TRANSFORM DOMIAN / CYCLIC CODE SHIFT KEYING SYSTEM ON AN URBAN MULTIPATH CHANNEL

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Chapter One

Introduction

1.1 Introduction and Purpose

In the last three decades, wireless communication has grown exponentially. This revelation has many requirements to remain accurate, such as experts, devices, maintenance, the orientation of the large number of users, and solving all the problems in the system due to this huge number of users. These problems are becoming more complicated, especially in areas that are crowded with electromagnetic wave signals. Moreover, the bandwidth is almost totally occupied in these areas, which yields interference between the users.

Because of these circumstances, wireless communications experts surprised the world with a new technique designed especially for solving the interference among users, as well as the problem of the bandwidth. This scheme is called "spread spectrum", where the bandwidth of the transmitted signal is much greater than the bandwidth of the information signal itself. Therefore, this technique generates other modern schemes and causes the competition to increase among the companies and individuals that have an interest in such subjects. Thus, the innovation of new techniques is happening every day;
however, inventing the code division multiple access --or what is called CDMA-- brings a furor among the wireless communications experts themselves. CDMA is a multiple access technique and is considered one of the spread spectrum applications. Since in CDMA many users can easily share the bandwidth, the bandwidth problem is almost solved using this technique. This topic will be discussed in a later chapter.

In 1992, Smallcomb [7] proposed a new scheme. This scheme is called transform domain cyclic code shift keying (TD/CCSK), which is one kind of CDMA. This research is related to this topic. In this research we are aiming to apply this technique in the environments which are crowded with users and electromagnetic wave signals. This environment is known as the urban environment.

Therefore, the goal of this research is to look for a good model that represents the urban multipath channel and then apply this model to the TD/CCSK system to study the performance, based on the bit error rate curve, in the multipath urban channel. Thus, by the end of this research we will know whether this system --TD/CCSK-- is capable of solving the urban multipath channel problems.

1.2 Fundamentals of Spread Spectrum (SS)

After the end of the Second World War, the demands of the new revolution of digital communication were incredibly insistent. Therefore, the communication experts invented a new modulation scheme where the bandwidth of the signal is much wider than the bandwidth of the information. In other words, they expanded the bandwidth of the signal without any change in the bandwidth of the information. This technique is called
spread spectrum (SS). Although spread spectrum was basically a military-based technology in the past, nowadays spread spectrum has both military and commercial applications. This increase in the use of spread spectrum multiplied the number of problems in communication systems, as seen by the many studies and papers done in this field. Figure (1.1) shows the simplified model of spread spectrum. Any system called a spread spectrum system must meet three conditions. First, the signal bandwidth must be much greater than the minimum information bandwidth. Second, the transmitted signal must be spread by using a code signal that is independent of the information signal. Finally, at the receiver, the received signal must be despread by using the synchronized replica of the spreading signal that was used to spread the transmitted signal to recover the information signal. There are many techniques used for spread spectrum. For instance, direct sequence (DS), frequency hopping (FH), time hopping (TH), hybrid (DS/FH), and hybrid (DS/TH). There are many desirable characteristics of spread spectrum; for example, low power spectral density, antijamming capability, multiple access capability, and robustness against multipath fading.

1.3 History of the Transform Domain (TD) Signal Processing

A signal is a function that represents data, typically about the state or behavior of a physical system [6]. In terms of transformation, Digital Signal Processing (DSP) deals with the signals that are discrete in both amplitude and time [6]. In the 1960s, a new digital signal processing (DSP) discarded the analog signal processing of electronic and mechanical devices and replaced it with digital computers.
Figure (1.1) Simplified Model of Spread Spectrum [4].

However, these digital computers were insufficient because of the huge amount of processing requirements of digital signal processing algorithms and the limited processing power of the microprocessors at that time. As a very important advantage of DSP, it is easier to deal with signals in the frequency domain than in the time domain. One way to make this transformation is to compute Discrete Fourier Transform (DFTs) of the analog signal. Cooley and Tukey have invented a very useful algorithm called Fast Fourier Transform (FFT). This algorithm is the best way to compute the (DFT) of the signal. Although this algorithm was invented in the mid 1960s, the development of real-time DSP implementations in the 1980s was possible because of the emergence of the integrated circuit. [7]

1.4 History of Rake Correlator

To combat the natural fading and multipath on High Frequency (HF) channel, Price and Green in 1958 did their best to introduce a solution to these problems. This solution is called the Rake Correlator, designed to solve multipath problems where a
multipath signals in company with noise. With different time delay, the Rake Correlator can isolate the arriving signal by transmitting the signal over a wide frequency band. This will give good performance; however, it is a waste of the bandwidth. [7]

1.5 Overview

This thesis contains seven chapters, which are described as follows. Chapter Two is background information that may be needed in future chapters, such as the discussion of the fundamentals of the direct sequence and multiple access spread spectrum. It then reviews the concepts of the code division multiple access (CDMA) and gives general information about the cyclic code shift keying (CCSK). Chapter Three is the implementation of the cyclic code shift keying (CCSK). This chapter, in which Smallcomb’s work is reviewed, is organized as follows. First, it begins with a description of the M-ary orthogonal signals. Then it discusses how to generate the CCSK waveforms, then it explains one method (BMLS) that is used to generate CCSK waveform. Finally, the implementation of this system is explained with its properties. At the end of this chapter, digital signal processing (DSP) and transform domain discussion is proposed and TD/CCSK is studied.

Chapter Four is mainly focused on the multipath channel. Specifically, it discusses in more depth the multipath urban channel. Then, it gives an explanation of the characteristics of the fading multipath channel. Finally, the model of the multipath urban channel is proposed in the last section of this chapter.

Chapter Five is the theoretical chapter, and it mainly discusses the theoretical performance of the TD/CCSK that has been discussed by Smallcomb (1992) [7]. This
chapter consists of three parts: first, the general theoretical performance of the TD/CCSK system. Second, the two cases (ideal and instantaneous) of the TD/CCSK with transmitted reference are reviewed, and finally, there is a comparison of the CCSK systems.

The computer simulation results and the discussion of this thesis are proposed in Chapter Six. Finally Chapter Seven is about the summary of the work, strengths and the weakness in the work, the conclusion, and the recommendations for future work.
Chapter Two

Background

2.1 Direct Sequence Spread Spectrum (DS)

Figure (2.1) Overall Direct Sequence System (Transmitter and Receiver) [3].

Direct sequence is one of the best-known spread spectrum techniques and the most widely used spread spectrum system. This is because direct sequence systems do not need a high-speed, fast settling frequency synthesizer, resulting in a simple implementation (see Figure 2.1). In direct sequence, the information signal is multiplied by a Pseudo Random Noise Code (PN Code). A PN-Code is the sequence that has two
has two values, either 1 or −1 (polar) randomly generated. Since the generation of the PN-Code needs only a number of shift registers, it is practical and easy to implement.

![Diagram of Direct Sequence Spreading Example](image)

Figure (2.2) Direct Sequence Spreading Example.

A direct example of spreading direct sequence is illustrated in Figure (2.2), with each bit of the information signal equivalent to 7 chips of the PN-Code. In the coded signal, we have 7 chips instead of one bit, so we spread the information signal. At the receiver, all we need to do to recover the information signal is to despread the received signal by multiplying it by the same (synchronized) PN-Code.
2.2 Multiple Access CDMA

MA/CDMA is the access that has many users sharing the same radio frequency (RF) bandwidth (BW) at the same time without any interference with one another, as illustrated in Figure (2.3).

![Figure (2.3) The Spectrum of CDMA [11].](image)

There are basically three common multiple access schemes. The first is Frequency Division Multiple Access (FDMA). The second scheme is what is called Time Division Multiple Access (TDMA). In these schemes, frequencies and time slots are used in FDMA and TDMA, respectively, to transmit and distinguish between signals. However, in the last scheme, Code Division Multiple Access (CDMA), the mathematical code is used. CDMA is one application of spread spectrum technique that has a very large bandwidth due to the use of the spreading code to spread the information signal. In CDMA, each user is assigned his own code (a unique code) sequence. These differences of the code sequences allows users to share the bandwidth at the same time. To get rid of the interference between users in the channel, the codes have to have zero-cross
correlation --or at least very low values of cross-correlation-- between any two different codes. Thus, the information signal can be easily recovered by multiplying the received signal by the synchronized de-spreading code (the matched spreading code) to de-spread. The remaining signals (users) are considered noise.

2.3 Cyclic Code Shift Keying (CCSK)

Cyclic code is very well known among the family of codes. There are many ways to encode them; however, a linear shift register is used. The linear shift register is considered the best among the encoding schemes because in this register there is no need for storage since the code words are generated by shifting and adding only.

Cyclic Code Shift Keying (CCSK) is one form of the Code Division Multiple Access (CDMA). Since CDMA is one of the applications of spread spectrum (SS), CCSK is a form of spread spectrum (SS). In CCSK, the signal is purposely spread in such a way that it is similar to direct sequence-spread spectrum (DSSS); however, in CCSK the signal is M-ary orthogonal or nears orthogonal signaling schemes.

There are several steps to attain CCSK. First, one must generate the spreading code base vector \(s_0\) that synchronized with the de-spreading code at the receiver. Second, one must convert the binary information to its decimal value. Then, by taking the cyclic shift of the base vector \(s_0\) nth time, where \(n\) is the equivalent value of the decimal information value, M-ary signal set \(\{s_i\}\) is generated. Two important criteria must be met to design a CCSK system [7]:

1- The circular auto-correlation peak must be the highest value.

2- The circular cross-correlation peak must be very low value.
Chapter Three

Implementation of CCSK

3.1 Binary and Orthogonal M-ary Signal

Binary is a method of storing information represented by a set of values using either number zero (negative) or one (positive). Any set of N linearly independent function \( \{ \Psi_i(t) \} \) is defined by N-dimensional orthogonal space. These orthogonal functions are called basis functions. The orthogonality aims to maximize the distance between the signals and minimize the energy. In addition, they are very helpful functions because any arbitrary function in the space can be represented by a linear combination of these basis functions. Any basis function must have the following conditions:

\[
\int_{0}^{T} \Psi_j(t) \Psi_k(t) dt = K_j \delta_{jk} \quad 0 \leq t \leq T
\]

(3.1)

where j, k=0,1,2,3…N, and \( \delta_{jk} \) is the Kronecker delta function and \( K_j \) is constant.

When the constant \( K_j \) is non-zero, the signal space is called orthogonal. However, when the basis functions are normalized where \( K_j \) equals one, the space is called orthonormal space. Each basis function \( \psi_j(t) \) of the set must be independent of every other member in the set. Also, it must not interfere with any other member of the basis function set. [2]
Therefore, M-ary orthogonal signals are a set of signals that have zero cross-correlation. The M-ary orthogonal waveform \( \{s_i(t)\} \) can be represented as a set of vectors:

\[
S_i = (s_{i1}, s_{i2}, s_{i3}, \ldots, s_{iN})
\]

That is

\[
S_i(t) = \sum_{j=1}^{N} S_{ij} \psi_j(t) \quad i=1,2,3,\ldots,M; \quad 0 \leq t \leq T
\]

### 3.2 Generation of CCSK Waveform

The base vector plays a very important role in CCSK systems because the signals that are transmitted using CCSK are nothing more than the circular shifted version of the base vector \( s_{0} \). Linear Frequency Modulation (LFM) signals, filter impulse trains based on Nyquist sampling theorem, and Binary Maximum Length Sequence (BMLS) are some of waveforms that can represent the base vector \( s_{0} \). Since BMLS waveforms are simple to generate with very few number of stages in shift registers, and since they have high peak in their circular auto-correlation properties, BMLS waveform are commonly used. Therefore, the next section will focus on BMLS.

#### 3.2.1 Binary Maximum Length Sequence (BMLS)

When BMLS register is applied to yielding CCSK, the result is a sequence that has several properties [1]:

1. In each period of the sequence, the total number of plus ones overcomes the total number of minus ones by exactly one.
2. In each period, half of the runs of the same sign have length one, one fourth of the runs have length two, one eight of the runs have length three, and so forth. In addition, the number of negative runs equals the number of positive runs.

3. The circular auto-correlation of a periodic sequence with length of N is two-valued.

4. Because by comparing the periodic sequence component by component with any cyclic shift of the sequence itself, the number of terms that have the same sign differ from the one that have different sign by almost one.

\[
C(k) = \sum_{n=1}^{N} a_n a_{n+k} \tag{3.4}
\]

\[= N \quad \text{for} \quad k = 0, N, 2N, \ldots \ldots \ldots \]

\[= -1 \quad \text{otherwise} \]

and for periodic sequence with period N,

\[a_{n+N} = a_n \tag{3.5}\]

where, \(a_n\) is the cyclic code in the n’th position, \(a_{n+k}\) is the cyclic code in the \((n+k)\)’th position, and N is the length of the sequence code.

<table>
<thead>
<tr>
<th>Initial Setting</th>
<th>Max. Length Sequence</th>
</tr>
</thead>
<tbody>
<tr>
<td>000</td>
<td>0000000</td>
</tr>
<tr>
<td>001</td>
<td>1001110</td>
</tr>
<tr>
<td>010</td>
<td>0100111</td>
</tr>
<tr>
<td>011</td>
<td>1101001</td>
</tr>
<tr>
<td>100</td>
<td>0011101</td>
</tr>
<tr>
<td>101</td>
<td>1010011</td>
</tr>
<tr>
<td>110</td>
<td>0111010</td>
</tr>
<tr>
<td>111</td>
<td>1110100</td>
</tr>
</tbody>
</table>
Figure (3.1) General M-Stage Shift Register with Linear Feedback [7].
Corresponding to a generating polynomial, BMLS are generated easily by using shift register with feedback loops, as illustrated in Figure (3.1).

From the example shown in Table (3.1), all output sequences are circular shifts of each other except for the first sequence. A binary base vector and a binary reference vector are denoted as $\mathbf{s}_b$ and $\mathbf{s}_r$, respectively. These two vectors are BMLS that are generated in the same manner, though from different generating polynomial [7].

For instance, in the case of the Global Position System (GPS) where the number of stages equals ten (the length of both sequences is 1023 i.e. $2^{10} - 1 = 1023$), $\mathbf{s}_b$, the binary base vector, is constructed with generating polynomial of \( g_1(x) = x^{10} + x^3 \), and $\mathbf{s}_r$, the reference vector, is constructed with generating polynomial of \( g_2(x) = x^{10} + x^9 + x^8 + x^6 + x^2 \). Figure (3.2) and Figure (3.3) show the shift registers with feedback connections for the binary vectors [9]. In this example, if the initial entries of binary base vector $s_0$ is [9]

\[
0 \ 0 \ 0 \ 0 \ 1 \ 1 \ 1 \ 1 \ 1 \ 1
\]

the output vectors is then

\[
1 \ 0 \ 0 \ 0 \ 0 \ldots \ldots \ldots \ 1 \ 1 \ 1 \ 1 \ 1
\]

\[
\text{Repeated vector with length 1023}
\]

By applying the same initial entries, different cyclic codes are generated from different generating polynomials. Figure (3.4) shows the circular auto-correlation of these codes [9]. More than one m-sequence can be used in a complete spread spectrum communication system. Therefore, the number of m-sequence is an important property, and it is given by

\[
N = \frac{1}{m} \phi (2^m - 1)
\]
Figure (3.2) Shift Register with Generating Polynomial $g_1(x) = x^{10} + x^3$ [9].

Figure (3.3) Shift Register with Generating Polynomial $g_1(x) = x^{10} + x^9 + x^8 + x^6 + x^3 + x^2$ [9].

Figure (3.4) Circular Auto-Correlation of BMLS of Length 1023 [7].
where $\phi(k)$ is the number of positive integers less than $k$ and relatively prime to $k$, $m$ is the number of register stages, and $N$ is the total number of sequences.

Several examples of possible sequences that can be generated from shift register are shown in Table (3.1). If $k$ is a prime number (i.e. $k=31$), all of the integers less than $k$ are prime to it. So, the number of sequences is much greater [1].

Table (3.2) Example of the Number of Sequence for Sequence of Different Length [1].

<table>
<thead>
<tr>
<th>Register Stages $(m)$</th>
<th>Sequence Length $(M=2^m-1)$</th>
<th>Total Number of Sequence $N=1/m \phi (2^m-1)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>4</td>
<td>15</td>
<td>2</td>
</tr>
<tr>
<td>5</td>
<td>31</td>
<td>6</td>
</tr>
<tr>
<td>8</td>
<td>255</td>
<td>16</td>
</tr>
<tr>
<td>10</td>
<td>1023</td>
<td>60</td>
</tr>
<tr>
<td>12</td>
<td>4095</td>
<td>144</td>
</tr>
<tr>
<td>15</td>
<td>16383</td>
<td>1800</td>
</tr>
<tr>
<td>16</td>
<td>65535</td>
<td>2048</td>
</tr>
</tbody>
</table>

3.2.2 Development of CCSK System

Because FFT operations cannot be applied to any sequence that does not have a length a power of two, such as BMLS, CCSK sequence cannot be applied to this kind of
operation to take advantage of transform domain. Although the addition of one term to
the sequence in BMLS will destroy the perfect circular auto-correlation property, it is
strongly needed to make CCSK sequence suitable for transform domain. Figure (3.5)
shows a high correlation peak for CCSK with a length of 1024, even though a perfect
auto-correlation property does not exist. Therefore, it can take advantage of transform
domain processing to apply to CCSK/TD system (which will be studied in the next
section). According to [9], the channel capacity will be maximized if the base vectors are
generated by a random uniform process method. The circular auto-correlation of random
uniform signals of length 1024 is illustrated in Figure (3.6). By comparing Figure (3.5)
with Figure (3.6), it is obvious that the two Figurers are almost the same. [7]

Figure (3.5) Circular Auto-Correlation of an Augmented BMLS of Length 1024 [7].
Figure (3.6) The Circular Auto-Correlation of Random Uniform Signals of Length 1024 [9].

Figure (3.7) Block Diagram of CCSK Transmitter [7].
Figure (3.8) Block Diagram of CCSK Receiver [7].

Figure (3.9) CCSK Demodulation Block Diagram [7].
3.3 Implementation of CCSK System

The CCSK system is like any communication system that has a transmitter and receiver model. Figure (3.7) and Figure (3.8) show the transmitter and receiver block diagram, respectively. At the transmitter, a suitable binary base vector, $s_o$, is selected to generate the binary transmitted signal, $s_m$, by taking cyclic shift of the base vector ($s_o$). The $m$'th circular shift version of $s_o$ is given by

$$s_m(n) = s_o(n-m) \quad n, m = 0, 1, 2, 3 \ldots M-1$$

(3.7)

where

$M$ is the sequence length.

The following example will show how Equation (3.7) can be applied.

Base vector $s_o$ is $0 \ 0 \ 1 \ 0 \ 0 \ 0 \ 0 \ 0$ (M=8)

Data is represented in binary $(1 \ 0 \ 0)_2$

Data is represented in decimal $(4)_{10}$

Then, the transmitted sequence $s_m$ is $0 \ 0 \ 0 \ 0 \ 0 \ 1 \ 0$ that representing data $(1 \ 0 \ 0)_2$ with length of $M=8$.

At the receiver, the receive signal $(r)$ is match filtered at base band. Because $M$-ary signal $\{s_i\}$ is all possible circular shifts of base vector $s_o$, the output of the CCSK demodulation $U$ is defined as cross-correlation of the received signal $r$ and the base vector $s_o$. Therefore, the highest value of this cross-correlation is taken as the decision [7].
3.3.1 Properties of CCSK

The orthogonality of the M-ary signal \( \{S_i\} \) enhanced the performance of the CCSK waveform. That is,

\[
\sum_{n=0}^{M-1} S_i^*(n) S_j(n) = E_s \quad \text{for } i=j
\]

\[
= 0 \quad \text{for } i \neq j \quad (3.8)
\]

for all values of \( i,j \).

where,

\( E_s \) is the energy per symbol and (*) denotes complex conjugation.

The circular auto-correlation of the base vector is denoted by \( R_s(\tau) \), and given by

\[
R_s(\tau) = \sum_{n=0}^{M-1} S_i^*(n) S_j(n+\tau) \quad (3.9)
\]

The orthogonality condition is satisfied for the circular auto-correlation because waveform \( \{S_i\} \) are all circular shifts of base vector \( s_o \) [7]. So,

\[
R_s(\tau) = \sum_{n=0}^{M-1} S_i^*(n) S_j(n+\tau) = E_s \quad \text{for } \tau=0
\]

\[
= 0 \quad \text{for } \tau=1,2,3,4\ldots M-1 \quad (3.10)
\]
By using M-ary signal, the spread spectrum processing gain PG is given by

$$PG = \frac{W_{ss}}{Rb} = \frac{MTc}{bTc} = \frac{M}{b}$$

(3.11)

Since $b = \log_2(M)$,

$$PG = \frac{M}{\log_2(M)}$$

(3.12)

(3.13)

where $W_{ss}$ is the spread spectrum Bandwidth, $M$ is the sequence length, $b$ is sources bit, $R$ is the data rate, and $Tc$ is chip rate.

### 3.4 DSP and FFT

In CCSK receivers, Transform Domain (TD) signal processing is very important because it reduces the complexity of the digital receiver and suppresses the narrow band interference. In CCSK demodulation, the received sequence is cross-correlated with the copy version of the base vector $\mathbf{s}_0$. And since both of the sequences have length $M$, the cross-correlation operation yields $M^2$ complex multiplications. In addition, the Rake correlator is a cross-correlation operation of stored base vector $\mathbf{s}_0$ with the received signal. This operation is also yielding $M^2$ complex multiplications. Thus, the total number of the complex multiplications needed is $(2M^2)$, which is a huge number.

However, by using FFT and IFFT algorithms, the total number will be reduced to $(2M(1+\log_2(M)))$. For example, assume we have sequence of length $(M=1024)$, then we have 11264 complex multiplications by using FFT and IFFT algorithms, whereas by
using the regular method the yield would be 1048566 complex multiplications. Therefore, it is possible for large M-ary CCSK signals to do real time digital demodulation [7].

3.5 TD/CCSK System

Comparing the TD/CCSK transmitter and receiver system with the conventional systems, TD implementation is exists in the CCSK demodulation, as shown in Figure (3.11). In addition, the implementation of the Rake correlation is added to TD/CCSK receiver to combine the components of the multipath and reduce the required DFT operation. Since this research is related to the urban channel, which is a kind of multipath channel, the multipath channel will be assumed in this research. After the received signal passes through the first two orientated stages, it yields base band vector \( \mathbf{r} \), and this vector is expressed as [7]

\[
\mathbf{r}(n) = \sum_{l=0}^{L-1} s_m(l) h_o(n-l) + z(n)
\]  

(3.14)

where \( n=0,1...L+M-1 \), \( h_o(n) \) is the channel impulse response, \( z(n) \) is an Additive White Gaussian Noise (AWGN), and \( L \) is the number of taps (diversity channel) channel model. Thus, the received sequence is transformed to the frequency domain to yield the TD received vector \( \mathbf{R} \). Since the TD received vector \( \mathbf{R} \) and TD base vector \( \mathbf{S}_o \) are both in frequency domain, the TDCCSK demodulation is implemented as multiplication of \( \mathbf{R} \) by the complex conjugate of the base vector \( \mathbf{S}_o \). Also, the output of the Rake Correlator \( \mathbf{C} \) is implemented as multiplication of the output of the TDCCSK demodulation by the
complex conjugate of TD channel impulse response estimate $H_o^*$, so the output is given by

$$C = R S_o^* H_o^*$$  \hspace{1cm} (3.14)

However, there is no need for the Rake correlator if these conditions have been met [7]:

1. The channel is single time-invariant path.
2. The channel has unity attenuation.
3. The receiver has perfect timing synchronization.

Figure (3.10) Block Diagram of TD/CCSK Transmitter with Transmitted Reference Signal [7].
Figure (3.11) Block Diagram of TD/CCSK Receiver with Transmitted Reference Signal [7].
In other words, the TD channel impulse response estimate $H_o$ is all ones ($H_o=1$).

Therefore, the output vector $C$ is

$$C = R \cdot S_o^*$$  \hspace{1cm} (3.15)

After taking the inverse Fast Fourier Transform (IFFT) of the output vector $C$, vector $\mathcal{e}$ in the time domain will be considered. And this output vector can be represented as

$$c(n) = \sum_{k=0}^{M-1} \left( \sum_{p=0}^{M-1} r(n+p+k) s_o^*(p) \right) h_o^*(k)$$  \hspace{1cm} (3.16)

where,

$n=0,1,2,\ldots,m-1$

$r$ is the received signal vector with length $M$, $s_o^*$ is the conjugate of the base vector of length $M$ in time domain, and $h_o^*(k)$ is the conjugate of the impulse response of the channel with length $M$.

Assume the case of one pulse at position $m$ has been sent and the remaining are zeros. In the case of the ideal channel, the output of the IFFT $\mathcal{e}$ should have a strong correlation peak at the position where the pulse has been sent. Thus, the optimum receiver as picking the largest positive real time of $U_i$ can choose the symbol decision.

$$U_i = \text{Re}[c(n) \delta(n-i)]$$  \hspace{1cm} (3.17)

There are several methods to determine the estimate channel impulse response $H_o$. In one method, the reference base vector $s_r$ is added before the shift register to the modulated CCSK base vector $s_m$, and transmits with $s_m$ at the same time (see Figure
(3.10)). The reference vector $g_r$ is not modulated; however, it is a CCSK base vector. Because the receiver knows the reference base vector $g_r$ a priori, a copy version of it is stored at the receiver. Therefore, no decision feedback is required to update the channel estimator.

There are three properties that have been chosen to be the reference base vector $g_r$ properties [7]:

1-Since the reference base vector $g_r$ is a CCSK base vector,

$$R_{sr}(\tau) = E_s \quad \text{for} \quad \tau = 0$$

$$= 0 \quad \text{for} \quad \tau = 1, 2, 3, \ldots, M-1$$

(3.18)

2-The magnitude spectrum of both reference vector $S_r$ and base vector $S_o$ are almost the same.

$$|S_o(k)| \equiv |S_r(k)|$$

(3.19)

where $k = 0, 1, 2 \ldots M-1$.

3-The reference base vector $g_r$ and the base vector $g_o$ are not correlated, so

$$R_{sosr}(\tau) \ll E_s \quad \text{for} \quad \tau = 0, 1, 2, \ldots, < M-1$$

And,

$$R_{sosr}(\tau) = \sum_{n=0}^{M-1} S_o(n) S_r(n+\tau)$$

(3.20)

where, $R_{sosr}(\tau)$ is the circular cross correlation of the $S_o(n)$ and $S_r(n)$.

\footnote{For more details see reference [7].}
Chapter Four

Urban Multipath Channel

4.1 Introduction

Any communication system consists of three parts. The first part is the transmitter, where information is modulated and sent to its destination. The second part is the receiver, where the received signal is demodulated and the transmitted data is recovered. The last part is the transmission medium between the previous two parts (transmitter and receiver). This transmission is called the channel. Therefore, the channel is the access between the sender and its destination, and this access varies as its environment changes. Channels can be classified into two categories: hardwire and softwire. There are several examples of hardwire channels, such as twisted pair telephone lines, coaxial cables, wave-guides, and fiber optic cables. Examples of softwire channels include air, vacuum, and seawater [5].

When there is more than one path between the sender and the receiver and when each path has its own amplitude and phase, interference between these signals takes place. So the arrived signal would confuse the receiver, and the main signal would be noisy. Thus the transmitted data would not be recovered in this way. The multiplicity of
paths is due to the object reflection or refraction. These can be natural, such as the atmosphere, mountains, trees, etc. or man-made, such as buildings, bridges, factories, etc. Moreover, the noise causes another problem in the channel. The channel noise is what causes the attenuation in the received signal. In term of sources, noise by itself is divided into two parts: natural sources (ionosphere, lighting, and the radiation of the sun, etc) and man-made (high-voltage transmission lines, automobile ignitions, electronic advice noise, etc).

4.2 Fading Multipath Channel

In a wireless communication system, even when there is one electromagnetic wave that represents the information, several paths arrive at the receiver. Each one of these may differ from the other by having time-delay and attenuation characteristics. For example, if one pulse has been transmitted via a time-varying multipath channel, the receiver would receive more than one signal --the main signal plus other attenuated and delayed versions of the original signal, as shown in Figure (4.1). In other words, if a radio signal is transmitted over multipath channels, the arrival signal would exhibit large fluctuations in both amplitude and phase. In the case of the line of site signals (LOS), the receiver would receive one direct signal and several reflected signals, with each signal represented in a path. The model of electromagnetic propagation energy plays a very important role in wireless communication. This electromagnetic propagation energy is affected by either scattering, by means of reflection off flat surfaces of buildings or any other man-made or natural obstacles, or by diffraction around objects.
Figure (4.1) Illustration of a Time-Variant Multipath [10].

Therefore, these indirect paths have different signal attentions and timed delays relative to the direct path. At the receiver input, they interfere with the original signal (direct path) and cause what is called multipath interference (multipath). The variation in the amplitude of the received signal is called fading.

Consequently, the arrival signal consists of the LOS path that represents the main signal and the timed delayed and distorted replicas of the LOS path. Figure (4.2) shows the geometry of the main signal of the multipath channel. Moreover, the multipath channel can be a Rayleigh-fading channel, if the sum of the phases has a uniform power density function (PDF) and the sum of the amplitude components can be modeled as a random process whose envelope has a Rayleigh PDF. The signal output has a Rician
PDF, if one of the amplitude components is much stronger than the others [10]. Generally, it is necessary to study multipath characteristics to enhance the digital communication system. Thus, the next section will focus on this topic.

Figure (4.2) Simplified Model of Multipath Channel [10].

4.3 Characterization of a Fading Multipath Channel

For a long time, ionosphere layers allowed electromagnetic signals to travel for thousand of miles or more by reflection phenomenon (see Figure (4.3)). The communication signals use the high frequency HF band (3-30 MHz) to take advantage of these layers. In terms of LOS path, ground waves are predictable up to 50 miles by
using HF band. Nevertheless, the sky waves can be reflected between ionosphere layers and the ground surface to travel further than the limited distance [11].

In the atmosphere, the layer themselves are formed by the solar electromagnetic radiation over the ultraviolet and x-ray spectra ionizing gases [7]. Therefore, the characteristics of the layer and their existence are dependent upon the changing of day.
and night. To study the effects of the multipath channel on the arrival signals, let us assume the input signal of the multipath channel is given as

$$S(t) = \text{Re}[U(t) e^{j2\pi f_c t}]$$  \hspace{1cm} (4.1)

So, the received signal that passed time-varying multipath channel is given by

$$x(t) = \sum_n \alpha_n(t) S[t-\tau_n(t)]$$  \hspace{1cm} (4.2)

where,

- $\alpha_n(t)$ is the attenuation factor for received signal.
- $\tau_n(t)$ is the propagation for input path.

The substitution for the input signal $S(t)$ into Equation (4.2) yields

$$x(t) = \text{Re}\{\sum_n \alpha_n(t) e^{j2\pi f_c t} U[t-\tau_n(t)] e^{j2\pi f_c t}\}$$  \hspace{1cm} (4.3)

So, the equivalent receive signal is [4]

$$r(t) = \sum_n \alpha_n(t) \exp[-j2\pi f_c \tau_n(t)] U[t-\tau_n(t)]$$  \hspace{1cm} (4.4)

Thus, the time-variant impulse response of the multipath channel described by [4]

$$c(\tau;t) = \sum_n \alpha_n(t) \exp[-j2\pi f_c \tau_n(t)] \delta[t-\tau_n(t)]$$  \hspace{1cm} (4.5)

If the received signal $c(\tau;t)$ is wide sense stationary (WSS), the auto-correlation of the received signal is given by

$$\phi_c(\tau_1,\tau_2;\Delta t) = \frac{1}{2} E[c^* (\tau_1;t) c (\tau_2;t+\Delta t)]$$  \hspace{1cm} (4.6)
At $\Delta t=0$, the auto-correlation function $\phi_c(\tau;0)$ is considered to be the average power output as a function of time-delay $\tau$. Therefore, $\phi_c(\tau)$ is called multipath intensity profile or the delay power spectrum of the channel. Figure (4.4) shows the measured function $\phi_c(\tau)$. The duration of multipath spread of the channel, which is denoted by $T_m$, is the range of the values of $\tau$ over which $\phi_c(\tau)$ is essentially non-zero.

\[ \phi_c(\tau) \]

\[ T_m \]

\[ \tau \]

**Figure (4.4) Multipath Intensity Profile [4].**

The transfer function of the impulse response $c(\tau; t)$ is given by taking the Fourier transform of $c(\tau; t)$,

\[ C(f; t) = \int_{-\infty}^{\infty} [c(\tau; t) \exp(-j2\pi f \tau)] \, d\tau \quad (4.7) \]

where $f$ is the frequency variable.

The transfer function is a WSS because it is the FT of the impulse response $c(\tau; t)$, and the $c(\tau; t)$ is a complex-valued, zero mean Gaussian random process in the $t$ variable. Thus, the auto-correlation of $C(f; t)$ is defined as [4]
According to [4], \( \phi_c(\Delta f) \) equals the Fourier transform of \( \phi_c(\tau) \), where \( \phi_c(\Delta f) \) is called an auto-correlation function in the frequency domain. In addition, \( \Delta f \) is known as a coherent bandwidth and it is given by [4]

\[
(\Delta f)_C \approx \frac{1}{T_m}
\]

Figure (4.5) Frequency Selective Channel Model [7].
So two sinusoids with frequency spectra greater than \((\Delta f)_C\) are influenced differently by the channel. If the bandwidth of the transmitted signal \((W_s)\) is less than the coherent bandwidth \((\Delta f)_C\), the channel is called frequency non-selective. However, for selective frequency channel, the transmitted signal bandwidth \((W_s)\) must be greater than the coherent bandwidth \((\Delta f)_C\). For this kind of channel, it can be modeled as tapped delay with tap spacing \(\frac{1}{W_s}\). Then the total number of taps is

\[
L = Tm \ W_s \tag{4.10}
\]

where, \(L\) is the number of taps in the channel.

In general, the multipath spread can be divided to \(L\) resolvable frequency terms spaced at intervals of \(\frac{1}{W_s}\). Frequency selective channel model is illustrated in Figure (4.5). In this model, the weights \(\{h(n, t)\}\) are assumed to be slowly fading, complex value Gaussian random process, zero mean [7].

### 4.4 Urban Multipath Channel

In each country in the world, there are a number of cities that contain tall buildings surrounded by a large number of other shorter buildings that have a uniform height and form square blocks (see Figure (4.6)). A passageway is located between these buildings, and this passageway has a width less than the buildings. In addition, these blocks are almost parallel, due to the street grid. In general, these are called urban channels. As illustrated in Figure (4.7), the propagation wave in these areas is more complicated and differs from non-urban channels.
Figure (4.6) General Look of an Urban Multipath Channel [13].

Figure (4.7) The Propagation of the Signals in the Presence of Buildings [8].
So the urban channel is due to the environment of modern cities. Since tall buildings, bridges, and transportation are dense in modern cites, the propagation waves are facing difficult paths due to the reflection and refraction off of these obstacles. In other words, the propagation of the signal passes through these objects, between them, or over them where they touch the rooftops of the buildings, which means part of the signal (weak or strong) goes toward the street. Therefore, the signals in these environments are usually accompanied by loss due to attenuation, reflection, and scattering by either an exterior or interior wall [8].

There are many factors that affect the path of the propagation waves: the distance between the base station and the mobile station, the frequency (the band) used to transmit the signal, the height of the base antenna, as well as the mobile antenna, the height of the buildings, the energy of the signal, noise (both natural and man-made), the material used for construction of the buildings, the shape of the rooftops of the buildings, the width of the streets, etc.

4.5 Urban Channel Model

After searching through numerous research and journal papers related to the urban channel, a good model was found that represents the model of the urban channel. This model should have two criteria: wide-band frequency and urban multipath channel. Figure (4.8) shows the model of the propagation in wide-band transmission. In this model, the multipath waves arrive to the receiver under the following conditions [12].
1. There are $N$ arriving signals; each signal has amplitude $A_i(f)$ and path length $L_i$. The amplitude and lengths are independent of each other and are distributed uniformly within a fixed range.

2. $A_0(f)$ is a line of site (LOS) direct signal amplitude, whereas the amplitude of non-direct signals (reflected or refracted) is $A_i(f)$, where $(i \geq 1)$. Also, $L_o$ is denoted as the minimum distance (path length) between the transmitter and receiver or LOS path.

3. The angles ($\theta_i$) of the arrival signals are distributed over $2\pi$ uniformly in a horizontal plane.

4. The amplitude of each signal is constant over $(f_c + \Delta F)$ and centered at a radio frequency of $f_c$.

5. The bandwidth of the receiver is $(2\Delta f)$ and centered at a radio frequency of $f_c$, where $\Delta f < \Delta F$.

The received power spectrum of the received signal as a function of the bandwidth $(2\Delta f)$ is found in [12] as

$$P(f) = 2\Delta f \left( \sum_{i=0}^{N} A_i^2(f) + \sum_{i=0}^{N} \sum_{j=0}^{N} \frac{A_i(f)A_j(f)}{K\Delta L_{ij} \Delta f} \times \left[ \cos(K\Delta L_{ij} f_c) \sin(K\Delta L_{ij} \Delta f) \right] \right)$$

where

$$K = \frac{2\pi}{c}, \text{ where } c \text{ is the velocity of light. And, } \Delta L_{ij} = L_i - L_j.$$

Also, $\sum_{i=0}^{N} \sum_{j=0}^{N}$ means $i \neq j$. 


Figure (4.8) The Model of the Propagation in Wide-Band Transmission [12].

The ratio of power spectrum density (PSD) $a^2$ is denoted as

$$a^2 = \frac{A^2_0(f)}{A^2_1(f)}$$

(4.12)

where $a$ is given by

$$a = 20 \text{ LOG}(a')$$

(4.13)
In Equation (4.11), the first term represents the LOS signal and the remaining represent multipath signals. Figure (4.9) illustrates the block diagram of this model, where $s_i(t)$ is representing the main signal (LOS) and the non-direct signals. Also, $\tau_i$ is representing the time delay where $i=0:N$; $N$ is the total number of signals which arrived.

Figure (4.9) The Block Diagram of the Urban Channel Model.
Chapter Five

Theoretical Performance of TD/CCSK Rake System

5.1 Generalized Theoretical Performance

The theoretical performance of CCSK and TD/CCSK Rake systems has been investigated by Smallcomb (1992) [7]. The comparative study between the theoretical performance of CCSK and TD/CCSK systems and the conventional Rake system will be studied in this section. There are two assumptions in this study: first, the system is an M-ary orthogonal signaling system; second the channel is a slow-fading Rayleigh channel. Basically, the theoretical performance of any communication system is measured in $P_{BE}$ or in probability of bit error. And $P_{BE}$ is a function of both received bit energy $E_b$ and average noise energy $N_0$, where $(P_{BE} = \frac{E_b}{N_0})$. In this study, the theoretical derivation for the probability of bit error will be divided into three parts: the derivation of a consistent definition of the signal to noise ratio ($E_b/N_0$), the decision variable $\{U_i\}$ probability Density Functions (PDFs) as a function of signal to noise ratio, and the derivation of bit error probability $(P_{BE})$ as a function of the PDFs.
The baseband transmitted CCSK signal \( s_m(t) \) is transmitted over a frequency selective HF skywave channel. In addition, there are some assumptions that need to be satisfied in this study [7].

1. There are \( L \) diversity channels, where \( L = T_m W_s \).
2. The sequence length \( M \) is much larger than \( L \), where \( L \) is much larger than 1.
3. Each diversity channel is assumed to be slowly fading with Rayleigh-distributed envelope statistics.
4. The fading processes of each channel are assumed to be mutually independent.
5. AWGN is added to the channel with zero mean.
6. For the simplicity, the intersymbol interference is ignored.

Let \( \gamma_s = \frac{E_s}{N_0} \)

So, by the definition,

\[
\gamma_s = \frac{\text{Received Signal Energy}}{\text{Received Noise Energy}} \tag{5.1}
\]

\[
\gamma_s = \frac{\text{Average Received Signal Power} \times \text{Symbol Duration}}{\text{Average Received Noise Power} / \text{Noise Bandwidth}} \tag{5.2}
\]

Where the symbol duration is \( NT_c \), the average signal power and noise power can be calculated from the received signal vector \( \mathbf{r} \), and noise bandwidth is the reciprocal of the sample rate, which is \( 1/T_c \). So, \( \gamma_s \) now is
\[
\gamma_s = \frac{\left( \frac{1}{N} E \left[ \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} s_m(l) h^*(n-l) \right]^2 \right) * NT_c}{\left( \frac{1}{N} E \left[ \sum_{n=0}^{N-1} z(n)^2 \right] \right) / (1/T_c)}
\]  

(5.3)

where \( E[\ ] \) is the expected value. For more simplicity, \( \gamma_s \) becomes

\[
\gamma_s = \frac{NE_s \sum_{l=0}^{L-1} E|h(l)|^2}{\sum_{n=0}^{N-1} E|z(n)|^2}
\]  

(5.4)

Let \( \sigma_z^2 \) and \( \sigma_h^2 \) be denoted as the Gaussian Variance for additive noise of channel and the attenuation of the Rayleigh fading channel. Because both \( |h(l)| \) and \( |z(n)| \) are Rayleigh-distributed, their squares are chi-square-distributed [7]. So,

\[
\gamma_s = \frac{2L_0 \sigma_h^2 E_s}{2\sigma_z^2}
\]  

(5.5)

The power spectrum density (PSD) of the Additive White Gaussian Noise (AWGN) is denoted as \( (N_0) \) and given by

\[
\sigma_z^2 = \frac{N_0}{2}
\]  

(5.6)
So, \( \gamma_s \) after substituting (5.6) in (5.5) becomes

\[
\gamma_s = \frac{2L\sigma_h^2 E_s}{N_o}
\]

(5.7)

The average received bit energy is the received symbol energy divided by the number of source bits (b).

\[
\frac{E_b}{N_o} = \frac{M\gamma_s}{\log_2(M)}
\]

(5.8)

At the receiver, the received signal is demodulated into a set of decision variables to take the decision. There are some assumptions that can be made for all decision variables [7].

1- The decision variables are Gaussian-distributed random variables, and mutually independent.

2- The decision variable has a nonzero mean corresponding to the transmitted symbol.

3- All other decision variables have zero means.

For now, let \( \bar{X} \) and \( \sigma_x^2 \) be the mean and the variance of the correct symbol decision variable, and all the remaining decision variables have zero mean and the same variance
\(\sigma_y^2\). When \(s_0(t)\) is transmitted symbol, the PDFs for the correct and incorrect variables are

\[
f_x(U_0) = \frac{1}{\sqrt{2\pi \sigma_x}} e^{-\frac{(U_0 - \bar{X})^2}{2\sigma_x^2}} \quad (5.9)
\]

\[
f_x(U_y) = \frac{1}{\sqrt{2\pi \sigma_y}} e^{-\frac{(U_y)^2}{2\sigma_y^2}} \quad (5.10)
\]

where \(y=1,2,3\ldots M-1\)

The probability that \(U_0\) exceeds the other entire decision variable is denoted as \(P_c\) and it is the probability that the receiver makes the correct decision. This probability is given by

\[
P_c = \int_{-\infty}^{\infty} P(U_1 < U_0, U_2 < U_0, \ldots, U_{M-1} < U_0 | U_0) dU_0 \quad (5.11)
\]

where the joint probability \((U_y)\), which is denoted as \(P(U_1 < U_0, U_2 < U_0, \ldots, U_{M-1} < U_0 | U_0)\) for \(y=1,2,3\ldots M-1\), are all less than \(U_0\), conditioned on \(U_0\) [5].

The probability that \(U_0\) be greater than one of the other decision variables conditioned on \(U_0\) is given by

\[
P(U_y < U_0 | U_0) = \int_{-\infty}^{U_y} f_y(U_y) dU_y
\]

\[
= \int_{-\infty}^{U_y} \frac{1}{\sqrt{2\pi \sigma_y}} e^{-\frac{U_y^2}{2\sigma_y^2}} dU_y
\]
where

\[ \text{erf}(\varepsilon) = \frac{2}{\sqrt{\pi}} \int_0^\varepsilon e^{-x^2} \, dx \]  

(5.13)

\[ \text{erfc}(\varepsilon) = 1 - \text{erf}(\varepsilon) \]  

(5.14)

The conditional probability can be written as Equation (5.15) since the probabilities for \( U_y \) are independent, identically distributed, and there are \( (M-1) \) of these decision variables.

\[
P(U_1 < U_0, U_2 < U_0 \ldots U_{M-1} < U_0 | U_0) = [1 - \frac{1}{2} \text{erfc}(\frac{U_0}{\sqrt{2}\sigma_y})]^{M-1} \]  

(5.15)

By solving the joint probability \( P_c \), we get for \( M \) symbols

\[
P_c = \frac{1}{\sqrt{2\pi}\sigma_x} \int_{-\infty}^{\infty} \left[ 1 - \frac{1}{2} \text{erfc}\left( \frac{U_0}{\sqrt{2}\sigma_y} \right) \right]^{M-1} e^{-\frac{(U_0 - \bar{X})^2}{2\sigma_x^2}} \, dU_0 \]  

(5.16.a)

\[
= \frac{1}{\sqrt{\pi}} \int_{-\infty}^{\infty} \left[ 1 - \frac{1}{2} \text{erfc}\left( \frac{\sigma_x}{\sigma_y} X + \frac{\bar{X}}{\sqrt{2}\sigma_y} \right) \right]^{M-1} e^{-X^2} \, dx \]  

(5.16.b)

The probability of the symbol error is given by
By applying this Equation into (5.16), we get

$$P_{SE} = 1 - P_c$$  \hspace{1cm} (5.17)

The conversion of symbol error to bit error performance is given by computing the average number of bit errors per symbol. In addition, all symbol errors occur with equal probability ($P_E$) due to the orthogonality of the symbols. Since all symbols have the same probability, the average number of bit errors per symbol ($\bar{B}$) is given by

$$\bar{B} = \frac{P_{SE}}{M - 1} \sum_{i=1}^{b} i \binom{b}{i}$$

$$= \frac{M b P_{SE}}{2(M - 1)}$$  \hspace{1cm} (5.19)

and

$$b = \log_2 (M)$$  \hspace{1cm} (5.20)

where $b$ is the number of bits per M-ary symbol.

The average bit error probability is computed by dividing the average number of bit errors per symbol by the number of bits per symbol. So,
When we substitute Equation (5.18) into (5.21), we get

\[
P_{BE} = \frac{M P_{SE}}{2(M - 1)}
\]  

(5.21)

From Equation (5.22) the average bit error probability is a function of two variables: first, \( \sigma_X/\sigma_Y \) and \( \frac{X}{\sqrt{2}\sigma_Y} \), which will be used as our figure of merit to compare the different signaling schemes, which are a function of \( \gamma_s \). In this study, assume the following:

\[
f_1(\gamma_s) = \frac{X}{\sqrt{2}\sigma_Y}
\]

(5.23)

and,

\[
f_2(\gamma_s) = \frac{\sigma_X}{\sigma_Y}
\]

(5.24)

Consequently, for the TD/CCSK systems, the average bit error probability \( P_{BE} \) as a function of \( \gamma_s \) is then

\[
P_{BE}(f_1(\gamma_s), f_2(\gamma_s)) = \frac{M}{2(M - 1)} \left[ 1 - \frac{1}{\sqrt{\pi}} \int_{-\infty}^{\infty} \left( 1 - \frac{1}{2} \text{erf} \left( \frac{f_1(\gamma_s)X + f_2(\gamma_s)}{\sqrt{2}\sigma_Y} \right) \right)^{M-1} e^{-X^2} dX \right]
\]

(5.25)
5.2 TD/CCSK with Transmitted Reference

5.2.1 Ideal Case

The vector $\textbf{s}_r$ is denoted as the reference signal vector and is transmitted with the modulated signal vector $\textbf{s}_m$ simultaneously over $L$ diversity channel. Moreover, they have the same conditions (i.e., equal energy). So, the baseband sampled received signal vector $\textbf{r}$ is

$$r(n) = \sum_{l=0}^{L-1} \left[ \frac{s_m(l) + s_r(l)}{\sqrt{2}} \right] h^*_o(n-1) + z(n)$$

(5.26)

where $n=0, 1, \ldots, L+M-1$

After one symbol period (M points), the received signal is truncated to let $n$ vary from (0 to M-1). So, the reference signal and the data would be distorted by this truncation. Thus, the loss due to this truncation is called processing loss and is denoted as ($\phi_t$) and is given to be [7]

$$\phi_t = 1 - \frac{L}{2M}$$

(5.27)

when $M \gg L$, the loss is negligible.

At the receiver, the received signal vector $\textbf{r}$, which is in the time domain, is transformed to frequency domain to yield $\textbf{R}$. Then the transformed $\textbf{R}$ is multiplied by both FFT of $\textbf{s}_o$ and $\textbf{s}_r$ to yield the output of the TD/CCSK demodulation ($S^*_o \textbf{R}$ and $S^*_r \textbf{R}$), respectively. Thus, the output of the reference TD/CCSK ($S^*_r \textbf{R}$) is used to estimate the
TD channel $\hat{H}_o$. For the ideal case, the channel impulse response $H_0$ is well known to the receiver, so $\hat{H}_o = H_0$.

When we expand the output of the data TD/CCSK demodulation, we get [7]

$$R(k)S_o^*(k) = \sqrt{\frac{\phi_t}{2}} \left( S_m(k)S_o^*(k) + S_r(k)S_o^*(k) \right) H(k) + z(k)S_o^*(k) \quad (5.28)$$

The second term of Equation (5.28) can be cancelled since $S_o$ and $S_r$ are known to the receiver priori. So, Equation (5.28) becomes

$$R(k)S_o^*(k) = \sqrt{\frac{\phi_t}{2}} \left( S_m(k)S_o^*(k) \right) H(k) + z(k)S_o^*(k) \quad (5.29)$$

According to [7], for ideal transmitted reference case, the decision variable PDFs is found as

$$U_i = \frac{\phi_t E_s}{2} \text{Re} \left[ R_h(i - m) \right] + \sqrt{\frac{\phi_t E_s}{2}} \text{Re} \left[ R_h(i) \right] \quad (5.30)$$

Since

$$\text{Re}[R_h(\tau)] = \begin{cases} G[2L\sigma^2_h,4L\sigma^4_h] & \text{for } \tau = m \\ G[0,2(L - |m - \tau|)\sigma^4_h] & \text{for } m - L < \tau < L = m \\ 0 & \text{else} \end{cases} \quad (5.31)$$

and since
\[ \text{Re} = [R_{\text{m}}(\tau)] = G[0, 2L\sigma_h^2\sigma_x^2] \]  

(5.32)

where \( G[\bar{X}, \sigma_x^2] \) is the probability density function and is defined as

\[ G[\bar{X}, \sigma_x^2] = f_X(x) = \frac{1}{\sqrt{2\pi\sigma_x^2}} e^{-\frac{(x-\bar{X})^2}{2\sigma_x^2}} \]  

(5.33)

Therefore, the statistics for the decision variable becomes

\[ U_i = \begin{cases} 
G\left[ \phi_i E_i L\sigma_h^2, \phi_i^2 E_i^2 L^2\sigma_h^4, \frac{\phi_i E_i N_o L\sigma_h^2}{2} \right] & \text{for } i = m \\
G\left[ 0, \frac{\phi_i^2 E_i^2 L\sigma_h^4 + \phi_i E_i L N_o \sigma_h^2}{2} \right] & \text{else} 
\end{cases} \]  

(3.34)

To find the figures of merit of this scheme, we need to calculate both of these function \( f_1(\gamma_s) \) and \( f_2(\gamma_s) \) where

\[ f_1(\gamma_s) = \frac{\bar{X}}{\sqrt{2\sigma_y}} = \frac{\phi_i E_i L\sigma_h^2}{\sqrt{\phi_i^2 E_i^2 L\sigma_h^4 + \phi_i E_i N_o L\sigma_h^2}} \]

\[ f_2(\gamma_s) = \sqrt{\frac{\phi_i \gamma_s}{\phi_i \gamma_s + 2}} \]  

(5.35)

and
In general, due to sending the transmitted reference, there is a loss called the "processing loss" ($\phi_r$). This loss is given by [7]

$$\phi_r = \frac{1}{2}$$ (5.37)

Therefore, the figure of merit ($f_1(\gamma_s)$) is given by

$$f_1(\gamma_s) = \frac{\sqrt{\phi_r \varphi_r \gamma_s}}{\phi_r \varphi_r \gamma_s + 1}$$ (5.38)

### 5.2.2 Instantaneous Case

In this case, the channel impulse response is the output of the reference TD/CCSK demodulation. However, the only different between this case and the previous one is the channel impulse response is no longer perfect. Thus, CCSK demodulated reference signal is taken as the channel estimator (i.e. $\hat{H}_o = RS^*$). For instantaneous case, the performance calculation for the transmitted reference requires the calculation of the variance of the received signal cross-correlation and the channel impulse estimator.
From Chapter Three, the output of the data and reference TD demodulation for this case is given by

$$C = RS_o^*(RS_o^*)^*$$  \hspace{1cm} (5.39)

Rewrite this equation to be

$$C(k) = R(k)S_o^*(k)[R(k)S_r^*(k)]^*$$

$$C(k) = |R(k)|^2 S_o^*(k)S_r(k)$$  \hspace{1cm} (5.40)

Now it is clear from Equation (5.40) that the output of the data and reference TD demodulation is taking a form of the auto-correlation of the received signal followed by a double CCSK demodulation to recover the data [7]. According to [7], if we assume that we have large L and the channel is consisting of L Rayleigh-fading diversity channels, the received $r$ can be approximated by an independent Gaussian sequence. Then the variance of this received signal is the sum of the variances associated with the average signal and the noise power, and it is given by

$$\sigma_r^2 = \varphi_e E_s \frac{L}{M} \sigma_h^2 + \sigma_z^2$$  \hspace{1cm} (5.41)

The probability density function for the received signal circular auto-correlation is

$$R_e[R, (\tau)] = \begin{cases} G[2M\sigma_r^2, 4M\sigma_r^4] & \text{for } \tau = 0 \\ G[0, 2M\sigma_r^2] & \text{else} \end{cases}$$  \hspace{1cm} (5.42)
By comparing the energy in the $R_r(\tau=0)$ with the other points of $R_r$ where the $(\tau=1:M-1)$, we get the energy at $R_r(0)$ relatively larger.

The noise reduction ($V$) due to the reduction of the potential noise variance on the decision variables can be implemented between the TD received signal auto-correlation operation and the TD/CCSK demodulation in frequency domain in the following formula:

$$V(k) = |R(k)|^2 - \frac{1}{M} \sum_{k=0}^{M-1} |R(k)|^2$$  \hspace{1cm} (5.43)

From the last two equations,

$$R_E[R_r(\tau)] = \begin{cases} 0 & \tau = 0 \\ G[0.2M \sigma_r^4] & \text{else} \end{cases}$$  \hspace{1cm} (5.44)

Therefore, the TD/CCSK demodulation becomes,

$$C(k) = V(k) S_r^*(k) S_r(k)$$  \hspace{1cm} (5.45)

Because of $R_h(\tau-m)$ being with the data impulse, the energy contained on this auto-correlation $R_h(\tau-m)$ is separated temporarily from $R_r(\tau)$ then added back after the CCSK demodulation operations. Thus, the correlation output yields

$$U_i = \text{Re}[C(i)] = \begin{cases} G[\phi_i E_s \log_2 2M \sigma_r^4] & \text{for } i = m \\ G[0.2M \sigma_r^4 - \frac{\phi_i E_s \log_2 2M \sigma_r^4}{2}] & \text{else} \end{cases}$$  \hspace{1cm} (5.46)

Substituting Equation (5.41) into Equation (5.46) to get
Now the figures of merit \( f_1(\gamma_s) \) and \( f_2(\gamma_s) \) can be calculated for instantaneous TD/CCSK scheme and is given by

\[
U_i = \begin{cases} 
G \left[ \frac{\varphi_i E_s L \sigma_h^2}{M} + \sigma_z^2 \right] & \text{for } i = m \\
G \left[ 0.2M \left( \frac{\varphi_i E_s L \sigma_h^2}{M} + \sigma_z^2 \right)^2 - \frac{\varphi_i^2 E_s^2 L \sigma_h^4}{2} \right] & \text{else} 
\end{cases}
\]

(5.47)

Now the figures of merit \( f_1(\gamma_s) \) and \( f_2(\gamma_s) \) can be calculated for instantaneous TD/CCSK scheme and is given by

\[
f_1(\gamma_s) = \sqrt{\frac{(\varphi, \varphi, \gamma_s)^2}{4 - \frac{1}{M}}} \left( \frac{4}{M} \right) \left( \varphi, \varphi, \gamma_s \right)^2 + \frac{4 \varphi, \varphi, \gamma_s + M}{L}
\]

(5.48)

and

\[
f_2(\gamma_s) = \sqrt{1 + \frac{f_1(\gamma_s)^2}{L}}
\]

(5.49)

By comparing this figure \( f_1 \) of merit with the same figure of merit that we got for the ideal case, we notice the following. For ideal case \( f_1 \) is function of \( \sqrt{\gamma_s} \); whereas, for the instantaneous case, \( f_1 \) is function of \( \gamma_s \). Therefore, a significant degradation of the performance is obviously clear. The instantaneous case is worse due to the noisy estimation of the channel [7].
5.3 The Comparison of CCSK Systems

Figure (5.1) shows the comparison of the bit error performance of the following systems: ideal TD/CCSK, ideal TD/CCSK with transmitted reference, and instantaneous TD/CCSK with transmitted reference. This study has been done with the following values [7]: $L=1024$, $M=8192$, $E_b/N_0 = \gamma_s / \log_2 (M)$. From Figure (5.1) we notice the following:

1- The ideal TD/CCSK system has better performance than the ideal BPSK Rake system.

2- Comparing the ideal TD/CCSK with transmitted reference by ideal TD/CCSK system, the degradation of 3 dB in the performance is a result of the addition of the transmitted reference $S_r$ to the modulated signal. The processing loss (3 dB) is due to the half of the power going to the transmitted reference signal $S_r$.

3- Comparing the instantaneous TD/CCSK with transmitted reference by ideal TD/CCSK system with transmitted reference, the degradation of 9 dB in the performance is a result of the instantaneous estimation of noisy channel.

Nevertheless, the technique of the transmitted reference provides the instantaneous estimate of the channel impulse response; thus, even on a highly disturbed HF channel, this technique is still reliable.
Figure (5.1) Bit Error Performance of Basic TD/CCSK Rake System [7].
Chapter Six

Computer Simulations and Results

6.1 MATLAB Simulation

The matrix laboratory called MATLAB, is a software package that has been designed to solve science and engineering problems. This package is very powerful in simulation and plotting function capability. It is available to be used in many computer systems, such as personal computer (PC), SUN, IBM, and UNIX. Therefore, all my computer simulations used to study the performance of the TD/CCSK system have been done using the MATLAB.

6.2 TD/CCSK System Computer Simulation

Since the goal of this research is to verify that the TD/CCSK system is a good solution for the urban channel problems, and since the urban channel is a multipath channel, then we first need to check whether this system is working in this multipath environment. To this end, the following steps were taken.
6.2.1 TD/CCSK in Ideal Channel

First, the code was generated for the transmitter and the receiver of TD/CCSK systems, as they are proposed in Figure (3.10) and Figure (3.11). Then, this code was tested by sending the information via the ideal channel, which means the information that we sent is the same information that received, without any corruption. So, if we sent the data 10101, for example, we would receive the same data 10101. Thus, to check whether TD/CCSK system is working, all that we need to do is send data and compare it with the received signal. If this system is working, they will be the same; otherwise they are not.

For this case, the input data was generated to be zeros at all the positions of this data except one position, which is chosen randomly. For example, if the length of the data is 1024, the position of the pulse may occur at any of the 1024 points. Figure (6.1) shows one of the inputs of the TD/CCSK system in my work. Also, the output of the same system is shown in Figure (6.2). By comparing the input data to the output data, the output has a peak at the position where the input pulse is occupied (located). Even though the output has a high peak in this position and these data were sent via an ideal channel, there is a low amplitude noise occurring in the output. This is because when we de-spread the transmitted signal ($S_m$), the reference signal ($S_r$) is considered as a noise, and vice versa. Therefore, from this comparison, it was found that this code is working for this system.
Figure (6.1) Input Signal with Length of 1024.

Figure (6.2) The Output of Rake Correlator TD/CCSK in Ideal Channel.
6.2.2 TD/CCSK in Non-Noisy Multipath Channel

In this case, no noise was added because we wanted to make sure that this system is working in multipath channel. Noise was added after it worked successfully in this situation. The code of the impulse response of multipath channel \( h(t) \) was generated. Then, it was plugged (applied) to the main program to study the effects of this channel on the TD/CCSK system. Thus by taking the IFFT for the output of the Rake correlator, the original data that has been sent was recovered. This was illustrated in Figure (6.3). By looking at the Figure (6.3), we found the output was almost the same as compared to the ideal channel. In addition, it had one high peak in the position where the input impulse has been located and low amplitude noise due to the same reasons discussed in the last section (6.2.1). Therefore, the TD/CCSK system is working successively in non-noisy multipath channel. Thus, in the next section we need to add the noise to multipath channel to reach the final decision of TD/CCSK system in multipath channel.

6.2.3 TD/CCSK in Noisy Multipath Channel

Comparing this with the previous section, the only --additive white Gaussian noise-- is added to the multipath channel. Although the AWGN is added to the channel, the output of the Rake correlator is still has the peak at the position of the input pulse with the low amplitude of the noise (see Figure (6.4)). This noise is due to the same reasons discussed in the previous section, as well as the addition of the noise to the channel.
Figure (6.3) The Output of Rake Correlator TD/CCSK in Non-Noisy Multipath Channel.

After all these steps are done perfectly, the TD/CCSK system is working successively in a noisy multipath channel. Therefore, to study the performance of this system, we need to plot the $P_e$ curve. This curve is plotted in Figure (6.5), using the following steps:

1- Specifies the position of the input pulse as well as the output's to be $i_x$ and $i_X$, respectively, in the code.
2- We ran the code and compare the received data with the input information by comparing the two positions of the ix with IX.

3- We ran the code with the following value (M=1024 bits, k=100 taps, \((E_b/N_0)_{dB}\) =0:10).

4- Maximum error is 70, \(P_e =\) (number of errors/ number of the iterations).

Figure (6.4) The Output of Rake Correlator TD/CCSK in Noisy Multipath Channel.

5- We run this code for each value of the signal to noise ratio and store these values in a matrix that is denoted as \(P_e\).
Finally, we plot the $P_e$ matrix against the signal to noise ratio to get the bit error curve then study the system.

By looking at the Figure (6.5) and comparing it to Figure (5.1) in the last Chapter, we find them similar to each other. Therefore, this system is working perfectly in the noisy multipath channel, and we can say that TD/CCSK system has very good performance in such environments.

After this step is taken, it gives us the green light to continue our work toward the main goal. Thus, we need remove the multipath channel and replace it with our urban multipath channel that has been discussed in Chapter Four of this research.
Figure (6.5) The Bit Error Performance of TD/CCSK in Multipath Channel
6.3 Power Spectrum Density of the Urban Multipath Channel

As described in Chapter Four, this model has been taken from [12]. This model gives me the PDF of the urban multipath channel. Thus, we get the PDF of the urban multipath channel by using the Equation (4.11) and the following values:

1- The number of arrival signal is N=50.

2- The carrier frequency fc=1.9Ghz.

3- The velocity of light is c=3e8, and k=2*pi/c.

4- The difference distance between the arrival signal and the one just next to it is varying from (0:5) and is denoted by ΔL.

5- The bandwidth of the received signal is denoted as DF, and it equals 90 MHz.

After we run the code of the urban multipath channel, which gives the percentage probability and the PDF of the channel (see Figures (6.6) and (6.7)), we compare them to the PSD and the percentage probability Figures in [12]. To make sure the code is running correctly, and it does so. Since this PDF works fine, we need to apply it to the main program to test the performance of the system. However, in the main code we need the impulse response of the channel, not its PDF. So, by taking the roots of the PDF function, it gives the transfer function of the channel. Then, by taking the inverse fast Fourier transform (IFFT) of the transfer function, it gives us the impulse response of the channel. Figure (6.8) and Figure (6.9) show the transfer function and the impulse response of the channel, respectively.
Figure (6.7) The Percentage Probability of the Urban Multipath Channel.
Figure (6.7) The Power Spectrum Density of the Urban Multipath Channel.
Figure (6.8) The Transfer Function of the Urban Multipath Channel.

Figure (6.9) The Impulse Response of the Urban Multipath Channel.
6.4 TD/CCSK System in Urban Multipath Channel

Since the impulse response of the urban multipath channel was found, and tested, we go back to our TD/CCSK system code and apply the impulse response of the urban multipath channel to the code instead of the multipath channel. By doing so, the TD/CCSK system performance can be studied, and then a decision of whether this system is a good solution of the urban channel problems can be made by looking at the bit error performance curve of this system. Figure (6.10) illustrates the bit error performance of the ideal TD/CCSK system with transmitted reference in the urban multipath channel. Finally, from all these results, we arrive at the decision that the TD/CCSK system in urban multipath channel is an applicable system in such channels. Therefore, the TD/CCSK system is a good solution for the urban multipath channel.
Figure (6.10) The Bit Error Performance of TD/CCSK System in Urban Multipath Channel.
Chapter Seven

Summary and Conclusion

7.1 Summary

This research was done to test the ability of the TD/CCSK system proposed by Smallcomb [7] (1992) to solve the urban multipath problems. In the previous chapter we tried to verify this statement as follows: first, we studied the SS technique, which is the basis of the CCSK system, and we spent one chapter on the CCSK system itself. In addition, transfer domain (TD) implementation has been discussed in the same chapter as well as the advantages of this implementation. Some examples of these advantages are the reduction of or the simplification in the hardware, suppression of the narrow band interference, and the enhancement of the overall system performance. After we finished with the TD/CCSK system studies, we moved to the next step --the urban channel-- and this topic was discussed in Chapter Four. In this chapter, we came up with the model for this channel, which is brought from Reference [12]. Even though this model gives us the power spectrum density, we derived the impulse response of the channel from this source. In Chapter Five, the theoretical performance of the TD/CCSK system compared with our simulation results. The simulation results are discussed in Chapter Six, and it is divided into five cases. The first case was for TD/CCSK system in ideal channel, the second was for non-noisy multipath channel, the third was noisy multipath channel, and
the fourth case was for the urban channel itself. The last case was for the TD/CCSK system in urban multipath channel.

By referring to the previous chapters and, especially, the last chapter and also by comparing between the theoretical performance with the simulation performance, we can conclude that the TD/CCSK system is a good solution for the urban multipath channel. Nevertheless, there are some restrictions that will be discussed in the next section.

7.2 Strengths and Weaknesses

The TD implementation provides the CCSK strength by reducing the complexity of the digital receiver implementation. In addition, this makes it practical for real time digital demodulation with a large M-ary CCSK system. By using the TD/CCSK system, which is a digital system, the high cost of implementation of the Rake correlator for a large number of the diversity channels in the analog is not considered a problem anymore. Therefore, the TD/CCSK system is a reasonable solution for the urban channel because it is a multipath channel.

However, the weakness in this research is in the calculation of the power spectrum density of the channel because it takes a very long time for a large number of the taps or arrival signals. For example, to calculate the PDF of the channel with number of arrival signals more than 100 would take more than 15 minutes. So, to calculate the PDF of the Channel, it will take a long time for a large number of the arrival signals.
7.3 Future Research

Different System – In this research we used the TD/CCSK system for solving the urban multipath channel. There are other systems which can be applied to solve this problem, such as a fiber optic CDMA system, and a local area network system, etc.

Impulse Response – In our research, we used the impulse response of the urban multipath channel found in Reference [15]. In the future, the derivation of the impulse response of the urban multipath channel is chosen to be practicable for real time for a large number of the arrival signals --N > 200 or more.

Another Application of TD/CCSK System – In this research, we found one application of the TD/CCSK system. However, future research should search for another application of this system.
References


Abstract

This research proposes a new application of the transform domain cyclic shift keying (TD/CCSK), which is one method of the spread spectrum. In this thesis, we applied the TD/CCSK system in the urban multipath channel and studied the performance of this system. The CCSK system was introduced, then we proposed the model for the urban multipath channel. Since the urban channel is a multipath channel, the computer simulation in this thesis was done first to the TD/CCSK in multipath channel. The performance of the system was based on the bit error rate for all cases. After we verified that the system was successfully working in the multipath channel, we applied this system to the urban multipath channel. Thus, the performance of this system in the urban multipath channel was studied and compared to the theoretical performance of the same system. By looking at all the simulation results, we can say that the TD/CCSK system is a good solution for the urban multipath channel.