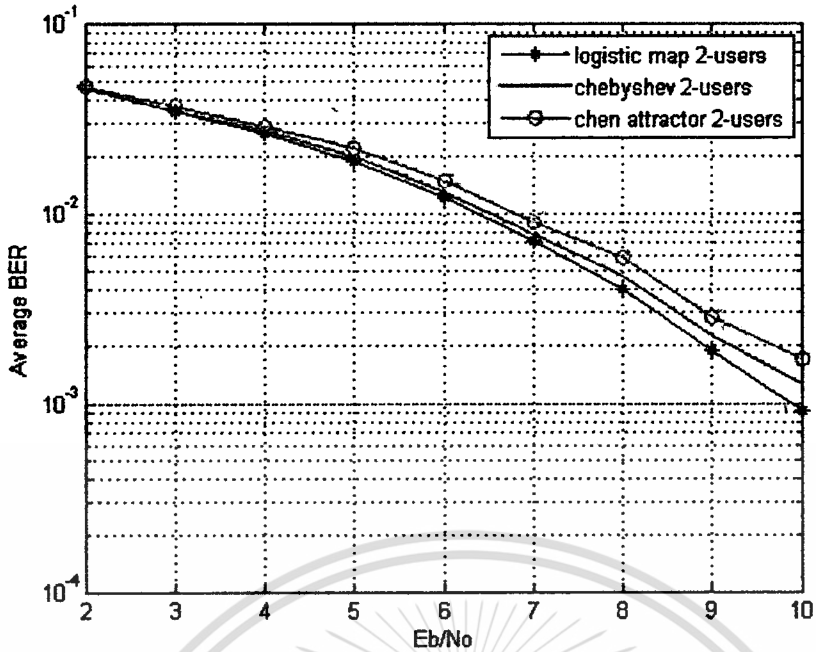

Fig. 4.10 The BER performance comparison for 2 users

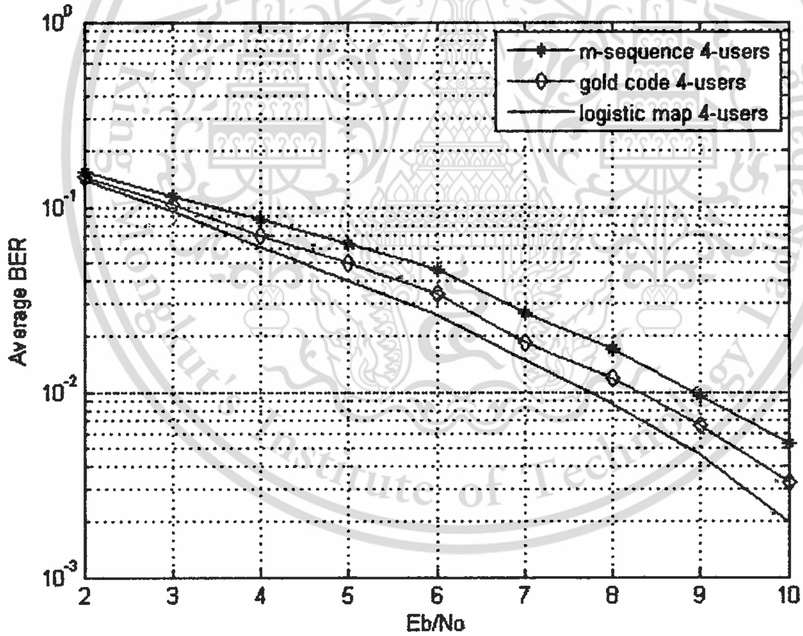

Fig. 4.11 The BER performance comparison using the *m*-sequence, the Gold code and the logistic map for 4 users

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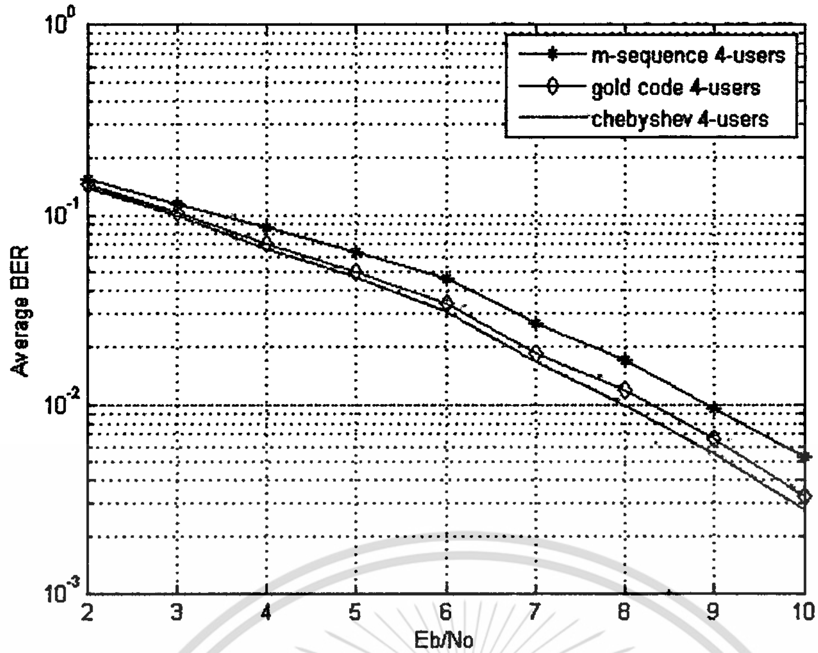

Fig. 4.12 The BER performance comparison for 4 users

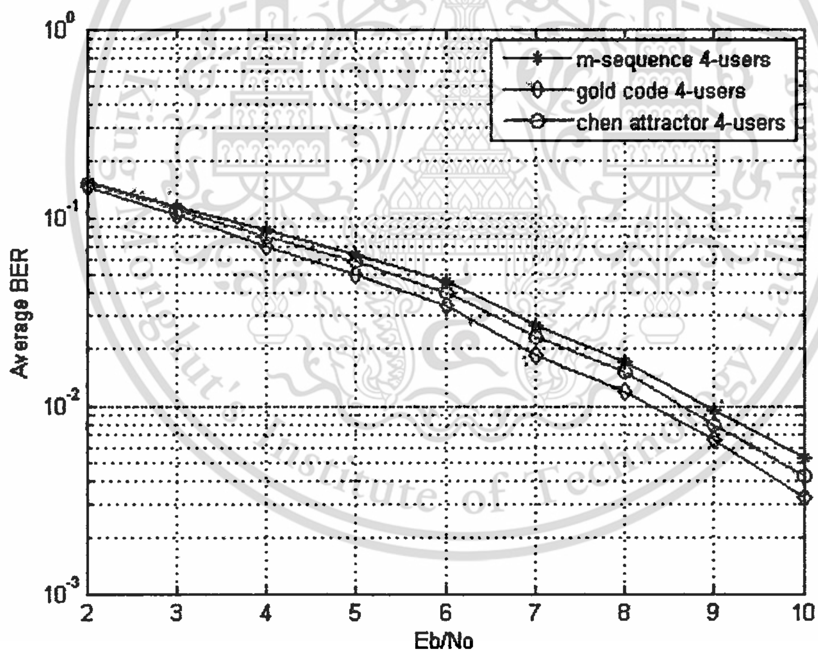

Fig. 4.13 The BER performance comparison for 4 users

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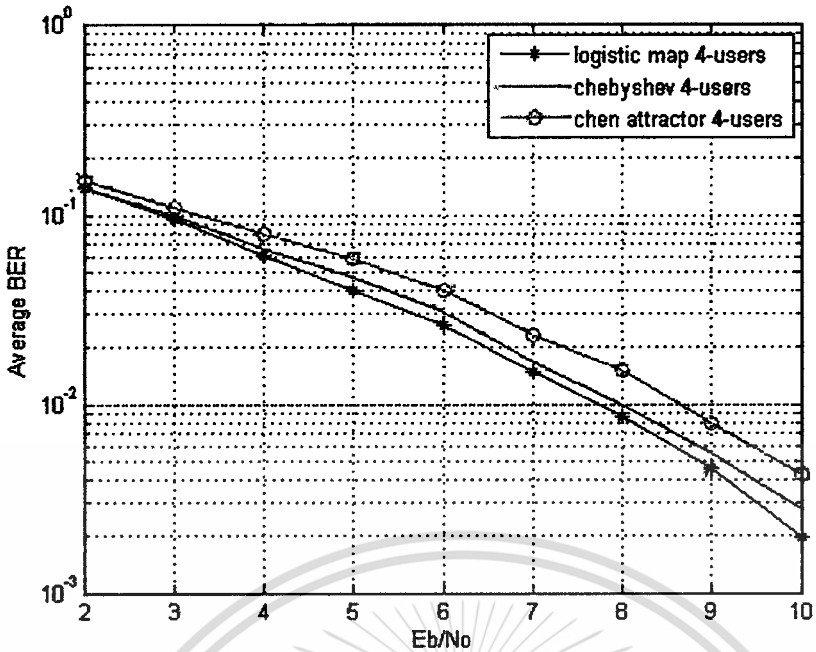

Fig. 4.14 The BER performance comparison for 4 users

In Figure 4.14, BER performance of the logistic map, the Chebyshev map, and the Chen's attractor are plotted. When the number of the users are increased, the performance of the chaotic is not significantly different. The logistic map still performs superior to the conventional code and the higher-dimensional chaotic map.

In this chapter, the performance of using various spreading sequences in the LSTC DS/CDMA adaptive RLS receiver is presented. The BER performance of different chaotic spreading sequences and conventional spreading sequences is evaluated. It can be seen that chaotic sequences based DS/CDMA performs outperforms the conventional sequences. But, in our case, when the higher-dimensional map is applied in this system, the performance is degrades slightly. Remark that the parameter for Chen's attractor used herein is the x parameter. This parameter is less sensitive compared to z parameter. It is recommended to carry out further investigation on other parameters, i.e. y and z before the final justification in term of BER performance.

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Chapter 5

Conclusions

5.1 Conclusions

The objective of this thesis is to investigate chaotic sequences as spreading codes in the layer space time coding (LSTC) direct sequence code division multiple access (DS/CDMA) system comparing with the conventional spreading code, the Gold and the m-sequence spreading code. The chaotic sequence in one-dimensional map, two-dimensional map, and three-dimensional map are proposed. The performance comparison of bit error rate (BER) performance in LSTC DS/CDMA communication system is carried out.

As the number of users in wireless communication system increased, the demand of spreading sequence in CDMA system is also increases. Since the technique in spread spectrum system employed a unique spreading code to be assigned to each user, the work is carried out under this assumption. In this thesis, the chaotic logistic map, the Chebyshev map, and the Chen's attractor described in Chapter 2 are presented to replace the conventional spreading code. In doing so, the large number of the sequence implied the higher capacity of the users can be accommodated.

The LSTC DS/CDMA transmitter and receiver are described to simulate the performance of the chaotic spreading code. The structure of an adaptive detection algorithm recursive least square (RLS) receiver is also given in Chapter 3 to mitigate the interference. Note that this interference occurs because of the co-channel interference (CCI) and multiple access interference (MAI).

The performance of chaotic sequences and gold sequences as spreading sequences with additive white Gaussian noise (AWGN) channels are presented in Chapter 4. From the simulation results, it can be shown that the chaotic spreading codes outperform the Gold sequence in terms of BER performance. But, the performance of the Chen's attractor is not as well as the lower-dimensional chaotic map i.e the logistic map and the Chebyshev map. This might be because the investigation is limited to x parameter only. Further investigation is needed to conclude the BER performance in other situation such as changes in y and z parameters, which is beyond the scope of work in this research and is suited for future study.

The use of chaotic sequences for LSTC DS/CDMA system has several advantages over the conventional methods. One advantage is the availability of an enormous number of different sequences of a given length compared to the PN sequences. The generation and regeneration of chaotic sequences

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are very simple and involve the storage of only a few parameters and functions even for very long sequences. Moreover, the code sequences are easily made independent from one information symbol to the next. The chaotic sequences make the transmitted signal looked like noise. As a result, chaotic transmissions have less risk of interception and hard to be detected by eavesdroppers. They possess low auto and cross correlations with spectral density similar to wideband noises.

5.2 Future Research Works

The details theoretical study of generalization chaotic system properties and its applications for a data transmission system LSTC DS/CDMA based on the properties of chaotic dynamics were conducted in this thesis. For the future research, the practical implementation of the chaotic spreading code in LSTC DS/CDMA system should be further investigated.

To enhance the performance of the higher-dimensional map, the improvement of synchronization is needed, so that the future investigation about synchronization technique will be considered in the future works. Furthermore, the implementation of chaotic in modulation technique will be a good topic to work in comparison with the performance of other techniques.

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Wireless Communication Systems: Chaotic system, MIMO systems, CDMA systems, OFDM system, OFDMA systems.

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List of International Conferences

Proceeding Papers

1. **Dewi Nugrahani**, Ken Umeno, Klomkarn Kitdakorn, and Pitikhate Sooraksa, "Performance Evaluation of CDMA System Based on PN Code and OCMBS Chebyshev Chaotic," *Information Communication Technology, International Workshop KMITL 2010*, Bangkok, Thailand, 2010.

2. **Dewi Nugrahani**, Teav Keovkolyan, Chakree Teekapakvisit, Klomkarn Kitdakorn, and Pitikhate Sooraksa, "Secure Data Communication in HDD Manufacture using Chen's Attractor," *The 3rd International Data Storage Technology Conference, DST-CON 2010 International Conference Proceedings*, BITEC, Bangkok, Thailand, 30 July – 1 August 2010.

3. Teav Keov Kolyan, Chakree Teekapakvisit, **Dewi Nugrahani**, Pitak Thumwarin, and Takenobu Matsuura, "An Adaptive Generalized RAKE CDMA Receiver for Layered Space Time Coded Systems in Multipath Fading Channels," *Electrical Engineering/Electronics Computer Telecommunications and Information Technology, ECTI-CON 2010 International Conference*, Chiang Mai, Thailand, pp. 56-60, 19-21 May 2010.

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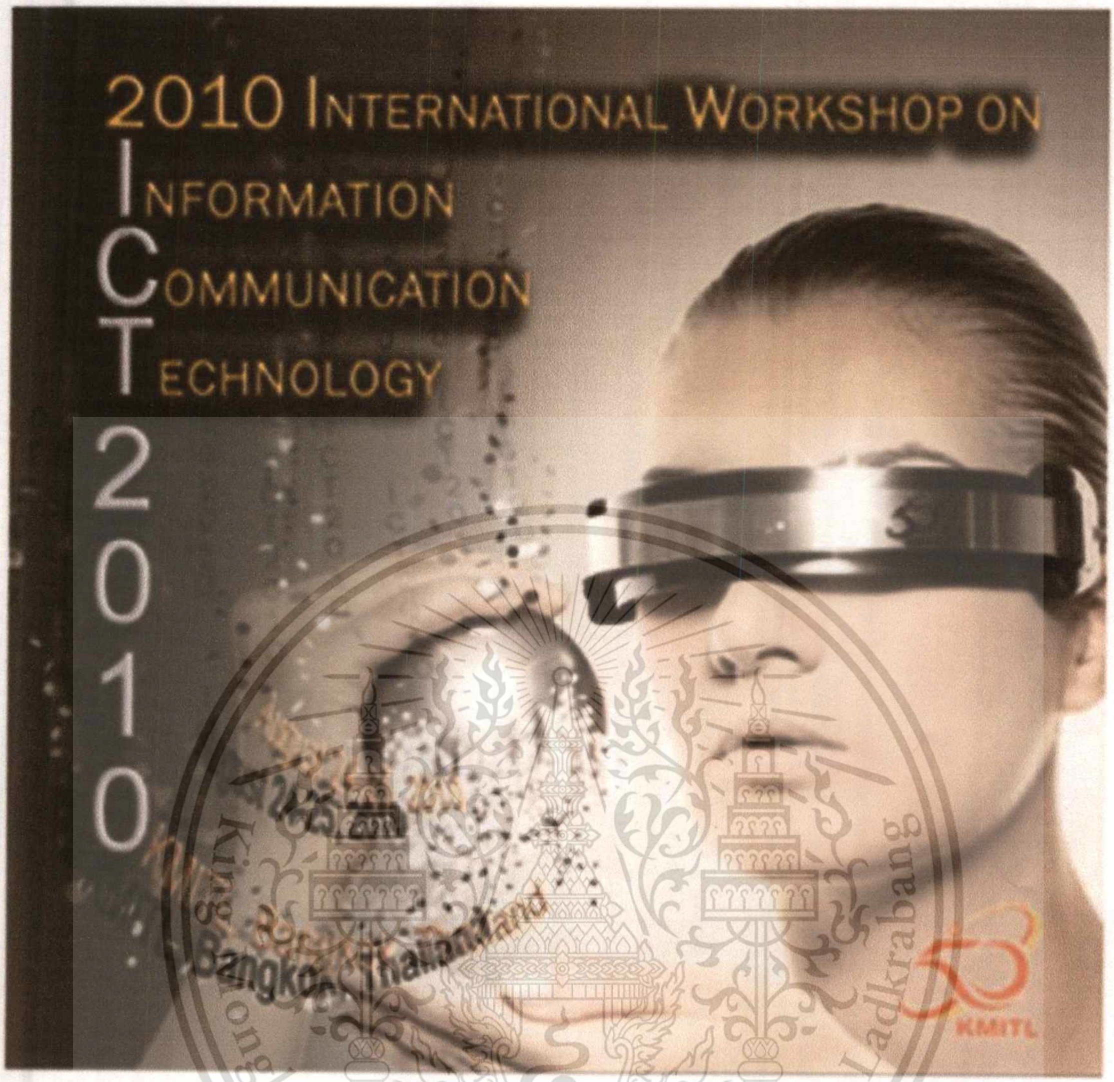

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Performance Evaluation of CDMA System Based on PN Code and OCMBS Chebyshev Chaotic

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***Abstract*—The growth of telecommunication technology need the high security and capacity to deliver data from source to destination. CDMA system with spreading sequences in spread spectrum system have many advantages to combat interference that is occurs during transmission. Optimal over-sampled chaotic binary sequences (OCMBS) Chebyshev chaotic as spreading sequences is proposed in this paper combined with LSTC adaptive receiver CDMA system to gain the high security for data transmission with their behavior in non-periodic, non-correlation and random performances. The performance of Chebyshev OCMBS then compare with gold code.**

I. INTRODUCTION

DS-CDMA as a spread spectrum technique have been widely used in wireless communication based on pseudo-noise (PN) sequences. In DS-CDMA, all of the users transmit their information using their own basic signals, which are orthogonal to each other. These basic signals can be pseudorandom (PN) sequences with good correlation properties. The PN sequences can be generated by Gold code, Kasami code, Walsh code, OVSF code and Maximal length shift register sequence (m-sequence).

Adaptive iterative receiver for Layered Space Time Coding (LSTC) CDMA was proposed in [1] to combat interference from adjacent layer and adjacent users in multiple user environments. Since the classical PN sequences is difficult to generate in a large number of sequences and because of their security issues, then chaotic spreading sequences were studied in [4]-[10], [12], [13]. There are chaotic as spreading sequences was introduced in spread spectrum communication system, e.g. Logistic map, Chen's attractor, Tent map, Chebyshev, to name a few. But, since the chaotic sequence is no longer binary, so digital encoding technique is used to convert continuous values to binary.

Based on adaptive iterative receiver for LSTC, The use of Chebyshev chaotic code is proposed in this paper to increase the performance of the spread spectrum CDMA system, especially in security and capacity of users. The main idea of using Chebyshev chaotic codes as a spreading sequences are because the generated signals have a good property of autocorrelation function and cross correlation function, they can generate a large number of orthogonal spread sequences, and easy to be tracked and acquired. For Chebyshev chaotic sequences in spread spectrum, Yang *et al.* was proposed the performance of chaotic with Optimal over sampled Chaotic Map Binary Sequenced (OCMBS) [14]. Furthermore, the OCMBS can enhance the security because it have several initial value, degree k and the other parameters which can produce many different value of spreading sequences and the SNR of system can improved by OCMBS.

This paper is organized as follows: system model about DS-CDMA based on Chebyshev code is presented in Section II, performance result is presented in Section III and finally, conclusion of this paper is given in Section IV. This document shows guidelines for preparing a final camera-ready manuscript in the proceedings of ICT 2010. The format here described allows for a graceful transition to the style required for that publication.

II. DS-CDMA BASED ON CHEBYSHEV

In a LSTC CDMA system model, all the signal of K user are transmitted simultaneously with N transmitter antennas and M receiver antennas. Fig. 1 shows structure of LSTC transmitter with Chebyshev chaotic codes as spreading sequences.

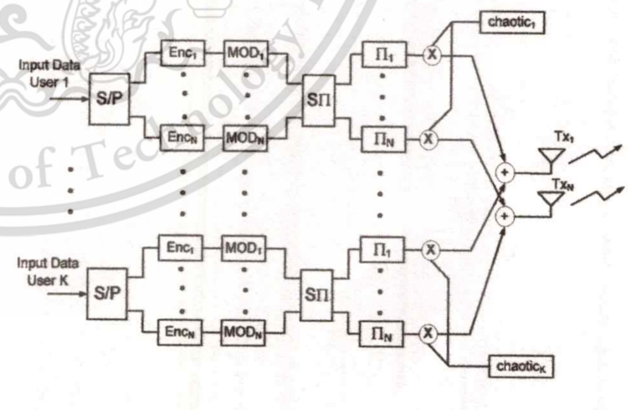

Fig. 1 DS-CDMA transmitter structure based on chaotic

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The binary information of each user is transmitted at a data rate of $r_b = 1/T_b$, where T_b is the bit interval. This data information is first converted into layered data information streams by a serial to parallel (S/P) converter. Before modulated, each information stream was encoded by a convolutional encoder to produce coded data stream for each layer. The layered coded data streams are then fed into a spatial time interleaver (∠S) and time interleaver (∠). Next, the coded data streams are spread by Chebyshev sequences using a spreading gain of L with $L = T/T_c$, where T is the symbol interval, and T_c is chip duration of spreading sequence. Finally, the spread symbols of all users are then combined together and transmitted simultaneously through N transmit antennas.

In this paper we proposed the second degree (Chebyshev polynomial) of chaotic Chebyshev map as spreading sequences that was studied in [2],[3]and [5]. The Chebyshev polynomial of degree k is described by:

$$x_{n-1} = cos(k \cos^{-1} x_n) \qquad x_n \in (-1,1) \tag{1}$$

With invariant density $\rho(x) = \pi^{-1}(1-x^2)^{-1/2}$, whose trajectories can behave "chaotically", from equation (2) we can get chaotic sequence with over-sampled chaotic map.

$$x_{n+1} = \underbrace{f\left(f \dots \left(f(x_n)\right)\right)}_{p} = f^{(k)}(x_n) \tag{2}$$

where $x_{n+1} = f(x_n)$ is a one-dimensional chaotic map and k is a natural number ≥ 2.

Since the chaotic sequence is no longer binary, so to obtain binary sequences from chaotic real-valued sequence $\{x_n\}_{n=0}^{\infty}$, over-sampled chaotic map is given as below :

$$\sigma_c(x) = \begin{cases} 0 & (|x| < c) \\ 1 & (|x| > c) \end{cases}, \quad (c \geq 0) \tag{3}$$

The threshold c are equal to the median value of chaotic real valued sequences for OCMBS.

Let **x** be the coded signal vectors transmitted by K users through N transmit antenna and be defined by :

$$\boldsymbol{x} = [x_1, x_2, \dots, x_p, \dots, x_k]^T \tag{4}$$

where $\boldsymbol{x_p} = [x_p^1, \dots, x_p^n, \dots, x_p^N]$ and b_p^n is the information bit of the p-th user for n-th transmit antenna with n=1, …, N and p=1, …, K. Let B represents the L x KN spread transmitted sequences of K users for n transmit antennas, as given by

$$\boldsymbol{B} = [b_1^1, \dots, b_1^N, \dots, b_K^1, \dots b_K^N] \tag{5}$$

where K is the number of K-user and N is the N-transmit antenna.

Fig. 2 shows the structure of adaptive iterative receiver with de-spreading Chebyshev map. Let $y_{t,j}^p$ be the received signal vector for the p-th user at the receiver antenna j, j=1, …, M, for the symbol t, and is given by :

$$y_{t,j}^p = BT_{t,j}^p x + n_{t,j}^p \tag{6}$$

$T_{t,j}^p$ represents the fading coefficient from j-th receive antenna to n-th transmit antenna of the p-user; $n_{t,j}^p$ is defined as an L x 1 noise factor at the receive antenna j of the p-th user. The assigned vector is given by $n_{t,j}^p = [n_{j,1}^p(t), \dots, n_{j,p}^p(t), \dots, n_{j,L}^p(t)]^T$ where $n_{j,p}^p(t)$ is Gaussian random variable with zero mean and two sided power spectral density $N_0/2$ per dimension. The received signals for all receive antennas of the p-th user is given by $Y_t^p = [r_{t,1}^p, \dots, r_{t,j}^p, \dots, r_{t,M}^p]^T$.

For a one-dimensional map $x_{n+1} = f(x_n)$, assume that the error at the trajectory point x_n is d_n

$$dx_{n+1} = f(x_n + dx_n) - f(x_n) \quad \approx \quad f'(x_n) \, . \, dx_n \tag{7}$$

Assuming the error dx_0 in the seed is given, the error x_n can be given by:

$$|dx_n| = dx_0 \prod_{i=0}^{n-1} |\, f'(x_i)| \tag{8}$$

Where $|f'(x_i)|$ is a local expanding factor.

The structure of the receiver consists of K adaptive iterative single user receivers with the adaptive detector. The received signal first forward to adaptive detector to estimate the channel coefficients and signature sequence for each layer of each user. The detector outputs are then fed to the time and spatial deinterleaver. And then the output of the spatial deinterleaver forward to the decoder to decode the signal with Maximum A Posteriori (MAP) decoder. The output of the MAP decoder as estimated soft symbol from all users are sent to the spatial and time interleaver and then fed back to the adaptive detector to cancel Co-Channel Interference (CCI) from adjacent antennas of the user and Multiple Access Interference (MAI) to cancel the interference from other users.

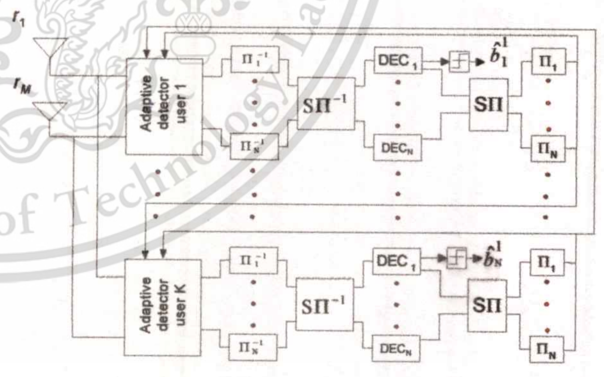

Fig. 2 DS-CDMA receiver structure based on chaotic sequences

The Chebyshev polynomials have the orthogonality as below as described in [11]:

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$$\int_{-1}^{1} T_i(x)\, T_j(x)\, \rho(x) dx = \delta_{i,j} \frac{(1+\delta_{i,0})}{2} \quad (9)$$

where $\delta_{i,j}$ is the kronecker delta function such that

$$\delta_{i,j} = \begin{cases} 1, & i = j, \\ 0, & i \neq j, \end{cases} \quad (10)$$

The autocorrelation and cross-correlation properties of Chebyshev OCMBS are described by equation (11) and (12), respectively. From Fig. 3 and Fig. 4 can be seen that Chebyshev based on OCMBS have good correlation properties, even though their initial values have a little difference.

Auto-correlation:

$$ac(m) = \frac{1}{N} \sum_{i=0}^{N-1} x_i x_{i+|m|} \quad (11)$$

Cross-correlation:

$$cc_{kl}(m) \begin{cases} \frac{1}{N}\sum_{i=0}^{N-1} x_{ki} x_{l(i+m)} & 0 < m \leq M \\ \frac{1}{N}\sum_{i=0}^{N-1} x_{ki} x_{li} & m = 0 \\ \frac{1}{N}\sum_{i=0}^{N-1} x_{k(l-m)} x_{li} & -M \leq m < 0 \end{cases} \quad (12)$$

III. PERFORMANCE RESULTS

As mentioned that the key point of spread spectrum system is about their correlation properties, OCMBS Chebyshev shows the good auto-correlation and cross-correlation properties. The auto-correlation property of Chebyshev map is shown in Fig. 3 and Fig. 4 for their cross-correlation.

This paper simulate about performance of OCMBS Chebyshev map as spreading sequences based on CDMA system compare with gold code as a classical spreading sequences. When the number of users increases, Chebyshev chaotic performance degrades slightly when compared to gold code.

IV. CONCLUSIONS

Chebyshev chaotic code is proposed in this paper, compared with gold code as spreading sequences. The performances result shown that OCMBS Chebyshev polynomial based on chaotic spreading sequences is inferior than gold code. The correlation function of chaotic shows the better performances than gold code and for chaotic can enhanced the security of system. Furthermore, the study of receiver structure will be improves to gain the better performance of chaotic spreading codes.

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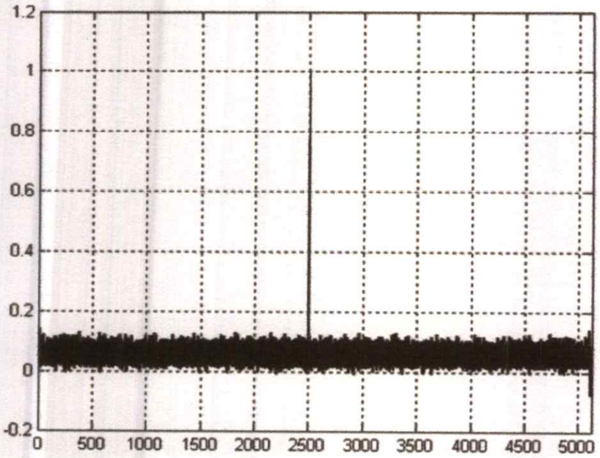

Fig. 3 Auto-correlation function for OCMBS Chebyshev chaotic

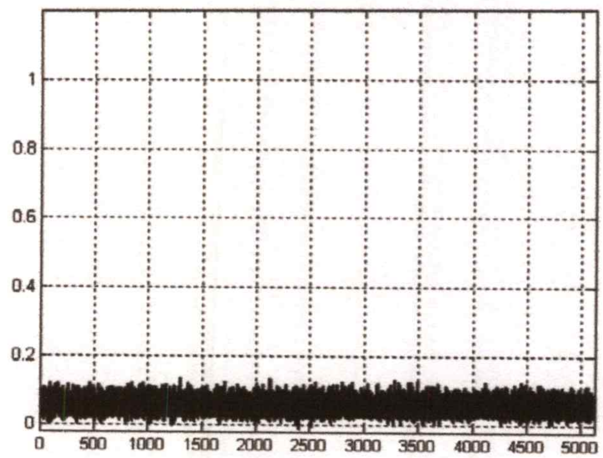

Fig. 4 Cross-correlation function for OCMBS Chebyshev chaotic

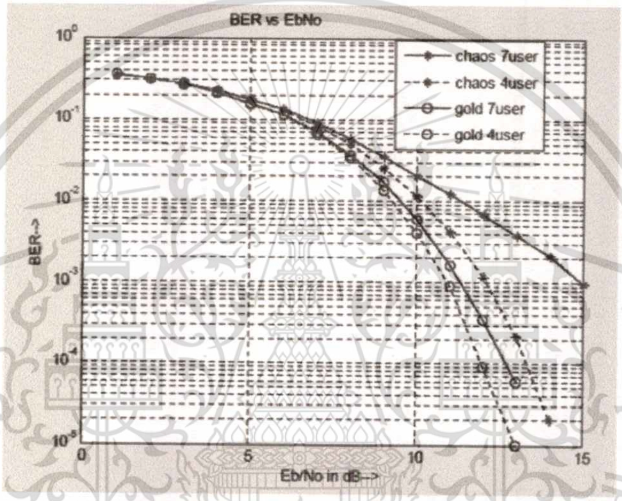

Fig. 5 BER performance comparison of OCMBS spreading codes and Gold codes in multipath channel

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Secure Data Communication in HDD Manufacture using Chen's Attractor

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Abstract—**In HDD manufacture, OCR is employed to identify each production lot for HGAs, HSAs, and other components. The obtained data can be sent via wireless communication to the production department for recording in the company's database. This paper presents DS-CDMA based on Chen's chaotic generator as pseudorandom (PN) sequences for secured communication. Digital encoding adopted to convert continuous sequence of the chaotic function's to binary sequence. Recent studies showed that chaotic sequence can enhance security and increase the capacity of the system using the behavior of the chaotic signals. In this paper we also presented the performance of adaptive iterative receiver in DS-CDMA system embedded with the chaos.**

Keywords-data communication; CDMA; chaos; PN sequence

I. INTRODUCTION

Hard disk drive industry has been employed OCR to identify serial number of each product such as HGA, HSA, and other components. However, utilization of the data recorded in the local computer may not suit the need for on-line tele-access. This can be solved by using RFID interface for communication arrangement. Nevertheless, the range of RFID communication is not appropriate for long distance. The modern communication for rich data such as 3-G technology can be employed for transferring the signals. To ensure the confidentiality, secured communication is needed. In this paper we propose a solution based on this demand using a technique of 3-G, namely, Direct Spread Spectrum Code Division (DS-CDMA) with chaotic signals to realize the system.

In DS-CDMA, all of the users transmit their information using their own basic signals, which are orthogonal to each other. These basic signals can be pseudorandom (PN) sequences with good correlation properties. The PN sequences can be generated by gold code, Kasami code, Walsh code, OVSF code and maximal length shift register sequence (m-sequence) as given in [1].

Implementation of Layered Space-Time Coding (LSTC) CDMA system has been studied in [2]. This system generates interference from adjacent layers and from the other users. Eventually the approach can degrade the system performance seriously. To mitigate this interference and increase the performance of the system, chaotic function is presented as PN sequences. In this paper, we purpose chaotic functions using Chen's system. Since the spreading sequence in chaotic spread spectrum is no longer binary, the application of the chaotic sequences in DS-CDMA is limited. A further attempt digital encoding technique is used to convert continuous values to binary.

In this work, chaotic function is used to generate PN-sequence because the characteristic of this signal can mitigate interference that is occurs during transmission in the receiver. Chaotic signals are non-periodic, non-converging and bounded, but deterministic and very sensitive to initial condition and control parameter as described in [3-5]. The behavior of the signal can increase the better performance of the ordinary DS-CDMA, especially for the security and capacity of the system.

There are chaotic functions that have been studied, e.g. Logistic map, Henon map, Lozi map, Chua's circuit, Lorenz system and Chen's system in [6,10], to name a few. In this paper, only Chen's system is employed as an example. The advantage of using chaotic sequences as spreading sequences is that infinite sequences exist with different initial conditions to carry different users, as meaning that the different user will spread the spectrum based on different initial condition as given by [7-8]. In doing so, the system can be traceable by the network administer and retain generating random sequences (PN) for the users.

This paper is organized as follows: system model is presented in Section II, performance result is presented in Section III and finally, conclusion of this paper is given in Section IV.

II. SYSTEM MODEL

Since the modern communication in HDD manufacture need the high data communication with high security, capacity and can cover the width area, so wireless communication i.e. DS-CDMA based on PN

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chaos is purposed to support the demand of secured data communication in HDD manufacture. In this chapter, transmitter and receiver structure of DS-CDMA system based on PN chaos is described.

A. Transmitter Structure

In a downlink DS-CDMA system, all user's signals are transmitted simultaneously with N transmitters and M receivers. Figure 1 shows the DS-CDMA transmitter structure with K users. This transmitted subsystem can be implemented and installed on site along the production line.

Data information from K users are converted from serial to parallel (S/P) converter, and then each of the information sequence is encoded by convolutional encoder to produce coded data stream for each layer and coded information stream. The stream is modulated by binary phase shift keying (BPSK) modulator. Before the coded information streams are spread by chaotic code, each of the coded information streams are fed into spatial time interleaver (SΠ) and time interleaver (Π). The spread symbol of all users are then combined together and transmitted into N transmitters. The chaotic function that is used in this system is Chen's system with 3 dimensional maps.

Let **x** be the coded signal vectors transmitted by K users through N transmit antenna and be defined by :

$$\boldsymbol{x} = [x_1, x_2, \dots, x_p, \dots, x_k]^T \tag{1}$$

where $\boldsymbol{x_p} = [x_p^1, \dots, x_p^n, \dots, x_p^N]$ and $\boldsymbol{b_p^n}$ is the information bit of the p-th user for n-th transmit antenna with n=1, …, N and p=1, …, K.

Let B represents the L x KN spread transmitted sequences of K users for n transmit antennas, as given by

$$\boldsymbol{B} = [b_1^1, \dots, b_1^N, \dots, b_K^1, \dots b_K^N] \tag{2}$$

where K is the number of K-user and N is the N-transmit antenna.

B. Receiver Structure

An adaptive receiver detector for multi-users was studied in [1]. This adaptive receiver is used to mitigate Co-Channel Interference (CCI) from the adjacent layers and Multiple Access Interference (MAI) from the users. The structure of the receiver of DS-CDMA is shown in Fig. 2. In this application scheme, the receivers can be installed remotely at the data collection office far away from the manufacturing sites.

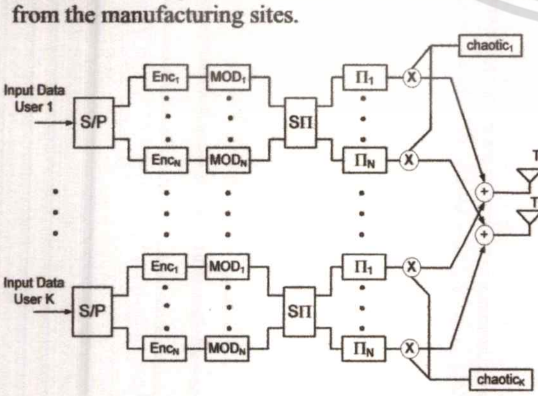

Fig. 1 DS-CDMA transmitter structure based on chaotic

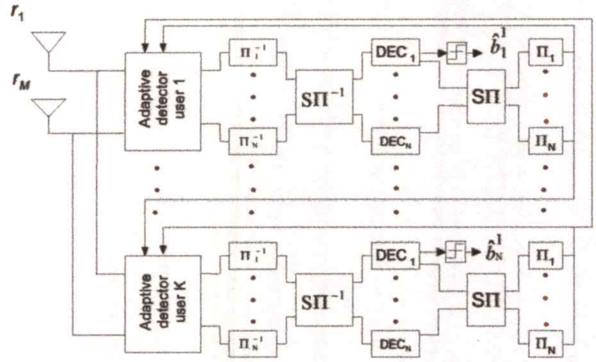

Fig. 2 DS-CDMA receiver structure based on chaotic sequences

Let $y_{t,j}^p$ be the received signal vector for the p-th user at the receiver antenna j, j=1, …, M, for the symbol t, and is given by :

$$y_{t,j}^p = BT_{t,j}^p x + n_{t,j}^p \tag{3}$$

$T_{t,j}^p$ represents the fading coefficient from j-th receive antenna to n-th transmit antenna of the p-user; $\boldsymbol{n_{t,j}^p}$ is defined as an L x 1 noise factor at the receive antenna j of the p-th user. The assigned vector is given by $\boldsymbol{n_{t,j}^p} = [n_{j,1}^p(t), \dots, n_{j,p}^p(t), \dots, n_{j,L}^p(t)]^T$ where $n_{j,p}^p(t)$ is Gaussian random variable with zero mean and two sided power spectral density $N_0/2$ per dimension. The received signals for all receive antennas of the p-th user are given by $Y_t^p = [r_{t,1}^p, \dots, r_{t,j}^p, \dots, r_{t,M}^p]^T$.

The structure of the receiver consists of K adaptive iterative single user receivers with the adaptive detector. The received signal first forward to adaptive detector to estimate the channel coefficients and signature sequence for each layer of each user. The detector outputs are then fed to the time and spatial deinterleaver. And then the output of the spatial deinterleaver forward to the decoder to decode the signal with Maximum A Posteriori (MAP) decoder. The output of the MAP decoder as estimated soft symbol from all users are sent to the spatial and time interleaver and then fed back to the adaptive detector to cancel Co-Channel Interference (CCI) from adjacent antennas of the user and Multiple Access Interference (MAI) to cancel the interference from other users.

C. Chaotic Chen's System

In this paper, we purposed Chen's system as a PN-sequence in DS-CDMA. Since the spreading sequence in chaotic sequences is no longer binary, the application of the chaotic sequences in digital communication is limited. Digital encoding technique is used to transform continuous values to binary sequences to adopt it in digital communication system. The studied of digital encoding is given in [9].

There are various methods to generate binary sequences from chaotic continuous sequences. Open flow coupled-chaotic sequence, threshold function and digitalization can be used to generate binary sequences from chaotic function [10].

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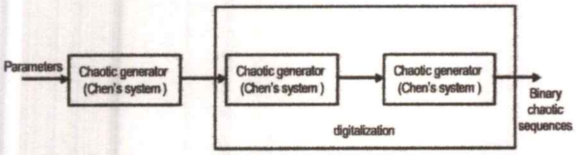

Fig. 3 Generation of binary chaotic sequences

Digitalization with quantizing and encoding process use to generate binary sequence in this paper. The block diagram of generation of the binary chaotic sequences is given in Fig. 3. The chaotic sequences are generated by chaotic function and then fed forward into quantizing and encoding block. The quantization performs an equal-interval quantization of the floating point input signal varying from -1 to +1. The output signal is quantized into whole units, the unit size is determined by the number of bits used in the binary representation. The coding block converts the quantized signal into a bit stream. And the sequence of the stream bits is a chaotic bit sequence.

Chen's system is purposed to generate the chaotic sequence. The chaotic Chen's system can be described as below:

$$\dot{x} = a(y-x)$$
$$\dot{y} = (c-a)x - xz + cy \quad (4)$$
$$\dot{z} = xy - bz$$

where a = 35, b = 3, and c = 28

Since good correlation is needed to achieve good performance in data communication, chaotic Chen's system presents the auto-correlation between PN-sequence generated at the transmitter and at the receiver and cross-correlation of the PN-sequence between different users at the same frequency and time [9]. The autocorrelation function is shown in Fig. 4. Figure 5 shows the cross-correlation function of the Chen's system. The auto correlation functions used to avoid the higher peak values which help in reliable recovery and provide optimal cross correlation functions, which minimize symbol error rate due to interference.

The normalized cyclic hamming auto correlation function of a sequence of length N can be defined as

$$R(\tau) = \frac{(n_\tau - d_\tau)}{N}, \quad 0 \le \tau \le N-1 \quad (5)$$

Similarly, the normalized cyclic hamming cross correlation function of an N sequence can defined as

$$R_{xy}(\tau) = \frac{(n_\tau - d_\tau)}{N}, \quad 0 \le \tau \le N-1 \quad (6)$$

Where and are number of agreements and disagreements respectively to an N - length sequence for any correlative distance with N = 5120 bits.

As the results from Fig. 4 and Fig. 5 show that the value of auto correlation and cross correlation is less than 0.2. It means that for the Chen's system have a minimum value of correlation properties. This eventually benefits to employ this system to the direct spread spectrum.

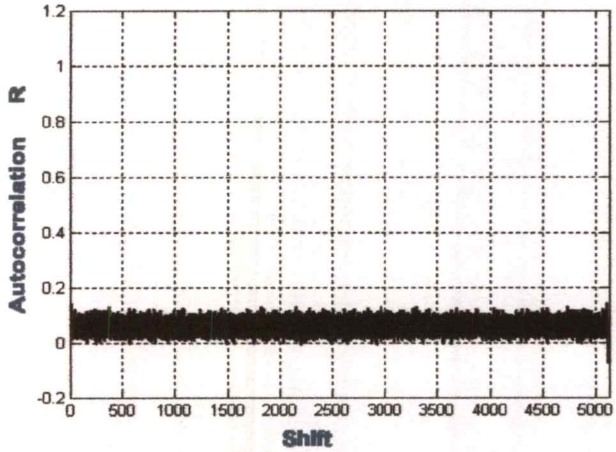

Fig. 4 Autocorrelation function of Chen's system

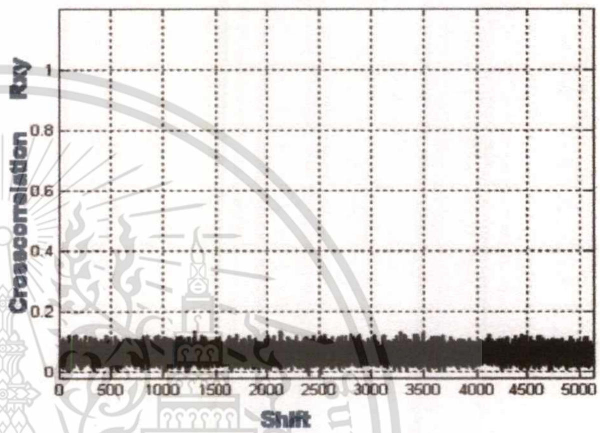

Fig. 5 Cross-correlation function of Chen's system

CDMA, because of having the low correlation properties. The displayed good auto correlation properties simplify the synchronization of such codes and the low cross correlation properties are useful in increasing the user's capacity in LSTC CDMA.

Chaotic sequences have also been proven for easy-to generate-and-store feature. Initial condition and control parameter of the chaotic function are needed for their generation, which means that there is no need for storage of long sequences. Moreover, a large number of different sequences can be generated by simply changing initial condition and control parameters.

III. PERFORMANCE RESULTS

This section presents simulation results for the Adaptive Minimum Means Square Error (MMSE) receiver. We use adaptive iterative downlink LSTC CDMA receiver over Additive White Gaussian Noise (AWGN) channel and multipath channel. Figure 6 shows the performance results of adaptive receiver for varying E_b/N_0 conditions. It is seen that when the number of users is 4, there is a 0.2 dB performance difference at a BER of 10^{-3} between chaos and gold methods. This difference is increased to almost 1 dB at a BER of 10^{-3} in case of 7 users. It also seen that there is 0.8 dB performance

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penalty at BER of 10^{-3} for chaotic sequences when users changed from 4 users to 7 users. So, as the number of users increases, gold based adaptive receiver performance increase slightly when compared to chaos based adaptive receiver. The users here can be represented by number of stations used for sending the data.

In Fig. 7, performance of adaptive receiver was investigated for carrying E_b/N_0 conditions over multipath channel. As can be seen, the performance of chaotic based on adaptive receiver is inferior to gold code based on adaptive receiver when the numbers of the users are increased. To achieve the same quality but retain encrypted data, the number of users during operation should be strictly limited for not exceeding 4 locations accessing at the same time.

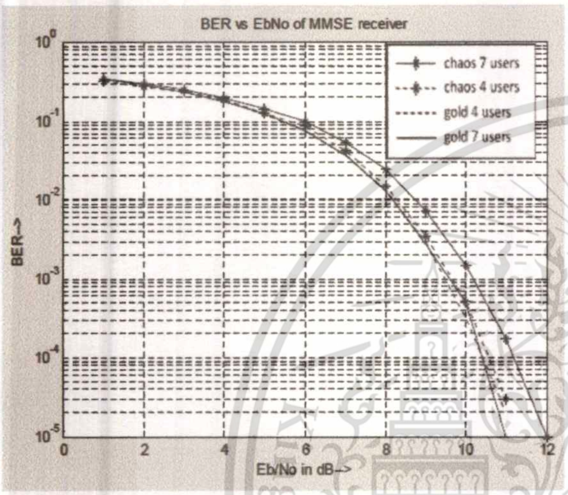

Fig. 6 BER performance of MMSE receiver for varying E_b/N_0 for 4 users and 7 users being active in the system in AWGN channel

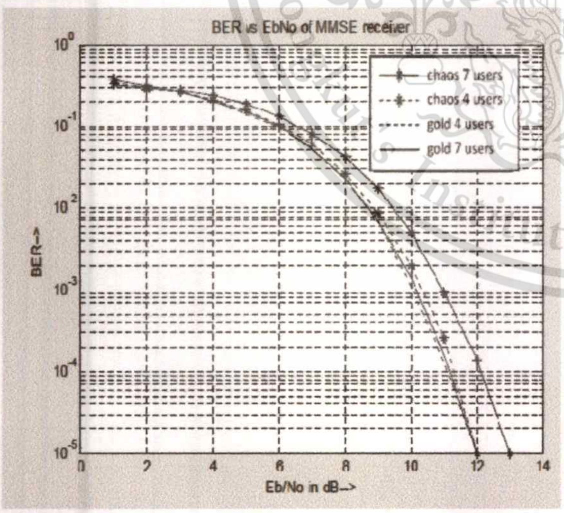

Fig. 7 BER performance of MMSE receiver for varying E_b/N_0 for 4 users and 7 users being active in the system in multipath channel

IV. Conclusion

In this paper we purposed chaotic Chen's system as a spreading code (PN-sequence) in DS-CDMA system to mitigate interference that is occurs at the Receiver during transmission system. The simulation is given to show the performance result of the chaotic sequences compared with gold code as PN-sequences. The correlation of the encrypted signals during transmissions reveals the effectiveness of the proposed scheme in generating the pseudorandom sequence. The HDD industry may be adapted the scheme to secure long distant data communication between the production line and the other offices. Although the generation of the chaotic sequences is simple for the transmitter and the intended receiver with the knowledge of parameters and functions involved, the exact regeneration is very difficult for the receiver that has to estimate them. Good and robust synchronization is required, which should be a topic for further research in implementation phase.

Acknowledement

This paper is supported partly by CR2009-2010 project, JICA, under AUN-SeedNet Program. The last two authors were also supported in part by NRCT Research Grant for the year 2553.

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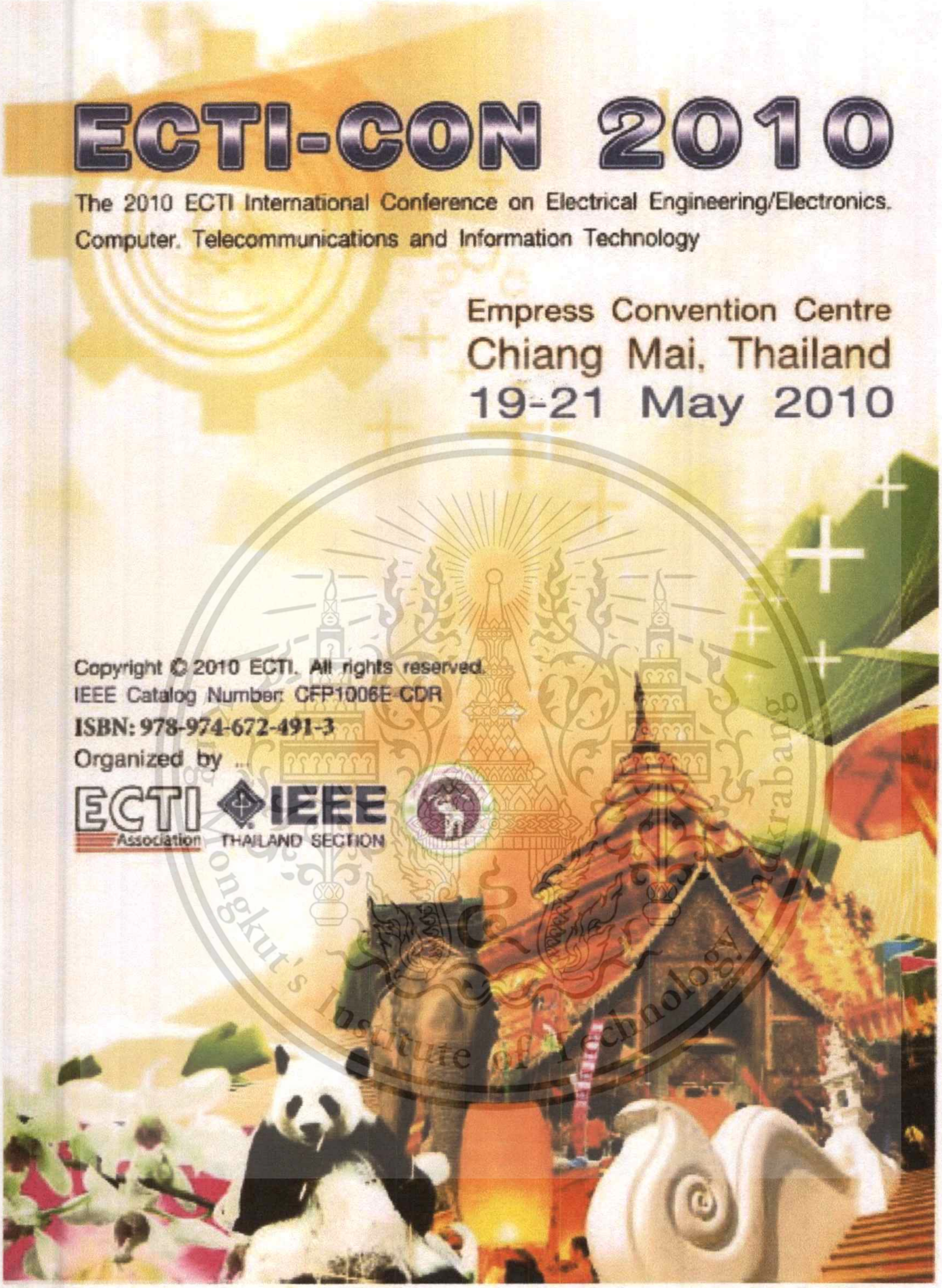

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An Adaptive Generalized RAKE CDMA Receiver for Layered Space Time Coded Systems in Multipath Fading Channels

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***Abstract*- An adaptive generalized RAKE receiver for Layered Space Time Coded CDMA (LSTC-CDMA) system has been proposed in this paper. The proposed Generalized RAKE receiver, based on a joint adaptive iterative detection and decoding algorithm, adaptively mitigates the effect of multipath fading and cancels the system interferences such as Co-Channel Interference (CCI) and Multiple Access Interference (MAI). The Partially Filtered Gradient Least Means Square (PFGLMS) algorithm is used for both feed-forward filter and feedback filter in the adaptive detection in order to determine the weight coefficient of each finger element. The performance of the system is evaluated by simulation results for the various numbers of RAKE fingers in multipath Rayleigh fading channels.**

I. Introduction

Wireless technology is experiencing spectacular developments according to the emergence of interactive and digital multimedia applications as well as rapid advances in the highly integrated systems. The MIMO technique proposed in recent years makes a full use of the space resources by employing multiple antennas at the transmitting and receiving terminal, thus largely increasing the channel capacity. Moreover, Direct-Sequence CDMA (DS-CDMA) has emerged as a predominantly multiple-access technique for 3G systems because of its efficient capacity and facility of network planning in a cellular environment.

To improve the throughput of this cellular system, the combination of Layered Space-Time Coding (LSTC) and CDMA, named as LSTC-CDMA, has been intensively studied in [1-3]. This implementation generates Co-Channel Interference (CCI) from the adjacent layers and Multiple Access Interference (MAI) from the users; hence, both will degrade the system performance seriously. To mitigate the aforementioned interferences controlled by Channel State Information (CSI), an adaptive Minimum Means Square Error (MMSE) receiver for a LSTC system have been proposed in [4]. In this work, CCI can be reduced, but MAI is still a huge problem, degrading the performance and yielding channel estimation inaccurate in a high interference environment. A non-linear adaptive iterative receiver has been studied in [5, 6], containing a feed-forward filter to suppress the system interference and a feedback filter to cancel the interference from adjacent antennas under an iterative format. In order to reduce the computational complexity of the feedback filter and improve the convergence speed of the above proposed systems, an adaptive iterative receiver for multiuser LSTC-CDMA system based on LMS and PFGLMS algorithms have been investigated in [7].

Due to the multipath fading of wireless channel, it is a difficult task to enlarge the network capacity and improve the quality of service. As the results, the generalized RAKE (G-RAKE) receivers for interference suppression and multipath mitigation have been developed in [8,9]. Compared to the conventional RAKE receiver, this generalized RAKE receiver has more fingers and has different combining weights.

In this paper, an adaptive iterative generalized RAKE LSTC-CDMA receiver for interference suppression and multipath mitigation has been proposed. Consequently, the proposed adaptive detector, based on the PFGLMS algorithm [10], was investigated in multipath Rayleigh fading channels. Moreover, the performance results show that the systems performances have been improved by increasing the number of RAKE fingers beyond the number of multipath components. The paper is organized as follows. System model is demonstrated in Section II which is mainly focus on proposed transmitter and receiver structure. The system performance is quantified through the Performance results from the computer simulation which is presented in Section III. Finally, section IV gives a conclusion of this work.

II. System Model

A CDMA system as depicted in Fig.1 is considered in this paper. Data from different users is modulated by different signature waveforms before being transmitted asynchronously through a wireless channel [11], which is modeled as multipath Rayleigh fading. At the receiver, the user data is retrieved adaptively base on the minimum mean square error criterion (MMSE).

Fig. 1. An adaptive CDMA system model.

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A. Transmitter Structure

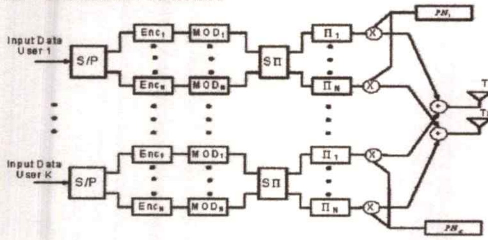

Fig. 2. LSTC-CDMA transmitter structure.

In a downlink LSTC-CDMA system, all user signals are transmitted simultaneously with N transmitters and M receivers antennas. Fig. 2 shows the LSTC-CDMA transmitter structure with K users. The binary information of each user is transmitted at a data rate of $r_b = 1/T_b$, where T_b is the bit interval. This data information is first converted into layered data information streams by a serial to parallel (S/P) converter. Before modulated, each information stream was encoded by a convolutional encoder to produce coded data stream for each layer. The layered coded data streams are then fed into a spatial time interleaver (SΠ) and time interleaver (Π). Next, the coded data streams are spread by its signature sequence using a spreading gain of L with $L = T/T_c$, where T is the symbol interval, and T_c is chip duration of spreading sequence. Finally, the spread symbols of all users are then combined together and simultaneously transmitted through N transmit antennas in a multipath Rayleigh Fading channels.

Let $\mathbf{b}$ be the coded signal vectors transmitted by K users through N transmit antennas and be defined by:

$$\mathbf{b} = [\mathbf{b}_1, \mathbf{b}_2, \ldots, \mathbf{b}_p, \ldots, \mathbf{b}_K]^T \quad (1)$$

where $\mathbf{b}_p = [b_p^1, \ldots, b_p^n, \ldots, b_p^N]$, and b_p^n is the information bit of the p-th user for n-th transmit antenna with $n=1,\ldots,N$ and $p=1,\ldots,K$.

Let $\mathbf{S}$ represents the $L \times KN$ spread transmitted sequences of K users for n transmit antennas, as given by:

$$\mathbf{S} = [\mathbf{s}_1^1, \ldots, \mathbf{s}_1^N, \ldots, \mathbf{s}_p^n \ldots \mathbf{s}_K^1, \ldots \mathbf{s}_K^N] \quad (2)$$

where $\mathbf{s}_p^n = [s_p^{n,1}, \ldots, s_p^{n,q}, \ldots, s_p^{n,L}]^T$, and $s_p^{n,q}$ is the q-th chip of a spreading sequence for the p-th user and n-th transmit antenna with $n=1,\ldots,N$, $p=1,\ldots,K$; and $q=1,\ldots,L$.

Let $\mathbf{A}$ is the received amplitude of the p-th user's signal for n-th transmit antenna with $n=1,\ldots,N$ and $p=1,\ldots,K$; which is represented by:

$$\mathrm{A} = \mathrm{diag}(\mathrm{A}_{1,1}, \ldots, \mathrm{A}_{1,N}, \ldots, \mathrm{A}_{p,n}, \ldots, \mathrm{A}_{K,1}, \ldots, \mathrm{A}_{K,N}) \quad (3)$$

Let $\mathbf{r}_{t,j}^p$ is the received signal vector for the p-th user at the receiver antenna j, $j = 1, \ldots, M$, for symbol t and delay time τ, is represented by:

$$r(t) = s(t-\tau)hAb + n(t)$$

and is equivalently by:

$$\mathbf{r}_{t,j}^p = \mathbf{S}\mathbf{H}_{t,j}^p \mathrm{A}\mathbf{b} + \mathbf{n}_{t,j}^p \quad (4)$$

where $\mathbf{r}_{t,j}^p = [r_{t,j}^{p,1}, \ldots r_{t,j}^{p,q}, \ldots, r_{t,j}^{p,L}]^T$, and $r_{t,j}^{p,q}$ is the received signal for the p-th user at the q-th chip of the t-th symbol for j-th antenna. $\mathbf{H}_{t,j}^p$ is defined by $\mathbf{H}_{t,j}^p = diag(\mathbf{h}_{t,j}^p, \ldots, \mathbf{h}_{t,j}^p)_{KN \times KN}$ where $\mathbf{h}_{t,j}^p = diag[h_{j,1}^p(t), \ldots, h_{j,n}^p(t), \ldots, h_{j,N}^p(t)]_{N \times N}$, and $h_{j,n}^p(t)$ represents the fading coefficient from j-th receive antenna to n-th transmit antenna of the p-th user. $\mathbf{n}_{t,j}^p$ is defined as an $L \times 1$ noise vector at the receive antenna j of the p-th user, given by $\mathbf{n}_{t,j}^p = [n_{j,1}^p(t), \ldots, n_{j,q}^p(t), \ldots, n_{j,L}^p(t)]^T$ where $n_{j,q}^p(t)$ is a Gaussian random variable with a zero mean and two sided power spectral density $N_o/2$ per dimension. The received signals for all receive antennas of the p-th user are given by $\mathbf{R}_t^p = [\mathbf{r}_{t,1}^p, \ldots, \mathbf{r}_{t,j}^p, \ldots, \mathbf{r}_{t,M}^p]^T$.

B. Receiver Structure

In downlink LSTC-CDMA system, we assume that the system has no knowledge of Channel State Information (CSI), spreading sequences, and fading coefficients except the training sequence. A block diagram of the proposed adaptive iterative generalized RAKE CDMA multiuser receiver for Layered Space Time Coded systems in multipath fading channels is shown in Fig. 3.

This structure consists of K adaptive iterative G-RAKE LSTC-CDMA single user receivers, each with an adaptive detector that followed by N parallel soft-input soft-output channel decoders. The received signals are first input to the adaptive detector in order to mitigate the effect of multipath fading. The detector outputs are then fed to the time and spatial deinterleavers, defined by Π^{-1} and $S\Pi^{-1}$, respectively. The deinterleavers outputs are then decoded by a Maximum A Posteriori (MAP) decoder. Next, the estimated soft symbols of MAP decoders from all users are sent to the spatial and time interleavers, and then fed back to the adaptive detector under iterative technique to cancel the interference from adjacent antennas of the user, called CCI, and the interference from other users, called MAI.

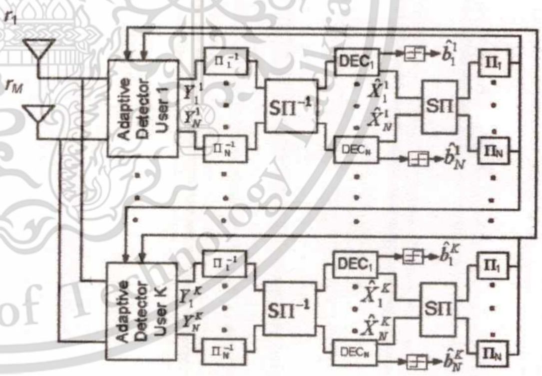

Fig. 3. Block diagram of G-RAKE LST- CDMA receiver.

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1. Proposed Adaptive Iterative Detector

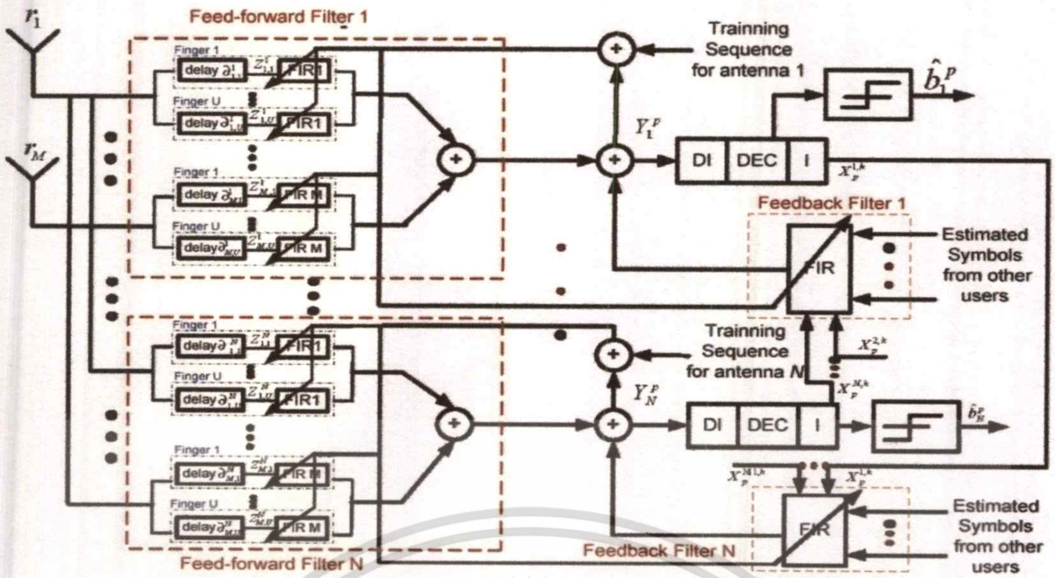

Fig. 4. Block diagram of the adaptive iterative G-RAKE LSTC-CDMA receiver.

The block diagram of the adaptive iterative generalized RAKE LSTC-CDMA receiver for the *p*-th user is shown in Fig. 4. In order to determine a weight coefficient of each finger element, the proposed detector employs an adaptive G-RAKE structure and antenna array processing. Unlike the conventional RAKE receiver, the G-RAKE receiver benefits from using more fingers than the number of multipath. In the proposed adaptive receiver structure, *N* sets of adaptive detector consist of *M* equalizers for the feed-forward filter and an equalizer for the feedback filter modules. An equalizer, employed in the proposed adaptive detector, is based on the PFGLMS algorithm. Moreover, the structure consists of a bank of *U* RAKE fingers, each correlating to a different delay of the received signal. The finger outputs are then combined to form a decision statistic. Importantly, the detector output for each layer is obtained from combining a feed-forward and a feedback filter output. In the iterative process, the feed-forward filter is compensated for the channel estimation error, and the feedback filter is used to cancel the interference from adjacent antennas and other users. In the first iteration, there are no estimated symbols from the decoders, and the feedback filter coefficients are zeros; thus, the feedback filter output is also zero. In the feed-forward filter, the *M* adaptive equalizers are used to estimate the channel coefficients and signature sequence for each layer of each user. The equalizer outputs from all receive antennas are added to obtain a feed-forward filter output signal for each transmit antenna as shown in Fig. 4.

Let $\mathbf{w}_j^{p,k}(t)$ be an Lx1 feed-forward tap coefficients vector for the *j*-th receive antenna of the *p*-th user during the *k*-th iteration at symbol interval *t* and be given by:

$$\mathbf{w}_j^{p,k}(t) = [w_{t,j}^{p,k}(0),\ldots,w_{t,j}^{p,k}(q),\ldots,w_{t,j}^{p,k}(L-1)]^T \quad (5)$$

where $w_{t,j}^{p,k}(q)$ is the feed-forward tap coefficient corresponding to the *q*-th chip of the spreading sequence.

Let $\mathbf{w}_b^{i,k}(t)$ be the feedback filter coefficients of all users, except the *i*-th antenna of the *p*-th user, at symbol interval *t* in time domain, is expressed as:

$$\mathbf{w}_b^{i,k}(t)=[\mathbf{w}_{b,1}^{1,k}(t),\ldots,\mathbf{w}_{b,1}^{N,k}(t),\ldots,\mathbf{w}_{b,p}^{i,k}(t),\ldots,\mathbf{w}_{b,K}^{1,k}(t),\ldots,\mathbf{w}_{b,K}^{N,k}(t)]^T \quad (6)$$

where $\mathbf{w}_{b,p}^{i,k}(t)=[w_{b,p}^{1,k}(t),\ldots,w_{b,p}^{i-1,k}(t),w_{b,p}^{i+1,k}(t),\ldots,w_{b,p}^{N,k}(t)] \quad (7)$

$\underline{\hat{\mathbf{x}}}_{t,p}^{i,k}$ is $(KN-1)\times 1$ vector of the estimated soft symbols, at the *k*-th iteration, from MAP decoders of every antenna of all users, except the *i*-th antenna of *p*-th user at symbol interval *t*, given by:

$$\underline{\hat{\mathbf{x}}}_{t,p}^{i,k} = (\hat{\mathbf{x}}_{t,1}^{1,k},\ldots,\hat{\mathbf{x}}_{t,1}^{N,k},\ldots,\hat{\mathbf{x}}_{t,p}^{i,k},\ldots,\hat{\mathbf{x}}_{t,K}^{1,k},\ldots,\hat{\mathbf{x}}_{t,K}^{N,k})^{\mathrm{T}} \quad (8)$$

where $\hat{\mathbf{x}}_{t,p}^{i,k} = (\hat{\mathbf{x}}_{t,p}^{1,k},\hat{\mathbf{x}}_{t,p}^{2,k},\ldots,\hat{\mathbf{x}}_{t,p}^{i-1,k},\hat{\mathbf{x}}_{t,p}^{i+1,k},\ldots,\hat{\mathbf{x}}_{t,p}^{N,k}) \quad (9)$

$\mathbf{z}_{t,j}^{p}$ is the finger outputs for *u*-th finger of *p*-th user at the receive antenna *j* and transmit antenna *i* for symbol *t* and be represented by:

$$\mathbf{z}_{t,j}^{p} = \mathbf{r}_{t,j}^{p}\,\delta(t-\partial_{j,u}^{i}) \quad (10)$$

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The detected symbol of the p-th user obtained at the output of the adaptive detector for the i-th antenna during the k-th iteration at symbol interval t, denoted by $y_{t,p}^{i,k}$, is defined as:

$$y_{t,p}^{i,k} = \sum_{j=0}^{M} \mathbf{w}_j^{i,k}(t)^H \mathbf{z}_{t,j}^p + \mathbf{w}_b^{i,k}(t)^H \hat{\underline{\mathbf{x}}}_{t,p}^{i,k} \qquad (11)$$

The detector soft output $y_{t,p}^{i,k}$ in the time domain is then compared to training symbol, denoted by $\mathbf{x}_{t,p}^{i,k}$ The difference between them is calculated as the detection error. The detection error for the p-th user in the k-th iteration at symbol interval t, for i-th antenna, denoted by $e_{t,p}^{i,k}$, is represented by:

$$e_{t,p}^{i,k} = y_{t,p}^{i,k} - x_{t,p}^{i,k} \qquad (12)$$

The detection error is then used to adapt the feed-forward filter and feedback filter tap coefficients in the time domain. After the Mean Square Error (MSE) approaches a specified value, the training mode is switched to the decision directed mode, in which the training sequence is replaced by the hard decision output of each user detector.

The values tap coefficients, $\mathbf{w}_j^{p,k}(t)$ in (5) and feedback filter tap coefficients, $\mathbf{w}_b^{i,k}(t)$ in (6) are calculated by minimizing the mean square error, defined as ζ, and given by

$$\zeta = mim\left(E\left|e_{t,p}^{i,k}\right|^2 \right) = \min\left(E\left[\left|y_{t,p}^{i,k} - x_{t,p}^{i,k}\right|^2\right]\right) \qquad (13)$$

And they can be determined recursively by partially filtered gradient LMS (PFGLMS) algorithm. PFGLMS algorithm based on an exponentially weighted least square error is utilized to improve the convergence speed with a slight increase in complexity [10].

The modified feed-forward and feedback coefficients of the PFGLMS algorithm can be expressed as:

$$\mathbf{w}_f^{i,k}(t+1) = \mathbf{w}_f^{i,k}(t) + \mu_f \mathbf{g}_f^{i,k}(t) \qquad (14)$$

and

$$\mathbf{w}_b^{i,k}(t+1) = \mathbf{w}_b^{i,k}(t) + \mu_b \mathbf{g}_b^{i,k}(t) \qquad (15)$$

where

$$\left.\begin{aligned} \mathbf{g}_f^{i,k}(t) &= e(t)\mathbf{z}_{t,j}^p + \hat{\mathbf{g}}_f^{i,k}(t) \\ \hat{\mathbf{g}}_f^{i,k}(t) &= \lambda_f \hat{\mathbf{g}}_f^{i,k}(t-1) + \gamma_f e(t)\mathbf{z}_{t,j}^p \end{aligned}\right\} \qquad (16)$$

and

$$\left.\begin{aligned} \mathbf{g}_b^{i,k}(t) &= e(t)\hat{\underline{\mathbf{x}}}_{t,p}^{i,k} + \hat{\mathbf{g}}_b^{i,k}(t) \\ \hat{\mathbf{g}}_b^{i,k}(t) &= \lambda_b \hat{\mathbf{g}}_b^{i,k}(t-1) + \gamma_b e(t)\hat{\underline{\mathbf{x}}}_{t,p}^{i,k} \end{aligned}\right\} \qquad (17)$$

where $\hat{\underline{\mathbf{x}}}_{t,p}^{i,k}$ and $\mathbf{z}_{t,j}^p$ are defined in (8) and (10), respectively. μ_f and μ_b are the step sizes for feed-forward and feedback adaptation. (γ_f, γ_b) and (λ_f, λ_b) are the forgetting factors and the scaling factors respectively, and $\hat{\mathbf{g}}_f^{i,k}(0) = \hat{\mathbf{g}}_b^{i,k}(0) = 0$.

2. Maximum A Posteriori Decoder

In the proposed receiver structure, detector output for user p, denoted by $y_{t,p}^{i,k}$ is decoded by Maximum A Posteriori (MAP) Decoder. The soft-output from the decoder is used to suppress the interference in the feedback filter in the next iteration. The process of this adaptive iterative detection/decoding is performed until the symbol estimation can converge to the optimal performance. The soft-output from the decoder in the last iteration is then fed into a decision device to produce a decision. For a Binary Phase Shift Keying (BPSK), the functions for the transmitted modulated symbols 1 and −1 can be written as [12]:

$$P\left(y_{t,p}^{i,k} \middle| x_{t,p}^{i,k} = \pm 1\right) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp\frac{-\left(y_{t,p}^{i,k} \mp 1\right)^2}{2\sigma^2} \qquad (18)$$

The Log-Likelihood Ratios (LLR) is determined in the k-th iteration for the i-th transmit layer of p-th user, denoted by $\lambda_{t,p}^{i,k}$:

$$\lambda_{t,p}^{i,k} = \log\left(\frac{P\left(x_{t,p}^{i,k} = 1 \middle| y_{t,p}^{i,k}\right)}{P\left(x_{t,p}^{i,k} = -1 \middle| y_{t,p}^{i,k}\right)} \right) \qquad (19)$$

The symbol A Posteriori Probabilities (APP) $P(x_{t,p}^{i,k} = q \mid y_{t,p}^{i,k})$, with q equal to 1 and -1, conditioned on the output variable which is defined as $y_{t,p}^{i,k}$ for p-th user, can be obtained as:

$$P\left(x_{t,p}^{i,k} = 1 \middle| y_{t,p}^{i,k}\right) = \frac{e^{\lambda_{t,p}^{i,k}}}{e^{\lambda_{t,p}^{i,k}} + 1}, \qquad P\left(x_{t,p}^{i,k} = -1 \middle| y_{t,p}^{i,k}\right) = \frac{1}{e^{\lambda_{t,p}^{i,k}} + 1} \qquad (20)$$

The soft-output symbols are estimated in the i-th layer and k-th iteration for p-th user, calculated as:

$$\tilde{x}_{t,p}^{i,k} = \frac{e^{\lambda_{t,p}^{i,k}} - 1}{e^{\lambda_{t,p}^{i,k}} + 1} \qquad (21)$$

III. PERFORMANCE RESULTS

This section presents simulation results of the adaptive G-RAKE receiver for downlink LSTC-CDMA systems over a quasi-static Rayleigh fading channel whereby the fading coefficient is constant within a frame, but changes independently from one frame to another. The constituent codes are nonsystematic convolutional codes with the code rate R of 1/2 and memory order of 3. The spreading sequence is a gold sequence with processing gain of 7. The proposed system is simulated with 2 transmit and 2 receive antennas with multipath fading of 3 and with 130 information bits in each frame per layer for each user. Each layer consists of 266 encoded symbols per frame. The data rate is 1 Mb/s at the carrier frequency, f_c, of 2 GH_z.

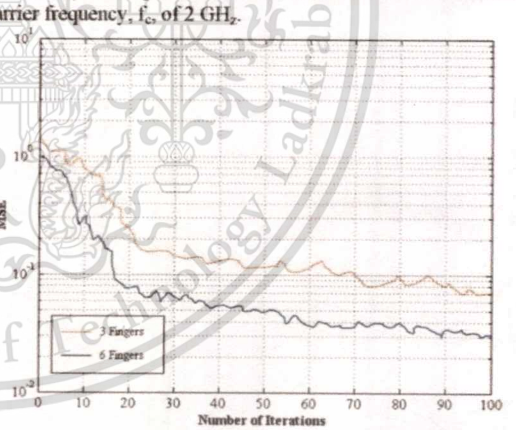

Fig. 5. The MSE performances of 3 and 6 fingers G-RAKE CDMA receiver.

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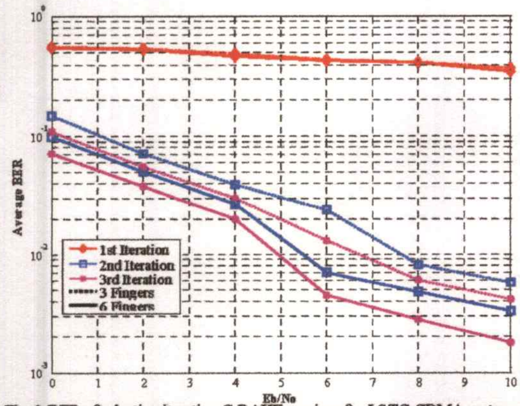

Fig.6. BER of adaptive iterative G-RAKE receiver for LSTC-CDMA systems.

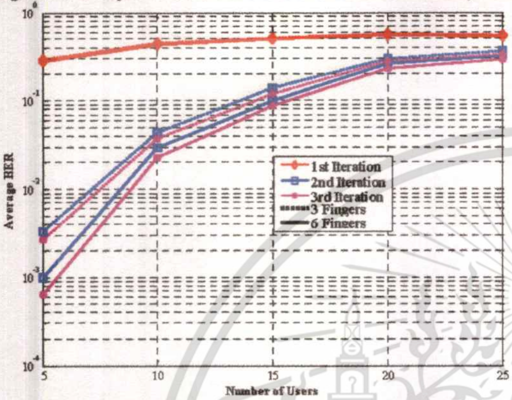

Fig.7. Number of users for time domain G-RAKE CDMA receiver.

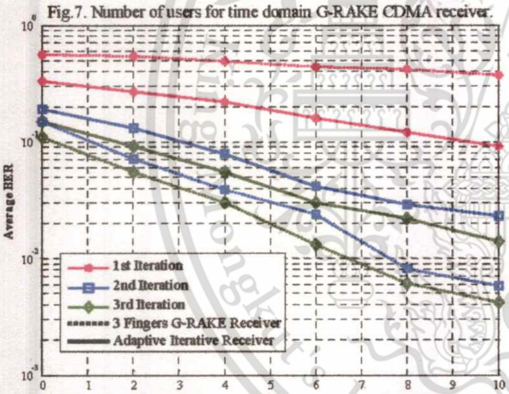

Fig. 8. Performances comparison between adaptive iterative receiver [7] and G-RAKE LSTC-CDMA receiver based on PFGLMS adaptive algorithm.

A comparison of MSE performances of adaptive G-Rake CDMA receiver for 3 and 6 fingers is shown in Fig. 5, whereby the MSE curve of 6 fingers RAKE receiver is lower than that of 3 fingers. This shows that the system performance improves when additional finger is placed at a delay. The performance of the 2 transmit and 2 receive antennas for 5 users under time domain approach is depicted in Fig. 6. The result shows that the performance of the G-RAKE receiver, based on PFGLMS algorithm, has a significant improvement in the first iteration and gradually improves for the higher iterations. The number of users for adaptive CDMA G-RAKE receiver is shown in Fig. 7. The simulation results show that the number of users is 15 at 3 iterations for a BER of 8.5×10^{-2}. These results also show that when the number of users increases, a higher number of iteration is required to achieve the same BER. In Fig. 8, we present the performance comparison between an adaptive LSTC-CDMA receiver [7] and a 3 fingers G-RAKE LSTC-CDMA receiver. The BER curves show that adaptive iterative receiver only performs better then G-RAKE receiver in the first iteration. This means that the proposed G-RAKE receiver have a better performance from second up to higher iteration.

IV. Conclusion

In this paper, we present an adaptive iterative generalized RAKE receiver for Layered Space-Time Coded CDMA systems, based on the Partially Filtered Gradient LMS algorithm, in order to mitigate the effect of multipath and suppress the interferences of the systems. Moreover, the system performance improved when the number of fingers was increased beyond the number of resolvable multipath. Due to the higher reliability of the Co-Channel Interference (CCI) and Multiple Access Interference (MAI) estimation, this performance results prove that the proposed G-RAKE CDMA receiver can effectively mitigate the CCI and MAI using the interference suppression and cancellation techniques.

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