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Design and Performance Analysis of a Multiuser OFDM Communication system Based Quadrature Chaos Shift Keying Spread Spectrum Modulation

Thesis

Submitted for fulfillment of the Master of science degree in Electrical Engineering

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بِسۡمِ ٱللَّهِ ٱلرَّحۡمَٰنِ ٱلرَّحِيمِ

قَالُواْ سُبۡحَٰنَكَ

لَا عِلۡمَ لَنَآ إِلَّا مَا عَلَّمۡتَنَآۖ

إِنَّكَ أَنتَ ٱلۡعَلِيمُ ٱلۡحَكِيمُ

سورة البقرة

الآيه ٣٢

The undersigned have examined the thesis entitled "**Design and Performance Evaluation of a Multiuser OFDM System Based on Differential Quadrature Chaos-Shift-Keying Spread Spectrum**" presented by **Eng. Mahmoud Osama El Beshry** , a candidate for Master in Engineering Sciences in Electrical Engineering and hereby certify that it is worthy of acceptance.

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Abstract

This work aims to propose a multiuser OFDM system with non-coherent differential quadrature chaos-shift-keying (OFDM-DQCSK) modulation. The proposed system is based on direct sequence spread spectrum where chaotic codes replace conventional spreading sequences. Orthogonal sets of chaotic spreading codes can be generated with excellent correlation properties. The proposed system uses duplicated reference subcarriers that are private and non-shared. These reference subcarriers carry the chaotic codes which are used to spread the modulated data symbols at the transmitter of a specific user. At the receiver, the reference subcarriers are used for the differential non-coherent de-spreading. The modulated data symbols are spread in time domain over shared subcarriers. Reference subcarriers are much less than data subcarriers to save bandwidth. At the receiver, neither complex channel estimators nor carrier synchronization is required for demodulation. The chaotic reference signals are averaged at the receiver so that the additive white gaussian noise (AWGN) on the reference signal is averaged to a negligible value. A simulation of the bit error rate performance is performed in the presence of AWGN, multipath Rayleigh flat-fading channels and multiple access interference. In conclusion, the proposed non-coherent system achieves a remarkable improvement due to the usage of the Reference Averaging Technique.

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Praise be to Allah the Lord of the worlds and may the blessings a nd peace of Allah be upon the most honored of messengers our master Muhammad and upon all his family and companion.

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This work is dedicated to the spirit of my father, may Allah have mercy on him, and to my mother. I am very grateful to my parents for their love, caring and sacrifices for educating and preparing me for my future. I want to thank my parents for their great support.

Finally, my thanks go to all the people who have supported me to complete the research work directly or indirectly.

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List of Abbreviations

4G	4^{th} Generation
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CDMA	Code Division Multiple Access
CDSK	Correlation Delay Shift Keying
CIR	Channel Impulse Response
CMOS	Complementary Metal Oxide Semiconductor
COOK	Chaos On-Off Keying
CSK	Chaos Shift Keying
DC	Direct Current
DCSK	Differential Chaos Shift Keying
DFT	Discrete Fourier Transform
DQCSK	Differential Quadrature Chaos Shift Keying
DS-SS	Direct Sequence Spread Spectrum
FFT	Fast Fourier Transform
FH-SS	Frequency Hopping Spread Spectrum

FM	Frequency Modulation
ICI	Inter-Carrier Interference
IDFT	Inverse Discrete Fourier Transform
IEEE	Institute of Electrical and Electronics Engineers
IF	Intermediate frequency
IFFT	Inverse Fast Fourier Transform
ISI	Inter-Symbol Interference
LPI	Low Probability-of-Intercept
LTE	Long-Term Evolution
MAI	Multiple Access Interference
MAX	Maximum
MC	Multi-Carrier
OFDM	Orthogonal Frequency Division Multiplexing
PN	Pseudo-Noise
PSK	Phase Shift Keying
QAM	Quadrature Amplitude Modulation
QCSK	Quadrature Chaos Shift Keying
QPSK	Quadrature Phase Shift keying
RF	Radio Frequency

SDR	Software Defined Radio
SER	Symbol Error Rate
SNR	Signal to Noise Ratio
SS	Spread Spectrum
TX	Transmitter
WiFi	Wireless Fidelity
WiMAX	Worldwide Interoperability for Microwave Access
WLAN	Wireless Local Area Network

Chapter 1
Introduction

1.1 Overview

The direct sequence spread spectrum (DS-SS) system supports multiple access communication among different users. It can combat multipath fading interference in the presence of a frequency selective channel. However, the performance of this system is limited by the inter-chip interference and the multiple access interference (MAI) [1]. In an OFDM system, the bandwidth of the modulated signal on each subcarrier is adjusted to be lower than the coherence bandwidth of the multipath fading channel. The channel is then converted to a non-frequency selective channel and a simpler channel equalizer can be used in the receiver. By the addition of an appropriate cyclic prefix, The ISI between successive OFDM symbols is completely removed. Therefore, Combining OFDM with the DS-SS system significantly reduces the inter-chip interference and becomes more resistant to multipath interference in frequency selective channels [2].

Therefore, a range of various techniques combining Direct sequence Code Division Multiple Access (DS-CDMA) and OFDM have been proposed, such as Multi-Carrier CDMA (MC-CDMA) in [2], and Multi-Carrier Direct Sequence CDMA (MC-DS-CDMA) in [2], [3].

One of the most important factors that affects multiple access interference (MAI) in a multiuser system is orthogonality among the spreading codes. Chaotic codes are non-periodic random like sequences and can give orthogonal sets of spreading codes. They are generated from nonlinear dynamical systems. A comprehensive comparison between chaotic and conventional PN codes is made in [4], [5]. The comparison shows that chaotic spreading codes are a strong competitor for the conventional spreading codes and outperform conventional spreading sequences in terms of ease of generation and flexibility in choosing the length of the sequence [6], [7].

Chaotic signals are suitable for spread-spectrum modulation because of their wideband characteristics, sharp autocorrelation property, and low cross-correlation values among each other. Chaotic signals usage increases data security due to their random-like nature. The chaotic sequence generator is very sensitive to initial condition values. Theoretically, this allows the generation of an infinite number of uncorrelated chaotic sequences with excellent autocorrelation property even if they are from the same chaotic map generator [6].

Several digital chaotic communication schemes have been proposed, including coherent chaos shift-keying (CSK) in [6], [8], and chaotic DS-CDMA in [9]. In chaotic DS-CDMA, the data symbols are spread using chaotic codes rather than conventional sequences. However, synchronization of chaotic signals at the receiver is not an easy operation. A non-coherent Differential Chaos Shift Keying (DCSK) is analyzed in [7], [10]. DQCSK is the QPSK version of the binary DCSK scheme. In this work, a non-coherent DQCSK modulation combined with OFDM is used. In the (OFDM-DQCSK) system, the demodulation process can be carried out without a spreading code synchronization or the usage of any channel estimators, which simplifies the system implementation.

Inspired by the system presented in [11], a combination of OFDM and direct sequence spread spectrum (DS-SS) based on differential quadrature chaos shift keying (DQCSK) is proposed. However, in this work, the proposed system remarkably outperforms the system in [11]. The proposed system allows multiuser communications and performs in the presence of multipath fading channels.

The BER performance is evaluated in the presence of AWGN, MAI interferences and multipath Rayleigh flat fading. BER simulation results are proposed and compared with the system in [11].

1.2 Related Works

In [12], a multicarrier system with DCSK modulation is proposed, that was upgraded in [10] to allow multiuser transmission capability. In the system presented in [12], a pilot signal is sent over a private subcarrier and contains a chaotic reference code, while multiple modulated data symbols are transmitted over the remaining subcarriers. The MC-DCSK system enhances energy efficiency compared to conventional DCSK, offers higher data rates, but it demands bandwidth to avoid co-channel interference. As the number of subcarriers increases, the system becomes more complex due to the high number of transmit and receive filters.

In [13], an OFDM system with DCSK modulation is proposed, which uses Fast Fourier Transform (FFT) algorithm rather than employing a bank of filters. This reduces the complexity of the system presented in [12] when using a high number of subcarriers. In [11], the author improves the system in [13] to support multiple access communication and analyzes the BER performance of a multi-user transmission in AWGN and multipath Rayleigh fading channels.

1.3 Thesis Objectives

- Design and analysis of a multi-user system that is a combination of OFDM and Direct sequence spread spectrum that achieves a considerable bit error performance in the presence multiple access interference, additive white gaussian noise and multipath fading channels.

- Exploiting the chaotic signals as an alternative and a strong competitor to conventional spreading codes such as PN codes.

- Enhancing the performance of the differential non-coherent spread spectrum-based systems that exhibits a large bit error performance degradation when using a large spreading factor

1.4 Thesis Contributions

- Introducing the proposed multiuser OFDM-DQCSK system that aims to improve the BER performance of the systems in [11], [12] and [10]. The proposed system allows multiuser communications and performs in the presence of multipath fading channels. The BER performance is evaluated in the presence of AWGN, MAI interferences and multipath Rayleigh flat fading channel. BER simulation results are proposed and compared with the benchmark system in [11].

- The proposed non-coherent system nearly reaches coherent performance by solving the problem of correlating data symbols with a noisy reference. The solution is to transmit more reference subcarriers replicas to be averaged at the receiver side so that the AWGN is averaged to a value almost zero according to the Strong Law of Large Numbers discussed in chapter 5. The multipath channel coefficients are also averaged to their mean value as well. As a result, the BER performance is improved remarkably.

- The author in [11] uses binary modulation. But using QPSK instead of binary modulation increases energy efficiency in this scheme because the reference code transmitted on the reference subcarrier is used to reference two bits instead of one. As a result, the BER performance is improved.

- The proposed system employs the comb-type pilot structure so that the reference signals can follow the changes in the channel coefficients.

- Designing a searching software to generate binary chaotic codes which are orthogonal and practical to be used in digital signal processors. A searching software is designed to find orthogonal chaotic sequences because it is observed that it is not guaranteed that all the sequences generated from a chaotic map to have a low cross-correlation value among each other. Binary orthogonal chaotic codes are generated which are practical to be used in digital signal processors.

1.5 Thesis Outline

The subsequent chapters of this thesis are organized as follows:

- Chapter 2: introduces fundamentals and basics concepts of the OFDM system.
- Chapter 3: reviews the background of the direct sequence spread spectrum.
- Chapter 4: presents the chaotic digital communication systems. The chaotic map generator system is presented briefly.
- Chapter 5: illustrates the proposed multiuser OFDM-DQCSK system structure and the key concepts behind this system. The transmitter and receiver structures of the multiuser OFDM-DQCSK system are explained in detail
- Chapter 6: Analysis and simulation results of the proposed systems.
- Chapter 7: Conclusions and summary of the results of this research.
- Future work: Recommendations and ideas to further improve this work.
- References

Chapter 2
Fundamentals of OFDM

2.1 Introduction

Orthogonal Frequency Division Multiplexing (OFDM) is a special kind of multicarrier modulation system that is the key behind many high-speed systems such as 4G mobile communications (LTE), WiFi (IEEE 802.11a, g, n, ac) and WiMAX (IEEE 802.16) [14]. In OFDM, the information is sent on multiple subcarriers within the available channel bandwidth rather than a single carrier. This is called multicarrier communication system. The main reason behind this, is to reduce the inter-symbol interference (ISI) distortion and therefore, enhance the performance to a large extent compared to single carrier modulation. OFDM can yield transmission rates very near to the channel capacity [1].

OFDM systems are particularly used when communicating over dispersive transmission media such as wireless channels. This is done without employing a complex channel equalizer to compensate for the frequency and time-domain channel quality variations of the wireless channel [2]. In the time domain, inter-symbol interference (ISI) between successive OFDM symbols is eliminated with the addition of an appropriate cyclic prefix. The ISI of the same symbol can be eliminated using a simpler equalizer. In the frequency domain, the frequency selective channel can be converted to a non-frequency selective channel so that the receiver operation can be simplified.

In non-ideal linear channels (e.g., multipath fading channels), the time dispersion of the channel is generally much greater than the reciprocal of the symbol rate. hence, ISI is introduced which degrades performance compared with the ideal channel. The frequency response characteristics of such a channel is not constant over the entire bandwidth of the transmitted signal. The degree of performance degradation depends on the frequency response characteristics of the non-ideal channel. Furthermore, as the span of the ISI increases, a more complex equalizer is necessary to compensate for the channel distortion which is not a trivial task. For a single carrier system, one solution to the problem of the ISI distortion is to lower the symbol rate (i.e., to increase the symbol time) and hence the span of the ISI decreases. In this way, a simpler equalizer can be used. But decreasing the symbol rate is undesirable. Multi-carrier modulation is an alternative approach to eliminate the inter-symbol interference and to design a bandwidth-efficient communication system in the presence of a non-ideal channel. This can be done by subdividing the available channel bandwidth into several subchannels of relatively narrow width so that each subchannel is almost ideal channel [1].

2.2 A Low Data Rate Signal in a Multipath Channel

Consider a low bit rate signal with only 100 kbps (i.e., the bit duration is $10us$). This signal is shown in Fig. 2-1.

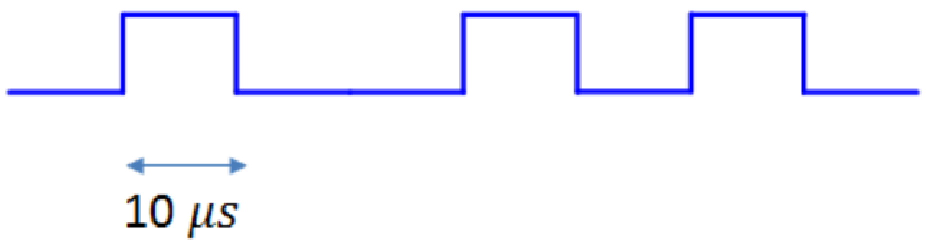

Fig. 2-1. A low bit rate baseband signal in ideal channel.

The signal in Fig. 2-1 is transmitted through a multipath fading channel. Assume that this signal arrives at the receiver from three different paths. The direct path is the first one. The second path is delayed by $1us$ after the direct path. The third path is delayed by $1us$ after the second path. The base band model is considered in this discussion. The three signals will arrive at the receiver side with different phase shifts and attenuations. These attenuations depend on the nature of the path the signal goes through. Phase shifts depend on the delay of the signal with respect to other multipath signals. The three multipath signals are shown in in Fig. 2-2.

Fig. 2-2. A low bit rate signal in multipath fading channel.

In Fig. 2-2, It is observed that the multipath components distort only 20% of bit duration of the original signal. If the bit duration is elongated to $20us$, then the multipath distortion would affect only 10% of bit duration of the original signal. One bit through its last path will interfere slightly with just one bit from another path to a little extent but not with any other bit farther than that. The inter-symbol interference among the bits themselves is considered negligible with longer bit durations. In conclusion, decreasing the bit rate and elongating the bit duration can reduce the multipath fading effect. A simple equalizer can handle this situation without heavy processing. The wireless channel does not add a serious problem to the Rx processing time in this low data rate scenario. In the case of destructive interference, a diversity technique can be employed to the transmitted signal to solve this problem. Diversity is to send one or more signal replicas along with the original signal in some form whether in frequency, time, space, etc. But elongating the bit duration is not recommended because a high bit rate is always needed to meet the requirements of today's technology.

2.3 A High Data Rate Signal in a Multipath Channel

Consider a bit rate of 10 Mbps is required in a high-speed communication system. This is equivalent to $0.1\ us$ bit duration. This signal is shown in Fig. 2-3.

Assume that the same channel used in section 2.2 is employed. This means that the same environment (i.e., the physical medium) is present. The environment is independent of the used bit rate. In other words, the second and the third multipath components will still arrive with delay of $1\ us$ and $2\ us$ respectively after the direct path.

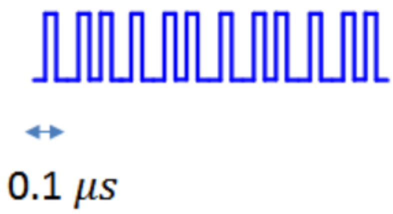

Fig. 2-3. A high bit rate signal in ideal channel.

In Fig. 2-4 this situation is explained, and the RF carrier is ignored for simplicity. As observed in Fig. 2-4, the initial bits of the direct path signals are interfering with many future incoming bits from the other paths with different attenuations and destructive and constructive interference through the signal span.

Fig. 2-4. A high bit rate signal in multipath fading channel.

As opposed to the case in section 2.2 where only a small portion of the bit has interference with certain phase shift. In some cases, this inter-symbol interference would extend to thousands of bits. Therefore, in this case, a more complex equalizer is needed. The design of such complex equalizers is not a trivial task and requires a high processing power. **In conclusion,** the same harmless channel for low bit rate communication system has become severe for the high bit rate system.

2.4 Equalizer Solution for ISI Elimination

An equalizer is a fundamental component that is employed at the receiver to make the communication system functional. An equalizer is not necessarily a physical device, but it can be lines of code in a digital signal processor following the concept of software defined radio (SDR).

An equalizer is a filter that is used mainly to eliminate the Inter-Symbol Interference (ISI) and combat the effects of multipath fading channel on the received signal. The input of the equalizer is the corrupted signal waveform (summation of all multipath components and the direct path signal) and the output is the clean desired bit stream as shown in

Fig. 2-5. Equalizer operation.

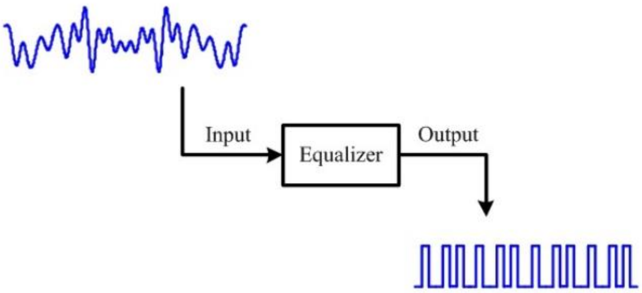

Fig. 2-5. Equalizer operation.

However, equalizer is the most intensive consuming component of the receiver processing resources especially for a high data rate system. It can use more than 75% of the processing power of the digital signal processor. Additionally, in a very high-speed communication system, the equalizer may be not fast enough to perform its operation in real time. it not practical to be busy processing a bit in a certain window of time while many future bits are arriving at the receiver. As a result, the buffer filling will be faster than being emptied and some messages will be dropped.

In conclusion, if the communication system uses a low data rate, the equalizer operation is simplified, and a low processing power is consumed. However, for a high bit rate stream, a heavy-duty receiver processor is required in addition to the complex design of the equalizer. Heavy processing requires high power consumption and therefore, not suitable for battery operated devices like smart phones.

2.5 Advantages of OFDM

- OFDM technique is mainly employed in the communication system to remove the inter-symbol interference (ISI) completely.

- Achieving a high communication rate with a simple equalizer that requires low processing resources and little power consumption, OFDM system is employed.

- Because the symbol period has been increased, the channel's delay spread becomes a significantly shorter fraction of a symbol period than in the serial system, potentially rendering the system less sensitive to channel-induced dispersion, than the conventional serial system.

- One of the most attractive features of this scheme is that the bandwidth of the subchannels is very narrow when compared to the communications channel's coherence bandwidth. Therefore, flat-fading narrowband propagation conditions apply. The sub-channel modems can use almost any modulation scheme, and QAM is an attractive choice in some situations.

- OFDM technique is mainly employed in the communication system to remove the inter-symbol interference (ISI) completely.

2.6 Causes of Inter-symbol Interference

ISI distortion may occur for many reasons. One of these reasons is the time dispersion of the transmitted symbols within a band-limited channel which has a non-ideal frequency response. Hence, the ISI would result when trying to transmit symbols at rates comparable to or exceeding the bandwidth of the channel [1]. Another source for the inter-symbol interference may come from a time-varying multipath fading channel. In multipath fading channels, time dispersion and, therefore, inter-symbol interference are the result of multiple propagation paths with different path delays. The number of paths and the relative time delays among the paths vary with time [1]. The fading channel can be frequency-selective or frequency nonselective. In frequency selective fading channels, the coherence bandwidth of the channel is smaller than the bandwidth of the signal (i.e., the channel is not flat). Therefore, the signal undergoes different channel gains and delays that results in a severe distortion. Another important usage of OFDM is that it can combat the severe distortion caused by the frequency selective multipath fading channel. This is done by converting the frequency selective fading channel into non-frequency selective flat fading channel. With the addition of an appropriate cyclic prefix, the ISI can be removed completely.

Without OFDM, a complex channel equalizer should be designed to compensate for the distortion caused by inter-symbol interference (ISI). Designing such an equalizer especially if the span of the ISI extends across many symbols is not a trivial operation. **From the previous discussion,** if the non-ideal channel can be converted to ideal channel, then the ISI can be reduced and eliminated with simplified operations.

2.7 Theory of Operation

The key idea behind an OFDM system is to subdivide the entire bandwidth of signal into many equally very narrow sub-bands so that the channel frequency response is considered constant (i.e., ideal) across each sub-band. This idea is illustrated in

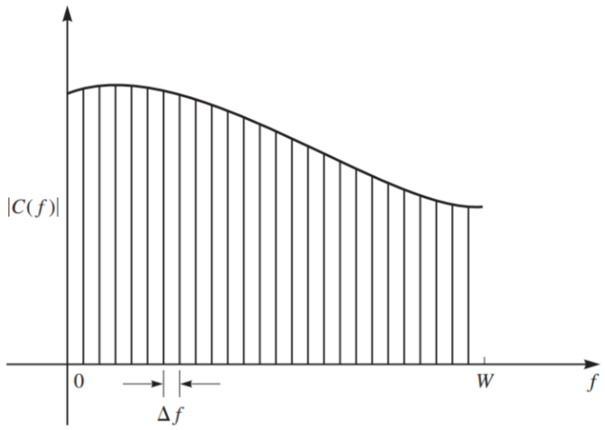

Fig. 2-6.

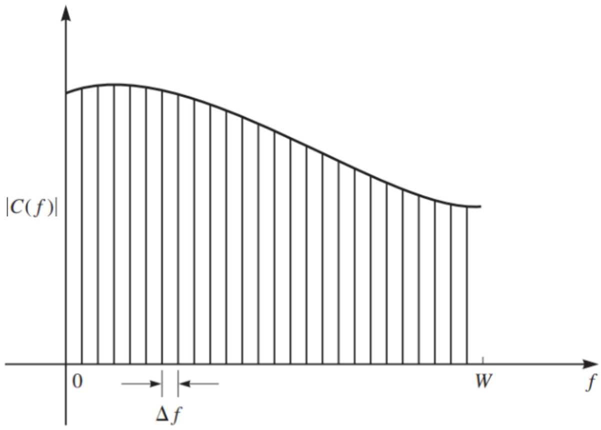

Fig. 2-6. Subdivision of the channel bandwidth into narrowband subchannels.

Given that W is the bandwidth of the original signal, Δf is the bandwidth of each sub-band and N is the number of sub-bands (i.e., subcarriers), then

$$\Delta f = \frac{W}{N} \tag{1.1}$$

Δf should be narrow enough so that the frequency response of the channel is constant across each sub-band and each subchannel is almost ideal. As shown in

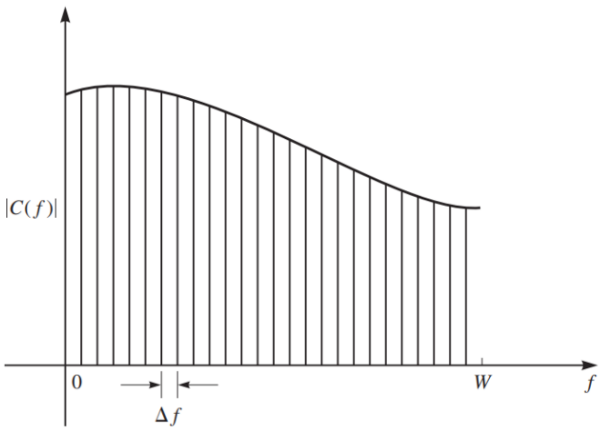

Fig. 2-6, Each subchannel becomes non-dispersive. Therefore, inter-symbol interference is negligible [1]. As a result, no need to add a complex channel equalizer at the receiver side. The transmitted signal within each sub-band may be separately modulated and coded at a synchronized symbol rate.

In the single carrier modulation system, the modulated signal occupies all the available bandwidth. The data symbols are passed directly to the modulator transmitting at a carrier frequency positioned at the center of the transmission band. If the coherence bandwidth of the channel is longer than the bandwidth of the transmitted signal, a multipath fading effect of a mobile channel can cause unacceptable burst of transmission errors especially if maximum multipath delay extends over the duration several bits. In contrast to the single carrier system, During the duration of the composite OFDM signal of N symbols, each subcarrier carriers only one symbol. Each of which has an N times longer duration. Therefore, with the same duration channel fade, only a fraction of the duration of each of the extended-length subcarrier symbols transmitted in parallel is affected. Therefore, the OFDM system may be able to recover all of the partially fading-contaminated N subcarrier symbols. Thus, while the serial system exhibits an error burst, no errors or few errors may occur using the OFDM approach. In conclusion, the OFDM system elongates the symbols duration and transmits the serial data in parallel orthogonal subcarriers to make the multipath fading effect negligible. As the number of the subcarriers increase, the more the system become more robust against the multipath effect and hence the ISI.

2.8 Orthogonality of OFDM Subcarriers

An essential feature of OFDM is that the subcarriers of the corresponding subchannels are mutually orthogonal [1]. Assume that within each sub band, a sinusoidal subcarrier signal is sent which has the form

$$s_i(t) = cos\,(2\pi f_i t), \qquad i = 0, 1, \ldots, N - 1 \tag{1.2}$$

where f_i is the subcarrier center frequency in the i^{th} subchannel.

To achieve orthogonality among subcarriers over a symbol interval T, the symbol rate $1/T$ in each of the subcarriers is selected to be equal to the frequency separation Δf of the adjacent subcarriers. Furthermore, the orthogonality does not depend on the relative phase relationship among subcarriers. That is,

$$\int_0^T \cos(2\pi f_i t + \phi_i)\cos(2\pi f_k t + \phi_k)dt = 0 \qquad (1.3)$$

where $f_i - f_k = \frac{n}{T}$, $n = 1,2,3, \dots \dots \dots N-1$, independent of the values of the phases ϕ_i and ϕ_k.

2.9 OFDM Operation in Time Domain Using Parallel Filters

In time domain, OFDM breaks one serial fast bit stream into many parallel slow bit streams. Then, these parallel slow bit streams are multiplied with orthogonal sinusoids [14]. The operation of OFDM in time domain is illustrated Fig. 2-7 in which traces the following steps:

1. Assume the bit duration is T and it is required to send 9 bits. In a conventional single carrier system, it will take $9T$ to transmit all bits in a fast serial bit stream communication system.
2. In an OFDM system, the first step in an is to break them down to 9 slow parallel bit streams each with a new bit duration of $9T$ seconds. This is equivalent to a fast serial bit stream of bit duration T.
3. Assume a fundamental frequency of $f_0 = \frac{1}{9T}$. Then, this sinusoid will be orthogonal to 8 other sinusoids with frequencies $0f_0, 2f_0, 3f_0, \dots, 8f_0$ according to equation (1.3) . This set of 9 sinusoids are called subcarriers which are shown in
4. Fig. 2-7.
5. Assume the bit duration is T and it is required to send 9 bits. In a conventional single carrier system, it will take $9T$ to transmit all bits in a fast serial bit stream communication system.

6. In an OFDM system, the first step in an is to break them down to 9 slow parallel bit streams each with a new bit duration of $9T$ seconds. This is equivalent to a fast serial bit stream of bit duration T.

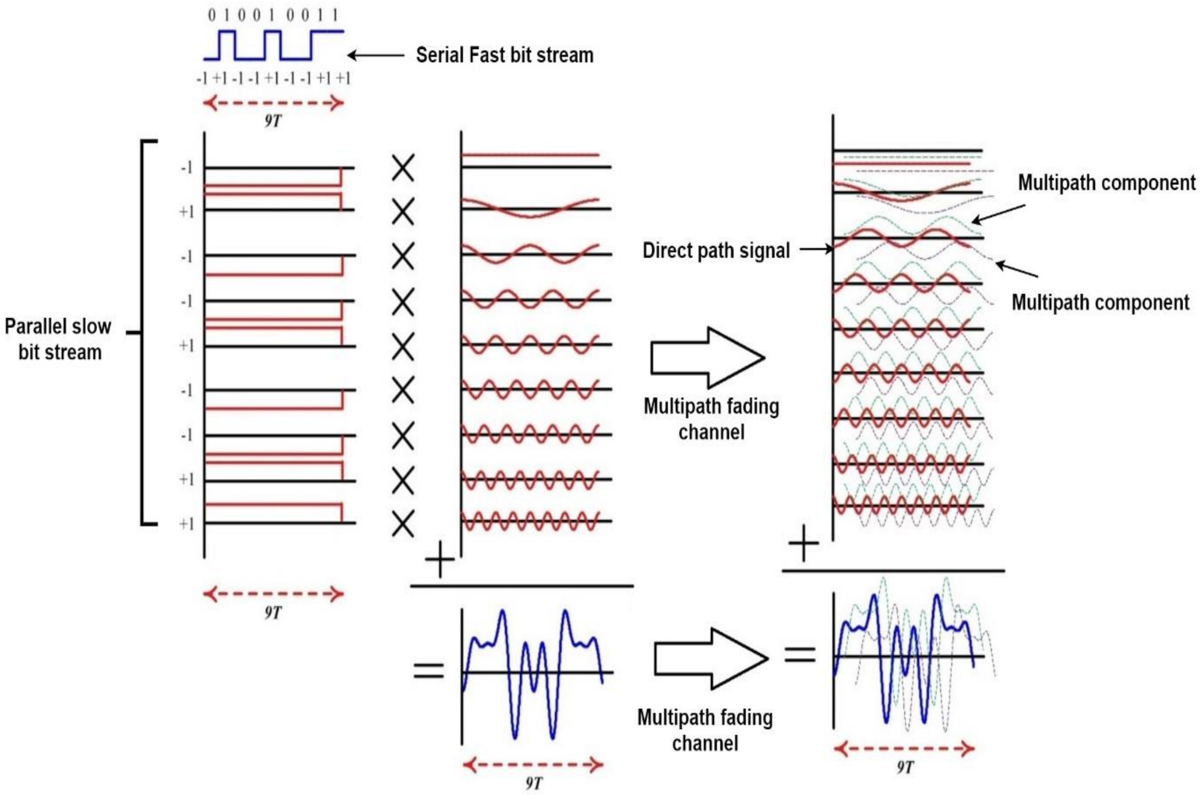

Fig. 2-7. OFDM symbol in time domain.

7. Assume a fundamental frequency of $f_0 = \frac{1}{9T}$. Then, this sinusoid will be orthogonal to 8 other sinusoids with frequencies $0f_0, 2f_0, 3f_0, \ldots, 8f_0$ according to equation (1.3) . This set of 9 sinusoids are called subcarriers which are shown in Fig. 2-7.
8. Next, we modulate each such sinusoid with $+1$ or -1 according to the data bit. A nine transmit filters are required.
9. Finally, all these amplitude-scaled sinusoids are added together to generate the desired signal.

Notice that this composite signal has a duration of $9T$ seconds (same duration as the original fast serial bit stream) but contains the information from all 9 bits [14]**.**

2.10 OFDM Block Diagram Using Parallel Filters

One way to implement OFDM system is shown in Fig. 2-8. This is the simplest version of an OFDM system. In this basic scheme, the system has N sub-bands with a suitable large guard band that separates the adjacent sub-bands to prevent interference between the nearby signals. A memory buffer stores symbols from the incoming input serial data stream with symbol rate $f_s = \frac{1}{\Delta t}$. when the buffer is filled with N symbols, a serial to parallel conversion is done to split the serial data in one traffic channel into N parallel subchannels. There should be another buffer to collect a new N symbols while the previous N symbols are being processed in the other buffer. This prevents the data from being overwritten. The data in each subchannel is applied to a modulator such that there are N modulators for the N subchannels with subcarrier frequencies $f_0, f_1, \dots, f_{N-1}$. In Fig. 2-8, The symbol duration in each sub-band of the parallel OFDM system is N times longer than that of the serial system and equal to

$$T = N \,.\, \Delta t \qquad (1.4)$$

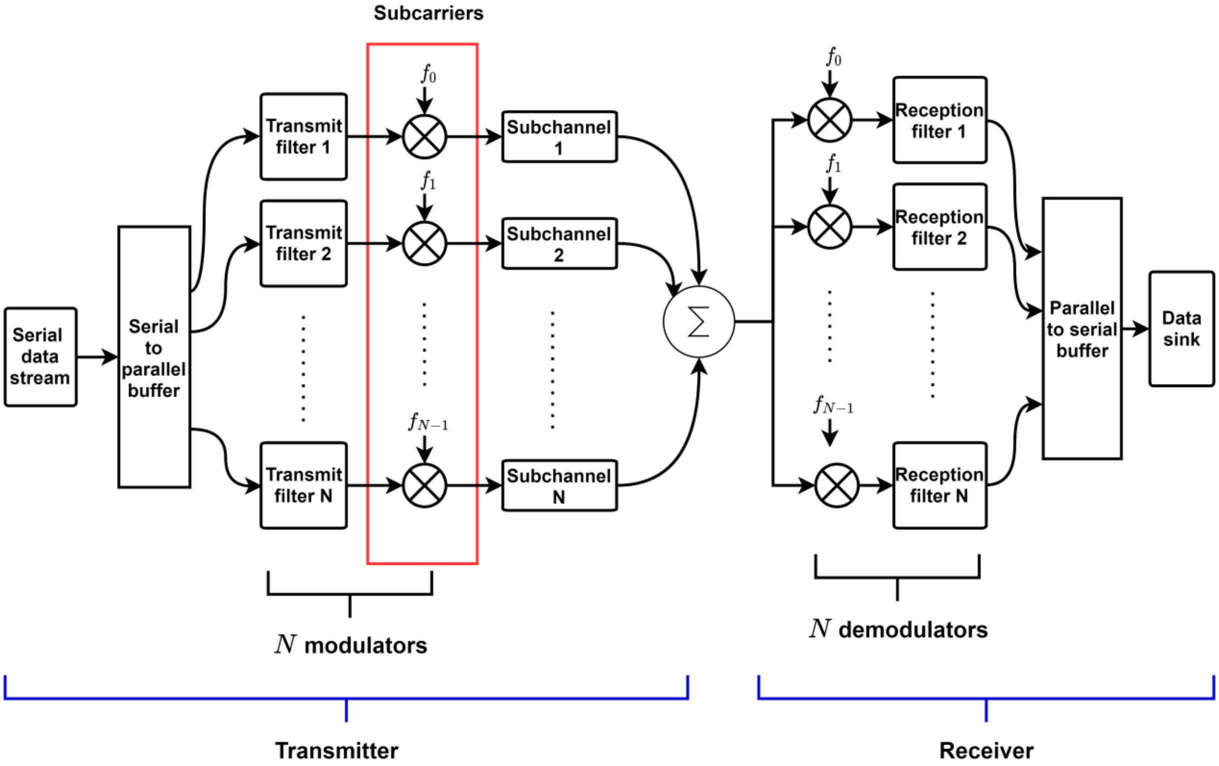

Fig. 2-8. Block diagram of OFDM modem using parallel filters

where T is the OFDM symbol duration. To achieve the orthogonality condition discussed in section 2.8, the subcarrier spacing must be equal to

$$\Delta f = 1/T \tag{1.5}$$

where Δf is the subcarrier spacing and T is the OFDM symbol duration. The sub-band carrier frequencies are given by the following equation

$$f_n = n \,.\, \Delta f \qquad n = 0,1,2,\ldots\ldots,N-1, \tag{1.6}$$

The total bandwidth of the OFDM transmitted signal is equal to

$$B = (N-1)\Delta f \tag{1.7}$$

Assume that the output of each sub-channel modulator is represented by $X_n(t)$ whose subcarrier frequency is given by (1.6). then these N modulated subcarriers are added so that an OFDM signal is generated. The OFDM signal can be represented by

$$y(t) = \sum_{n=0}^{N-1} X_n(t) \qquad n = 0,1,2,\ldots\ldots,N-1, \tag{1.8}$$

At the receiver side, the composite baseband received OFDM signal is passed through N parallel correlators to extract the N parallel symbols. A parallel to serial operation is performed. Then the serial symbols are demodulated to give a binary data stream. At the receiver side, the composite baseband received OFDM signal is correlated with N parallel subcarriers to extract the N parallel symbols. In each correlator, the data of one of the subcarriers will be extracted while the contribution of the other subcarriers will be suppressed and cancelled to zero utilizing their orthogonal property. Then a parallel to serial operation is performed. the serial symbols are demodulated to give a binary data stream.

In practice, there is some spectral spillage due to adjacent frequency sub-bands which reduces this efficiency. Spectral spillage due to the sub-bands at the top and bottom of the overall frequency band requires a certain amount of guard space between adjacent users. Furthermore, spectral spillage between OFDM sub-bands due to the imperfections of each of the sub-band filters requires that the sub-bands be spaced further apart than the theoretically required minimum amount, decreasing spectral efficiency. In order to obtain the highest efficiency, the block size should be kept high and the sub-band filters made to meet stringent specifications.

A major disadvantage of the scheme shown in Fig. 2-8 is the increased complexity compared to the conventional serial single carrier system. Because it is required to employ a bank N modulators at the transmitters and N matched filters and demodulators at the receiver. As the number of subcarriers N increases, the more complex the scheme in Fig. 2-8. This complexity can be reduced by using the discrete Fourier Transform (DFT). The Fast Fourier Transform is a special case of the DFT which can further reduce the complexity of the scheme in Fig. 2-8 if the number of the subcarriers is a power of two. Instead of implementing a bank of sub-channel modems, it can be simplified using a single Fast Fourier Transformer (FFT).

2.11 OFDM Using IFFT/FFT Algorithm

The OFDM scheme shown in Fig. 2-8 has a major problem. To achieve a good performance against multipath channel, the number of subcarriers N must be very high to make each subchannel narrow enough compared to the coherence bandwidth of the channel. This requires a large number of subchannel parallel modems which complicate the design. The discrete Fourier transform (DFT) is a solution to simplify the design when using a large number of subcarriers N. The Fast Fourier Transform gives the same result as the DFT but faster and more simplified than the DFT. FFT can be employed only when N is an integer power of two.

Mathematically, it is proved that by taking the Inverse Discrete Fourier Transform of the N symbols and transmitting the IDFT coefficients serially is exactly the same as the transmitter operation in Fig. 2-8 [2]. Significant hardware simplifications can be made with OFDM transmissions if the bank of sub-channel modems is implemented using the computationally efficient pair of inverse fast Fourier transform and fast Fourier transform (IFFT/FFT).

When using IFFT/FFT operations, it is convenient to use the notation of sampling rate instead of symbol rate and sampling period instead of sampling period. The sampling rate is chosen according to the Nyquist theorem to represent analog signals in digital domain adequately [2]. To use the IFFT/FFT operation, The Nyquist criterion must be achieved as the following

$$f_s > 2\,(N-1)\,\Delta f, \tag{1.9}$$

where f_s is the sampling rate of the original symbol stream before IFFT operation, Δf is the subcarrier frequency spacing. Practically, to achieve orthogonality, the sampling rate and the subcarrier spacing relationship is calculated as the following

$$f_s = M\,\Delta f, \tag{1.10}$$

implying that the sampling rate of the original serial sequence f_s must be an integer multiple of the subcarrier spacing Δf with the following condition

$$M > 2(N-1), \tag{1.11}$$

According to the sampling theorem, The sampled signal repeats itself at multiple of the sampling rate f_s.

2.11.1 OFDM Transmitter Structure Using IFFT/FFT Algorithm

The block diagram of a basic OFDM transmitter is shown in Fig. 2-9.

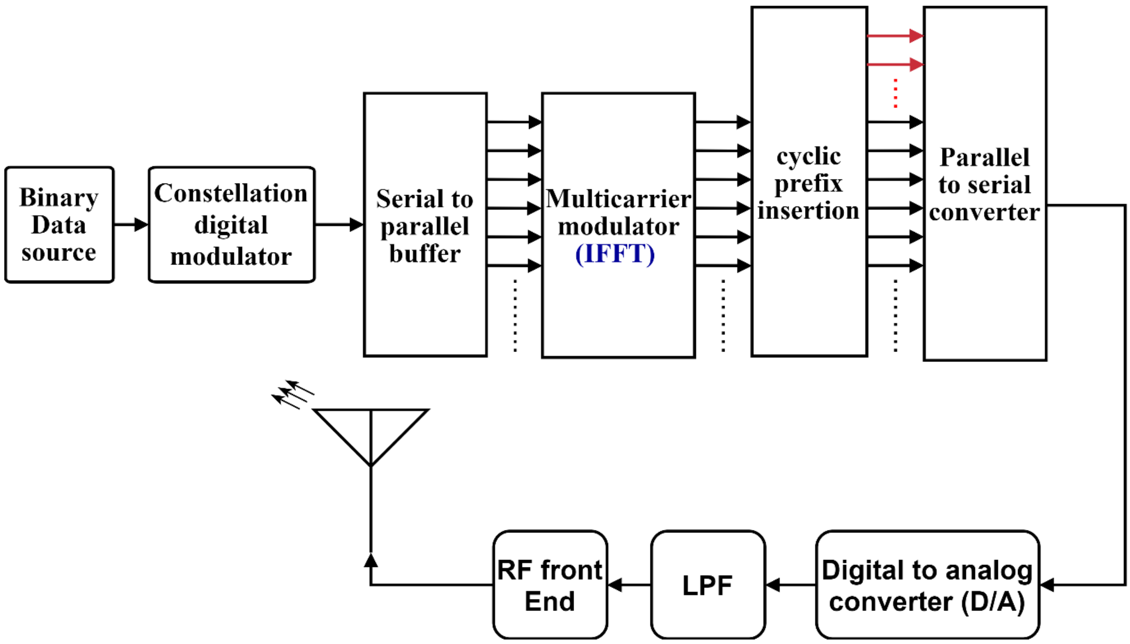

Fig. 2-9. OFDM transmitter block diagram

The serial data digital binary stream is applied to a digital modulator (maybe a QAM modulator). The modulated samples are then stored in a buffer until it is filled with N samples. After that, an IFFT operation is done to the N samples to perform an orthogonal multicarrier modulation. There should be another serial to parallel buffer that is ready to store the other incoming serial modulated samples while the other buffer contents are applied to the IFFT operation. This should prevent the buffer contents from being overwritten during the IFFT operation that may take some time depending on the processor speed. Cyclic prefix samples are added to the parallel data after the IFFT operation to further elongate the OFDM symbol time and remove completely the inter-symbol interference (ISI). Next, a parallel to serial operation is performed to extend the OFDM symbol period by $(N + N_{cp})$ compared to the symbol period in a single carrier modulation where N is the FFT size (i.e., the number of subcarriers) and N_{cp} is the number of cyclic prefix samples. So far, the signal that is synthesized in the previous operations is a digital signal which is not suitable for transmission through a physical medium. Therefore, the output of the serial to parallel converter is fed to a D/A converter to transform the signal into analog form. However, this analog signal may contain some residual images of the original spectrum at multiple of the sampling rate due to the periodicity of the digital signal. A low pass filter with certain specifications should attenuate these images. The low pass filter is also called a reconstruction filter. After the low pass filter, the signal is fed to the radio frequency front end. Some of the most important operations at this stage are the following: firstly, the signal is up converted to the required center frequency maybe using a local oscillator. Then the signal is amplified with a suitable gain using a power amplifier so that the signal can travel a considerable distance with high signal to noise ratio. Particularly in the OFDM modulation system, the signal after the D/A should be maintained below the saturation level of the power amplifier input or undesired harmonics can distort the operation of the transmitter. A harmonic suppression filters can be inserted after the power amplifier to attenuate the harmonics to a considerable level. The signal is then fed to the antenna to be transmitted on the physical medium.

2.11.2 OFDM Receiver Structure Using IFFT/FFT Algorithm

The block diagram of a basic OFDM receiver is shown in Fig. 2-10. Before performing any processing on the signal received from the antenna, the received signal enters the RF

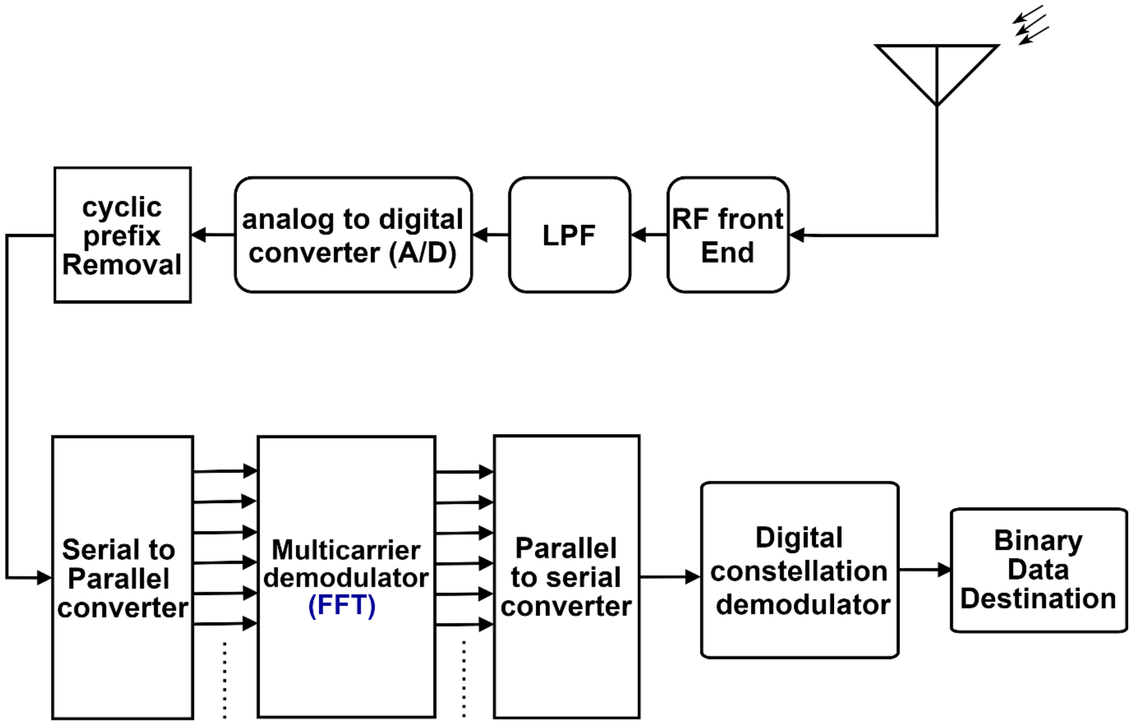

Fig. 2-10. OFDM receiver block diagram

front end stage. In this stage, a band pass filter is tuned to the desired frequency band in which the signal has its center frequency for many reasons. One of these reasons is to reject any other undesired signals in the other channels that is added to the desired signal in the time domain. Another reason is to reduce the additive white noise power that enters the receiver that is theoretically distributed equally across the frequency spectrum. Then a low noise amplifier is employed to give the received signal a considerable gain. After that, the signal is down converted using an analog local oscillator to a suitable center frequency convenient to be sampled by the base band processor. It is recommended not to down convert to the DC frequency. Semi-conductors such as CMOS technologies make very much of flicker noise which occurs at low frequency especially at DC [15]. Instead, down converting to a suitable intermediate center frequency (IF frequency) is performed. The signal is then passed through analog anti-aliasing filter before going through the analog to digital converter (A/D) to remove any undesired signal more than one half of the sampling rate according to the sampling theory of Nyquist. To begin processing on the signal, it must be converted to a digital form that the base band processor can work with. This done using an analog to digital converter (A/D). In this stage, the signal is now sampled and buffered by the base band processor. Firstly, the cyclic prefix samples are removed from the OFDM symbol. These samples are redundant samples to eliminate the inter-symbol interference (ISI). After that, the incoming stream of samples of are buffered until the entire N samples of an OFDM symbols are received. Then a serial to parallel operation is performed to the buffered N sampled to be ready for an FFT operation. There should be another buffer that collects the next N samples while the FFT operation is done to the contents of the other buffer. This is because the FFT operation may take some time according to the base band processor speed. As a result, this should prevent the contents of the buffer from being overwritten during the FFT operation. An FFT operation is done to every block of N samples. The FFT operation is equivalent to a multicarrier demodulator with parallel matched filters for the data subcarriers. After the FFT operation, the modulated symbols are now recovered with any possible distortion like AWGN, multipath fading, multiple access interference, etc. the distorted modulated symbols are applied then to a constellation demodulator (i.e., detector) to decide which symbol was actually transmitted. A good detector gives a low symbol error probability (SER). The SER depends on many factors line the signal to noise ratio in the analog front end or equivalently the energy per symbol to the power spectral density of the noise (E_s/N_0) in the digital environment. Besides the SER criterion, the bit error probability (BER) is a standard criterion to measure the performance of the digital system. The lower the BER, the better the performance of the system. It is usually measured according to the bit energy per power spectral density of the noise (E_b/N_0). Finally, each symbol is mapped to its binary equivalent to be ready for usage in the destination device.

2.12 Sources of Interference in OFDM

There are two major sources of interference that can be detected when using OFDM.

1. The first one is the inter-symbol interference (ISI). ISI is described as the crosstalk between signals of successive FFT frames in the same subchannel separated by an OFDM signaling interval. ISI is a time domain distortion. The proper usage of a suitable cyclic prefix may completely remove the ISI.

2. The second type of interference is the inter-channel interference (ICI). ICI is defined as the crosstalk between adjacent subchannels of the same FFT frame. If the orthogonality between the OFDM subcarriers is lost, ICI would happen and causes a significant BER performance degradation.

2.13 Cyclic OFDM Symbol Extension

A cyclic prefix (cyclic extension) is a guard interval of N_{cp} samples that is employed to combat the dispersive channels and eliminate the inter-OFDM symbol interference. The time dispersion may occur due to a multipath fading channel or an inherently non ideal channel with non-ideal frequency response.

For an IFFT multicarrier modulation, N frequency domain samples produce N time domain samples. Practically the bandwidth of the OFDM spectrum should be finite. As a result, the time domain signal is assumed to be of infinite duration. Under this assumption and based on the theory of the IFFT operation, it is assumed that the time domain signal should be periodically repeated for an infinite duration. In practice, however, it is sufficient to repeat the time domain signal periodically for the duration of the channel's memory, i.e. for a duration that is comparable to the length of the channel impulse response (CIR). Therefore, to combat the ISI due to the transmission through the time dispersive channels, the OFDM symbol is extended by a cyclic extension in the time domain. The insertion of a cyclic prefix (i.e. a cyclic extension) is shown in Fig. 2-11.

The samples of the cyclic extension are added after the IFFT operation. They are copied from the end of the N useful data samples. The samples of the cyclic extension are copied from the end of the time domain OFDM symbol. Assume that the time domain signal samples after the IFFT operation are $\{s_0, s_1, s_2, \dots \dots \dots \dots, s_{N-1}\}$. Then after adding the cyclic prefix, the signal becomes $\{s_{N-N_{cp}-1}, \dots \dots \dots \dots., s_{N-1}, s_0, s_1, s_2, \dots \dots \dots \dots, s_{N-1}\}$. At the receiver, the cyclic extension samples are ignored and removed. The length of the cyclic prefix depends only on the channel memory (i.e., the maximum time dispersion of the channel).

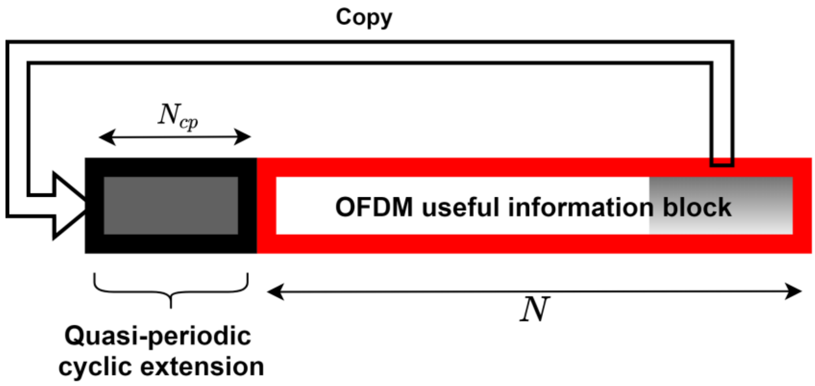

Fig. 2-11. Cyclic extension in an OFDM symbol

The length of the cyclic prefix should be at least equal to the maximum delay spread of the multipath fading channel or the channel memory to remove the ISI as shown in Fig. 2-12.

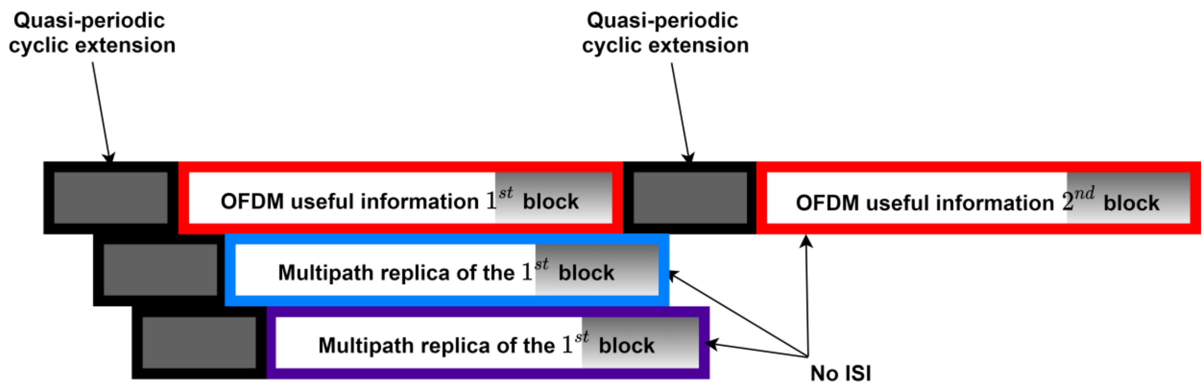

Fig. 2-12. ISI elimination due to cyclic prefix in a multipath fading channel

Increasing the length of the cyclic prefix beyond this limit should be considered carefully because the cyclic prefix has many drawbacks. One drawback for the usage of the cyclic prefix to combat the time dispersive channels is the reduction of the OFDM transmission rate by $N/(N+N_{cp})$. A degradation in the signal to noise ratio is observed. The degradation of the signal-to-noise (SNR) ratio would be

$$SNR_{lost} = -10\ \log_{10}\left(\frac{N}{N+N_{cp}}\right), \tag{1.12}$$

As the length of the cyclic extension increases, the energy efficiency becomes lower. One way to increase the energy efficiency is to employ a high number of subcarriers N with taking into consideration the inter carrier interference (ICI). Practically a cyclic extension of not more than 10% of the OFDM symbol's duration is used.

2.14 Summary

In time domain, OFDM converts one serial **fast** bit stream into many parallel **slow** bit streams. In frequency domain, OFDM segments one **wide** spectrum into many **narrow** spectra. OFDM converts a frequency selective channel into a non-frequency selective channel that requires a simple equalizer. The use of a cyclic prefix in OFDM systems eliminates completely the inter symbol interference. Therefore, OFDM system eliminates the need to design a complex equalizer to fight the ISI. The use of the discrete Fourier transform algorithm significantly reduces the integration complexity of the OFDM system especially when a high number of subcarriers are employed. An IDFT operation replaces a bank of parallel transmit filters and modulators at the transmitter and a DFT operation replaces a bank of correlators and demodulators at the receiver. The fast Fourier transform provides a simpler and fast operation than the discrete Fourier transform, however, the fast Fourier transform required the number of subcarriers to be an integer power of two. The inter carrier interference (ICI) should be considered carefully when a doppler spread is recognized. The doppler frequency shift destroys the orthogonality between the frequency subcarriers leading to a significant degradation in the BER performance.

Chapter 3
Spread Spectrum Modulation

3.1 Introduction

Spread spectrum signals are distinguished by an important feature. Their bandwidth which is used for the transmission of digital information is much more than the bit rate. The excess bandwidth redundancy inherent in spread spectrum signals is required to combat many sources of interference that are encountered in the transmission of digital information over some radio and satellite channels [1]. a spread-spectrum signal spreads its power over a much wider bandwidth. Therefore, the average power spectral density becomes much lower and the signal is easily hidden in the background noise [6].

The spreading factor L_c should be much greater than unity. The most efficient method for introducing redundancy and expanding the bandwidth of the original signal is the usage of coding since coded waveforms are characterized by a bandwidth expansion factor greater than unity. It follows that coding is an important element in the design of spread spectrum signals and systems [1].

The second important key feature that characterize spread spectrum signals is pseudo randomness which makes the signals appear like random noise and difficult to demodulate by receivers other than the intended ones [1].

There are two major techniques used to design a spread spectrum modulation system. One of them is called Direct sequence spread spectrum (DS-SS) and the other is frequency hop spread spectrum (FH-SS).

3.2 Spread Spectrum Modulation Advantages

- Combat and suppress the severe effects of the ISI (i.e., self-interference) due to multipath propagation. Resolvable multipath components resulting from time-dispersive propagation through a channel may be viewed as a form of self-interference.

This type of interference may also be suppressed by the introduction of a pseudorandom pattern in the transmitted signal [1].

- Reduce the interference due to jamming and interference arising from other users of the channel [1].

- Hiding a signal by transmitting it at low power and, thus, making it difficult for an unintended listener to detect in the presence of background noise achieving message privacy in the presence of other listeners. A message may be hidden in the background noise by spreading its bandwidth with coding and transmitting the resultant signal at a low average power. Because of its low power level, the transmitted signal is said to be “covert.” It has a low probability of being intercepted (detected) by a casual listener and, hence, is also called a low probability-of-intercept (LPI) signal [1].

- Interference from the other users arises in multiple-access communication systems in which several users share a common channel bandwidth. At any given time, a subset of these users may transmit information simultaneously over the common channel to corresponding receivers. Assuming that all the users employ the same code for the encoding and decoding of their respective information sequences, the transmitted signals in this common spectrum may be distinguished from one another by superimposing a different pseudorandom pattern, also called a code, in each transmitted signal. Thus, a particular receiver can recover the transmitted information intended for it by knowing the pseudorandom pattern, i.e., the key, used by the corresponding transmitter. This type of communication technique, which allows multiple users to simultaneously use a common channel for transmission of information, is called code division multiple access (CDMA) [1].

- Finally, message privacy may be obtained by superimposing a pseudorandom pattern on a transmitted message. The message can be demodulated by the intended receivers, who know the pseudorandom pattern or key used at the transmitter, but not by any other receivers who do not have knowledge of the key [1].

3.3 Direct Sequence Spread Spectrum (DS-SS) Principles

(DS-SS) aims to make the information bearing signal occupy a bandwidth that is much more than the minimum bandwidth necessary to transmit it. To achieve this principle, two stages of modulation are used. First, the incoming data sequence is used to modulate a wideband code. This code transforms the narrowband data sequence into a noise like wideband signal. The resulting wideband signal undergoes a second modulation using a phase-shift keying technique [16]. Therefore, this technique relies on the availability of a noise like spreading code. In fact, there are many variations of the spreading codes that can be used such as pseudo-noise sequence (PN sequence), Gold codes, Walsh codes, chaotic codes, etc.

The desired spread spectrum modulation is achieved by applying the data signal and the code signal to a product modulator or multiplier. We know from Fourier transform theory that multiplication of two signals produces a signal whose spectrum equals the convolution of the spectra of the two component signals. Thus, if the message signal is narrowband and the code signal is wideband, the product signal will have a spectrum that is nearly the same as the wideband code signal. In other words, the code sequence performs the role of spectrum spreading [16]. By multiplying the information-bearing signal by the code signal, each information bit is “chopped” up into a number of small-time increments, as illustrated in the waveforms of Fig. 3-1. These small-time increments are commonly referred to as chips. The DS-SS system is designed such that the chip rate is much faster than the data rate. Therefore, the bandwidth of the resulting "spread signal" is determined by the chip rate, which is much larger than the binary data bandwidth [6].

Assume that the code signal waveform is represented by $c(t)$ and the phase shift keying modulated data signal is $m(t)$ and the spread data signal is s(t). the chip period is T_c and the symbol period is T_s then for a spread spectrum signal, the spreading factor (i.e., the bandwidth expansion factor) is calculated as follows

$$L_c = \frac{T_s}{T_c} \tag{3.1}$$

From equation (3.1), the period of the code waveform is equal to T_s (the symbol period).

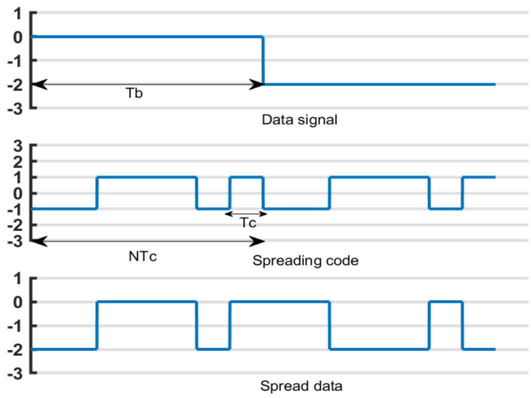

Fig. 3-1. Spread spectrum waveforms

3.4 Direct Sequence Spread Spectrum (DS-SS) Modem Block Diagram

Fig. 3-2, illustrates the block diagram of the baseband modem of the (DS-SS) modulation. Binary data stream is applied to a phase shift keying (PSK) modulator. Then the PSK modulated symbols are passed into a spread spectrum product modulator. The PSK symbols are multiplied by a wide band code with a certain spreading factor calculated using equation (3.1). The transmitted DS-SS signal may be expressed as the following

$$s(t) = m_i(t)c_i(t) \tag{3.2}$$

The subscript i represents a specific user. Assume that the channel distorts the transmitted signal with interference $i(t)$ and an AWGN. This interference maybe another user using the same channel with another different code or self-interference from a multipath fading signal. For user i, the received signal $r(t)$ may be expressed as in the following equation

$$r(t) = m_i(t)c_i(t) + \sum_{\substack{j=1 \\ j \neq i}}^{n} c_j(t)m_j(t) + n(t) \tag{3.3}$$

The second term represents interference from other users in a CDMA system using different spreading codes. The third term represents the AWGN term. Hence, to recover the original message signal $m(t)$, the received signal $r(t)$ at the receiver side of user i is applied to a demodulator that consists of a multiplier followed by an integrator (correlator), and a decision device as shown in Fig. 3-2. The multiplier is supplied with a locally generated code sequence that is an exact replica of that used in the transmitter. Moreover, we assume that the receiver operates in perfect synchronism with the transmitter, which means that code sequence in the receiver is lined up exactly with that in the transmitter.

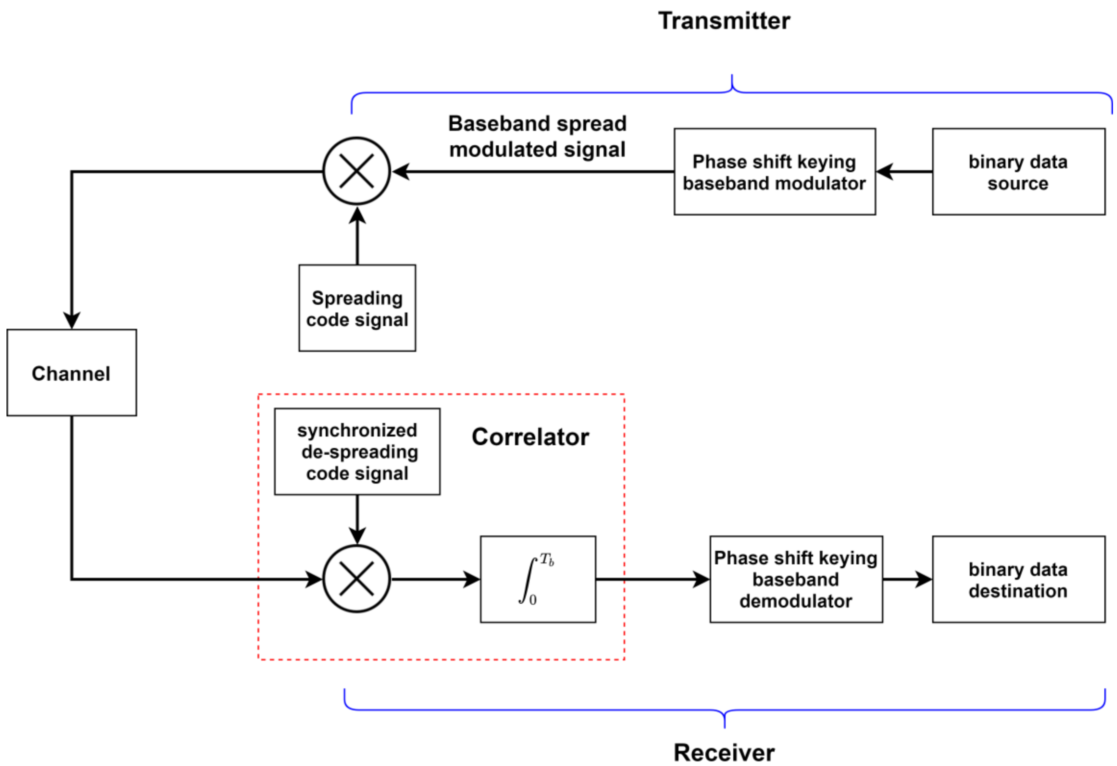

Fig. 3-2. Spread spectrum baseband modem

The multiplier output in the receiver is therefore given by

$$y_i(t) = c_i(t)r(t) \tag{3.4}$$

$$= c_i^2(t)m(t) + c_i \sum_{\substack{j=1 \\ j\neq i}}^{n} c_j(t)m_j(t) + c_i(t)n(t) \tag{3.5}$$

Equation (3.5) shows that the data signal $m(t)$ is multiplied twice by the PN signal $c_i(t\ )$ and with the integration block, it represents the code auto-correlation function of user i and gives a unity delta function, the second term of equation (3.5) represents the interference from other users which is an unwanted signal and is multiplied only once by $c_i(t\ )$. This interference after the integration block represents the cross-correlation function between user i and all other users one to one. An efficient code set should have these cross-correlation values very low and zero if they are orthogonal. If they are random like codes, there auto correlation function should be a unity delta function. Accordingly, we may simplify Equation (3.5) with the assumption that the codes for different users are orthogonal

$$y_i(t) = m(t) + c_i(t)n(t) \tag{3.6}$$

The second term in equation (3.6) represents the AWGN. It is proved that the noise at the digital constellation de-modulator remains a white Gaussian with zero mean and the spectral level of the AWGN also remains unchanged even if it is multiplied by the wideband code. Thus, the performance analysis carried out for coherent QAM and PSK detections can be applied directly and the BER probability is the same given that the spreading codes are orthogonal [17].

In the receiver shown in Fig. 3-2, the low pass filtering action is performed by the integrator that evaluates the area under the signal produced at the multiplier output. The integration is carried out for the entire symbol interval, providing a sample value to the decision device. Finally, a decision is made by the receiver using a predefined threshold according to the digital modulation type [16].

3.5 Pseudo-Noise Spreading Sequences (PN Codes)

Conventional Pseudo-Noise (PN) sequences are binary sequences, which exhibit noise-like properties. Maximal length sequences (m-sequences), Gold sequences and Kasami sequences are well known PN sequences. Other important PN sequences are considered in [18]. The characteristics of the spreading sequences play an important role in terms of the achievable system performance.

3.5.1 Properties

- *A spreading sequence should have wide-band frequency characteristics and a noise like properties.*

  A spread spectrum communication system spreads the original information signal using user specific signature code sequences. The receiver then correlates the synchronized replica of the signature sequences with the received signal, to recover the original information. Due to the noise-like properties of the spreading sequences, "eavesdropping" is not straightforward.

- *A sharp delta auto correlation function.*

  DS-CDMA exploits the code's autocorrelation properties to combat multipath fading effects and using the rake receiver to optimally combine the multipath signals of a particular user.

- *Very low cross correlation function among different code sequences.*

  The different users' codes should exhibit a low cross-correlation, which can be exploited for separating each user's signal. MC-CDMA also relies on this cross-correlation property in supporting multi-user communications.

3.5.2 Maximal Length Sequences (m-sequences)

A maximum length shift register sequence, or *m*-sequence, has length $n = 2^m - 1$ bits and is generated by an *m*-stage shift register with linear feedback as illustrated in Fig. 3-3.

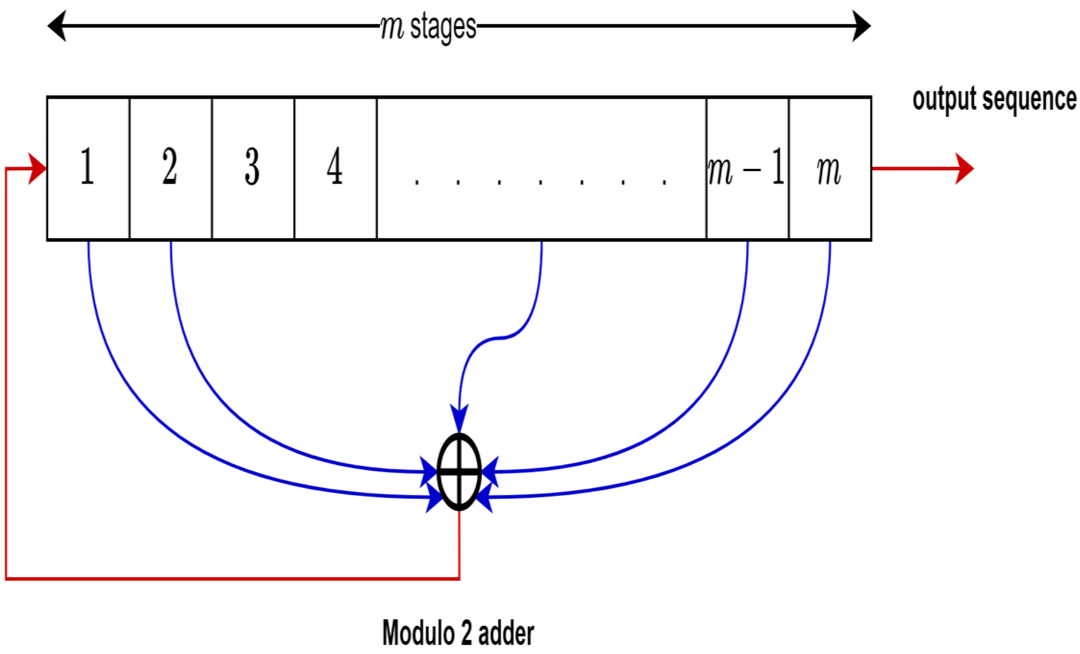

Fig. 3-3. Linear feedback shift register block diagram

A feedback shift register consists of an ordinary shift register made up of *m* flip-flops (two-state memory stages) and a logic circuit that are interconnected to form a multiloop feedback circuit. The flip-flops in the shift register are regulated by a single timing clock. At each pulse (tick) of the clock, the state of each flip-flop is shifted to the next one. With each clock pulse the logic circuit computes a Boolean function of the states of the flip-flops [16]. The result is then fed back as the input to the first flip-flop, thereby preventing the shift register from emptying. The PN sequence so generated is determined by the length *m* of the shift register, its initial state, and the feedback logic.

A feedback shift register is said to be linear when the feedback logic consists entirely of modulo-2 adders. In such a case, the zero state (e.g., the state for which all the flip-flops are in state 0) is not permitted. For a specified length *m*, this Boolean function uniquely determines the subsequent sequence of states and therefore the PN sequence produced at the output of the final flip-flop in the shift register. Each period of the sequence contains 2^{m-1} ones and $2^{m-1} - 1$ zeros such that it can be considered a noise like sequence [1].

Only some certain configurations of the feedback logic can generate an *m*-sequence. Other sequences generated with any feedback combination may not give an *m*-sequence and will give a sequence that is not very similar to a truly random sequence with a period lower than 2^m. Table 3-1 provides examples of *m*-sequences logic generator.

Table 3-1. examples of m-sequences

Length (*m*)	**Feedback logic**
2	$x^2 + x + 1$
3	$x^3 + x + 1$
4	$x^4 + x + 1$
5	$x^5 + x^2 + 1$ $x^5 + x^4 + x^2 + x + 1$ $x^5 + x^4 + x^3 + x^2 + 1$

In DS spread spectrum applications the binary sequence with elements {0, 1} is mapped into a corresponding bipolar sequence with elements {-1, 1}. A certain pulse shape is assigned to this sequence [1]. These pulses maybe rectangular, root raised cosine, chaotic, etc. For example, consider a maximal-length sequence using a linear feedback-shift register of length $m = 5$. For feedback taps, we select the polynomial $x^5 + x^2 + 1$ from Table 3-1. The corresponding configuration of the code generator is shown in Fig. 3-4.

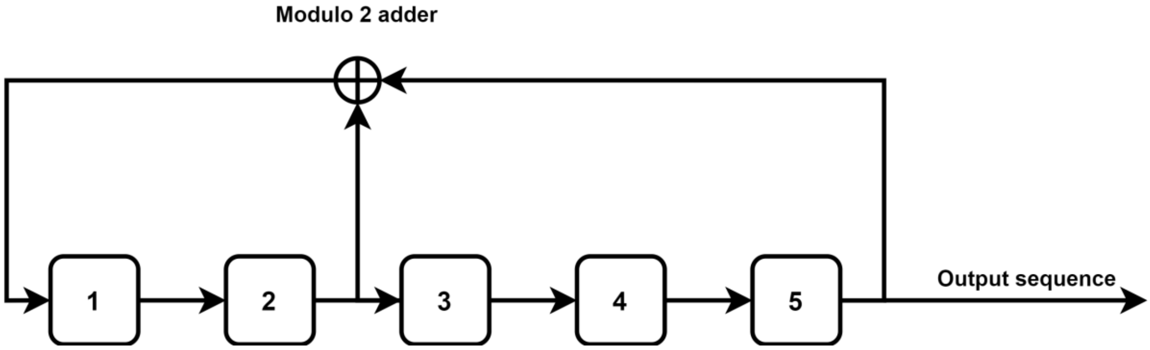

Fig. 3-4. A linear feedback shift register with five stages.

Assuming that the initial state is 10000, the evolution of one period of the maximal-length sequence generated by this scheme is shown in Table 3-2, where we see that the generator returns to the initial 10000 after 31

Table 3-2. Generating a maximal length sequence using feedback shift register.

State of the shift register					Output symbol
1	0	0	0	0	
0	1	0	0	0	0
1	0	1	0	0	0
0	1	0	1	0	0
1	0	1	0	1	0
1	1	0	1	0	1
1	1	1	0	1	0
0	1	1	1	0	1
1	0	1	1	1	0
1	1	0	1	1	1
0	1	1	0	1	1
0	0	1	1	0	1
0	0	0	1	1	0
1	0	0	0	1	1
1	1	0	0	0	1
1	1	1	0	0	0
1	1	1	1	0	0
1	1	1	1	1	0
0	1	1	1	1	1

0	0	1	1	1	1
1	0	0	1	1	1
1	1	0	0	1	1
0	1	1	0	0	1
1	0	1	1	0	0
0	1	0	1	1	0
0	0	1	0	1	1
1	0	0	1	0	1
0	1	0	0	1	0
0	0	1	0	0	1
0	0	0	1	0	0
0	0	0	0	1	0
1	0	0	0	0	1
0	1	0	0	0	0
1	0	1	0	0	0
0	1	0	1	0	0
1	0	1	0	1	0
1	1	0	1	0	1
1	1	1	0	1	0
0	1	1	1	0	1
1	0	1	1	1	0
1	1	0	1	1	1
0	1	1	0	1	1
0	0	1	1	0	1

The code generated from Table 3-2 is **0000101011101100011111001101001**

3.5.2.1 Autocorrelation function

Ideally, a pseudorandom sequence should have an autocorrelation function with the property that $R(0) = n$ and $R(j) = 0$ for $1 \leq j \leq n-1$, where n is the PN sequence periodic length. In the case of m −sequences, the normalized periodic autocorrelation function is

$$R(j) = \begin{cases} 1 & j = 0 \\ -\dfrac{1}{n} & 0 < j < n-1 \end{cases} \tag{3.7}$$

For large values of n, i.e., for long m −sequences, the size of the off-peak values of $R(j)$ relative to the peak value $R(j)/R(0) = -1/n$ is small and, from a practical viewpoint, inconsequential. Therefore, m sequences are almost ideal when viewed in terms of their autocorrelation function.

3.5.2.2 Cross correlation function

The cross-correlation properties of PN sequences are as important as the autocorrelation properties in some applications like CDMA. The periodic cross-correlation function between any pair of m sequences of the same period can have relatively large peaks. Such high values for the cross correlations are undesirable in CDMA because each user is assigned a particular PN sequence. Ideally, the PN sequences among users should be mutually orthogonal so that the level of interference experienced by any one user from transmissions of other users adds on a power basis. However, the PN sequences used in practice exhibit some correlation.

To be specific, we consider the class of m-sequences made by (Sarwate and Pursley, 1980). Table 3-3 lists the peak magnitude R_{max} for the periodic cross correlation between pairs of m-sequences for $3 \leq m \leq 12$. The table also shows the number of m-sequences of length $n = 2^m - 1$ for $3 \leq m \leq 12$. As we can see, the number of m-sequences of length n increases rapidly with m. We also observe that, for most sequences, the peak magnitude R_{max} of the cross-correlation function is a large percentage of the peak value of the autocorrelation function.

Table 3-3. Peak cross correlation of m-sequences

m	$n = 2^m - 1$	Number of possible m-sequences	Peak cross correlation R_{max}	Normalized cross correlation R_{max} / n
3	7	2	5	0.71
4	15	2	9	0.6
5	31	6	11	0.35
6	63	6	23	0.37

7	127	18	41	0.32
8	255	16	95	0.37
9	511	48	113	0.22
10	1023	60	383	0.37
11	2047	176	287	0.14
12	4095	144	1407	0.34

3.5.3 Gold Codes

PN sequences with better periodic cross-correlation properties than *m*-sequences have been given by Gold (1967, 1968) and Kasami (1966). They are derived from *m*-sequences as shown in Fig. 3-5. Gold and Kasami proved that certain pairs of *m*-sequences of length n exhibit a three-valued cross-correlation function with values $\{-1, -t(m),\ t(m) - 2\}$, where

$$t(m) = \begin{cases} 2^{(m+1)/2} + 1 & if\ odd \\ 2^{(m+2)/2} + 1 & if\ even \end{cases} \tag{3.8}$$

For example, if $m = 10$, then $t(10) = 26 + 1 = 65$ and the three possible values of the periodic cross-correlation function are $\{-1, -65, 63\}$. Hence the maximum cross correlation for the pair of m sequences is 65. Pair of m-sequences that satisfy equation (3.8) are called preferred sequences. On the other hand, the peak for the family of 60 possible sequences generated by a 10-stage shift register with different feedback connections is R_{max} = 383 about a sixfold difference in peak values. Thus, preferred sequences are defined as two *m*-sequences of length n with a periodic cross-correlation function that takes on the possible values $\{-1, -t(m), t(m) - 2\}$ [1].

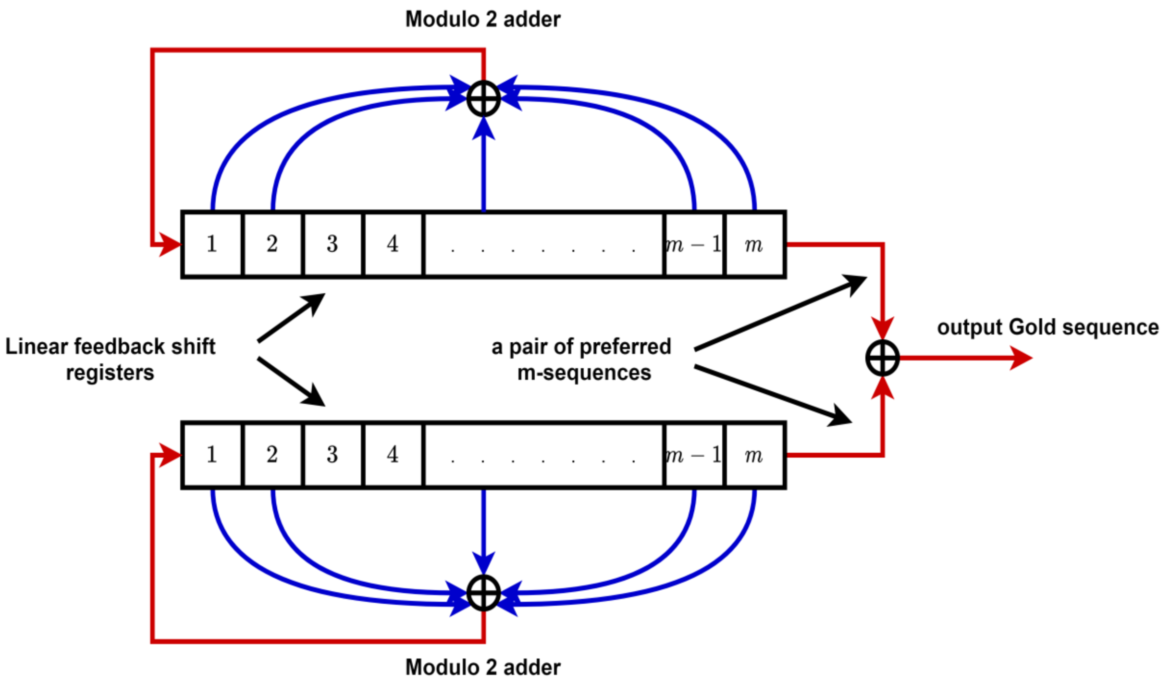

Fig. 3-5. Gold code generator.

3.5.3.1 Method of generation

As shown in Fig. 3-5, from a pair of preferred sequences, say a = $[a_1\ a_2\ \cdots a_n]$ and $b = [b_1\ b_2\ \cdots\ b_n]$, we construct a set of sequences of length n by taking the modulo-2 sum of a with the n cyclically shifted versions of b or vice versa. Thus, we obtain n new periodic sequences with period $n = 2^m - 1$. We may also include the original sequences a and b, and, thus, we have a total of $n + 2$ sequences. The $n + 2$ sequences constructed in this manner are called Gold sequences.

Except for the sequences a and b, the set of Gold sequences is not comprised of maximum-length shift-register sequences of length n. Hence, their autocorrelation functions are not two-valued. Gold (1968) has shown that the off-peak autocorrelation function for a Gold sequence takes on values from the set $\{-1, -t(m), t(m) - 2\}$. Hence, the off-peak values of the autocorrelation function are upper bounded by t(m).

The values of the off-peak autocorrelation function and the peak cross-correlation function, i.e., t(m), for Gold sequences and m-sequence is listed in Table 3-4. Also listed are the values normalized by $R(0) = n$.

Table 3-4. Comparison between *m*-sequences and gold codes.

m	$n = 2^m - 1$	Peak cross correlation of gold sequences $t(m)$	Normalized cross correlation for gold sequence $t(m) / n$	Normalized cross correlation for m-sequence R_{max}/ n
3	7	5	0.71	0.71
4	15	9	0.60	0.6
5	31	9	0.29	0.35
6	63	17	0.27	0.37
7	127	17	0.13	0.32
8	255	33	0.13	0.37
9	511	33	0.06	0.22
10	1023	65	0.06	0.37
11	2047	65	0.03	0.14
12	4095	129	0.03	0.34

3.6 Summary

The use of a spreading code (with pseudo-random properties) in the transmitter produces a wideband transmitted signal that appears noise like to a receiver that has no knowledge of the spreading code. The longer we make the period of the spreading code, the closer will the transmitted signal be to a truly random binary wave, and the harder it is to detect. Naturally, the price we must pay for the improved protection against interference is the increased transmission bandwidth, system complexity, and processing delay

Chapter 4
Chaotic Signals in Digital Communication Systems

4.1 Introduction

The past decade has seen heightened interest in the exploitation of chaos for useful applications in engineering systems. One application area that has attracted a great deal of attention is communications. Chaotic signals, by virtue of their wideband characteristic, are natural candidates for carrying information in a spread-spectrum communication environment. The use of chaotic signals in communications thus naturally inherits the advantages that are currently being offered by conventional spread-spectrum communication systems, such as robustness in multipath environments, resistance to jamming, low probability of interception, etc. In addition, chaotic signals are easy to generate and hence offer a potentially low-cost solution to spread spectrum communications [6].

4.2 Chaos theory

Chaotic signals are non-periodic, random-like signals derived from nonlinear dynamical systems [6] [7]. To understand how chaotic signals are generated, it suffices to consider discrete-time representations of dynamical systems. Essentially, when a system is described in discrete time, its state variables are sampled at fixed time intervals and its dynamics is described by an iterative function which expresses the state variables at one sampling instant in terms of those at the previous sampling instant [6]. The iterative function can be expressed as the following

$$x_n = f(x_{n-1}, \mu) \tag{4.1}$$

where x_n is the vector of state variables sampled at the n^{th} sampling instant, $f(.)$ is the iterative function that describes the dynamics of the system and μ is the vector of parameters that affect the system's dynamics.

The discrete-time representation allows a dynamical system to be studied in terms of an iterative function which takes the current state variables as the input and generates new state variables at discrete times. Thus, the state variables are updated at intervals determined by the sampling period. When the sampling period tends towards zero in the limit, the iterative function resembles a continuous-time representation which takes the form of differential equations [6].

Chaotic systems are dynamical systems whose state variables move in a bounded, non-periodic, random-like fashion. They are also characterized by a special property known as sensitive dependence upon initial conditions. To show the sensitive dependence on initial conditions, two chaotic waveforms are plotted with slightly different initial values, as shown in Fig. 4-1.

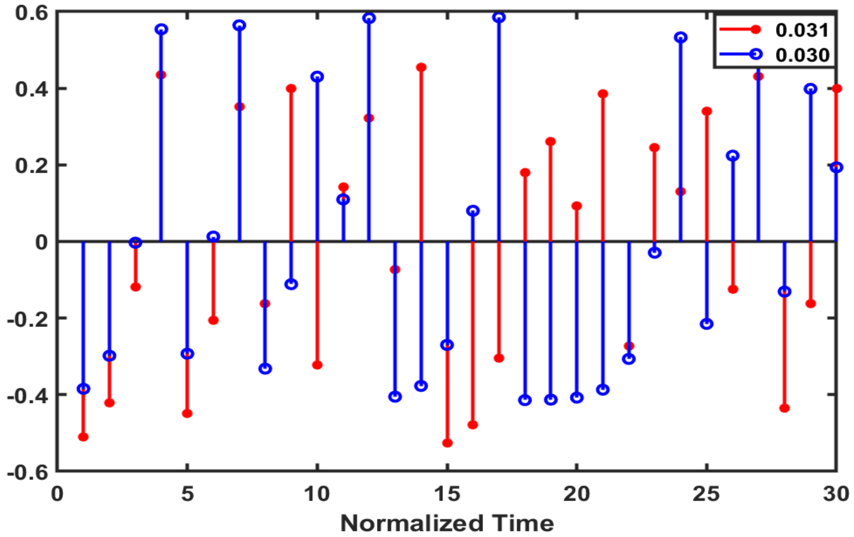

Fig. 4-1. Two chaotic signals with a slightly different initial value.

which essentially means that any two nearby starting conditions can lead rapidly to two entirely uncorrelated motions or trajectories of the state variables. This property allows us to generate, theoretically, an infinite number of uncorrelated chaotic signals from the same system using different initial values [6].

Here, apart from observing the aperiodic and random nature of the signal, we further confirm that the signal is bounded within the range [-1, +1]. Furthermore, it is worth noting that because of their random property, chaotic signals have impulse-like auto-correlation functions and white wideband power spectra. Also, cross-correlation of chaotic signals has a very small value as shown in

Fig. 4-2 and Fig. 4-3 [6] [11].

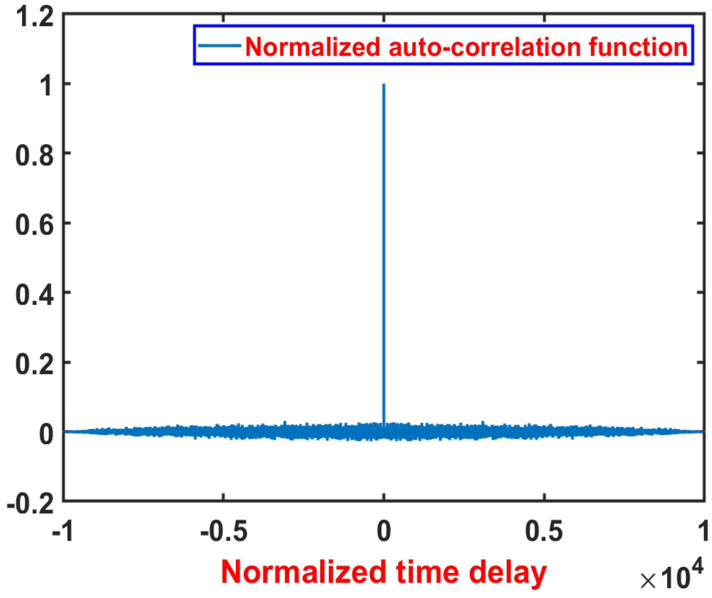

Fig. 4-2. Normalized autocorrelation function for chaotic signal.

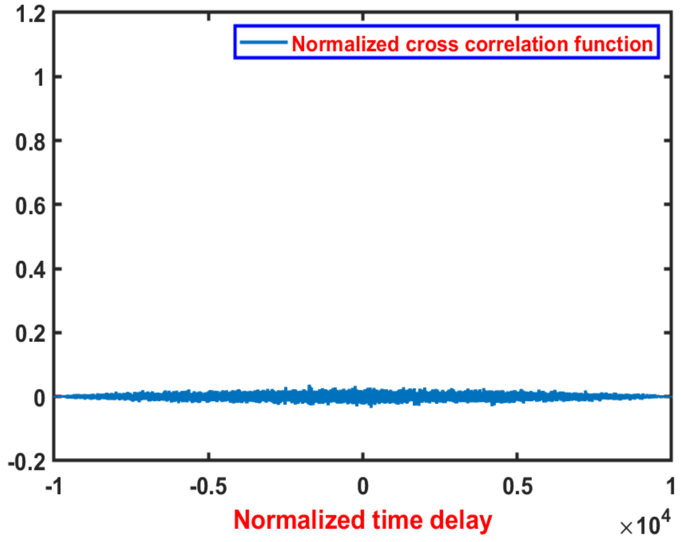

Fig. 4-3. Normalized cross correlation function for chaotic signal.

4.3 Application of Chaos to Communications

4.3.1 Direct-Sequence Spread-Spectrum.

Chaotic signals, with their inherent wideband characteristic, are natural candidates for spreading narrowband information. Thus, when using chaotic signals to encode information, the resulting signals are spread-spectrum signals having larger bandwidths and lower power spectral densities. They enjoy all the benefits of spread-spectrum signals such as difficulty of uninformed detection, mitigation of multipath fadings, anti-jamming, etc., as mentioned in the previous section. Moreover, many spreading waveforms can be produced easily because of the sensitive dependence upon initial conditions and parameter variations. Thus, chaos provides a low-cost and versatile means for spread-spectrum communications.

The basic principle is to replace the conventional binary spreading sequences, such as m-sequences or Gold sequences, by the chaotic sequences generated by a discrete-time nonlinear map. It has been shown that the performance of the new system is comparable to that of the conventional systems using binary spreading sequences [18]. The advantages of using chaotic spreading sequences are that an infinite number of spreading sequences exist and that the spread signal is less vulnerable to interception.

4.3.2 Digital Modulation

Several schemes have been proposed in the past for encoding digital information with chaotic signals. In most proposed methods, the basic principle is to map digital symbols to nonperiodic chaotic basis signals. For instance, chaotic switching or chaos shift keying (CSK) maps different symbols to different chaotic basis signals, which are produced from a dynamical system using different values of a bifurcation parameter or from several different dynamical systems. If synchronized copies of the chaotic basis signals are available at the receiver, detection can be achieved using a conventional correlator-type detector [19]. This class of detection is known as coherent detection. Moreover, if synchronized copies of the chaotic basis signals are not available at the receiver, detection has to be done by non-coherent means [20].

Another widely studied modulation technique for encoding digital information is based on a differential keying approach. Known as differential chaos shift keying (DCSK) [21], this technique essentially fabricates a special structure of the information bit that allows detection to be done in a non-coherent manner, i.e., without the presence of the synchronized copies of the chaotic signals at the receiver. Specifically, in the binary case, every transmitted symbol is represented by two chaotic signal sample sets. The first one serves as the reference sample set and the second as the data sample set. Dependent upon the symbol being sent, the data sample set is either an exact or inverted copy of the reference sample set. Demodulation can be done in a straightforward manner by correlating the two chaotic sample sets. The binary symbols can be differentiated by comparing the correlator output with a threshold value.

A few other digital modulation schemes derived from CSK and DCSK have also been proposed, e.g., chaotic on-off-keying (COOK), frequency-modulated DCSK (FM-DCSK)], correlation delay shift keying (CDSK), symmetric and quadrature CSK [22].

4.4 Differential Chaos-Shift-Keying (DCSK) System

The DCSK system was first proposed by Kolumban, Vizvari, Schwarz and Abel in [21]. The basic principle in the DCSK system is to make use of a special structural arrangement of each bit to carry digital information. Specifically, each bit duration is divided into two equal time slots and every transmitted symbol is represented by two sets of chaotic signal samples sent in the two slots. The first sample set serves as the reference (reference sample) while the second one carries the data (data sample) [6, 21].

Fig. 4-4 shows the block diagram of a basic single user differential chaos shift keying (DCSK) modulator. If a "$+1$" is to be transmitted, the data sample will be identical to the reference sample, and if a "-1" is to be transmitted, an inverted version of the reference sample will be used as the data sample [6].

Let 2β be the spreading factor in DCSK system, defined as the number of chaotic samples sent for each bit, where β is an integer. During the i^{th} bit duration, the output of the transmitter $e_{i,k}$ becomes

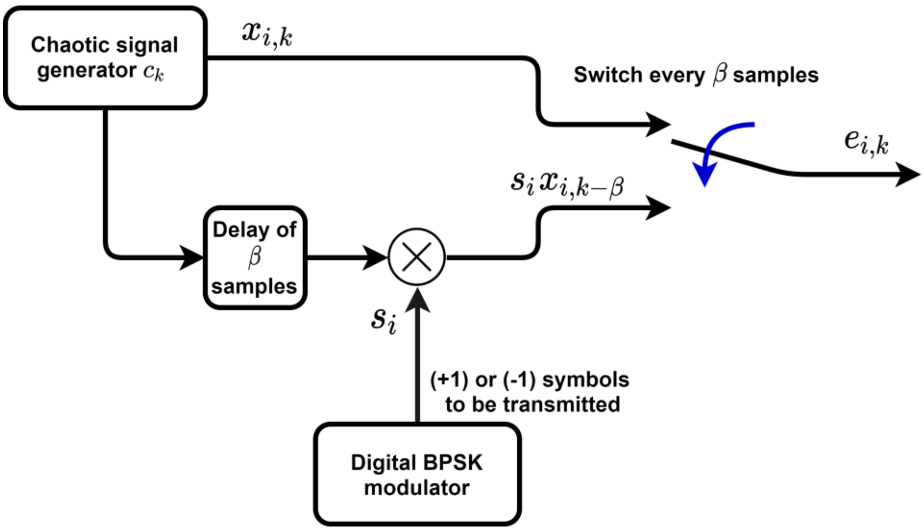

Fig. 4-4. Differential chaos shift keying (DCSK) modulator.

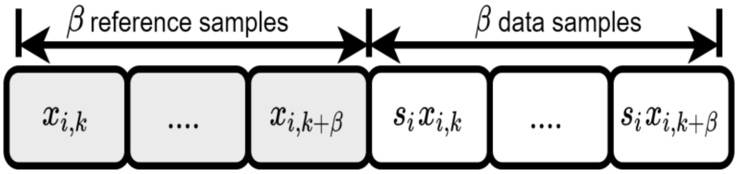

Fig. 4-5. Differential chaos shift keying (DCSK) symbol.

$$e_{i,k} = \begin{cases} x_{i,k} & for\ 1 \leq k \ \leq \beta \\ s_i x_{i,k-\beta} & for\ \beta < k < 2\beta \end{cases} \tag{4.2}$$

where x_k is the chaotic sequence used as reference, $x_{k-\beta}$ is the delayed version of the reference sequence x_k and each baseband BPSK modulated symbol is represented by $s_i = \{-1, +1\}$. Fig. 4-5 shows the structure of the Differential Chaos Shift Keying (DCSK) symbol.

Fig. 4-6 illustrates that the received signal r_k is correlated to a delayed version of the received signal $r_{k+\beta}$ and summed over a half bit duration T_b (where $T_b = 2\beta T_c$ and T_c is the chip time) to demodulate the transmitted bits. The received bits are estimated by computing the sign of the output of the correlator.

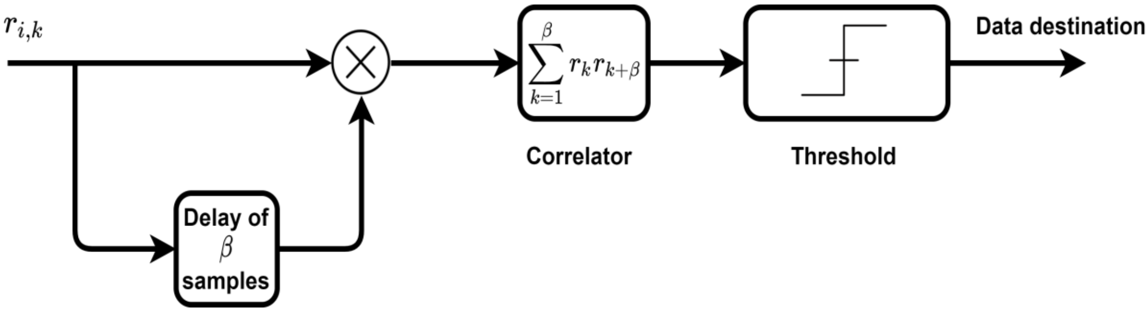

Fig. 4-6. Differential chaos shift keying (DCSK) demodulator.

As shown in Fig. 4-5, half of the transmitted energy and half of the bit duration time are spent sending a non-information bearing reference. Therefore, the data rate of this architecture is seriously reduced compared to other systems using the same bandwidth, leading to a loss of energy and spectral efficiency [11].

4.5 Examples of Chaotic generators

4.5.1 First order chaotic generators

- Logistic Map

$$c_{k+1} = 1 - 2c_k^2 \tag{4.3}$$

- Cubic Map

$$c_{k+1} = 4c_k^3 - 3c_k \tag{4.4}$$

- Bernoulli-shift map

$$c_{k+1} = \begin{cases} 1.2c_k + 1 & c_k < 0 \\ 1.2c_k - 1 & c_k > 0 \end{cases} \tag{4.5}$$

4.5.2 Second order chaotic generators

- Hénon map

$$y_{k+1} = 0.3c_k \tag{4.6}$$

$$c_{k+1} = 1 + y_k - 1.4c_k^2 \tag{4.7}$$

- Duffling map

$$y_{k+1} = c_n \tag{4.8}$$

$$c_{n+1} = -0.2y_k + 2.75c_n - c_n^3 \tag{4.9}$$

Chapter 5

Implementation of the OFDM-DQCSK system

5.1 Introduction

This chapter aims to illustrate and analyze the OFDM-DQCSK system. The block diagram of the system is presented. This system is designed to perform in frequency selective multipath fading channels in the presence of multiple access interference and Additive white gaussian noise. No synchronization or complex channel estimators are required. This due to the non-coherent nature of the system. Although this system uses a non-coherent differential modulation, it is proved that the bit error rate performance nearly reaches the coherent performance with the addition of an averaging block. This improvement comes at the expense of using some additional pilot subcarriers. The bandwidth efficiency does not get worse too much. This because the system is designed so that the pilot subcarriers are much fewer than the data subcarriers. The use of chaotic signals as an orthogonal spreading codes enhances the performance considerably when it is compared with the usage of conventional PN codes like m-sequence and gold codes. Not all chaotic sequences are orthogonal. A searching software is designed to find the orthogonal sets from a certain chaotic generator. This system is considered an improvement to the OFDM-DCSK system introduced in [11].

5.2 Chaotic map generator

In this system, a one-dimensional logistic map generator is employed [7]

$$x_{n+1} = 4\,x_n(1 - x_n) \qquad 0 < n < L_c \tag{5.1}$$

where x_n designates the current chaotic sample (i.e., a chip in a spreading sequence) and L_c is the spreading factor. The logistic map is used because of the easy way it generates a chaotic sequence and the excellent performance [5], [6]. There are many other chaotic maps proposed in [6], [7]. The initial condition is the first sample of a chaotic sequence. By varying the initial condition, different uncorrelated sequences are generated. A chaotic map is very sensitive to the very small differences among initial conditions to the extent of less than 10-8.

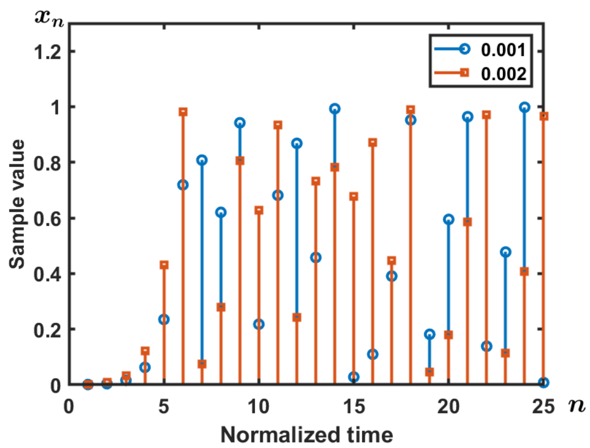

Fig. 5-1. Generating chaotic sequences from logistic map.

Fig. 5-1 shows two chaotic sequences generated from the logistic map in equation (5.1) with a very small difference between initial conditions. As shown in Fig. 5-1, the chaotic sample can be any value from the range of $[0, 1]$, therefore, they are not necessarily binary and theoretically take on continuous values.

The initial conditions of the chaotic sequences can be very small fractional numbers. Because of the high sensitivity of the chaotic map to initial conditions, any rounding to these numbers will lead to a completely different sequence, which is not practical in many hardware platforms. Therefore, in this work, chaotic sequences generated from equation (5.1) are quantized to a binary sequence using equation (5.2) according to the quantization model in [5].

$$c_n = 2(\lfloor 2\, x_n \rfloor) - 1 \qquad 0 < n < L_c \tag{5.2}$$

where $\lfloor\ \rfloor$ is the integer part and $c_n \in \{-1, 1\}$ is a bipolar binary code sample following the chaotic behavior. The quantization operation is shown in Fig. 5-2. However, not all spreading sequences generated from chaotic maps are orthogonal. A software program is designed to find these orthogonal sets with a searching algorithm. Firstly, an initial condition is selected from the range of [0,1] to substitute in equation (5.1). After that, the sequence generated from equation (5.1) is used in equation (5.2) to quantize the chaotic sequence if a binary sequence is required.

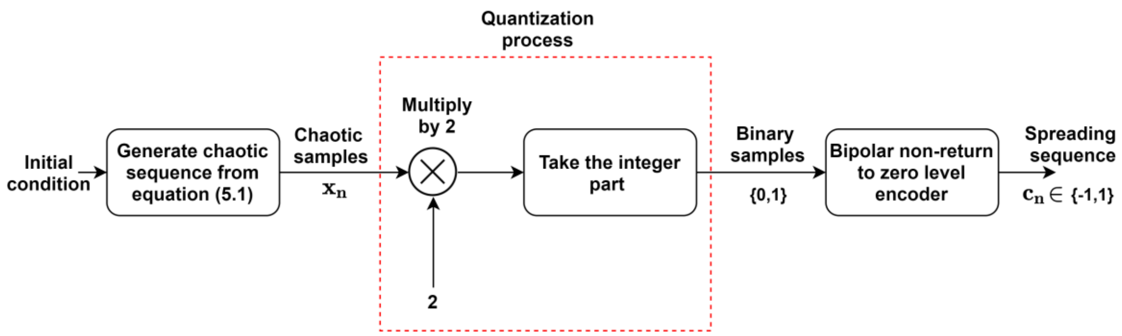

Fig. 5-2. Quantization of chaotic sequences.

Then the cross-correlation is tested between the sequence generated from the selected initial condition and each sequence generated from all the other possible initial conditions along the range of [0 , 1] until an orthogonal sequence is found. This searching operation can be done with a resolution of 10^{-8} in the range of [0 , 1]. The lower the resolution, the more orthogonal sequences can be found. Repeat the operation with the two orthogonal sequences until a third orthogonal sequence is found and so on. It should be noted that these sequences are not unique. If a different initial condition is selected from the beginning, another different orthogonal set is obtained. For example, in Table 5-1, sixteen different orthogonal binary chaotic sequences are generated with a spreading factor $L_c = 16$ using equations (5.1), (5.2) and a searching operation. The first initial condition is 0.098765. The resolution step is 10^{-8}. By the proper design of the searching software, only the orthogonal sequences are collected and used.

Table 5-1. Orthogonal quantized chaotic sequences.

Initial condition	**Quantized chaotic sequence**															
0.098765	-1	-1	1	-1	1	1	1	-1	-1	-1	-1	1	-1	1	-1	1
0.00000007	-1	-1	-1	-1	-1	-1	-1	-1	-1	-1	-1	-1	1	1	1	-1
0.0000149	-1	-1	-1	-1	-1	-1	-1	-1	1	1	1	1	-1	-1	-1	1
0.00395937	-1	-1	-1	-1	1	1	1	1	-1	1	1	-1	-1	-1	1	-1
0.01287891	-1	-1	-1	1	1	-1	1	1	1	1	-1	-1	1	1	-1	1
0.04333328	-1	-1	1	1	-1	-1	1	1	-1	-1	1	1	1	-1	1	1
0.05507052	-1	-1	1	1	-1	1	-1	1	1	1	-1	1	-1	1	1	-1
0.06139986	-1	-1	1	1	1	1	-1	-1	1	-1	1	-1	1	-1	-1	-1
0.15869982	-1	1	1	-1	-1	-1	1	1	1	-1	1	-1	-1	1	-1	-1
0.16673226	-1	1	1	-1	-1	1	1	-1	1	1	-1	-1	1	-1	1	1
0.21427516	-1	1	1	-1	1	-1	-1	1	-1	1	-1	1	1	-1	-1	-1
0.22359151	-1	1	1	1	1	-1	-1	-1	-1	1	1	-1	-1	1	1	1
0.3335199	-1	1	-1	1	-1	1	1	-1	-1	1	1	1	1	1	-1	-1
0.34318189	-1	1	-1	1	-1	1	-1	1	-1	-1	-1	-1	-1	-1	-1	1
0.38379479	-1	1	-1	1	1	-1	1	-1	1	-1	-1	1	-1	-1	1	-1
0.44386466	-1	1	-1	-1	1	1	-1	1	1	-1	1	1	1	1	1	1

5.3 Multiuser OFDM-DQCSK system architecture.

5.3.1 The OFDM-DQCSK modulator structure

In this section, a multiuser (OFDM-DQCSK) baseband modulator is illustrated. The modulator block diagram is shown in Fig. 5-4. The QPSK signal mapper block converts the incoming serial binary stream into N_S parallel baseband modulated QPSK symbols. Each symbol is spread in time using the same chaotic code sequence with a spreading factor L_c. Besides being a spreading code in the modulator, this chaotic code is added as a reference code to be sent along with the spread modulated symbols to be used later by the receiver to perform differential non-coherent de-spreading. More replicas of the reference chaotic code can be added and distributed among the spread modulated symbols according to the coherence bandwidth of the channel. Therefore, N_R reference code replicas are sent along with the N_S

spread modulated symbols. This will allow the reference codes to follow the changes of the channel coefficients and eliminate the need for a channel estimator. An IFFT operation is done to every group of $N_T = N_R + N_s$ chips such that there are L_c number of IFFT operations that is required to send the entire OFDM-DQCSK symbol. In other words, there are L_c number of OFDM symbols in every OFDM-DQCSK symbol. After each IFFT operation, the parallel N_T chips are then converted to serial stream, and a cyclic prefix is added to further extend the chip time to remove the inter-chip interference.

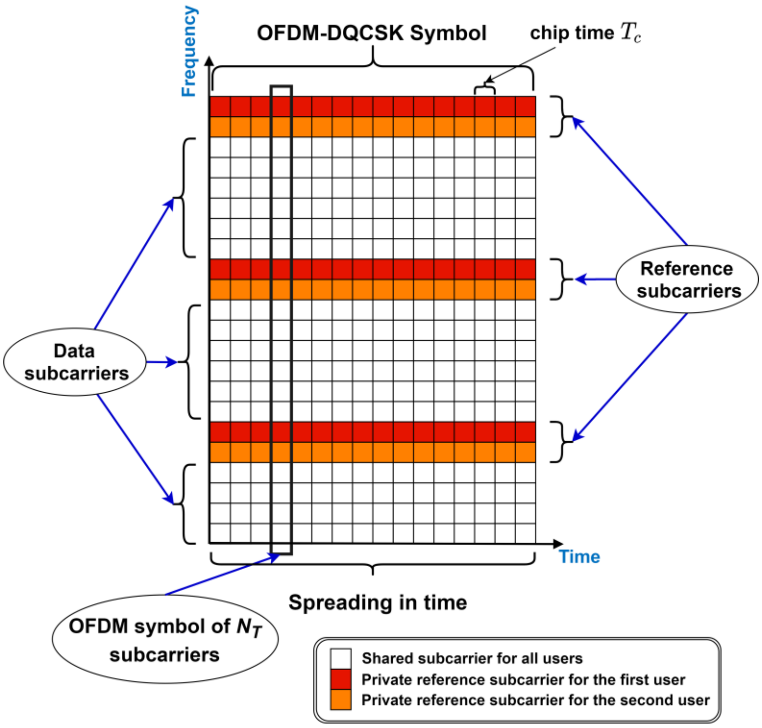

Fig. 5-3. OFDM-DQCSK signal structure for two users

As shown in Fig. 5-4, only the reference signals of different users are separated in pre-defined private frequencies to allow multiple access communications. The data symbols of users are transmitted on shared subcarriers and distinguished by the private chaotic reference codes on the private subcarriers of each user. The reference signals follow the changes in the channel coefficients that may occur during a chip period. As a result, the reference code signals are used for direct de-modulation and no need for channel estimators or a spreading code synchronization.

In Fig. 5-3, OFDM-DQCSK signal structure is divided into N_T subcarriers. Let U be the number of users and N_R be the number of private reference subcarriers for each user. UN_R private frequencies out of N_T subcarriers are used to transmit different reference signals of different users. The remaining N_S subcarriers are shared to transmit the spread data symbols.

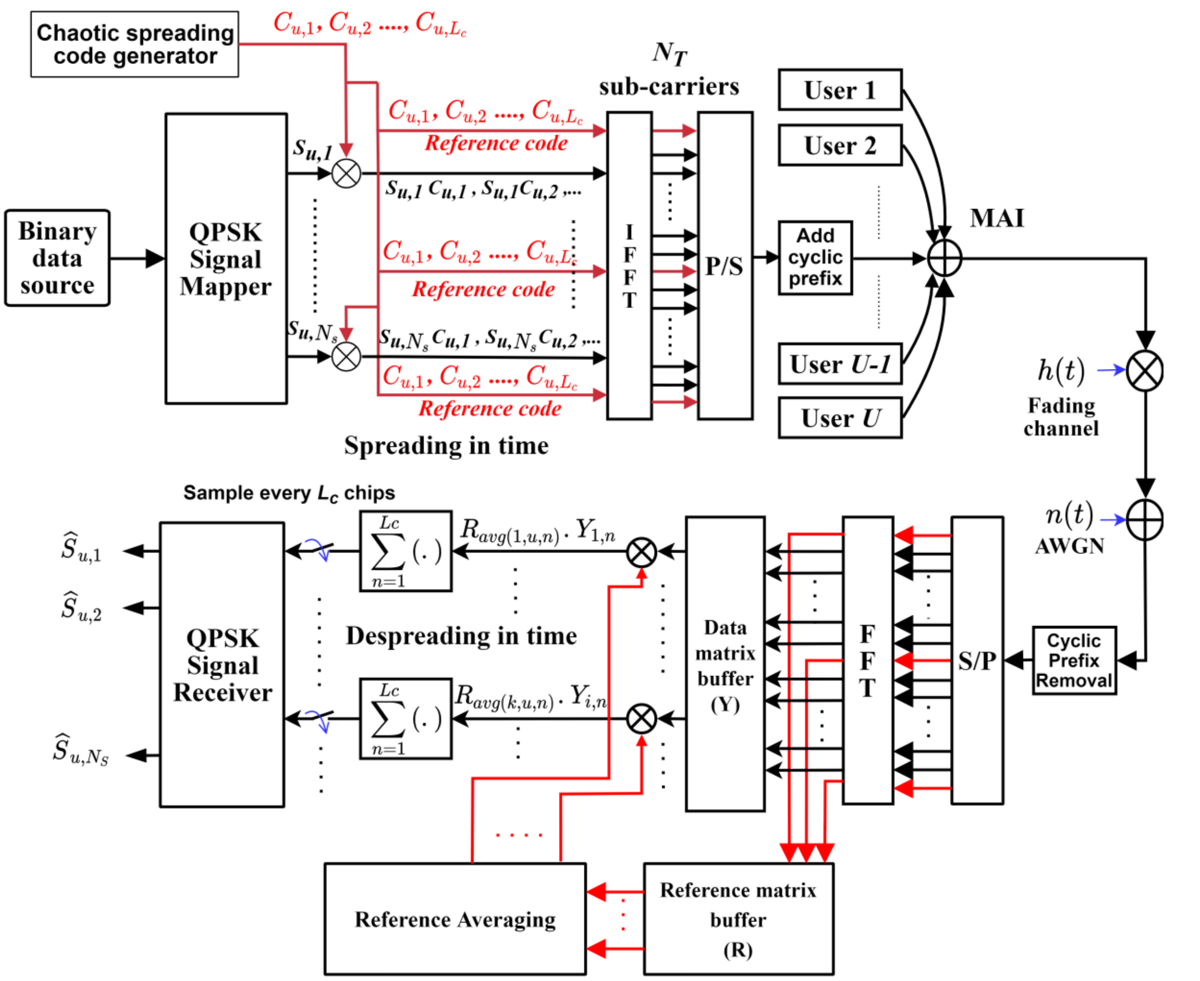

Fig. 5-4. Transmitter and the non-coherent receiver structure for a specific user.

Each user sends nulls in the other $UN_R - N_R$ private subcarriers of the other users. The total number of N_R subcarriers for each user depends on the coherence bandwidth of the channel.

Fig. 5-3 shows that the distribution of the reference signals follows the comb-type pattern design. In a comb-type arrangement, the reference signals are always present in a few subcarriers of all OFDM symbols [23]. The comb-type structure is employed to allow the reference codes to follow the channel coefficients when the channel varies in time from one OFDM symbol to another. In fast fading channel environments, the BER for the comb-type arrangement is better than other types such as block-type [24] or 2D-grid type in [25].

As shown in Fig. 5-3, A time-domain spreading operation is done for the modulated symbols where L_c is the spreading factor for each symbol. Transmission of an OFDM-DQCSK symbol requires L_c number of IFFT operations to send the N_s spread symbols. Besides, since each user shares a part of his bandwidth with the other users, this reduces the total required bandwidth but increases MAI. However, MAI can be reduced by the proper usage of orthogonal spreading codes and increasing the spreading factor length [1], [16].

As shown in Fig. 5-3, the OFDM-DQCSK symbol duration T_s is given by

$$T_s = L_c T_{OFDM} \tag{5.3}$$

where T_{OFDM} is the time duration of the OFDM symbol and L_c is the spreading factor.

As shown in Fig. 5-3 and Fig. 5-4, the chaotic signal $C_u(t)$ is the spreading code for the N_S symbols of user u that is transmitted over N_R frequencies,. Hence, the N_S symbols of user u are spread due to multiplication in time with the same chaotic spreading code $C_u(t)$

$$C_u(t) = \sum_{n=1}^{L_c} c_{u,n}\, \mathrm{g}(t - nT_c) \qquad 0 < t < T_c \tag{5.4}$$

where L_c is the spreading factor, $C_{u,n}$ is a chaotic chip sample for user u, and $\mathrm{g}(t)$ is the pulse shaping filter. T_c is the chip period. Let $\mathrm{g}(t)$ be a rectangular pulse shape throughout the entire analysis.

As shown in Fig. 5-4, the transmitted OFDM-DQCSK symbol of the u^{th} user consists of L_c OFDM symbols that is given by

$$Z_u(t) = \sum_{n=1}^{L_c} q_{u,n}\,(t) \tag{5.5}$$

where $Z_u(t)$ denotes the entire (OFDM-DQCSK) spread symbol to be transmitted for a specific user number u, L_c is the spreading factor and $q_{u,n}$ represents OFDM symbol after the n^{th} IFFT operation. Rewriting equation (5.5),

$$Z_u(t) = \sum_{n=1}^{L_c}\left(\sum_{i=1}^{N_S} s_{u,i}c_{u,n}e^{j2\pi\left(f_{S_{u_i}}\right)t}\,\mathrm{g}(t-nT_c) + \sum_{k=1}^{N_R} c_{u,n}e^{j2\pi\left(f_{R_{u_k}}\right)t}\,\mathrm{g}(t-nT_c)\right) \tag{5.6}$$

where $f_{R_{u_k}}$ is the k^{th} private subcarrier used to send the reference spreading code of the u^{th} user. $f_{S_{u_i}}$ is the shared public subcarrier frequency to transmit the i^{th} symbol of the N_s spread modulated symbols. Therefore, each user is allowed to send the data symbols on a maximum of N_S shared subcarriers. N_s is calculated from the following equation

$$N_S = N_T - U.N_R \tag{5.7}$$

where N_R is the number of private frequencies used to send the reference signals for each user.

For N_T subcarriers, the bandwidth BW_{N_T} of an OFDM symbol is given by

$$BW_{N_T} = subcarrier\ spacing * N_T \tag{5.8}$$

The minimum number of reference subcarriers $N_{R_{min}}$ that can cover the entire OFDM spectrum to follow the changes in the channel coefficients, depends on the coherence bandwidth of the channel. It can be calculated using the following equation

$$N_{R_{min}} = \frac{BW_{N_T}}{coherence\ bandwidth} \tag{5.9}$$

where the coherence bandwidth is approximately equal to the inverse of the maximum delay spread of the channel [1]. The subscript minimum is used because more replicas can be added to reduce the AWGN using the reference averaging technique, as shown later in section 4. It is assumed that the OFDM-DQCSK signal is transmitted over a multipath fading channel, the equivalent lowpass channel for the u^{th} user is given by the time-variant impulse response [1]

$$I_u(t;\tau) = \sum_{n=1}^{L_c} \left(h_{u,n}(t;\tau)\right) \tag{5.10}$$

where $h_{u,n}(t;\tau)$ is the time-variant impulse response at the n^{th} OFDM symbol for a specific user u, and L_c is the spreading factor. Rewriting equation (5.10),

$$I_u(\tau\,;\,t) = \sum_{n=1}^{L_c} \left(\sum_{l=1}^{L_p} \alpha_{u,l,n}(t)\delta\big(\tau - \tau_{u,l}(t)\,\big) \right) \tag{5.11}$$

where L_p is the number of channel paths, $\alpha_{u,l,n}$ is a complex random variable that represents the n^{th} multipath channel coefficient (i.e., the channel gain) for the signal received on the l^{th} path and $\tau_{u,l}$(t) is the propagation delay of the l^{th} path for a specific user u. In this work, the multipath complex channel coefficients follow Rayleigh distribution [26] given by

$$f(\alpha|\sigma) = \frac{\alpha}{\sigma^2} e^{\frac{-\alpha^2}{2\sigma^2}} \qquad ,\alpha \geq 0 \tag{5.12}$$

where σ > 0 is the scaling parameter of the distribution and $E(\alpha^2) = 2\sigma^2$. The multipath channel coefficients $\alpha_{u,l,n}$ is assumed to change every OFDM symbol.

The received multiuser OFDM-DQCSK signal over the wireless channel is given by

$$r(t) = \sum_{u=1}^{U} Z_u(t) \otimes I_u(t) + n(t) \tag{5.13}$$

where U is the number of users, a convolution operation is represented by ⊗, and n(t) is a complex AWGN with zero mean and power spectral density of N_0.

5.3.2 The non-coherent OFDM-DQCSK demodulator structure

In this section, the non-coherent (OFDM-DQCSK) baseband demodulator is illustrated. The demodulator block diagram is shown in Fig. 5-4. In the beginning, the cyclic prefix redundancy is removed, then a serial to parallel operation is done for every N_T different samples. An FFT operation is then performed to every N_T parallel symbols. The output of the FFT operation is stored in two matrix memories R and Y dedicated for the reference and data signals, respectively. Based on the strong law for large numbers (SLLN) [26] [27], The current received reference signal is averaged using the reference averaging method presented in section 5.4 to reduce the AWGN to a negligible value. After L_c successive FFT operations, a non-coherent de-spreading operation is performed using the averaged received reference signal $R_{av}(k,u,n)$ to recover the transmitted symbols. Therefore, after the n^{th} FFT operation, the multi-user spread data chip $Y(n,f_{S_{u_i}})$ and the reference chip of the u^{th} user $R(n,f_{R_{u_k}})$ can be represented as the following by taking the FFT to the received signal in equation (5.13),

$$Y\left(n,i,f_{S_{u_i}}\right) = \sum_{u=1}^{U} s_{u,i}c_{u,n}H_{u,n,f_{S_{u_i}}} + W(S) \qquad 1 < n < L_c \ and \ 1 < i < N_S \qquad (5.14)$$

$$R\left(n,f_{R_{u_k}}\right) = c_{u,n}H_{u,n,f_{R_{u_k}}} + W(R) \qquad 1 < n < L_c \ and \ 1 < k < N_R \qquad (5.15)$$

where L_c is the spreading factor, $R\left(n,f_{R_{u_k}}\right)$ is the n^{th} reference sample of user u recovered from the private frequency $f_{R_{u_k}}$. $Y\left(n,i,f_{S_{u_i}}\right)$ is the n^{th} the multi-user sample of the i^{th} symbol transmitted over the shared $f_{S_{u_i}}$ frequency, $W(S)$ and $W(R)$ represents the AWGN after the FFT operation. $H_{u,n,f_{S_{u_i}}}$ is the channel frequency response of i^{th} public shared subcarrier $f_{S_{u_i}}$. $H_{u,n,f_{R_{u_k}}}$ is the channel frequency response of the k^{th} private subcarrier $f_{R_{u_k}}$.

The channel frequency response of the private subchannel at the subcarrier $f_{R_{u_k}}$ is given by

$$H_{u,n,}f_{R_{u_k}} = \sum_{l=1}^{L_p} \alpha_{u,l,f_{R_{u_k}}} e^{-2\pi j f_{R_{u_k}} \tau_{u,l}} \qquad 1 < n < L_c \ and \ 1 < k < N_R \qquad (5.16)$$

It is assumed that the maximum delay spread for a given user u is much lower than the OFDM symbol time, i.e., $\tau_{u,l_{max}} \ll N_T T_c$. In this case, the channel is considered a quasi-static flat fading non-frequency selective channel for each set of OFDM subcarriers. Therefore, under the condition of low maximum delay spread, the term $e^{-2\pi j f_{R_{u_k}} \tau_{u,l}} \approx 1$. Hence, the frequency response of the private subchannel reduces to

$$H_{u,n,}f_{R_{u_k}} = \sum_{l=1}^{L_p} \alpha_{u,l,f_{R_{u_k}}} \qquad (5.17)$$

Similarly, the frequency response of the public subchannel at the subcarrier $f_{S_{u_i}}$ is given by

$$H_{u,n,}f_{S_{u_i}} = \sum_{l=1}^{L_p} \alpha_{u,l,f_{S_{u_i}}} \qquad (5.18)$$

Moreover, the spread spectrum modulation makes the channel introduces a negligible amount of inter-symbol interference [1].

For low values of maximum delay spread, the channel is assumed to be flat in frequency over an adjacent group of subcarriers according to the following condition

$$\alpha_{u,f_{R_{u_k}}} \approx \alpha_{u,f_{S_{u_i}}} \qquad (5.19)$$

Finally, the data of user u are de-spread using the of the averaged reference signal $R_{av}\left(n, f_{R_{u_k}}\right)$, then decoded by computing the decision variable as follows

$$D_{u,i} = \sum_{n=1}^{L_c} Y\left(n, i, f_{S_{u_i}}\right) . R_{av}\left(n, f_{R_{u_k}}\right)^* \qquad 1 < u < U \qquad (5.20)$$

5.4 Reference Averaging Technique

It is used to improve the BER performance that is deteriorated because of correlating the data symbols with a noisy reference. The received reference code, at the OFDM-DQCSK receiver described in section 3, is corrupted by AWGN, and multiplied by the multipath flat fading channel coefficients. Correlating a noisy reference with the data symbols is the reason why the differential non-coherent de-modulation gives a lower BER than coherent performance.

The non-coherent performance can reach the coherent performance if **the Strong Law for Large Numbers (SLLN) theorem** [26] is exploited which states the following, for independent and identically distributed (iid) random variables, the average (the sample mean) of a large number of random variables should be close to the expected value (true mean) and will eventually become equal to the expected value with a probability of one as the number of random variables is infinity. This concept is illustrated in Fig. 5-5.

Let $x_1, x_2, \ldots\ldots, x_N$ be a sequence of iid random variables with an expected value (true mean) $E(x) = \mu$ and variance $var(x) = \sigma^2$. The sample (observed) mean M_N of the sequence is given by

$$M_N = \frac{1}{N}\sum_{j=1}^{N} x_j \tag{5.21}$$

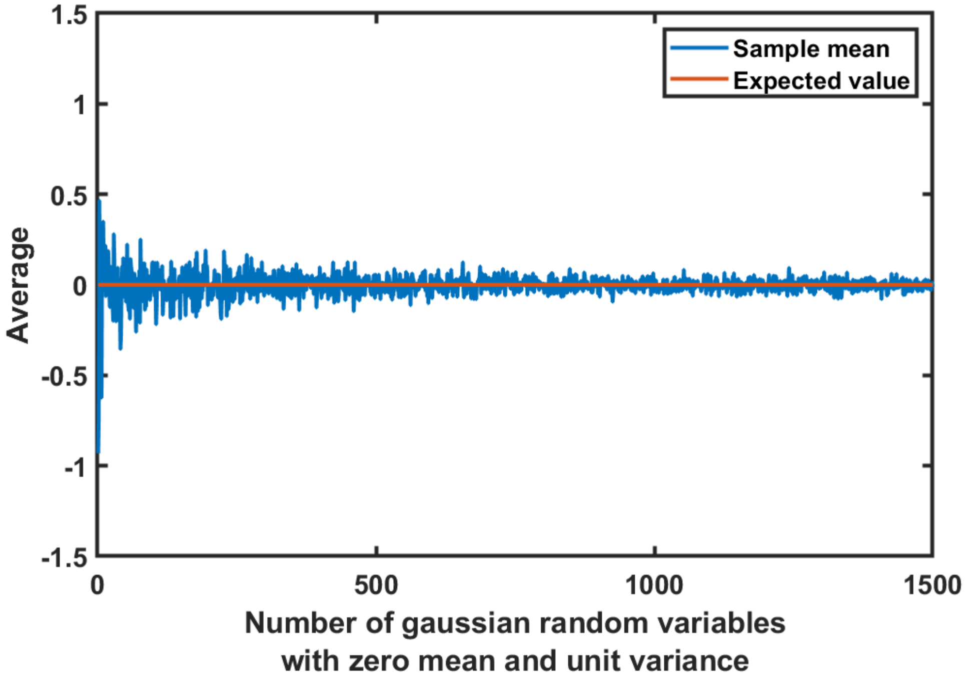

Fig. 5-5. The application of the SLLN theorem to gaussian random variables.

The variance of the sample mean ($\boldsymbol{M_N}$) equals

$$var(M_N) = \frac{1}{N^2} var(x_1 + x_2 + \ldots\ldots, + x_N) = \frac{1}{N^2} . N\,\sigma^2 = \frac{\sigma^2}{N} \qquad (5.22)$$

Equation (5.22) states that the variance of the sample mean approaches zero as the number of random variables is increased. This implies that the probability that the sample mean is close to the true mean (expected value) approaches one as n becomes very large.

The mathematical description of the SLLN [26] is given by

$$P\left(\lim_{N\to\infty} (M_N) = \mu\right) = 1 \qquad (5.23)$$

5.4.1 Method 1: Similar reference code in each OFDM-DQCSK

In this case, the user will use the same reference chaotic code in every (OFDM-DQCSK) symbol.

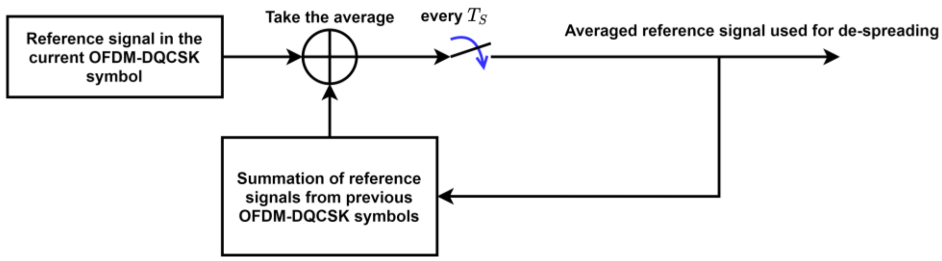

Fig. 5-6. Averaging of reference signals using method 1.

Fig. 5-6 illustrates the averaging procedure. From equation (5.15) and (5.17), the received reference signal after the FFT operation is expressed as

$$R\left(n, f_{R_{u_k}}\right) = \alpha_{u,n,f_{R_{u_k}}} c_{u,n} + W(R) \qquad 0 < n < L_c \tag{5.24}$$

where$R\left(n, f_{R_{u_k}}\right)$ is the n^{th} reference sample of user u recovered from the private frequency $f_{R_{u_k}}$. $c_{u,n}$ is the n^{th} code chip of the spreading sequence. $W(R)$ represents the AWGN after the FFT operation. $\alpha_{u,n,f_{R_{u_k}}}$ is the n^{th} multipath channel coefficient of the k^{th} private subcarrier $f_{R_{u_k}}$. According to the SLLN, the current received reference signal in (5.24) should be added to the previous one and use their average to de-spread the data symbols. Their summation should be stored to added to the next reference signal in the next OFDM-DQCSK symbol and so on. The averaging operation can be expressed as the following using equation (5.24)

$$R_{av}\left(n, f_{R_{u_k}}\right) = \left\{\sum_{p=1}^{P} \left\{c_p(u,n). \alpha_p\left(u, n, f_{R_{u_k}}\right)\right\} + \sum_{p=1}^{P} W_p(R)\right\} \Big/ P \tag{5.25}$$

where P is the number of reference signals to be averaged. $c_p(u,n)$ is a deterministic constant so it is not affected by the averaging process. Rewriting equation (25)

$$R_{av}\left(n, f_{R_{u_k}}\right) = c_{u,n}\left\{\sum_{p=1}^{P} \alpha_p\left(u, n, f_{R_{u_k}}\right)\right\}/P + \left\{\sum_{p=1}^{P} W_p(R)\right\}/P \qquad (5.26)$$

According to the (SLLN) and the result deduced from Fig. 5-5, after nearly 1000 received reference signals (i.e., after nearly $P = 1000$ OFDM-DQCSK symbols), the noise term $\sum_{p=1}^{P}\{W_p(R)\}/P$ can be reduced to the mean of the AWGN which is zero. In this way, the AWGN on the reference code is averaged to a negligible value almost zero. The term $\sum_{p=1}^{P} \alpha_p(u, n, f_{R_{u_k}})/P$ which represents the multipath fading reduces to the mean of the Rayleigh distribution. Assume that the channel average power gain $E(\alpha_u^2) = 1$ then according to equation (5.12), the expected value will be nearly 0.88. In conclusion, the performance nearly reaches coherent performance. Simulation results show the advantage of the averaging technique. One reference subcarrier is used to demodulate a group of adjacent data subcarriers according to the coherence bandwidth of the channel. The total number of reference subcarriers in this method is $N_R = N_{R_{min}}$ to cover the entire bandwidth.

5.4.2 Method 2: Different reference code in each OFDM-DQCSK

If the user uses a different reference code each OFDM-DQCSK, the scheme in Fig. 5-6 cannot be used. As shown in Fig. 5-7 and Fig. 5-8, a certain number of reference replicas are sent over multiple subcarriers then they are averaged at the receiver side. In contrast to the scheme in Fig. 5-6, more than one reference replica are used to reference a group of adjacent data subcarriers (i.e. $N_R \geq N_{R_{min}}$). This is illustrated in Fig. 5-8 for $N_R = 2N_{R_{min}}$. The more replicas are used, the better performance is achieved according to the Strong Law for Large Numbers. The drawback of this method is the decreased bandwidth efficiency.

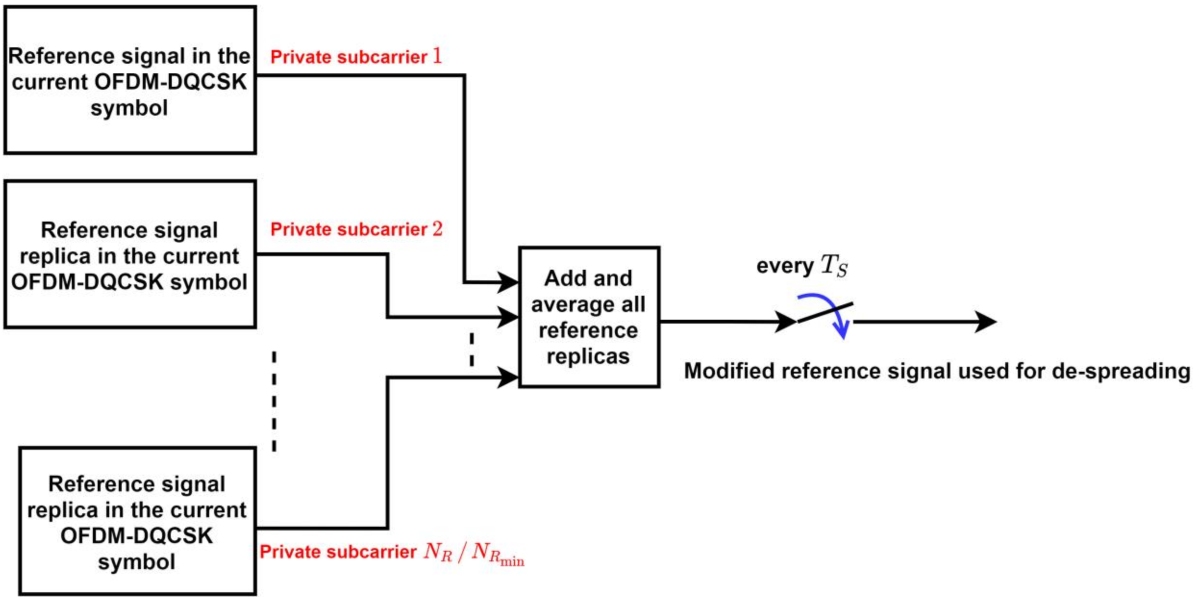

Fig. 5-7. Averaging of reference signals using method 2.

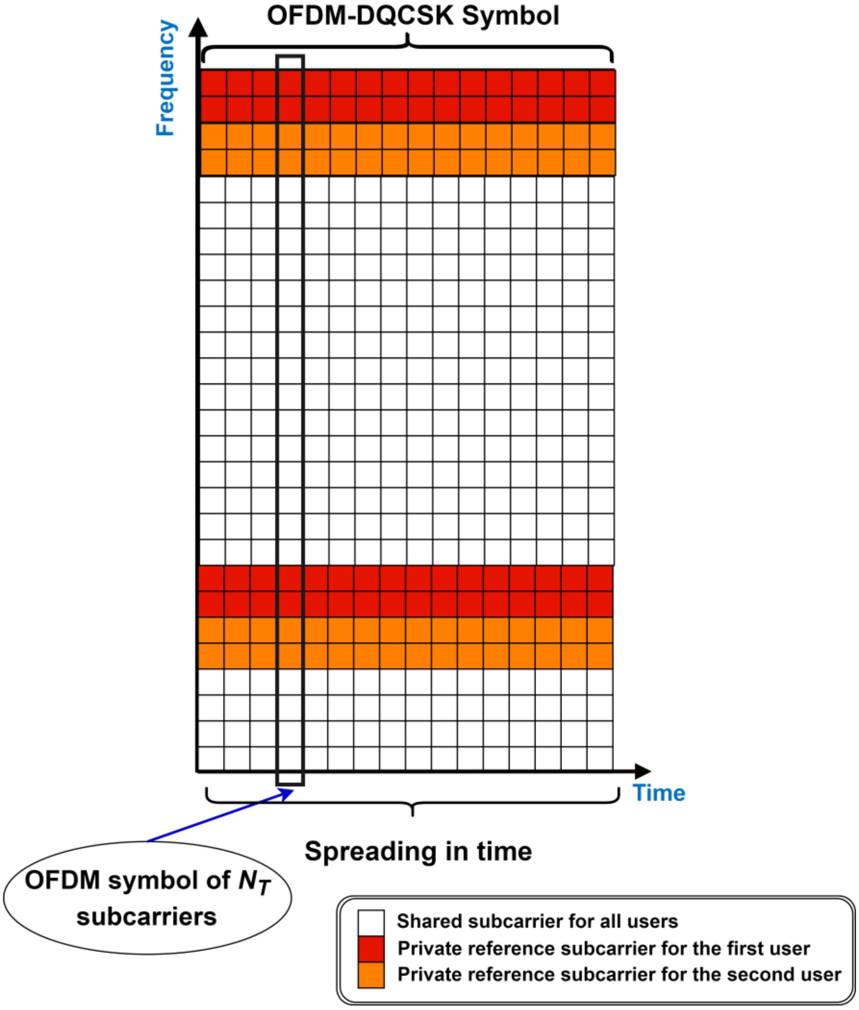

Fig. 5-8. The transmitted signal structure for two users with using method 2.

5.5 Summary

The OFDM-DQCSK is designed to combine OFDM with spread spectrum technique. The system uses chaotic sequences as a spread spectrum signals to replace conventional PN codes. The system uses differential modulation to overcome many practical problems. No need for synchronization or complex channel estimators. Although the non-coherent nature of the system, it is proved that with the addition of the AWGN averaging block, the performance is improved to the extent that it may reach the coherent performance. The system can perform in multipath fading channels in the presence of MAI and AWGN. Simulation results in the next chapter shows the advantages of the hybrid system.

Chapter 6
Analysis and simulation results of the OFDM-DQCSK system

6.1 Introduction

This chapter aims to present the simulation results of the OFDM-DQCSK system that is implemented in chapter 5. A detailed discussion for the simulation results is included and explains the behaviour of the system in each case. the bit error rate performance of the proposed non-coherent multi-user OFDM-DQCSK system is evaluated in the presence of AWGN and Rayleigh fading channels for a different number of users U. Various lengths of the spreading factor L_c are used. Different numbers of transmitted data symbols subcarriers N_S are tested.

6.2 Simulation parameters

The simulation parameters of the proposed system follow the IEEE 801.11a standard [28], where about one-fourth of the total subcarriers N_T are unused. A cyclic prefix of 4 samples is employed. The FFT size is chosen according to the coherence bandwidth, the total number of users U, and the required number of data subcarriers N_s in the entire OFDM symbol. To achieve accurate simulation results, the unused subcarriers are nullified so as not to degrade the performance of the system. Assuming an indoor environment, Let the maximum multipath delay spread of the channel is $\tau_{L,max} = 0.1\ \mu s$, which makes the frequency response of the channel flat over a bandwidth of 10 Mhz. Assuming N_T (FFT size) is equal to 128 and a subcarrier spacing of 312.5K sample/sec. According to equation (5.9), a minimum of three reference signals (i.e., $N_{R_{min}} = 3$) can properly cover the entire OFDM spectrum and make the demodulation possible. An AWGN channel without multipath fading is considered first in section 5.1 then a Rayleigh multipath fading channel in the presence of AWGN is analyzed next in section 5.2.

6.3 Simulation results

6.3.1 AWGN channel without multipath fading

6.3.1.1 Testing the OFDM-DQCSK system without the proposed reference averaging technique and without MAI (single user)

A simulation of the BER performance in the presence of only an AWGN channel for a single user (i.e., $U = 1$) is shown in Fig. 6-1, without using the reference averaging technique proposed in section 5.4. A total number of subcarriers (i.e., an FFT size) equals to $N_T = 128$, reference subcarriers $N_R = 3$, data subcarriers $N_S = 49$ for the user to transmit an OFDM-DQCSK symbol. Various spreading factor lengths $L_c = 16, 64, 120, 150$ are tested. The coherent OFDM-QCSK is used as a benchmark.

For the non-coherent OFDM-DQCSK system, it is observed in Fig. 6-1 that a performance degradation occurs when the spreading factor L_c is increased. The reason for this problem is that the reference code is distorted by AWGN and is not noise-free as in the case of the coherent receiver. At the receiver, data symbols are correlated with a noisy reference code. As the reference code length increases, the more noise affects the correlator. As a result, the performance of the non-coherent differential modulation-based systems is deteriorated to a large extent [6], [29]. However, in the case of the coherent system, whatever the value of the spreading factor, the BER performance is the same. This is because, at the receiver, a reference code replica is regenerated without the AWGN, but this would require a robust synchronization between the transmitter and the receiver. It is observed in

Fig. **6-1** that the coherent version of the proposed system does not reach the theoretical QPSK performance. This is due to the insertion of the cyclic prefix, which causes an overall degradation of system performance.

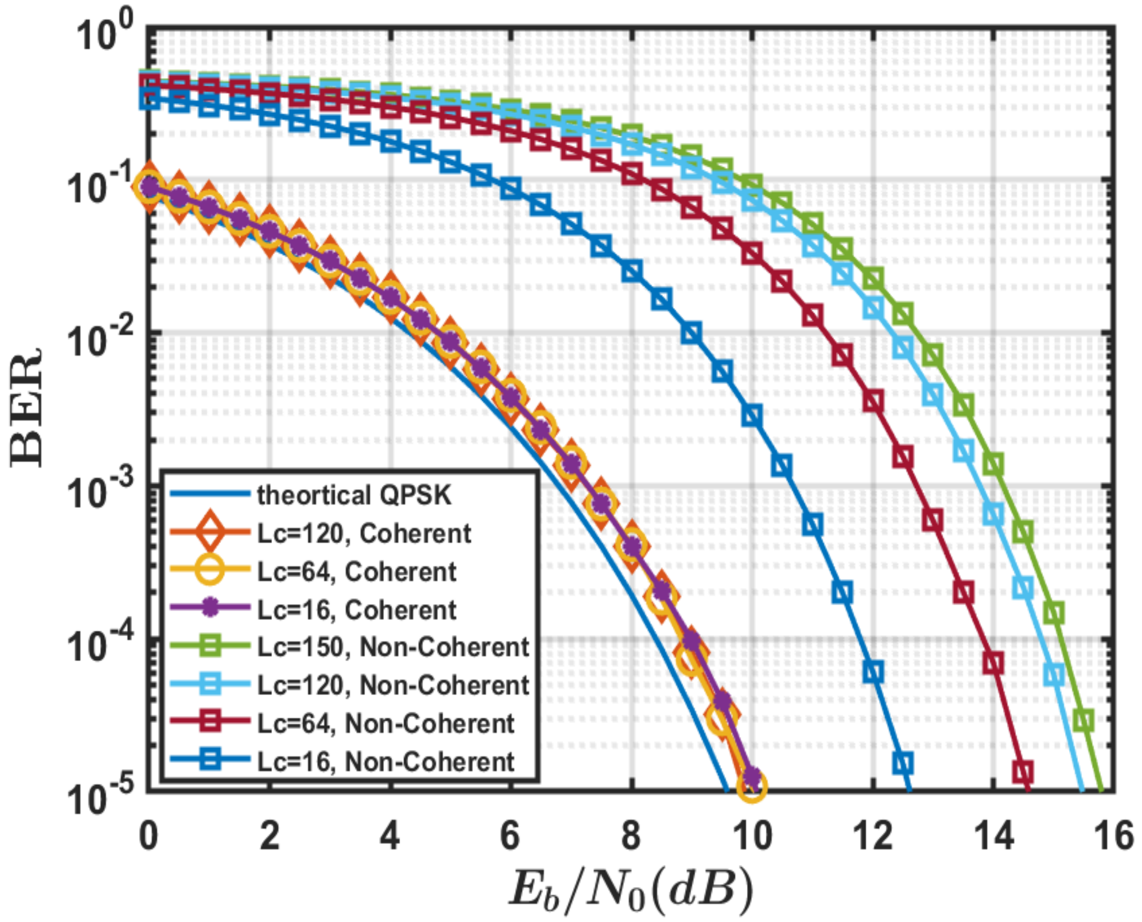

Fig. 6-1. BER of the proposed system for a single user in AWGN channel.

6.3.1.2 Testing the OFDM-DQCSK system in a multiuser scenario without the proposed reference averaging technique

In Fig. **6-2**, the cross-correlation properties of the proposed chaotic codes are investigated. The simulation parameters are an FFT size equals to $N_T = 128$, $N_s = 49$ data subcarriers, and $L_c = 64, 120$ for a single user ($U = 1$) and multi-user ($U = 5$). The simulation in Fig. **6-2** is done without using the proposed reference averaging technique.

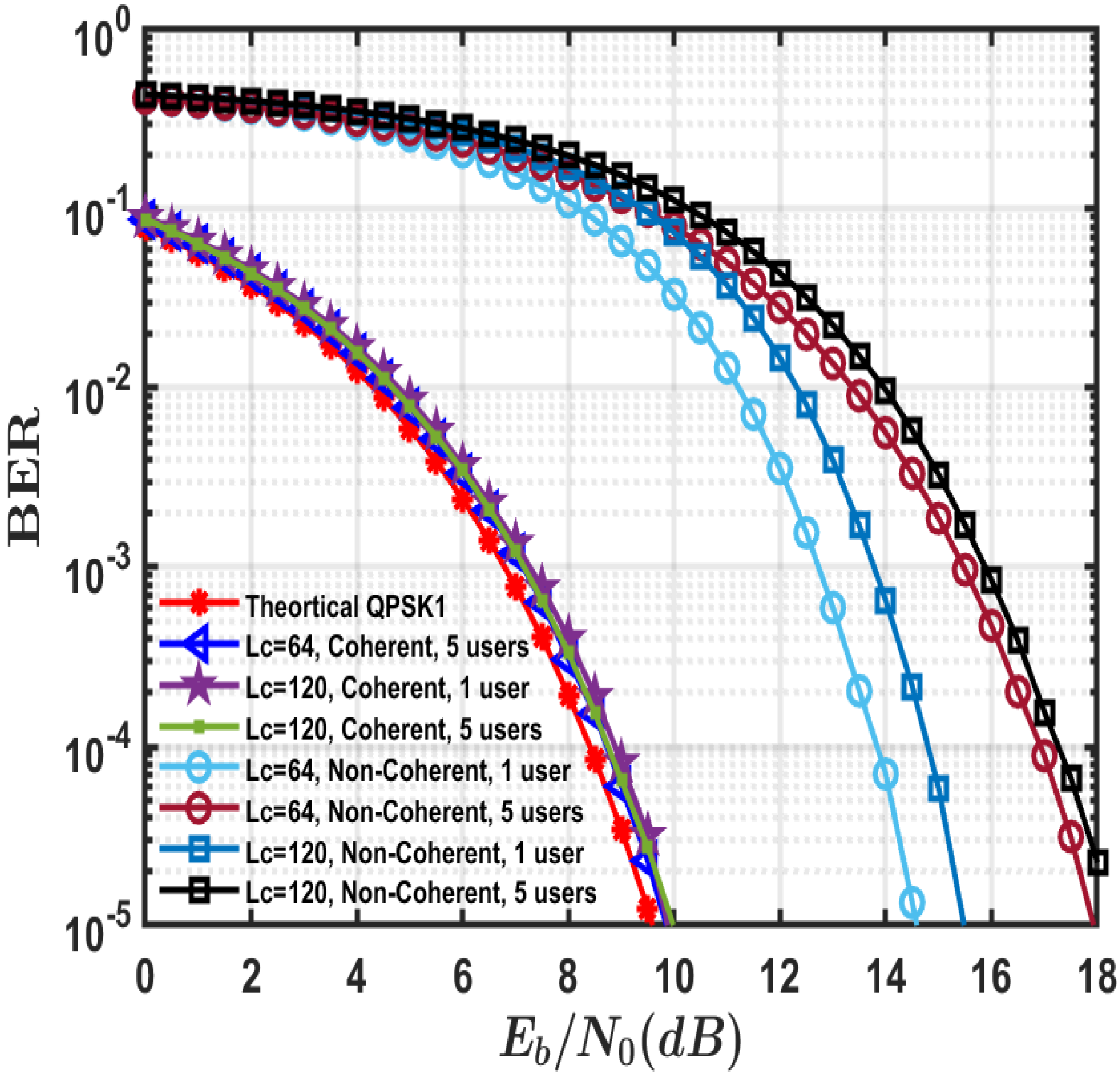

Fig. 6-2. BER performance of the proposed system for multiuser in AWGN channel.

It is observed that, for the coherent version of the proposed system, whatever the number of users, the BER performance is nearly the same which proves the orthogonality and the excellent correlation properties of the chaotic codes generated by the searching software proposed in this work. In the non-coherent model, the orthogonality of the chaotic codes is not maintained because the reference code is corrupted by AWGN which destroys the orthogonality between the chaotic codes to some extent and the MAI is increased. However, as shown in Fig. 6-3, although the orthogonality of the chaotic codes is lost to some extent, they still give better performance than other non-orthogonal codes such as conventional maximal length codes.

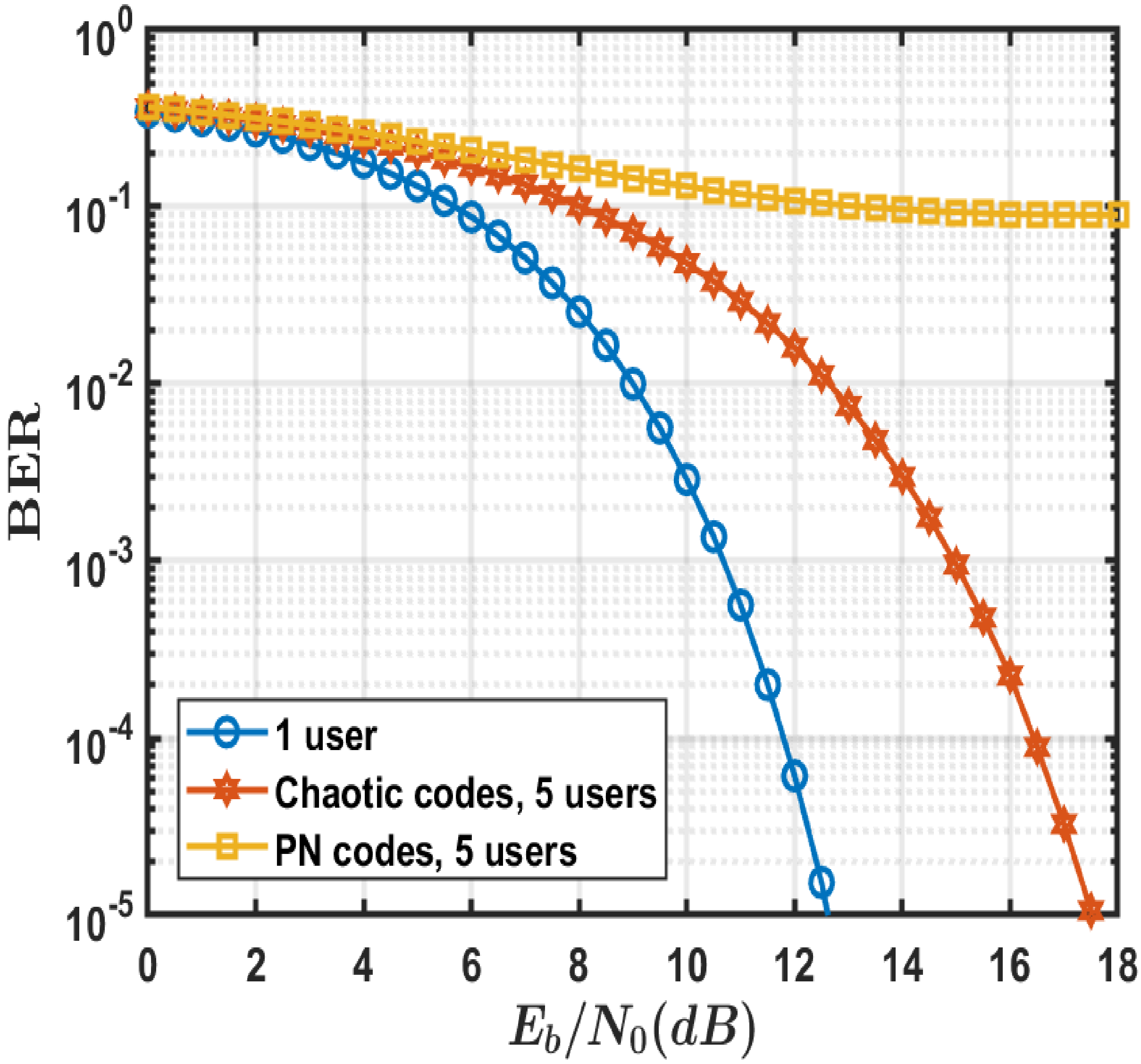

Fig. 6-3. A Comparison between chaotic and maximal length codes.

6.3.1.3 Testing the feasibility of using DQCSK system instead of DCSK in a single user scenario without the proposed reference averaging technique

In *Fig. 6-4*, the effect of using QPSK instead of BPSK in the proposed system is studied. For an FFT size equals to $N_T = 128$, $N_s = 10$, 30 data symbols per the entire OFDM-DQCSK symbol, $L_c = 120$ and N_R= 3. In the normal case, they should give the same BER performance. But in *Fig. 6-4* it is observed that, for the same number of data subcarriers, the QPSK modulation gives a better BER performance than BPSK. This is because the reference code chip can reference two bits instead of one, which leads to better energy efficiency.

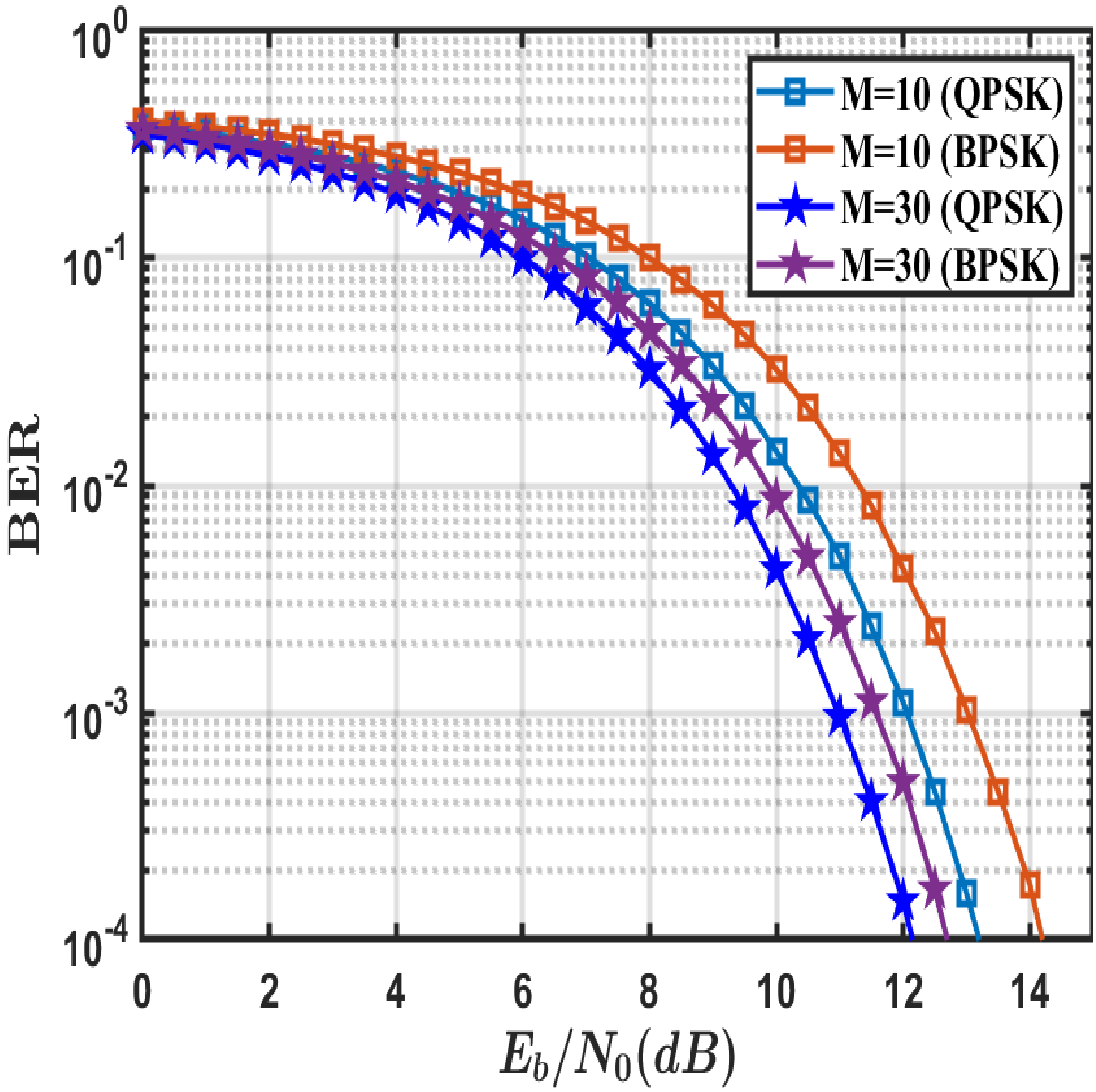

Fig. 6-4. BER comparison between BPSK and QPSK in the proposed system

6.3.1.4 Testing method two of the proposed reference averaging technique using the proposed OFDM-DQCSK system in a multiuser scenario

In *Fig. 6-5,* to improve the performance that is deteriorated because of correlating data symbols with a noisy reference, the proposed averaging technique using method two in *section 4.2* is employed. The simulation parameters are an FFT size equals to $N_T = 256$, $N_S = 49$ data symbols per the entire OFDM-DQCSK symbol, and $L_c = 120$ for multiuser ($U = 5$). The BER performance is improved as the number of reference replicas is increased (i.e., $N_R > N_{R_{min}}$). This is because the AWGN in the reference subcarrier is averaged to a negligible value. The simulation is done for a different number of reference subcarriers N_R.

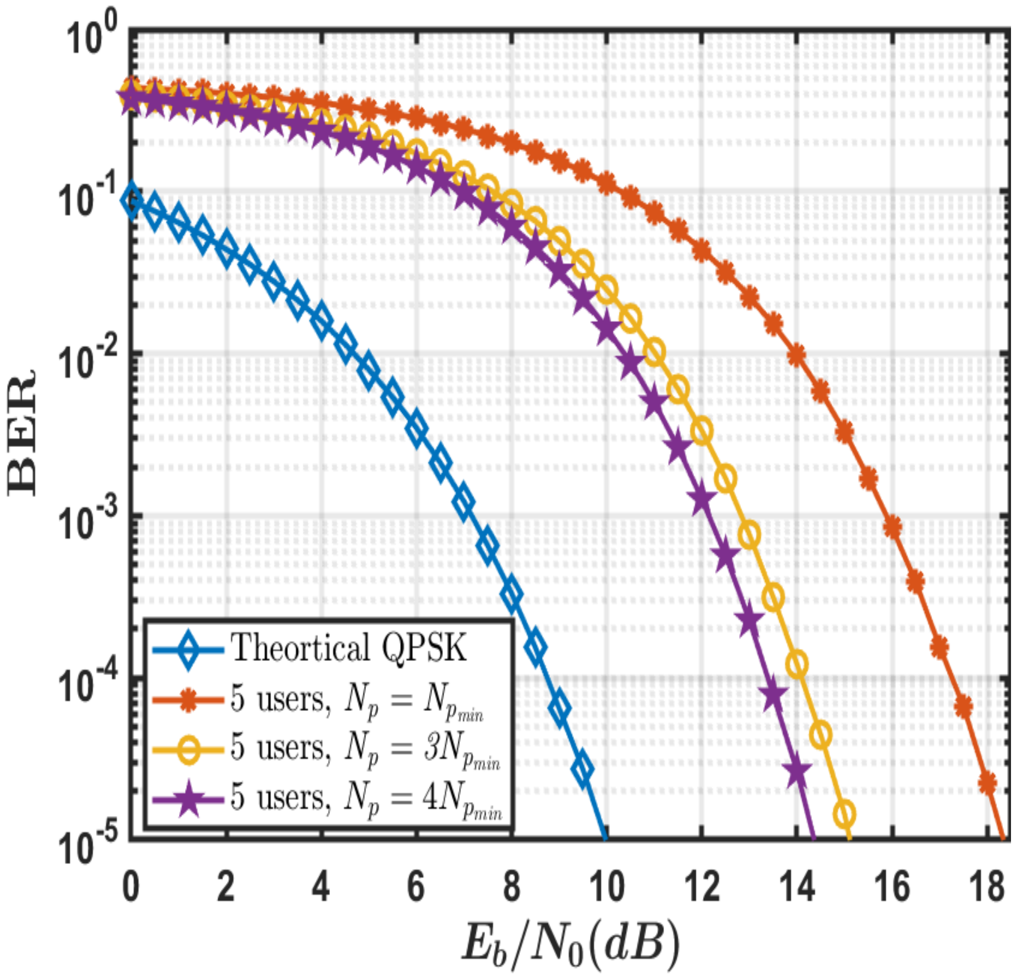

Fig. 6-5. BER of the proposed system in AWGN channel using method two.

6.3.1.5 Testing method one of the proposed reference averaging technique using the proposed OFDM-DQCSK system and comparing the performance with the OFDM-DCSK benchmark system in [11] in a multiuser scenario

Method one in section 5.4.1 of the proposed averaging technique is employed in Fig. 6-6 to improve the performance degradation because of correlating data symbols with a noisy reference. A comparison is held in Fig. 6-6 between the proposed system and the benchmark system in [11]. The BER performance of the proposed system nearly reaches the coherent one.

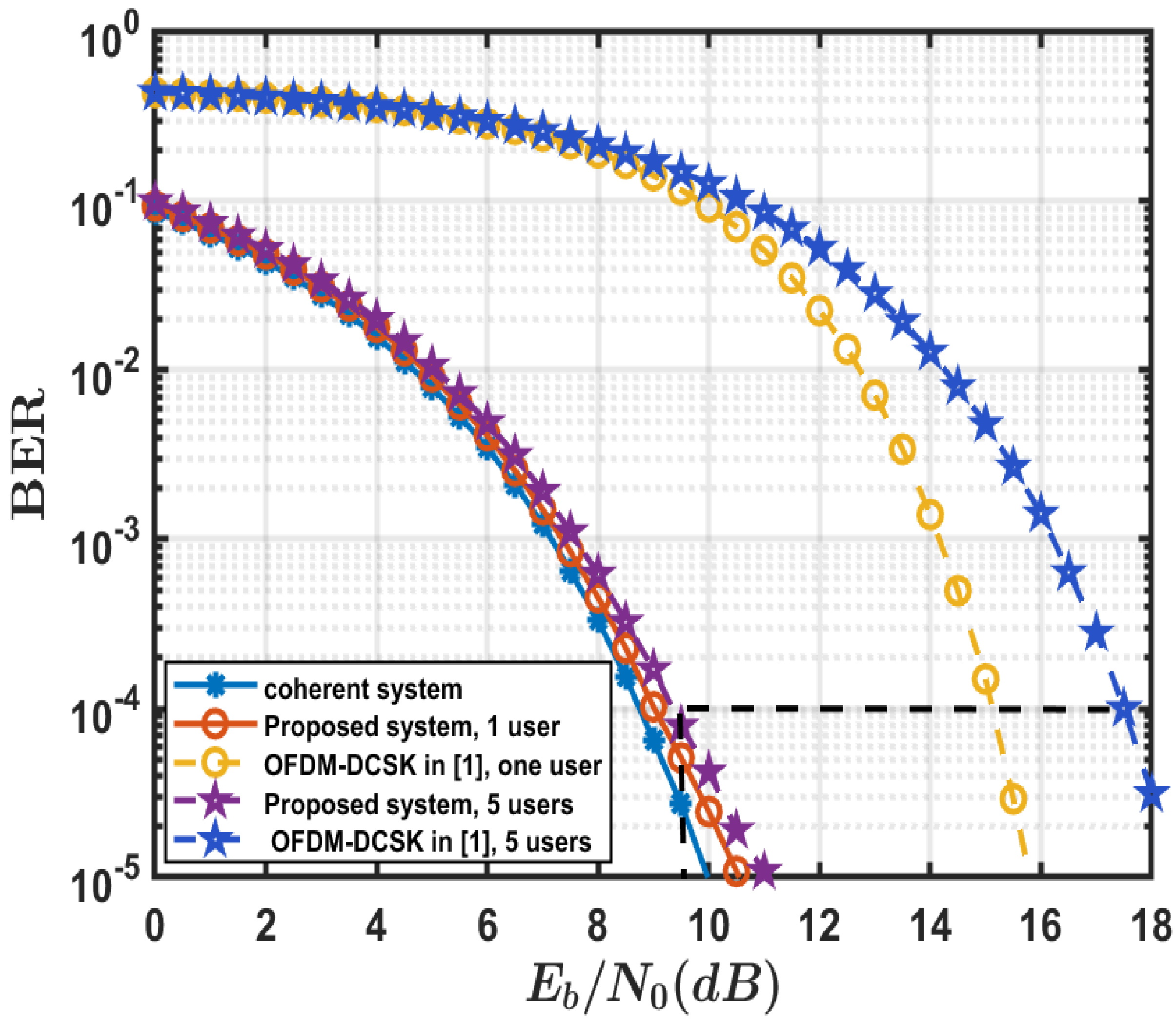

Fig. 6-6. BER comparison between the proposed and benchmark systems over AWGN channel.

This is because the AWGN on the reference signal is nearly eliminated. The simulation is done in the presence of an AWGN channel for five users. It can be shown that the proposed OFDM-DQCSK system outperforms the system OFDM-DCSK in [11] by more than 8 dB for the same BER. The simulation parameters in Fig. 6-6 are an FFT size equals to $N_T = 128$, $N_S = 49$ and $L_c = 120$. The number of users $U = 1, 5$.

6.3.2 Multipath Rayleigh flat fading channel in the presence of AWGN

6.3.2.1 Testing the performance of the proposed OFDM-DQCSK system using method one of the proposed Reference Averaging Technique

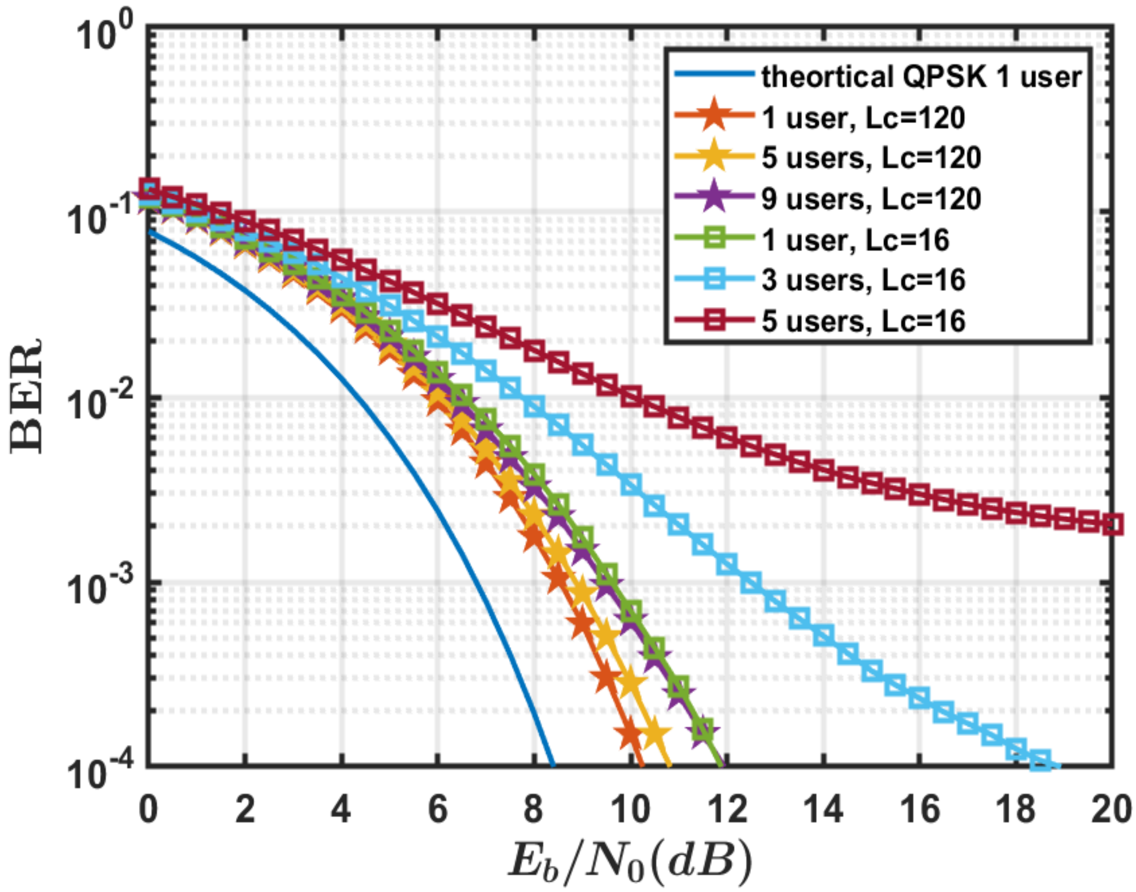

Fig. 6-7. BER performance of the proposed system in AWGN and multi-path Rayleigh flat fading channel.

The performance of the proposed OFDM-DQCSK in AWGN and multipath Rayleigh flat fading channels is evaluated and presented in Fig. 6-7. The channel fades every OFDM symbol. For $N_S = 70$ symbols, an FFT size $N_T = 128$ and various values of the spreading factor (i.e., $L_c = 16$ and $L_c = 120$) for single and multi-user scenario. The channel average power gain $E(\alpha_u^2) = 1$. Method one of the reference averaging technique is used. Therefore, the scheme shown in Fig. 5-6 is employed. It is observed that increasing the spreading factor length is very effective in combating the multipath fading distortion.

6.3.2.2 Comparing the BER performance of the proposed OFDM-DQCSK system with the OFDM-DCSK benchmark system in [11] in multi-path Rayleigh flat fading channel in the presence of AWGN for a multiuser scenario.

In Fig. 6-8, a BER performance comparison is made between the OFDM-DQCSK system proposed in this work and the benchmark OFDM-DCSK system in [11]. The simulation parameters are an FFT size $N_T = 128$, $N_s = 49$ data symbols, and a spreading factor $L_c = 120$ over a multipath Rayleigh flat fading channels in the presence of AWGN. Channel fades every three OFDM symbols. The AWGN averaging method in Fig. 5-6 is used in the proposed system. Also, the simulation is done for a multiuser situation $U = 6$. The channel average power gain $E(\alpha_u^2) = 1$.

Simulation results show that our proposed OFDM-DQCSK system outperforms the OFDM-DCSK in [11] by more than 8 dB at a BER of 10^{-4}. The reason for this improvement comes from three factors. The first factor is the use of the proposed orthogonal chaotic codes. The second factor is the usage of the reference averaging technique proposed in this work in section 5.4.1. The last factor is the use of quadrature modulation that increases the energy efficiency compared to the binary modulation in this special case.

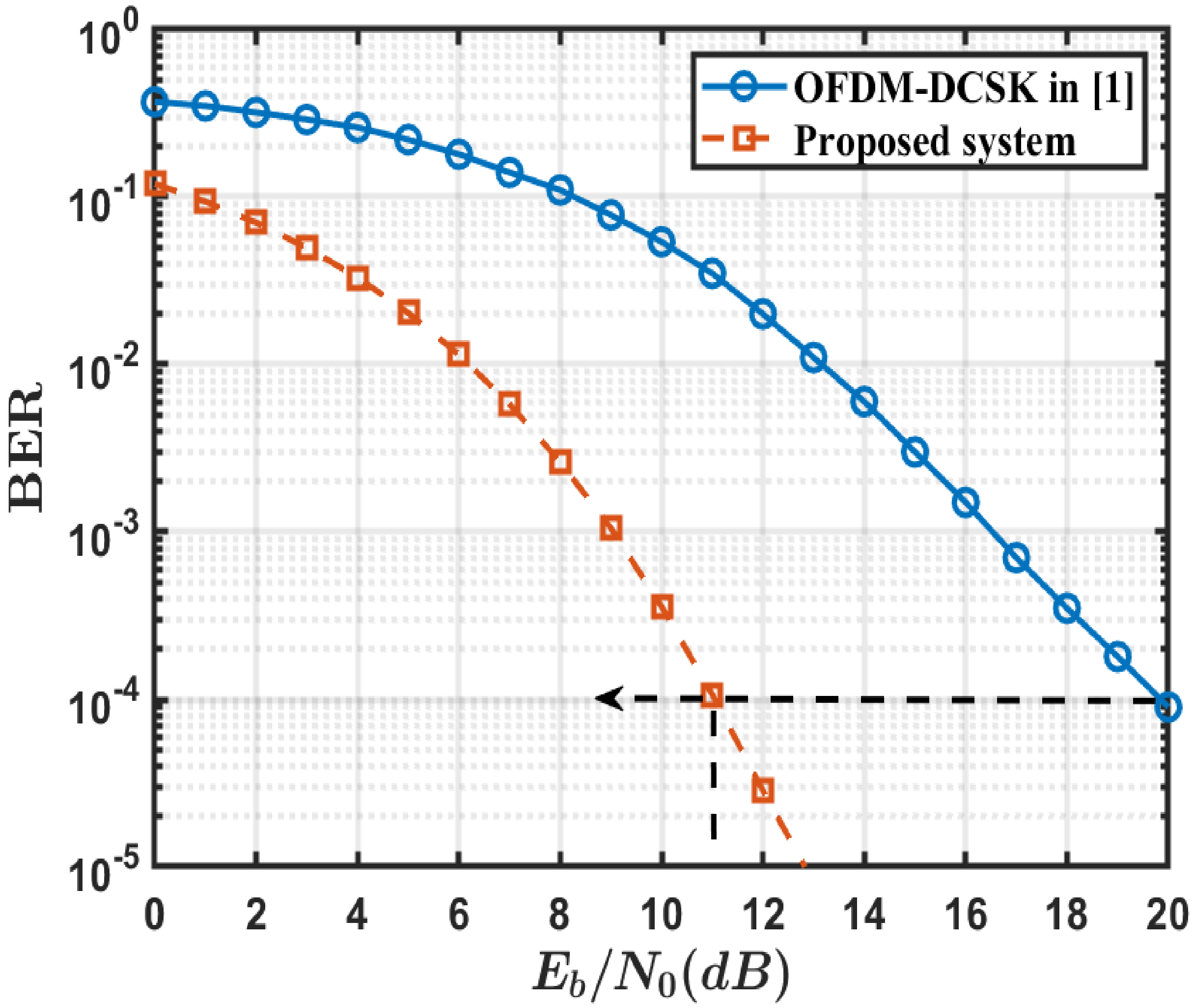

Fig. 6-8. BER comparison between the proposed and benchmark systems over multipath Rayleigh fading channel.

Chapter 7

Conclusions and Summary

A multiuser OFDM-DQCSK system is presented in this work. The proposed system aims to improve the BER performance of the systems in [11], [12] and [10]. Three contributions are the source of these improvements. Firstly, the chaotic reference signals are averaged at the receiver so that the additive white gaussian noise (AWGN) on the reference signal is averaged to a negligible value almost zero. Also, the multipath channel coefficients are averaged to their mean. In this way, the problem of correlating data symbols with a noisy reference is solved, and as a result, the proposed non-coherent system nearly reaches coherent performance. The second contribution is the generation of binary chaotic codes which are orthogonal and practical to be used in digital signal processors. Finally, the use of QPSK modulation instead of binary modulation increases energy efficiency because the reference code chip is used to reference two bits instead of one. The proposed system allows multiuser transmission in the presence of AWGN and multipath fading channels. The proposed system employs the comb-type pilot structure so that the reference signals can follow the changes in the channel coefficients. As a result, no need for a complex channel estimator or a spreading code synchronization. Simulation results of the bit error rate performance in the presence of an AWGN and multipath Rayleigh flat fading channels are presented. By comparing the performance of the proposed system to that in [11], [12] and [10], the proposed system achieves a considerable performance enhancement.

Future work

- Using a higher order chaotic map instead of first order chaotic logistic map may give a chance to extract more orthogonal chaotic codes using an appropriate searching algorithm. As a result, more users can use the same channel with less interference.

- Using a quadrature component for the reference signal may increase the bandwidth efficiency.

References

[1] J. G. Proakis and M. Salehi, *Digital Communications,* New York: The McGraw-Hill Companies, Inc, 2008.

[2] L. Hanzo, M. Munster, T. Keller and B. J. Choi, *OFDM and MC-CDMA for Broadband Multi-User Communications, WLANs and Broadcasting,* Chichester, England: Wiley, 2003.

[3] S. Kondo and B. Milstein, "Performance of Multicarrier DS CDMA Systems," *IEEE Transactions on Communications,* vol. 44, no. 2, pp. 238-246, Feb. 1996, doi: 10.1109/26.486616.

[4] L. Novosel and G. Šišul, "Performance evaluation of chaotic sequences for DSSS multiple access on software defined radio," *EURASIP Journal on Wireless Communications and Networking,* pp. 1-4, 2017, doi: 10.1109/ICECom.2016.7843896.

[5] C. Fatima and D. Ali, "New chaotic binary sequences with good correlation property using logistic maps," *IOSR Journal of Electronics and Communication Engineering (IOSR-JECE),* vol. 5, pp. 59-64, Jan. 2013, doi: 10.9790/2834-0535964.

[6] F. C. M. Lau and C. K. Tse, *Chaos-Based Digital Communication Systems, Operating Principles, Analysis Methods and Performance Evaluation,* New York: Springer, 2003.

[7] K. T. Alligood, T. D. Sauer and J. A. Yorke, *CHAOS: An Introduction to Dynamical Systems,* New York, NY, USA: Springer, 1996.

[8] L. Lei, F. Xu, Y. Xu and Y. Wu, "A Chaotic Direct Sequence Spread Spectrum Communication System in Shallow Water," in *International Conference on Control, Automation and Systems Engineering (CASE)*, Singapore, 2011.

[9] A. P. Kurian, S. Puthusserypady and S. M. Htut, "Performance enhancement of DS-CDMA system using chaotic complex spreading sequence," *IEEE Trans. Wireless Commun.*, vol. 4, no. 3, pp. 984-989, May 2005, doi: 10.1109/TWC.2005.847028.

[10] G. Kaddoum, F. Gagnon and F.-D. Richardson, "Design of a secure multi-carrier DCSK system," *in Proc. 9th Int. Symp. Wireless Commun.Syst. (ISWCS)*, p. 964–968, Jun. 2012, doi: 10.1109/ISWCS.2012.6328511.

[11] G. Kaddoum, "Design and Performance Analysis of a Multiuser OFDM Based Differential Chaos Shift Keying Communication System," *IEEE Transactions On Communications*, vol. 64, no. 1, pp. 249-260, Jan. 2016, doi: 10.1109/TCOMM.2015.2502259.

[12] G. Kaddoum, F. Richardson and F. Gagnon, "Design and analysis of a multi-carrier differential chaos shift keying communication system," *IEEE Trans. Commun.*, vol. 61, no. 8, p. 3281–3291, Aug. 2013, doi: 10.1109/TCOMM.2013.071013.130225.

[13] S. Li, Y. Zhao and Z. Wu, "Design and analysis of an OFDM-based differential chaos shift keying communication system," *J. Commun.*, vol. 3, pp. 199-205, 2015, doi: 10.12720/jcm.10.3.199-205.

[14] C. Qasim, "dsprelated," 1 May 2017. [Online]. Available: https://www.dsprelated.com/showarticle/1046.php.

[15] A. M. Niknejad, "Introduction to Mixers," 2005. [Online]. Available: http://rfic.eecs.berkeley.edu/~niknejad/ee142_fa05lects/pdf/lect15.pdf.

[16] S. Haykin, *Communication Systems,* 4th ed., New York: John Wiley & Sons, Inc., 2001.

[17] B. P. Lathi and Z. Ding, *Modern digital and analog communication systems,* 4th ed., New York: Oxford University Press, 2010.

[18] G. Heidari-Bateni and C. McGillem, *Chaotic sequences for spread spectrum: An alternative to PN-sequences,* Vancouver, Canada: Proc. IEEE International Conference on Selected Topics in Wireless Communications, 1992.

[19] G.Kolumban, M.P.Kennedy and L.O.Chua, "The role of synchronization in digital communications using chaos - Part II: Chaotic modulation and chaotic synchronization," *IEEE Transactions on Circuits and Systems Part I,* pp. 1129-1140, 1998, doi: 10.1109/81.735435..

[20] G.Kolumban and M.P.Kennedy, *Digital communication using chaos, Controlling Chaos and Bifurcation in Engineering Systems,* Boca Raton FL: CRC Press, 2000.

[21] G. Kolumbán, B. Vizvári and W. Schwarz, "Differential chaos shift keying: A robust coding for chaos communication," Seville, Spain, January 1996.

[22] Z. Galias and G. Maggio, "Quadrature chaos shift keying," Sydney, Australia, 2001.

[23] Y. Shen and E. Martinez, "Channel estimation in OFDM systems," Freescale Semiconductor, Austin, TX, USAUSA, 2006.

[24] S. Coleri, M. Ergen, A. Puri and A. Bahai, "Channel estimation techniques based on pilot arrangement in OFDM systems," *IEEE Trans. Broadcast,* vol. 48, no. 3, pp. 223-229, 2002, doi: 10.1109/TBC.2002.804034.

[25] T. A. Lin and C. Y. Lee, "Predictive equalizer design for DVB-T system," *IEEE Int. Symp. Circuits Syst.*, vol. 2, pp. 940-943, 2005, doi: 10.1109/ISCAS.2005.1464744.

[26] Leon-Garcia and Alberto, *Probability, Statistics, and Random Processes for Electrical Engineering*, Upper Saddle River, NJ, USA: Pearson Prentice Hall, 2008.

[27] W. Feller, *An Introduction to Probability Theory and Its Applications*, New York: John Wiley & Sons, Inc., 1970.

[28] IEEE Standards Association, "802.11a-1999 - IEEE Standard for Telecommunications and Information Exchange Between Systems," [Online]. Available: https://standards.ieee.org/standard/802_11a-1999.html.

[29] M. Sushchik, L. S. Tsimring and A. R. Volkovski, "Performance analysis of correlation-based communication schemes utilizing chaos," *IEEE Transactions on Circuits and Systems I: Regular Papers*, vol. 47, no. 12, pp. 1684 -1691, Dec. 2000, doi: 10.1109/81.899920.

Publications

A. Y. Hassan, M. O. El-Beshry, K. T. Harb and H. El-Hennawy, "Design and Performance Evaluation of a Multiuser OFDM System Based on Differential Quadrature Chaos-Shift-Keying Spread Spectrum," in *38th National Radio Science Conference (NRSC 2021)*, Mansoura, Egypt, 2021.

٢- باستخدام طريقة اخذ المتوسط للإشارات المرجعية المبنية على قانون الأعداد الكبيرة، تم التغلب على مشكلة الأداء الضعيف للأنظمة المعتمدة على الإشارات المرجعية وأصبح أداء معدل الخطأ في البيانات يصل تقريبا الى أداء الأنظمة المعتمدة على المزامنة بين المُرسِل والمُستقبِل. وتم تسجيل جميع النتائج التي تثبت ذلك.

٣- تم تصميم برنامج يبحث عن القيم الأولية في مولدات الإشارات العشوائية لكي تعطى إشارات عشوائية متعامدة والتي عند استخدامها لتوسيع النطاق الترددي نتج عن ذلك أداء أفضل من استخدام الإشارات المعتادة.

٤- النظام المستخدم في هذا العمل يستعمل إشارات مرجعيه تُرسَل مع إشارات البيانات لكي تقوم باستخلاصها بدون الحاجة الى مزامنه بين المُرسل والمُستقبِل نهائيا وبدون الحاجة الى تقدير نوع التشويش الحادث في القناة.

٥- تم رصد تحسن ملحوظ في كفاءة استخدام الطاقة عند استعمال النظام الجديد مقارنة بالأنظمة السابقة.

الخلاصة

يكاد النظام الجديد المقترح الذي لا يعتمد على المزامنة يصل إلى أداء النظام الذي يعتمد على المزامنة على حساب زيادة قليله من النطاق الترددي.

تتميز مولدات الإشارات العشوائية بأنها حساسة جدا للقيم الأولية حتى وإن كان الفارق بين هذه القيم قليل جدا. هذا يعنى انه نظريا يمكن توليد عدد لا نهائي من الإشارات العشوائية بدون أي ترابط حتى ولو كانوا جميعا من نفس المولِّد العشوائي.

عملية مزامنة الإشارات المستخدمة في نظام توسيع النطاق الترددي عملية ليست سهله وتحتاج الى مُستقبِل أكثر تعقيدا. ولذلك عمليا يتم استخدام نظام لا يعتمد على المزامنة ويعتمد على إرسال إشارة مرجعيه مع إشارة البيانات. هذه الإشارة المرجعية قد استخدمت في جهاز الإرسال لتوسيع النطاق الترددي لإشارة البيانات. سيتم إرسال نفس هذه الإشارة المرجعية مع إشارات البيانات لكي تستخدم في المُستقبِل لاستخلاص إشارة البيانات من الإشارة المستقبلة، وبالتالي لن تحتاج هذه العملية الى مزامنه بين المُرسل والمُستقبِل ولن تحتاج أيضا الى محاوله تقدير التشويش الحادث للإشارة المستقبلة.

أهداف الرسالة

١- تصميم وتحليل نظام مكون من دمج نظام توسيع النطاق الترددي (spread spectrum) مع نظام التقسيم المتعامد للترددات (OFDM) لكي يحقق أداء ممتاز في معدل الخطأ في البيانات عند الإرسال في قناة بها تشويش (Gaussian) المجموع على كل الترددات ووجود التداخل الناتج عند استخدام العديد من المستخدمين نفس القناة ووجود عدة مسارات لنفس الإشارة.

٢- استغلال الخواص الممتازة للإشارات العشوائية كبديل للإشارات المعتادة لتوسيع النطاق الترددي وتقليل التداخل بين المستخدمين لنفس القناة والنطاق الترددي وبالتالي إمكانية زيادة عدد المستخدمين لنفس النطاق الترددي.

٣- تحسين أداء أنظمه توسيع النطاق الترددي التي تستخدم الإشارات المرجعية بدلا من المزامنة والتي تبين أنها تعاني من تدهور كبير في الأداء خاصة عند استخدام معامل توسيع كبير.

نتائج الرسالة

١- تصميم نظام (OFDM-DQCSK) والذي يهدف الى تطوير الأنظمة السابقة التي تعتمد على الإشارات المرجعية التي لا تحتاج الى مزامنه. النظام الجديد يعطى أداء اعلى بكثير في وجود العديد من المستخدمين لنفس النطاق الترددي ويسمح بوجود عدد أكبر من المستخدمين، وكل ذلك في ظروف قاسيه من التشويش سواء من وصول الإشارة من عدة مسارات مختلفة أو التشويش المجموع على كل الترددات (AWGN) وتم عرض النتائج والمقارنة مع الأنظمة السابقة.

الملخص العربي

نبذة مختصرة

الهدف من هذا العمل هو تصميم نظام OFDM ليكون متعدد المستخدمين مدمجاً مع نظام توسيع النطاق الترددي (DS-SS) باستخدام الإشارات العشوائية (CHAOS). يستبدل النظام الجديد المقترح أنظمة توسيع الترددات القديمة بنظام توسيع التردد العشوائي. يمكن إنشاء مجموعات غير مترابطة ومتعامدة من الإشارات العشوائية بخصائص ارتباط ممتازة. يستخدم النظام الجديد المقترح مجموعة قنوات اتصال بحيث تكون خاصة لكل مستخدم وغير مشتركه بينهم. تحمل هذه القنوات المساعدة الكود الخاص بكل مستخدم مستعملا الإشارات العشوائية التي تُستخدم كمرجع لقنوات الاتصال التي تحمل المعلومات لمستخدم معين. أما بالنسبة للمعلومات، فيتم إرسالها على قنوات اتصال مشتركه باستخدام نظام توسيع النطاق الترددي. تم تحديد عدد قنوات الاتصال الخاصة المساعدة ليكون أقل بكثير من عدد قنوات الاتصال المشتركة التي تحمل المعلومات حفاظا على عدم إهدار النطاق الترددي. ليست هناك حاجة إلى مزامنة بين الجهاز المُرسِل والمُستقبِل أو معالجات معقدة في الجهاز المستقبل لإزالة التشويه الذي قد يحدث للمعلومات. تم اخذ المتوسط لقنوات الاتصال الخاصة المساعدة بحيث يتم تقليل تأثير الإشارات غير المرغوب فيها (AWGN) المضافة على قنوات الاتصال الخاصة المساعدة إلى الحد الأدنى بحيث يصبح هذا التشويش قيمة ضئيلة جدا قد تساوى صفرا مقارنة بالإشارة المرجعية. تم إجراء محاكاة لأداء معدل الخطأ في المعلومات في وجود AWGN وقناة اتصال متعددة المسارات باستخدام نموذج Rayleigh وفى وجود تداخل من أكثر من مستخدم. في الختام، يكاد النظام الجديد المقترح الذي لا يعتمد على المزامنة يصل إلى أداء النظام الذي يعتمد على المزامنة.

مقدمه

يستطيع أكثر من مستخدم الإرسال على نفس النطاق الترددي من خلال نظام توسيع النطاق الترددي(spread spectrum) . هذا النظام يقاوم التشويش الناتج عن وصول نفس الإشارة من عدة مسارات مختلفة (multi-path fading). وبالرغم من هذه المميزات، إلا أن أداء هذا النظام يتأثر بشدة من تداخل الإشارات مع نفسها عند زيادة سرعة معدل نقل البيانات في الثانية الواحدة ويتأثر أيضا بالتشويش الناتج من المستخدمين الأخرين لنفس النطاق الترددي. لمعالجة هذا التأثير يتم دمج نظام توسيع النطاق الترددي مع نظام التقسيم المتعامد للترددات (OFDM) عند الرغبة في استخدام معدل نقل بيانات عالي جدا.

كلما زاد عدم الارتباط بين الإشارات المختلفة للمستخدمين، كلما كان إمكانية زيادة عدد المستخدمين لنفس النطاق الترددي أكثر في نظام توسيع النطاق الترددي. ولذلك في هذا العمل، تم استخدام الإشارات العشوائية (chaotic signals) لما تبين عنها من عدم الترابط بينها بشكل ممتاز. وتتميز أيضا بسهولة توليدها بمقارنتها بالإشارات الأخر

الموقعين أدناه قاموا بمراجعة الرسالة العلمية بعنوان **"تصميم وتحليل أداء نظام الاتصالات للترددات المتعامدة لأكثر من مستخدم معتمداً على التغيير العشوائي المتعامد للطيف الترددي الممتد"** مقدمة من **المهندس/ محمود أسامه البشرى عبد التواب إلى** كلية الهندسة ببنها – جامعة بنها كجزء من متطلبات الحصول على درجة الماجيستير في العلوم الهندسية في الهندسة الكهربية من كلية الهندسة ببنها.

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جامعة بنها

كلية الهندسة ببنها

قسم الهندسة الكهربية

تصميم وتحليل أداء نظام الاتصالات للترددات المتعامدة لأكثر من مستخدم معتمداً على التغيير العشوائي المتعامد للطيف الترددي الممتد

رسالة مقدمة من

محمود أسامه البشرى عبد التواب

كجزء من متطلبات الحصول على درجة الماجيستير في العلوم الهندسية في الهندسة الكهربية

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