Applying diversity to OFDM

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Applying Diversity to OFDM

A thesis submitted in fulfilment of the requirements for the award of the degree

Doctor of Philosophy

from

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by

Ibrahim Samir Raad
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AND TELECOMMUNICATIONS ENGINEERING
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Abstract

In today’s world, wireless communications has become an essential part of everyday life. An example of this is the exchange and transmission of data in many forms. Multi-user access systems provide a method to allow multiple users to transmit and exchange this type of information concurrently. Due to its orthogonality, Orthogonal Frequency Division Multiplexing (OFDM) has been used in Ultra Wide Band (e.g. MB-OFDM), WLAN (such as IEEE802.11a and IEEE802.11g) and mobile broadband systems (such as 3GPP LTE) as an efficient scheme to achieve the expected outcomes for today’s society needs for communications.

Although OFDM achieves an excellent transmission rate and its application can be seen in everyday life, it still suffered from corruption especially in indoor wireless environments in applications such as Wireless Local Area Networks (WLANS) in business offices, universities and shopping centers as an example. In these types of environments OFDM suffers the greatest in degradation of performance. This degradation is due to multipath and fast frequency channels. Many solutions for this performance degradation have been proposed and the application of different types of diversity has been used.

This thesis proposes three applications of three different types of diversity to improve the OFDM system performance in terms of Bit Error Rate (BER).

Firstly, a new spreading matrix called the Rotation Spreading matrix used to introduce frequency diversity to OFDM is proposed. This new spreading matrix employs the use of a rotation angle to increase the correlation between the transmitted symbols to ensure in an indoor environment the system maintains
an excellent performance. This thesis provides many studies through experiments and simulations of this new spreading matrix against other well known matrices such as the Hadamard and the Rotated Hadamard. This includes the introduction of methods to increase the size of this new matrix to ensure it is scalable to higher order matrices.

Secondly, time diversity is employed through the use of delaying the block symbols of Block Spread OFDM (BSOFDM) to allow each symbol of an OFDM packet to be transmitted across independent channels.

Thirdly, a new scheme called Parallel Concatenated Spreading Matrices OFDM is presented which employs coding gain to improve the overall BER performance of Block Spread OFDM in frequency selective channels.

As a direct result of the solutions and methods proposed in this thesis to improve the OFDM system, 15 international peer reviewed publications have been achieved. Two of these include book chapters.
Statement of Originality

This is to certify that the work described in this thesis is entirely my own, except where due reference is made in the text.

No work in this thesis has been submitted for a degree to any other university or institution.

Signed

Ibrahim Samir Raad
30 November, 2008
Acknowledgments

I would firstly like to thank Associate Professor Xiaojing Huang for all his support and guidance during this period. I believe his technical knowledge is amongst the best. This work could not have been completed (let alone started) without his help. I would like to give a special mention to Darryn Lowe. His help, discussion and advice ensured that the ideas became true. To Professor Salim Bouzerdoum I would like to thank him for encouraging me to apply for the scholarship that helped ensure I continued and finished this thesis.

I would like to thank my parents and brothers for their support and believing in me. Their encouragement really made a difference.

Finally, to my wife Hoda for her support and consistent encouragement to get this thesis finished. Thank you for your patience.

I would like to dedicate this work to Abbass I. Raad.
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<td>BS-OFDMA</td>
<td>Block Spread - OFDMA</td>
</tr>
<tr>
<td>CCSS</td>
<td>Complete Complementary Sets of Sequences</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>CIR</td>
<td>Carrier to Interference Ratio</td>
</tr>
<tr>
<td>CJR</td>
<td>Carrier - to - Jammer Ratio</td>
</tr>
<tr>
<td>CP</td>
<td>Cyclic Prefix</td>
</tr>
<tr>
<td>D-BSOFDM</td>
<td>Delay BSOFD</td>
</tr>
<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
</tr>
<tr>
<td>DPLL</td>
<td>Digital Phase Locked Loops</td>
</tr>
<tr>
<td>DS-CDMA</td>
<td>Direct Sequence - CDMA</td>
</tr>
<tr>
<td>DS-SS</td>
<td>Direct Sequence Spread Spectrum</td>
</tr>
<tr>
<td>FAF</td>
<td>Floor Attenuation Factor</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency Division Multiple Access</td>
</tr>
<tr>
<td>FH</td>
<td>Frequency Hopping</td>
</tr>
<tr>
<td>FO</td>
<td>Frequency Offset</td>
</tr>
<tr>
<td>FP-MAP</td>
<td>Fincke Pohst Maximum a Posteriori Algorithm</td>
</tr>
<tr>
<td>GLCP</td>
<td>Grouped Linear constellation pre-coding</td>
</tr>
<tr>
<td>Hz</td>
<td>Hertz</td>
</tr>
<tr>
<td>ICD</td>
<td>ICI Cancelling Demodulation</td>
</tr>
<tr>
<td>ICI</td>
<td>Inter-carrier Interference</td>
</tr>
<tr>
<td>IFFT</td>
<td>Inverse Fast Fourier Transform</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter-symbol interference</td>
</tr>
<tr>
<td>LDPC</td>
<td>Low Density Parity Check</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
</tr>
<tr>
<td>--------------</td>
<td>--------------------------------------------------</td>
</tr>
<tr>
<td>LSD</td>
<td>List Sphere Decoder</td>
</tr>
<tr>
<td>MAI</td>
<td>Multiple Access Interference</td>
</tr>
<tr>
<td>MC-CDMA</td>
<td>Multicarrier - CDMA</td>
</tr>
<tr>
<td>MLE</td>
<td>Maximum Likelihood Estimation</td>
</tr>
<tr>
<td>MMSE</td>
<td>Minimum Mean Square Error</td>
</tr>
<tr>
<td>M-PSK</td>
<td>M - Phase Shift Keying</td>
</tr>
<tr>
<td>MRC</td>
<td>Maximum Ratio Combining</td>
</tr>
<tr>
<td>OCDMA</td>
<td>Orthogonal CDMA</td>
</tr>
<tr>
<td>OFDMA</td>
<td>Orthogonal Frequency Division Multiple Access</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplex</td>
</tr>
<tr>
<td>PAP</td>
<td>Peak to Average Power</td>
</tr>
<tr>
<td>PCSM-OFDM</td>
<td>Parallel Concatenated Spreading Matrices</td>
</tr>
<tr>
<td>PER</td>
<td>Packet Error Rate</td>
</tr>
<tr>
<td>PHN</td>
<td>Phase Noise</td>
</tr>
<tr>
<td>PN</td>
<td>Pseudo Noise</td>
</tr>
<tr>
<td>P/S</td>
<td>Parallel to Serial</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>QoS</td>
<td>Quality of Service</td>
</tr>
<tr>
<td>SC</td>
<td>Subchannels</td>
</tr>
<tr>
<td>SD</td>
<td>Selection Diversity</td>
</tr>
<tr>
<td>SHO</td>
<td>Soft Handoff</td>
</tr>
<tr>
<td>SFBC</td>
<td>Space Frequency Block Codes</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
</tr>
<tr>
<td>S/P</td>
<td>Serial to Parallel</td>
</tr>
<tr>
<td>TDMA</td>
<td>Time Division Multiple Access</td>
</tr>
<tr>
<td>TDM</td>
<td>Time Division Multiplex</td>
</tr>
<tr>
<td>TEQ</td>
<td>Time Domain Equalizer</td>
</tr>
<tr>
<td>UMTS</td>
<td>Universal Mobile Transmission System</td>
</tr>
<tr>
<td>UWB</td>
<td>Ultra-wideband</td>
</tr>
<tr>
<td>WAF</td>
<td>Wall Attenuation Factor</td>
</tr>
<tr>
<td>WH</td>
<td>Walsh Hadamard</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
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<td>--------------</td>
<td>----------------------------------</td>
</tr>
<tr>
<td>WLAN</td>
<td>Wireless Local Area Networks</td>
</tr>
<tr>
<td>ZF</td>
<td>Zero Forcing</td>
</tr>
</tbody>
</table>
Chapter 1

Introduction

In today’s world, it has become extremely important to continue to develop wireless communications to maintain the continues economic growth. This is only achievable by ensuring that businesses and their customers have the best possible communications available. It is very important to remember that many businesses have invested large amounts of capital into the existing communication systems and as such it is not possible to deploy new systems. Therefore, to achieve better use of existing solutions and make use of the existing bandwidth becomes the priority. A number of wireless solutions for modulating symbols across frequency selective channels exist. One of these solutions is called Orthogonal Frequency Division Multiplexing (OFDM).

OFDM is a method used to implement mutually orthogonal signals and this is done by setting up multiple carriers at a suitable frequency separation and modulating each symbol stream separately [1]. By increasing the number of carriers the data rate per carrier can be reduced for a given transmission.

The symbol streams do not interfere with each other because of the carriers being mutually orthogonal. It is possible to mitigate fading through suitable interleaving and coding. One method of ensuring the signals are independent of each other is to select the frequency separation between each signal in a manner which will achieve orthogonality over a symbol interval. This can be seen in Figure 1.1. A simplified block diagram of an OFDM system is presented in
Figure 1.2.

Figure 1.1 Transmitter for multi-carrier modulation [1].

While OFDM will combat the effect of multipath transmission, other methods need to be utilized to mitigate the effect of fading. One way of achieving this is called Diversity Transmission. Diversity transmission can be used to reduce or remove the effect of fading by the transmitted signal power being split between two or more subchannels that fade independently of each other, then the degradation will most likely not be severe in all subchannels for a given binary digit. Then when all the outputs of these subchannels are recombined in the proper way the performance achieved will be better than the single transmission. There
are a number of ways to achieve this diversity and the main methods include transmission over spatially different times (space diversity), at different paths (time diversity) or with different carrier frequencies (frequency diversity) [8].

Block Spread OFDM (BSOFDM), also known as pre-coded OFDM, has been used to achieve frequency diversity and has shown significant improvement over conventional OFDM in frequency selective channels. This is done by dividing the $N$ subcarriers into $M$ sized blocks and spreading them by multiplying these blocks with spreading codes such as the Hadamard matrix. A block diagram representation of BSOFDM channel for a block length of two is shown in Figure 1.3.

This thesis contributes a number of methods to improve the OFDM system which are listed below.

\subsection{1.1 Publications and Contributions}

The list below is the direct contributions from this PhD thesis. This includes 15 publications in IEEE peer-reviewed conferences and two of these are book chapter contributions.

1. I. Raad, X. Huang and R. Raad, “A New Spreading Matrix for Block


5. I. Raad and X. Huang, “A New Approach to BSOFDM-Parallel Concatenated Spreading Matrices OFDM,” the 7th IEEE International Symposium on Communications and Information Technologies, ISCIT07 2007, Sydney Australia, (16 - 19 October) [13].


7. I. Raad, X. Huang and D. Lowe, “Higher Order Rotation Matrix for Block Spread OFDM,” the 14th International conference on telecommunications (ICT)and 8th International conference on Communications (MICC), Penang Malaysia 2007, (14th -17th May) [15].

8. I. Raad, X. Huang and D. Lowe, “A Study of different Decoders for Block Spread OFDM in UWB channels,” the 14th International conference on
telecommunications (ICT), 8th International conference on Communications (MICC), Penang Malaysia 2007, (14th -17th May) [16].


11. I. Raad, X. Huang and D. Lowe, “Higher Order New Matrix for Block Spread OFDM,” 14th International conference on telecommunications (ICT), 8th International conference on Communications (MICC), Penang, Malaysia 2007 (14th -17th May) [19].


1.2 Descriptions of Chapters

This thesis has 8 chapters and a description of the main chapters is presented below.

Chapter Two and Three present the literature review on the fundamentals of wireless communications and the Block Spread OFDM respectively.

Chapter Four presents a new spreading matrix which can be used with Block Spread OFDM or pre-coded OFDM. This new spreading matrix is called the Rotation Spreading matrix. This makes use of the frequency diversity to improve the BER performance in frequency selective channels. This is studied and compared to other spreading matrices in the same system. Different OFDM systems use different decoders at the receiver. While it is common knowledge that the best of the present decoders is the Maximum Likelihood (ML) Decoder but due to its complexity it is not used in practical systems. Minimum Mean Square Error (MMSE) decoder is a good alternative. A study of different decoders which include ML, MMSE and Zero Forcing (ZF) is carried out while using the new spreading matrix in the BSOFDM or pre-coded OFDM system and presented in this chapter.

In order to ensure that this new spreading matrix is scalable for larger block sizes for BSOFDM, two methods to expand the size of the Rotation Spreading matrix are presented in Chapter Four.

Chapter Five presents Time delayed BSOFDM and this is where the $M$ sized blocks are delayed by a time $\tau$. This exploits time diversity to improve OFDM systems. It was discovered that in certain environments, such as slow fading channels, frequency diversity does not improve the system performance. By employing time diversity this problem is overcome.
Chapter Six presents the Parallel Concatenated Spreading Matrices OFDM (PCSM-OFDM). The simulation and experimental results show that this system outperformed the normal case BSOFDM by greater than 4 dB. Studies are presented in this chapter based on the new system which includes a study when the number of parallel streams are increased.

Finally, the conclusion chapter highlights the main results and contributions of this thesis, followed by the references.
Chapter 2

Literature Review

2.1 Introduction

In this work the improvement to OFDM is explored through different types of schemes and methods. A new spreading matrix to improve the overall BER of the OFDM system through frequency diversity is presented, then followed by time delay to improve the OFDM through time diversity. Finally, coding gain is explored to help improve the overall OFDM system.

This chapter will present fundamental theories in wireless communication, which includes a brief discussion on OFDM and different types of diversity. It would be very appropriate then to begin by discussing a very well known and established theory in wireless communications such as Shannon’s theory.

2.2 Shannon’s Information Theory

This theory was published in one of the most popular papers in 1948 and it is still used in today’s communication theory. This introduced two concepts which the authors in [4] discuss.

Efficient encoding of a source signal and its reliable transmission over a noisy channel are the first concept that this paper discussed.

This source-coding theorem is motivated by two important facts.
1. A common characteristic of information-bearing signals generated by physical sources (e.g. speech signals) is that they contain a certain amount of information that is *redundant*, the transmission of which is wasteful of primary communication resources, namely, transmit power and channel bandwidth.

2. For efficient signal transmission, the redundant information should be removed from the information-bearing signal prior to transmission.

The theorem basically says that the average code-word length for a distortion less source encoding scheme is upper bounded by the entropy if the given source is discrete memory less and is characterized by a certain amount of entropy.

Entropy in information theory is a measure of the average information content per symbol emitted by the source. According to this theorem, entropy represents a fundamental limit on the average number of bits per source symbol necessary to represent a discrete memoryless source, in that the number can be made as small as, but no smaller than, the entropy of the source.

So, the efficiency of a source encoder can be expressed as

\[ \eta = \frac{H(S)}{\bar{L}} \] (2.1)

where \( H(S) \) is the entropy of the source with source alphabet \( S \) and \( \bar{L} \) is the average number of bits per symbol used in the source-encoding process. Entropy can be defined as

\[ H(S) = \sum_{k=0}^{K-1} p_k \log_2 \left( \frac{1}{p_k} \right) \] (2.2)

where \( p_k \) is the probability that a certain symbol \( s_k \) is emitted by the source. With the base of the logarithm in this definition equal to 2, the entropy is measured in bits, a basic unit of information. Which then leads to a discussion about
the measure of the useful information in relation to the useless information (i.e. noise).

### 2.3 Signal-to-Noise Ratio in a Digital Communication System

The probability of error for each of the different schemes can be expressed in terms of the parameter \( \frac{E}{N_0} \) [24]. The unit of energy \( E \) (energy/bit) can be expressed in terms of the signal power, \( S \), and the bit duration \( T \) as

\[
E = ST
\]  

(2.3)

and the parameter \( \frac{E}{N_0} \) can be described as

\[
\frac{E}{N_0} = \frac{ST}{N_0}
\]  

(2.4)

where the data rate \( R \) is equal to \( \frac{1}{T} \), which allows the equation above to be re-written as

\[
\frac{E}{N_0} = \frac{S}{N_0R}
\]  

(2.5)

The equation above can be re-written if the signal bandwidth can be defined as \( B \) Hz as

\[
\frac{E}{N_0} = \frac{SB}{N_0RB} = \left( \frac{S}{N} \right) \left( \frac{B}{R} \right)
\]  

(2.6)

(2.7)

where \( N = N_0B \) and \( R/B \) has units of bps/Hz. It can be seen that \( \frac{E}{N_0} \) and \( \frac{S}{N} \) are linearly related and sometimes are interchangeable.
The quantity that relates these two ratios, energy/bit-to-noise, is the inverse of the bandwidth efficiency, defined in $\text{bps/Hz}$. If a certain modulation format can transmit more $\text{bps/Hz}$ of available bandwidth for the same signal power at the same performance level, the format that provides a higher value of $R/B$ ($\text{bps/Hz}$) will be more bandwidth efficient. The performance of the modulation formats can be compared in terms of the value of $\frac{E}{N_0}$ required to maintain a fixed probability of error. A modulation scheme that requires a lower value of $\frac{E}{N_0}$ to maintain a certain bit error rate has a better power efficiency than a scheme requiring a higher value of $\frac{E}{N_0}$ to maintain the same bit error rate [24].

This is important information to know as it will be the primary method in which the contribution will be judged by. In other words the way in which the performance will be measured, Bit Error Rate (BER) versus Signal-to-Noise ($\text{SNR}$). The less the $\text{SNR}$ used, the lower the BER is, the more efficient the overall system performance will be.

Now it is also important to discuss simple modulation and demodulation since these simple modulation schemes such as Binary Phase Shift Keying (BPSK) and Quadrature Phase shift Keying (QPSK) are used as the basis for further improvement in the OFDM systems.

### 2.4 Simple Modulations Schemes

#### 2.4.1 Binary Phase Shift Keying

In BPSK, the phase is what holds the information. The carrier phase is zero during the transmission of a one, while the transmission of a zero means the carrier phase takes a value of $\pi$. The following expression describes the BPSK modulation,

$$S_{\text{PSK}}(t) = \begin{cases} \sqrt{\frac{2E}{T}} \cos(2\pi f_0 t) & 1 \\ -\sqrt{\frac{2E}{T}} \cos(2\pi f_0 t) & 0 \end{cases}$$  \hspace{1cm} (2.8)
Figure 2.1 BPSK constellation plot.

\[ 0 \leq t \leq T_s. \]

Figure 2.1 depicts the scatter plot for BPSK modulation.

### 2.4.2 Quaternary Phase Shift Keying

BPSK is not suitable for wireless applications due to very poor bandwidth efficiency [24]. BPSK requires a transmission bandwidth of \(2B\) Hz for a data rate of \(B - \text{bits/s}\). So it is important to use schemes which will reduce the bandwidth required for transmission. \(M - ary\) modulation schemes, where each symbol has more than one bit, can be used for this purpose. Quaternary Phase Shift Keying (QPSK) modulation is an example of such scheme.

In QPSK the phase can take any one of the four values \(0, \pi/2, \pi\) or \(3\pi/2\). Where each symbol consists of two bits, an \(in - phase\) component and a \(quadrature\) component. So a pair of bits will correspond to one of the four unique values [24]. The following gives the QPSK signal expression,

\[ S_{QPSK}(t) = \sqrt{\frac{2E}{T_s}} \cos(2\pi f_0 t + \phi_n) \]  \tag{2.9}  

where

\[ 0 \leq t \leq T_s \]
and the phase $\phi_n$ can take any one of the four phase values depending on the bit pairs. The symbol duration, $T_s$ is $2T$. The phase constellation associated with QPSK is shown in Figure 2.2.

The constellation can be shifted by $\pi/4$ degrees, resulting in phases of $\pi/4$, $3\pi/4$, $5\pi/4$ and $7\pi/4$. The waveform resulting in this shifted version of QPSK can be expressed as

$$S_{QPSK}(t) = \sqrt{2E/T_s} \cos[(2\pi f_0 t + \pi/4) + \phi_n]$$

where

$$0 \leq t \leq T_s.$$

(2.10)

### 2.5 Types of Diversity

Although diversity is a form of redundancy, it is seen by many as a very useful solution to the problem of multipath fading in wireless communications. In basic terms, diversity is transmission of several replicas of the same information transmitted simultaneously over independent channels [25], [26], [27], [28], [29], [30], [31], [32].
There are three types of diversity that have been studied, discussed and implemented in many different systems and include

1. Frequency diversity.

2. Time (signal-repetition) diversity.

3. Space diversity.

For frequency diversity, carriers are spaced sufficiently apart from each other so the system can provide independently fading versions of the channel which are used to transmit the signal. An example of this is frequency hopping; another is the use of spreading matrices. This thesis contributes to frequency diversity in OFDM by introducing a new spreading matrix called the Rotation Spreading matrix.

By transmitting the signal in different time slots the system can achieve time diversity. The interval between successive time slots is set to be equal or greater than the coherence time of the channel. The performance of the system will degrade if the interval is less than the coherence time of the channel, although the system may still achieve diversity.

To ensure independence for possible fading events, space diversity, which is usually multiple transmit and receive antennas where the space between adjacent antennas chosen, can be utilized by a system. If the correlation is as high as 0.7, the systems will potentially lose a 0.5 dB in performance. This can be seen in the results section for the time delay Block Spread OFDM in Chapter 6 when one compares with the ideal case for correlation of zero (the delay BSOFDM is not space diversity - since it does not use multiple antennas at the transmitter and receiver - but is a case where it shows the correlation in channels and if the correlation is at 0.7 then the system will lose a performance of 0.5).

There are different applications of space diversity which involve combinations of antennas at the transmitter or the receiver.
1. Receive diversity, which involves the use of a single transmit antenna and multiple receive antennas.

2. Transmit diversity, which involves the use of multiple transmit antennas and a single receive antenna.

3. Diversity on both transmit and receive, which combines the use of multiple antennas at both the transmitter and receiver.

Clearly this third form of space diversity includes transmit diversity and receive diversity as special cases. This leads into modes of propagation as these types of diversity requires the designer to know the types of channels that the system is experiencing.

2.6 Modes of Propagation

Any mode of propagation can contribute to the losses witnessed by wireless communications systems. There are three modes [4] used for propagation and they include the following,

1. Free space propagation is where the power decreases as the square of the distance from the transmitter.

2. Reflection is where the received power decreases as the fourth power of distance.

3. Diffraction introduces a constant attenuation that depends on the proportion of the direct path that is blocked. For a terrestrial radio link, the signal may be diffracted a number of times along its path.

2.6.1 Median Path Loss

In any transmission, the received signal is the sum of several versions of the transmitted signal received over different transmission paths [4].
The total electric field can be presented in the following equation,

\[
\hat{E} = \hat{E}_d \sum_{k=1}^{N} L_k e^{j\phi_k}
\]  \hspace{1cm} (2.11)

where \(\hat{E}\) represents the total received electrical field. \(\hat{E}_d\) is the electrical field of an equivalent direct path, \(N\) are different paths between the transmitter and receiver due to different reflections. Finally, the \(L_k\) represents the relative losses for the different paths with the \(\phi_k\) representing the relative phase rotations. If the \(L_k = 1\) and \(\phi_0 = 0\) then this means a direct path exists.

A general propagation model for median path loss is produced with the following form,

\[
\frac{P_R}{P_T} = \frac{\beta}{r^n}
\]  \hspace{1cm} (2.12)

where the path loss exponent \(n\) typically ranging from 2 to 5 depending on the environment and \(P_R\) and \(P_T\) are transmitted and received packets. \(\beta\) represents a loss that is related to frequency and that may also be related to antenna heights and other factors. The right hand side of the same equation can also be written in a logarithmic form with the following format,

\[
L_p = \beta_0(dB) - 10n\log_{10}\left(\frac{r}{r_0}\right)
\]  \hspace{1cm} (2.13)

where \(\beta_0\) represents the measured path loss at the reference distance \(r_0\), which is typically one meter. Table 2.1 presents some sample of path-loss exponents.

**2.6.2 Local Propagation Loss**

The previous section presented a model for predicting the median path loss. But for any particular site, there will be a variation from this median value.
which depends on the local environment and its characteristics. In [4], based on several researches work, it presents a model for local propagation loss which can be modelled as a log-normal distribution. The probability density function can be presented as

\[
f(x_{dB}) = \frac{1}{\sqrt{2\pi}\sigma_{dB}} e^{-\frac{(x_{dB}-\mu_{50}/2\sigma_{dB}^2)^2}{2}}
\]

where the \(\mu_{50}\) is the median value of the path loss in \(dB\) at a specified distance \(r\) from the transmitter and \(x_{dB}\) is the distribution of observed path losses at the distance.

Equation 2.14 is also known as the log-normal model for local shadowing.

### 2.6.3 Indoor Propagation

This topic is of interest to this thesis as the use of OFDM and its’ applications are primarily in the indoor environment. Some examples include the IEEE802.11 a and g modulation for wireless access points.

It has become important to take into account when designing wireless communications systems the propagation characteristics in high density locations such as shopping malls, airports and densely populated cities. Many view this area of study as a growth area due to the increase in demand as populations grow.

Other important developments in this area are the implementations and applications of Local Area Networks with the wireless aspects at universities and

<table>
<thead>
<tr>
<th>Environment</th>
<th>n</th>
</tr>
</thead>
<tbody>
<tr>
<td>Free Space</td>
<td>2</td>
</tr>
<tr>
<td>Flat rural</td>
<td>3</td>
</tr>
<tr>
<td>Rolling rural</td>
<td>3.5</td>
</tr>
<tr>
<td>Suburban, low rise</td>
<td>4</td>
</tr>
<tr>
<td>Dense urban, sky scrappers</td>
<td>4.5</td>
</tr>
</tbody>
</table>

*Table 2.1 Sample path-loss exponent.*
elsewhere to eliminate the cost of wiring. Understanding the effects of indoor propagation on services such as these, so optimal performance is achieved, has become important.

In [4] the authors provide a statistical approach and a simple model for the indoor path loss which can be seen below,

\[ L_p(dB) = \beta(dB) + 10\log_{10}\left(\frac{r}{r_0}\right)^n + \sum_{p=1}^{P} WAF(p) + \sum_{q=1}^{Q} FAF(q) \]  \hspace{1cm} (2.15)

where the distance separating the transmitter is \( r \), \( r_0 \) is the nominal reference distance which is typically 1m. \( n \) is the path-loss exponent. The wall attenuation factor is defined as \( WAF(p) \), the floor attenuation factor is defined as \( FAF(q) \) and finally \( P \) and \( Q \) are the number of walls and floors, respectively, separating the transmitter and the receiver.

### 2.6.4 Local Propagation Effects with Mobile Radio

Mobile communication systems are mainly used in large population areas. This usually means that the antennas are below buildings. This would mean that the transmitted and received signals are scattered and diffracted over and around the buildings. These multiple propagation paths, also known as multi-paths, introduce slow or fast fading channels [4], [33], [34], [35].

That is why it is important to study and discuss solutions to this problem. This thesis sets out with a number of solutions to ensure, through different types of diversity, that the overall system performance is improved. The two points below discuss the two types of fading.

1. Slow Fading is due mostly to the large reflectors and diffracting objects along the transmission path and are distant from the terminal.

2. Fast Fading is the rapid variation of signal levels when the user terminal moves short distances.
The second point is the primary concern of this thesis and the contributions discussed in the proceeding chapters. This thesis’s contributions are an attempt to ensure that Fast Fading and its affects’ are minimized and in some instances removed totally.

2.6.5 Rayleigh Fading

A communication device which can transmit or receive data while only stationary but can be also be moved is called a portable terminal.

Figure 2.3 can be used to illustrate the basic concept of a stationary receiver defined as $I_1$ or $I_2$, which can be a wireless access device using the OFDM modulation to transmit or receive.

If one was to characterize the amplitude distribution of the received signal over a variety of positions, then the model needs to be done with the case in which the transmitted signal reaches a stationary receiver via multiple paths where difference are due to only local reflections.

Equation 2.16 provides the complex phaser of the $N$ signal reflections (also commonly known as rays. In this thesis $H2$ is used to define a two ray channel),
where the electric field strength is represented by $E_n$ of the $n^{th}$ path and $\theta_n$ is the relative phase. A random variable representing the multiplicative effects of the multipath channel is represented by $\tilde{E}$. It has been established that small differences in path length can make large differences in phase. Since the reflections can arrive from any direction, it can be assumed that the relative phases are independent and uniformly distributed over $[0, 2\pi]$.

### 2.6.6 Rician Fading

The Rayleigh distribution assumes that all paths are relatively equal, which means there is no dominant path. This is a popular method being used by researchers to test and experiment with new ideas in the wireless communication field. Even though in reality this is not always true, and in most cases there will be always a dominant path in one way or another. Also, there might be in the received path a direct line of sight from the transmitter to the receiver and this is also true for indoor propagation. Therefore, a different model is required to take into account this direct line of sight and Equation 2.17 gives the complex envelope as

$$\tilde{E} = E_0 \sum_{n=1}^{N} E_n e^{j\theta_n}$$

where $E_0$ is the constant term represents the direct path and the summation represents the collection of reflected paths. This is known as the Rician fading model and is common in the Ultra-wideband (UWB) CM1 channel, which is used to carry out experimental results for the contributions in this thesis.

A key factor in the analysis is the ratio of the power in the direct path to the power in the reflected paths. This is referred to as Rician K factor, which can be defined as [4]
\[ K = \frac{s^2}{\sum_{n=1}^{N} |E_n^2|} \]  

(2.18)

where \( s^2 = |E_0|^2 \). This factor is often expressed in \( dB \).

The amplitude density function in the Rician fading can be expressed as [4]

\[ f_R(r) = \frac{r}{\sigma^2} e^{-r^2/2\sigma^2} I_0(r\sigma^2) \]  

(2.19)

where \( I_0 \) is the modified Bessel function of the zeroth order.

### 2.6.7 Doppler

One of the contribution chapters presents a method which utilizes time diversity to improve the overall system performance of BSOFDM. It relies on the concept known as the Doppler effect. If one was to look at a very simple example of a train whistle which appears to have a different pitch if it is moving away from you or towards you. This is also true of radio waves which have the same phenomenon.

If a receiver is moving towards the source, then the zero crossings of the signal appears to be faster and the received frequency is higher. The opposite effect occurs if the receiver is moving away from the source. The resulting change in frequency is known as the Doppler shift.

Figure 2.4 depicts an illustration of the phenomenon known as Doppler shift. This shows a fixed transmitter and receiver moving at a constant velocity away from the transmitter.

If the complex envelope of the signal emitted by the transmitter is \( Ae^{j2\pi f_0 t} \), then the signal at a point along the \( x \)-axis is given by,

\[ \tilde{r}(t, x) = A(x)e^{j2\pi f_0(t-x/c)} \]  

(2.20)
where the $A(x)$ represents the amplitude as a function of distance and $c$ is the speed of light. Equation 2.20 shows that the signal has a phase rotation that depends on the distance of the signal from the source. If $x$ represents the position of the constant-velocity receiver, then it can be written as,

$$x = x_0 + vt \quad (2.21)$$

where $x_0$ is the initial position of the receiver and $v$ is its velocity away from the source. The signal at the receiver is given below,

$$\tilde{r}(t) = A(x_0 + vt)e^{-j2\pi f_0 x_0/c}e^{j2\pi f_0(1-v/c)t}. \quad (2.22)$$

The received frequency can be derived and can be seen below,

$$f_r = f_0 \left(1 - \frac{v}{c}\right) \quad (2.23)$$

where $f_0$ is the carrier transmission frequency. The Doppler shift is given by,

$$F_D = f_r - f_0 \quad (2.24)$$

$$= -f_0 \frac{v}{c}. \quad (2.25)$$

Figure 2.4 Illustration of Doppler effect [4].
It can be seen that the relationship between the observed Doppler frequency shift and the velocity away from the source is,

\[ \frac{v}{c} = -\frac{f_D}{f_0} \]  

Equation 2.26 describes the case when the direction of terminal motion and the radiation are co-linear. The more general expression for the shift is given as follows,

\[ f_D = -\frac{f_0}{c} c \cos \Psi, \]  

if the terminal motion and the direction of the radiation are at an angle \( \Psi \).

### 2.6.8 UWB Channels

Ultra-wideband (UWB) systems occupy - by definition - a signal spectrum of more than 500 MHz or more than 20% with regards to their centre frequency. The application of such large bandwidths enables communication systems with unique novel properties, like high-precision indoor positioning [36], [37], [38], [39], [40].

Based on the Saleh-Valenzuela model for indoor multipath radio propagation channels, a set of statistically UWB channel models have been produced by the IEEE.802.15.3a task group [7]. Line of Sight (LOS) and Non-Line of Sight (NLOS) cases can be modelled using these standard channel models. The four different scenarios are summarized in Table 2.2. Each are identified from CM1 to CM4. These channels are used in our study of different contributions.

### 2.7 Multiple Access Systems

Since this thesis primarily focuses on OFDM, the following section briefly discusses similar multiple access systems and highlights the advantages and dis-
advantages of the OFDM system. This section begins with the fundamental concepts of multiple access systems.

2.7.1 Narrow Channelized Systems

Carrier frequency defines a large number of relatively narrow radio channels, which make up the total spectrum of a channelized system. The radio channel is made up from a pair of frequencies.

2.7.1.1 Forward and Reverse Channel

The frequency used for transmission from the base to the mobile unit is called the forward channel and the frequency used for the transmission from the mobile unit to the base station is called the reverse channel.

During the call the user is assigned both of the channels/frequencies mentioned above. To keep the interference to a minimum from both the two channels, the pair of frequencies are assigned widely separated frequencies. By confining the transmission of a mobile unit to a narrow bandwidth, this avoids interference to adjacent channels [5].

2.7.2 Auto and Cross Correlation

Auto correlation is a property which is a measure of a similarity between \( f(t) \) and a \( \tau \)-second time shifted replica of it self. The autocorrelation functions is a plot of autocorrelation over all shifts \( (t–) \) of the \( f(t) \) signal [1].
\[ R_a(\tau) = \int_{-\infty}^{\infty} f(t) f(t - \tau) dt. \] (2.28)

This principle is used in this thesis to study how the blocks for the BSOFDM encountered different channels when these blocks are delayed. The whole purpose of delaying the blocks is to ensure that the blocks encountered un-identical channels.

Another property called the Cross Correlation is defined as the correlation between two different signals \( f(t) \) and \( g(t) \) which is defined as

\[ R_c(\tau) = \int_{-\infty}^{\infty} f(t) g(t - \tau) dt. \] (2.29)

By computing the number of agreements (\( A \)) minus the number of disagreements (\( D \)) we can obtain autocorrelation and cross-correlation. This is done by comparing codes bit by bit for every discrete shift \( \tau \) in the field of interest.

### 2.7.3 Spectral Efficiency

Spectral efficiency is the measure of how well the spectrum available is used. This can be achieved by reducing the channel bandwidth with information compression. Using a multiple access scheme is a very important factor in improving or obtaining spectral efficiency. By knowing the modulation and the multiple access spectral efficiencies separately the overall efficiency of a mobile communications system can be estimated \[5\], \[41\].

The spectral efficiency with respect to modulation is defined as

\[
\eta_m = \frac{\text{total number of channels available in the system}}{(\text{Bandwidth})(\text{total coverage})}
\] (2.30)

\[
\eta_m = \frac{B_c \times N}{B_w \times N_c \times A_c}
\]

\[
\eta_m = \frac{1}{B_c \times N \times A_c}
\]
where $\eta_m$ is modulation efficiency ($\text{channels/MHz/km}^2$), $B_w$ is bandwidth of the system (MHz), $B_c$ is channel spacing (MHz), $N_c$ is number of cells in a cluster, $N$ is frequency reuse factor of system and $A_c$ is area covered by a cell ($\text{km}^2$).

Equation 2.30 shows that spectral efficiency of modulation does not depend on the bandwidth of the system, it only depends on the channel spacing, the total coverage area and the frequency reuse factor.

The ratio of the total time frequency domain dedicated for voice transmission to the total time frequency domain available to the system defines the multiple access efficiency. Thus, the multiple access spectral efficiency is a dimensionless number with an upper limit of unity.

### 2.7.4 Frequency Division Multiple Access

In Frequency Division Multiple Access (FDMA) each signal is assigned different frequencies for all users. Guard Bands are used between adjacent bandwidth to minimize interference, also known as cross talk. By reducing the information bit rate and using efficient digital codes a capacity increase is achievable. The improvement available in capacity depends on operation at a reduced S/I ratio [5]. Figure 2.5 depicts the FDMA channel architecture.

### 2.7.5 Time Division Multiple Access

In Time Division Multiple Access (TDMA) the data from each user is conveyed in time intervals called slots, which are made up of a preamble plus information bits addressed to various stations. In turn, several slots make up frames. The preamble is used to provide identification and incidental information which allows synchronization of the slot at the intended receiver. Guard times are used between each user’s transmission to minimize crosstalk/interference between channels which reduces the overall bandwidth efficiency. Figure 2.6 depicts the TDMA/FDD channel architecture. Figure 2.7 depicts the divisions in a TDMA frame and divisions within each slot.
Figure 2.5 FDMA/FDD channel architecture [5].

Figure 2.6 TDMA/FDD channel architecture [5].
2.7.6 CDMA and Extensions

Code Division Multiple Access (CDMA) is a well known multiple access scheme which is utilized in the IS – 95, the North American cellular and satellite systems. CDMA was originally developed for military communications and was based on direct - sequence spread spectrum (DS – SS) which has two interesting properties listed below,

1. The low intercept probability.
2. Robustness to narrow band jammers.

There are two basic CDMA schemes [6] which are based on DS-SS. The first, utilizes orthogonal spreading sequences such as the Walsh - Hadamard (WH) sequences, is known as Orthogonal CDMA (OCDMA). $N$ users can be facilitated in CDMA due to the fact that there are exactly $N$ orthogonal sequences of length $N$ when the channel bandwidth is $N$ times that required by a single user, and the spectral efficiency of this technique is identical to that of TDMA.

The second form of CDMA employs pseudo-noise (PN) sequences with low cross-correlation which is referred to as PN-CDMA in the sequel. This has different properties from OCDMA. PN-CDMA involves multi-user interference (MUI)
which grows linearly with the number of users. Therefore the number of users that can be accommodated is directly related to the performance degradation that one is prepared to accept.

2.8 Orthogonal Frequency Division Multiplexing

It is appropriate to discuss Orthogonal Frequency Division Multiplexing (OFDM) since all the contributions in this thesis are based on and used for OFDM. OFDM is currently used in high speed DSL modems over copper based telephone access lines. Since this thesis is primarily a contribution to the wireless communications then it is important to discuss where this modulation scheme is used in this field. OFDM has been standardized as part of the IEEE802.11a and IEEE802.11g for a high bit rate 54Mbps data transmission over wireless LANs (WLANs) [42], [43], [44], [45], [46], [47], [48], [49], [50], [51], [52].

In today’s world there is an increased need for higher bit rate and higher bandwidth data transmission over radio based communication systems. As the transmission bandwidth increases, frequency selective fading and other signal distortions occur. One of the major signal distortions that does occur in digital transmission is inter-symbol interference.

OFDM, by dividing the signal transmission spectrum into narrow segments and transmitting signals in parallel over each of these segments, mitigates this effect. If the bandwidth of each of these frequency spectrum segments is narrow enough, flat or non-frequency-selective fading will be encountered and the signal transmitted over each segment will be received non-distorted. For this reason OFDM has been widely used in wireless applications. Figure 2.8 depicts the simple procedure of OFDM.

In Figure 2.8 the $R$ is the rate in bps of the transmission of the binary digits. The $B$ is the bandwidth required to transmit these bits defined to be $R(1 + r)$, where $r$ is the Nyquist roll off factor, or the order of R Hz. Also, the parameters $a_k$, $1 \leq k \leq N$, represent the successive bits stored, while the frequencies
transmit $N$ carriers in parallel every $T = \frac{N}{R}$

serial - to - parallel conversion

(a) OFDM transmitter

(b) OFDM spectrum

Figure 2.8 OFDM (a) serial to parallel conversion (b) OFDM spectrum.
\( f_k, 1 \leq k \leq N \), represent the \( N \) carrier frequencies transmitted in parallel. \( N \) is a sequence of these bits stored for an interval \( T_s = \frac{N}{f_s} \). This interval is known as the OFDM symbol interval. Serial-to-parallel is then carried out, with each of the \( N \)-bits stored used to separately modulate a carrier. All \( N \) modulated carrier signals are then transmitted simultaneously over the \( T_s \) long interval. This description provided in [42] is the essence of OFDM.

In Figure 2.8, part b, it depicts the resultant OFDM spectrum. It is important to ensure that the various carrier frequencies are orthogonal to each other, and this is done by spacing the \( \Delta f \) between the carriers equal to \( \frac{1}{T_s} \). So the bandwidth can now be described as follows,

\[
B = N \Delta f
\]  

(2.31)

which means that effectively the \( \Delta f \) is the bandwidth of each of the \( N \) parallel frequency channels.

Flat fading occurs on each of the frequency channels by ensuring that \( N \) is large enough when the transmission bandwidth of each of the signals transmitted in parallel by a factor of \( N \). This over comes any frequency selective fading that might have occurred without this serial to parallel process.

Flat fading occurs if the transmission bandwidth \( B \) and delay spread \( \tau_{av} \) are related by,

\[
\tau_{av} = \frac{1}{2\pi B}.
\]  

(2.32)

Therefore the following condition is met,

\[
\Delta f \leq \frac{1}{2\pi \tau_{av}}.
\]  

(2.33)

In equivalent terms of time when using \( \Delta f = \frac{1}{T_s} \), the following is achieved,
In other words to avoid inter-symbol interference the OFDM symbol interval must be much larger than the delay spread. This is always assumed when presenting experimental results in this thesis.

Although OFDM achieves a good performance during transmission, this is not always the case in an indoor environment. When frequency selective channels occur, such as UWB, the performance degrades when coding is not applied.

There are implementation issues with OFDM when using the parallel transmission of \( N \) orthogonal carriers as depicted in Figure 2.8. This process has been replaced by a discrete Fourier Transform (DFT) calculation to solve these implementation problems. This can be seen in Figure 2.9.

Now \( a(t) \) is sampled at intervals \( \frac{T_s}{N} \) apart, which is at the rate of \( R \) samples per seconds. Therefore, in place of \( a(t) \) there is \( a(n) \), which is the sampled function, where \( t \) is replaced by \( \frac{nT_s}{N} \) and \( n = 0, \ldots, N - 1 \). \( a(n) \) can be written as follows,
\[ a(n) = \sum_{k=0}^{N-1} a_k e^{j2\pi kn/N} \] (2.35)

knowing that \( \Delta f_t = 1 \) and \( n = 0, \ldots, N - 1 \). At the receiver, the reverse process of that described above is carried out on each symbol interval \( T_s \) after the demodulation of the received modulated carrier signal. The DFT is calculated and the \( N \) coefficients \( a_k \) with \( k = 0, \ldots, N - 1 \) are recovered. Parallel to serial conversion is used to generate the desired output bit stream.

OFDM is required to be used with some sort of coding or pre-coding to ensure that the received signal is not distorted. In this thesis a new spreading matrix called the Rotation Spreading matrix is used in a scheme called Block Spread OFDM (BSOFDM), discussed in the next chapter.

### 2.9 Orthogonal Frequency Division Multiple Access

#### 2.9.1 Overview of OFDMA

Although OFDM was primarily introduced approximately 30 years ago, OFDMA is a relatively new idea [2]. OFDMA is presented in 1996 in [53]. Utilizing OFDM as a multiple access scheme, depicted in Figure 2.11, OFDMA is achieved. This is made possible by dividing the total FFT space into a number of subchannels (SC), where the SC is a subset of available subcarriers that are assigned to a user for data exchange. Based on quality of service (QoS) profiles and system loading characteristics users may or can occupy more than one SC. Figure 2.10 depicts two possible scenarios for establishing subcarrier groups. The first scenario, which is less advantageous, places subcarriers in the same frequency range in each subchannel. The second scenario has the subchannels spread out over the total bandwidth. Since a deep narrow-band fade can affect only a fraction of subcarriers in each subchannel, this approach is more advantageous especially in frequency selective fading channels. This leads to
the argument that frequency diversity is a very useful concept to utilize with OFDM.

There are other methods which can enhance OFDMA’s performance and they include random frequency hopping (FH), adaptive frequency hopping, and adaptive modulation. A brief description is given below. All these have advantages and disadvantages and can be compared to frequency diversity.

2.9.2 Random Frequency Hopping

It is known that OFDMA assigns each user a predetermined subset of available subcarriers but the received signal power on some subcarriers may drop down randomly due to interference or channel fading. So the user who experiences this low signal power can end up losing connection to the network or have a very poor quality connection. This problem can be solved by using hopping in-use subcarriers in certain time intervals. To transmit the subcarrier groups

\[ \begin{array}{c}
\text{Pilot} \\
\vdots \\
\text{User A} \quad \text{User B} \quad \text{User Z} \\
\text{Pilot} \\
\vdots \\
\text{User A} \quad \text{User B} \quad \text{User Z}
\end{array} \]
in short time slots and frequency-hop them randomly to ensure that it is very unlikely to assign a low-SNR subcarrier to a particular user during a long period of time. This also improves frequency diversity, because each user utilizes the whole system bandwidth. This might also mean that one or a number of users can potentially end up with low SNR subcarriers.

This has other benefits as well, which include the ability to remove intra-cell interference completely and is done by using orthogonal hopping sequences. A general rule in generation of hopping is when the system has $N$ subcarriers, it is possible to construct $N$ orthogonal hopping patterns. In [2] the authors of this overview paper point to [54] and [55] which give rules for generation of hopping patterns. These are outside the scope of this thesis and will not be discussed further.
2.9.3 Adaptive Modulation

Although each carrier in a multiuser OFDM system has the ability to have a different modulation scheme depending on the channel conditions, most OFDM systems use a fixed modulation scheme over all carriers. Any coherent or differential, phase or amplitude modulation scheme can be used, these include BPSK, QPSK, 8PSK, 16QAM, 64QAM and a number of other schemes. Frequency selective fading can result in large variations in the received power or each carrier which can be as much as 30dB for a channel with no direct signal path. To maximize the overall spectral efficiency some have suggested the use of adaptive modulation where the carrier modulation is matched to the SNR. The carrier modulation must be designed to achieve a reasonable bit error rate (BER) under the worst conditions in systems that use a fixed modulation scheme [56], [57], [58], [59], [60], [61], [62], [63], [64], [65], [66], [67], [68], [69], [70], [71], [72].

This results in most systems using BPSK or QPSK, which give poor spectral efficiency (1 − 2 bits/s/Hz) and provide an excess link margin most of the time. Using adaptive modulation, where the remote stations can use a much higher modulation scheme when the radio channel is good, should be used. Modulation can be increased from 1 bits/s/Hz (BPSK) up to 4 − 6 bits/s/Hz (16QAM-64QAM) as a remote station approaches the base station, which increases the spectral efficiency of the overall system. System capacity can potentially be doubled using adaptive modulation in cellular networks.

The down side to this method is increased overhead, since both the transmitter and receiver need to know what modulation is currently being used, overhead information need to be transferred, causing some limitation to using adaptive modulation. Again, this modulation process needs regular updates as the mobility of the remote station increases, which in turn increases the overhead even further. Finally, accurate knowledge of the radio channel is required for adaptive modulation. Any errors in this knowledge can cause large increases in BER.
2.9.4 Adaptive Frequency Hopping

This scheme is based on the channel conditions and discussed in detail in [56]. Using the characterization of the radio channel, carriers are allocated to each user based on the best SNR for that particular user. The fading pattern is different for each remote station due to the varying location of the user. Since the strongest carriers for each user are different, the allocation for each user has minimal clashes. Research has shown adaptive hopping’s ability to achieve a dramatic improvement in received power (5 – 20dB) in a frequency selective channel. In some cases adaptive hopping virtually eliminates frequency selective fading but when the velocity is too high this scheme returns to normal random hopping.

Again, this improvement comes at an increase of complexity of the transmitter and receiver. Forward error correction and time interleaving is also used to improve the BER.

2.9.5 OFDMA System Performance

Some of the advantages and disadvantages of the OFDMA system are as follows,

- Using FFT and IFFT OFDMA transceivers are implemented easily.
- There is no need for a complex RAKE structures at the receiver due to its ability to combine nearly all multipath elements.
- The frequency selective fading channels are transformed to large number of flat fading subchannels making it easier to equalize.
- The signal is divided into a large number of more or less independent channels, which will provide the flexibility needed for future multimedia services.
- OFDMA signals have a near rectangular shape, especially when a large number of subcarriers are used. Also known as sharp edges in the frequency domain.
The two weaknesses listed in [2] are synchronization and power amplifier design which if designed incorrectly, would have a serious impact on the performance of the entire system.

2.9.6 Implementation Issues

2.9.6.1 Cyclic Prefix

Cyclic Prefix (CP) is used to remove the impact of the multipath channel on the performance of the overall system and this is done by copying a fraction of the OFDM symbol from its tail to its head. This is also known as the guard interval. The CP is required to be longer than the channel delay spread to completely stop inter-symbol interference (ISI) and inter-channel interference (ICI). Since this is subject to the interference caused by the multipath channel it has no usable information, which makes it a redundant part of the transmitted signal and reduces the data throughput. When the spread is so large, using a standard CP would cause a large drop in system throughput. The systems in 4G cellular networks use a shorter than the delay spread to minimize the resulted interference power by means of a time domain equalizer (TEQ).

2.9.6.2 Peak to Average Power (PAP)

Due to the naturally large PAP of the sum of a number of independently modulated sub-carriers, which is what an OFDM signal is composed of, D/A and A/D converters are more complex. There are techniques required for these and come in three categories:

1. Signal distortion methods. E.g. clipping, windowing and peak removal;

2. Using channel coding algorithms which reduce the probability of generating high PAP symbols with error control, and

2.9.6.3 Soft Handoff (SHO)

Identical sets of sub-carriers need to be used simultaneously in neighbouring cells for SHO to be used in OFDMA in the base stations involved.

2.9.6.4 Synchronization

There are three consecutive processes in synchronization in OFDMA, they are the following:

1. Symbol timing acquisition.
2. Frequency offset estimation.
3. Phase tracking.

This process is shown in Figure 2.12. The autocorrelation function of the symbol has a sharp peak at its starting time sample, since the guard interval of each OFDM symbol is repeated at its head and tail. This first step is done in the time domain. The next step is for the Frequency Offset (FO) of every sub-carrier to be estimated and compensated for. A bit error in frequency domain synchronization may result in complete loss of orthogonality. “The synchronizer should start the phase tracking process, to track the FO changes caused by the time varying channel, also minimizing the residual error of the primary estimation” [2]. Using Digital Phase locked Loops (DPLL) for tracking the FO variations is popular.

2.9.6.5 Subcarrier and power allocation

Time (timeslots), bandwidth (subcarriers), signal space size (number of bits carried by each subcarrier) and transmission power are all part of the resources of an OFDMA system. Resource allocation algorithms are widely available and are outside of the scope of this thesis.
Figure 2.12 Simplified structure of an OFDM receiver. Numbered blocks are involved in synchronization process [2].

<table>
<thead>
<tr>
<th></th>
<th>UMTS</th>
<th>IEEE802.16</th>
</tr>
</thead>
<tbody>
<tr>
<td>System bandwidth</td>
<td>100kHz-1.6MHz</td>
<td>6MHz</td>
</tr>
<tr>
<td>Number of subcarriers</td>
<td>240 / 100kHz</td>
<td>2048</td>
</tr>
<tr>
<td>Subcarrier spacing</td>
<td>4.6kHz</td>
<td>3.35kHz</td>
</tr>
<tr>
<td>Subcarriers/Band-unit</td>
<td>24 SB/Bandslot</td>
<td>53 SB/Subchannel</td>
</tr>
<tr>
<td>Modulation time</td>
<td>240µs</td>
<td>298µs</td>
</tr>
<tr>
<td>Guard time</td>
<td>38µs (pre) and 8µs (post-guard)</td>
<td>38µs</td>
</tr>
<tr>
<td>Symbol Time</td>
<td>288µs</td>
<td>340µs</td>
</tr>
<tr>
<td>Resource allocation unit</td>
<td>1 band slot and 1 time slot (1symbol)</td>
<td>1 sub-channel and 1 time slot</td>
</tr>
<tr>
<td>Modulation</td>
<td>QPSK, 8PSK (differential and coherent)</td>
<td>QPSK, 16QAM, 64QAM</td>
</tr>
<tr>
<td>Channel Coding</td>
<td>Convolutional (1/3, 2/3)</td>
<td>Turbo (2/3)</td>
</tr>
<tr>
<td>Max. Data throughput</td>
<td>2Mbps</td>
<td>54Mbps</td>
</tr>
</tbody>
</table>

Table 2.3
OFDMA system parameters in the UMTS and IEEE 802.16 standards [2].

2.9.7 OFDMA Application

There are a number of industry backed applications of OFDMA system including the European standard for 3G cellular mobile communications Universal Mobile Transmission System (UMTS) and a broadband wireless access standard for metropolitan area networks (MAN) called IEEE802.16. Table 2.3 describes the parameters for both the standards mentioned.

2.10 Comparison of OFDMA, TDMA and CDMA

Studies have been carried out to investigate which multiple access systems are more robust in the presence of Narrowband Interference and according to [6]
OFDMA has more potential than CDMA and TDMA. The following are the conclusions derived from their work:

1. The BER is independent in TDMA and PN - CDMA. The BER in TDMA, for a given $CJR_{tot}$ (carrier - to - jammer ratio for the $n^{th}$ user), does not depend on the jammer frequency ($F_J$) or on the (maximum) number N of users.

2. The BER in PN - CDMA is independent of $F_J$, but increases with the ratio $\frac{(M-1)}{N}$ of the number of interfering users and the spreading factor.

3. PN - CDMA is worse than TDMA even in the absence of interfering users due to the different statistics of the jammer term at the input of the decision device.

4. The BER for a given $CJR_{tot}$ is a function of the user index, the maximum number of users N and $F_J$ in OFDMA and OCDMA.

5. In TDMA and PN-CDMA, when $CJR_{tot}$ is smaller than some threshold value, the fraction of users operating at a BER less than $10^{-3}$ is zero (resp. one).

6. The threshold value in PN-CDMA depends on $\frac{(M-1)}{N}$ and is always larger than that in TDMA.

7. In OFDMA and OCDMA, the fraction of users with a BER less than $10^{-3}$ is a gradually increasing function of $CJR_{tot}$.

8. OFDMA and OCDMA systems have 50% users with BER $< 10^{-3}$ at values of $CJR_{tot}$ far below the threshold values at which TDMA and PN-CDMA systems completely break down.

9. Finally, OFDMA performs significantly better than OCDMA.

Figure 2.13 depicts the percentage of users with bit error rate less than $10^{-3}$ versus carrier - to - jammer ratio in dB for OFDMA, TDMA and PN - CDMA.
OFDMA can be seen in Figure 2.13 outperforming the other two multiple access schemes. OFDMA uses less dB to achieve a higher percentage of users that have a BER \( < 10^{-3} \).

### 2.11 Spreading Matrices

This section will discuss existing spreading matrices that are used, or can be used, to introduce frequency diversity in OFDM and specifically the system known as Block Spread OFDM (BSOFDM) or pre-coded OFDM. Some of the popular and known spreading matrices include the Hadamard matrix and the Rotated Hadamard.

#### 2.11.1 Hadamard Matrix

By selecting as codewords the rows of a Hadamard matrix, it is possible to produce Hadamard codes. An \( N \times N \) matrix of 1’s and 0’s is a Hadamard matrix such that each row differs from any other row in exactly \( \frac{N}{2} \) locations. One row contains all zeros with the remainder containing \( \frac{N}{2} \) zeros and \( \frac{N}{2} \) ones [8].

The minimum distance for these codes is \( \frac{N}{2} \). An example for a \( N = 2 \), the Hadamard matrix \( A \) is,

\[
A = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix}. \tag{2.36}
\]
In a BSOFDM system, after the modulated data is multiplied by the Hadamard matrix, a higher order modulation scheme is created which increases the correlation between the transmitted symbols, therefore achieving a better system performance. The new modulation scheme scatter plot is shown in Figure 2.14. The new scheme shows it has increased from the four QPSK symbols to nine, where the modulation at the transmission was carried using QPSK. This is an example of employing frequency diversity to achieve better performance for OFDM.

2.11.2 Rotated Hadamard

The rotated Hadamard matrix is a Hadamard matrix with the rotation described in Equation 2.37 applied.

\[
U = \frac{1}{\sqrt{N}} H_{M \times M} \text{diag}(\exp(j\pi m/C)) \quad (2.37)
\]
where $C$ is the rotation value which the modulation is rotated back on to itself. $H$ is the Hadamard matrix and $M$ is the size of the matrix.

The modulation data is multiplied by $U$ and the rotation takes place producing a higher modulation scheme. This can be seen in Figure 2.15 depicting the modulated data after the rotated Hadamard matrix. The rotated Hadamard is capable of achieving 16QAM. So as can be seen this rotated Hadamard produces a higher order scheme than the traditional Hadamard. This is directly translated into a better BER performance in BSOFDM system of rotated Hadamard over Hadamard. This will be seen in the following contribution chapters.

This leads into the following chapter where these types of spreading matrices can be used in such systems as BSOFDM or pre-coded OFDM.
2.12 Conclusion

This literature review chapter on the fundamentals of the wireless communications field began with presenting the building blocks of a wireless communications system. It discusses some of the affects in the physical layer regarding channel properties in particular concentrating on indoor propagation. OFDM and some other multi-carrier system is discussed and compared. Spreading matrices are also presented which are used for applying frequency diversity to OFDM systems.
Chapter 3

Block Spread Orthogonal Frequency Division Multiplexing

3.1 Introduction

This chapter focuses on Block Spread OFDM (BSOFDM) and discusses work carried out in literature based on the concept of block spreading in a multicarrier system to gain diversity advantage over multipath fading channels. It concludes by presenting similar research areas.

3.1.1 Optimal Binary Spreading for Block OFDM

In [3] a study into an “optimal” block spreading code for OFDM is presented, where the main idea of BSOFDM is to split the full set of subcarriers into smaller blocks and spread the data symbols across these blocks via unitary spreading matrices in order to gain multipath diversity across each block at the receiver. They found that the spreading code was optimal for the quadrature amplitude modulation (QAM), binary phase shift keying (BPSK) and quadrature-phase shift keying (QPSK) modulation. For 8PSK the optimality was not there. This paper points out that any unitary matrix will achieve full diversity, except highly patterned spreading codes such as the Hadamard and DFT matrices. This is also supported in [73].

The output of the receiver’s FFT processor is depicted in Figure 3.1 and de-
where \( y \) is the FFT output. \( q \in A^N \) is the vector of transmitted symbols, each drawn from an alphabet \( A \), \( C \) is a diagonal matrix of complex normal fading coefficients and \( n \) is a zero mean complex normal random vector. Equalization of the received data is done through multiplication by \( C^{-1} \) and then “quantized independently on each subcarrier to form the soft or hard decision \( \hat{q} \), which may be further processed if the data bits are coded” [3]. There is no loss in performance when the detection is performed independently on each carrier due to the noise being independent and identically distributed with fading being diagonal.

In this work they consider the use of block spreading matrices to improve the performance on the fading channel. The block spreading matrices are used to introduce dependence among the subcarriers. \( N \) subcarriers are split into \( N/2 \) for blocks of size 2, then each of the blocks are multiplied by a \( 2 \times 2 \) unitary matrix \( U_2 \). The resulting length 2 output vectors are then interleaved to separate the entries in each block as far as possible across the frequency band so that they
will encounter independent fading channels. The transmitter’s IFFT has the interleaved data passed through it and this data is sent across the frequency selective channel. The data is passed through an FFT processor at the receiver and deinterleaved before using block by block processing. The BSOFDM channel model is shown in Figure 3.1.

In analyzing the BSOFDM performance, the author of [3] derives, after a lengthy calculation, the resulting asymptotic bound for the BSOFDM channel as

\[
P_e = \Sigma_{\epsilon_k \in \epsilon} \frac{3E(k)}{(e_k^* U_2^* AU_2 e_k)(e_k^* U_2^* BU_2 e_k)}
\]  

(3.2)

where \(A = \text{diag}\{1, 0\}\) and \(B = \text{diag}\{0, 1\}\). While \(\epsilon\) is the set of equivalence classes of nearest neighbour error vectors corresponding to the 2-dimensional constellation points. The \(E(k)\) is enumeration and finally \(e_k\) is the unique component for the modulation schemes used. They use QAM, MPSK, QPSK, 8PSK and 32PSK modulation in their search for an optimal spreading code. In which they find that the QAM modulation scheme with their designed “optimal” spreading code is the optimal solution.

The author of this paper states that finding the “optimal” spreading matrix, \(U_2\), is done by minimizing the bound in Equation 3.2 over the set of 2 unitary matrices, \(U(2)\).

They produce an extra gain of 7dB comparing OFDM against BSOFDM with their “optimal” spreading. While with other modulation schemes such as BPSK and 8PSK they compare their “optimal” spreading code and BSOFDM against BSOFDM with rotated Hadamard spreading code, which does not improve significantly. As an example, for BPSK they achieve the same gain, 7dB, as that of the rotated Hadamard code, with 8PSK an extra coding gain of 0.2dB at 17dB SNR. While for 32PSK an extra coding gain of 0.8dB is achieved against the rotated Hadamard code. No comparison for the 16QAM is done, but they compare just the normal OFDM, which shows an extra coding gain of 7dB. So one can say that this “optimal” spreading code has no significant improvement.
on the coding gained over the Rotated Hadamard codes.

The Rotation Spreading matrix (This is different to the Rotated Hadamard matrix) which is a contribution of this thesis introduces improvements for all the modulation schemes.

### 3.1.2 Coded Block OFDM for the Frequency Selective Fading Channel

In [74] the same study mentioned in Section 3.1.1 is carried out except this time the author applies correcting code with a large interleaver depth, like the Low Density Parity Check (LDPC) codes, to spread the information bits across frequency and time. They use the BSOFDM as an inner code together with a powerful error correcting code as an outer code. According to the results presented and discussed in this paper, this process outperforms the standard coded OFDM, as expected, and the coded BSOFDM with rotated DFT.

They incorporate an inner code followed by the constellation mapping and block spreading. At the output of the decoder they can append a hard decision, or a soft decision at the BSOFDM detector (or demapper) by using the MAP detection rule and a soft-decision decoder. This can possibly be coupled with a turbo iterative receiver which passes information between the demapper and the decoder.

The following is the MAP rule which forms soft decisions to pass to the decoder.

\[
\hat{c} = \arg \max_c f_{c|z}(c|z)
\]  

(3.3)

where \( c \in \{0, 1\}^{M\log_2 Q} \) are the coded bits which choose the \( M \) symbols and \( f_{c|z}(c|z) \) is the conditional likelihood function for \( c \) given the channel output \( z \).

For the BSOFDM channel the MAP processor can be written as

\[
\hat{c} \approx \arg \min_c (\|z - \sqrt{\mu}CUq(c)\|^2 - \Sigma \log p(c_k))
\]  

(3.4)

where the prior probabilities \( p(c_k) \) are set to \( \frac{1}{2} \) for the initial demapping and
are given by the extrinsic information output by the decoder for subsequent
super-iterations at the receiver and $q(c)$ is the Q-ary modulation set.

The demapper can also supply extrinsic information of its own to the decoder,
with the (extrinsic) likelihoods for each bit generated via

$$f_k^1 = \sum_{c: c_k = 1} f(z|c) \prod_{l \neq k} p(c_l)$$

$$f_k^0 = \sum_{c: c_k = 0} f(z|c) \prod_{l \neq k} p(c_l).$$

The author of this work suggests using the list sphere decoder (LSD) or the
Fincke-Pohst Maximum a posteriori algorithm (FP-MAP) so the the complexity
of the search at the demapper is reduced. The LSD tries to limit the number
of potential symbols over which the search of Equations 3.4, 3.5 and 3.6.

3.2 Analytical Study of BSOFDM Systems

The following is an analytical study for BSOFDM and its performance using
MMSE equalization.

This work (as an analysis in this thesis which was presented in [21]) shows that
regardless of the unitary spreading matrix used, in these conditions you will
always have the same performance and the block size is the main contributing
factor to the improved system performance.

Referring to the transmitter model shown in Figure 3.2, let $x[i]$ where $i =
0, 1, ..., MN - 1$, denote $MN$ data symbols ($M$ and $N$ are integer powers of 2),
which are modulated from the information data bits after BPSK, QPSK or
any other QAM constellation mapping.

Before pre-coding, the $MN$ data symbols are firstly divided into $N$ groups of
size $M$ with the $n^{th}$ group denoted as a vector where $n = 0, 1, ..., N - 1$ and $(.)^T$
denotes matrix transposition,

$$x_n = (x[nM], x[nM + 1], ..., x[nM + M - 1])^T.$$
This is then expressed as a vector after serial-to-parallel conversion \((S/P)\),

\[
x = \begin{pmatrix} x_0 \\ x_1 \\ \vdots \\ x_{N-1} \end{pmatrix}.
\]  

(3.8)

The pre-coding process is to apply an \(M \times M\) unitary matrix \(U\), which satisfies the property shown in Equation 3.9, where \((.)'\) denotes transposition and complex-conjugation operation. The \(I\) is the identity matrix of order \(M\), to each vector \(x_n\) to produce a precoded vector where each element is a linear combination of the symbols in vector \(x_n\).

\[
UU' = U'U = I.
\]  

(3.9)
Each vector $x_n$ produces a precoded vector where each element is a linear combination of the symbols in vector $x_n$.

The precoded symbols are mapped onto subcarriers equally spaced across the transmitted bandwidth to better exploit frequency diversity. This is equivalent to a block interleaving operation among $N$ precoded vectors $Ux_n$, where $n = 0, 1, ..., N - 1$, and then performing IFFT of length $MN$ on the resulting precoded and interleaved vector $Y$.

A time domain sequence $y[i]$ where $i = 0, 1, ..., MN - 1$ is produced after IFFT and parallel to serial conversion (P/S). Either a cyclic prefix (CP) or zero padding (ZP) of sufficient length (longer than the maximum channel multipath delays in samples) are added to $y[i]$ to form a precoded OFDM symbol. This is done to avoid interference between adjacent precoded OFDM symbols and turn the linear convolution of the transmitted signal with the channel impulse response into a circular one.

The precoded OFDM signal is then transmitted over a frequency selective multipath fading channel and received at the receiver baseband.

By removing the CP or performing an overlap-add operation, $MN - \text{point}$ received precoded OFDM samples $r[i]$ where $i = 0, 1, ..., MN - 1$, will be produced. After FFT and de-interleaving, the discrete-time received signal can be expressed in the frequency domain as

$$R_n = H_n Ux_n + V_n$$  \hspace{1cm} (3.10)

where $n = 0, 1, ..., N - 1$ and

$$R_n = (R[n], R[N + n], ..., R[(M - 1)N + n])^T$$  \hspace{1cm} (3.11)

is a vector of $M$ elements which are decimated from $R[k]$, the $MN - \text{point}$ discrete Fourier transform (DFT) or $r[i]$, by a down-sampling factor $N$. 
\[ H_n = \text{diag}(H[n], H[N + n], ..., H[(M - 1)N + n]) \] (3.12)

is an \( M \times M \) diagonal matrix with diagonal elements decimated from \( H[k] \), the \( MN - \) point DFT of the normalized discrete channel impulse response \( h[i] \), and \( V_n \) is a zero-mean Gaussian noise vector with covariance matrix \( E\{V_nV_n^\prime\} = \sigma_v^2 I \), where \( E\{} \) denotes ensemble average.

To recover the transmitted data vector \( x_n \), equalization and detection must be performed on the received signal \( R_n \). Due to the complexity of Maximum Likelihood Decoder (ML) which is not considered to be practical in many systems, only the Minimum Mean Square Error decoder (MMSE) is considered since it can simply use a one tap equalizer for each subcarrier in the frequency domain. The same author does study ML and other linear decoders, but are out of scope for this discussion and can be included for future studies for the new Rotation Spreading matrix.

The equalization and detection process can be described as follows. Let \( C[k] \) denote the one tap equalizer coefficient to be applied to the received signal \( R[k] \) on the subcarrier \( k \) and the following denotes an \( M \times M \) diagonal matrix,

\[ C_n = \text{diag}(C[n], C[N + n], ..., C[(M - 1)N + n]) \] (3.13)

with diagonal elements \( C[lN + n] \), where \( l = 0, 1, ..., M - 1 \). First, applying \( C_n \) to \( R_n \) produces the equalized precoded data vector \( C_nR_n \). Secondly, using the \( U' \) to remove the precoding yields the decision variable vector \( d_n = U'C_nR_n \). Finally, an estimate of the transmitted data vector \( x_n \) is obtained after hard decision. Repeating the above process for \( n = 0, 1, ..., N - 1 \), all the transmitted data symbols are retrieved.

### 3.2.1 Performance of MMSE Equalization

The first step the post-equalization signal to noise ratio (SNR) is derived as a function of the equalizer coefficients \( C[lN + n] \) for the received signal vector \( R_n \).
The decision variable vector, according to the equalization process described above, can be expressed as

\[
d_n = U' C_n R_n \\
= U' C_n H_n U x_n + U' C_n V_n
\]  
(3.14)

(3.15)

Assume that the data symbols in \(x_n\) are independent with average power \(\sigma_x^2\) so that \(E\{x_n x'_n\} = \sigma_x^2 I\). The covariance matrix of \(d_n\) can be derived as

\[
E\{d_n d'_n\} = \sigma_x^2 U' C_n H_n H'_n C'_n U + \sigma_v^2 U' C_n C'_n U.
\]  
(3.16)

Suppose that we want to decide the \(mth\) data symbol \(x[nM + m]\) in \(x_n\) from the \(mth\) element in \(d_n\). The useful signal component can be found from the first term on the right hand side of Equation 3.14 as,

\[
\sum_{l=0}^{M-1} C[lN + n] H[lN + n] |u_{l,m}|^2 \cdot x[nM + m]
\]  
(3.17)

where \(u_{l,m}\) is an element of \(U\), the spreading matrix at the \(lth\) row and the \(mth\) column, and thus the useful signal power after equalization is as

\[
\left| \sum_{l=0}^{M-1} C[lN + n] H[lN + n] |u_{l,m}|^2 \right|^2 \sigma_x^2 = q_0[n, m]
\]  
(3.18)

The average power of the \(mth\) element in \(d_n\) can also be found from Equation 3.16 as follows,

\[
\sigma_x^2 \sum_{l=0}^{M-1} |C[lN + n] H[lN + n] u_{l,m}|^2 \\
+ \sigma_v^2 \sum_{l=0}^{M-1} |C[lN + n] u_{l,m}|^2 = q_1[n, m].
\]  
(3.19)
Therefore, the output SNR after equalization can be expressed as

\[
\gamma[n, m] = \frac{q_0[n, m]}{q_1[n, m] - q_0[n, m]}.
\] (3.20)

According to the MMSE criterion, \(C_n\) should be designed so that the following,

\[
E\{(d_n - x_n)(d_n - x_n)\} = E\{(Ud_n - Ux_n)(Ud_n - Ux_n)\} = E\{(C_nR_n - Ux_n)(C_nR_n - Ux_n)\}
\] (3.21)

is minimized. Using the orthogonality principle, we have

\[
E\{(C_nR_n - Ux_n)R_n'\} = 0
\] (3.23)

and consequently,

\[
C_n = E\{Ux_nR_n'\}(E\{R_nR_n'\})^{-1}
\] (3.24)

\[
= UE\{x_nx_n'\}U' H_n' (H_n U E\{x_nx_n'\}U' H_n' + E\{V_nV_n'\})^{-1}
\] (3.25)

\[
= H_n' (H_n H_n' + \frac{1}{\gamma_{in}}I)^{-1}
\] (3.26)

where \(\gamma_{in} = \frac{\sigma_i^2}{\sigma_v^2}\) is the input SNR before equalization.

From Equation 3.24, the diagonal element is found to be,

\[
C[lN + n] = \frac{H_n^* [lN + n]}{\|H_n^* [lN + n]\|^2 + \frac{1}{\gamma_{in}}}.
\] (3.27)

Substituting Equation 3.27 into Equation 3.18 and 3.19 and using Equation 3.20, the output SNR after MMSE equalization is finally expressed as,
\[ \gamma[n,m] = \frac{\sum_{l=0}^{M-1} \frac{|H[lN+n]|^2}{|H[lN+n]|^2 + \gamma_{in}}}{1 - \sum_{l=0}^{M-1} \frac{|H[lN+n]|^2}{|H[lN+n]|^2 + \gamma_{in}}} \]  

(3.28)

If we consider a class of Unitary matrices that satisfy \(|u_{l,m}| = \frac{1}{M}\), such as the Hadamard matrix and the Rotated Hadamard matrix, Equation 3.28 can be simplified as,

\[ \gamma[n] = \frac{1}{M} \frac{1}{\sum_{l=0}^{M-1} \frac{1}{|H[lN+n]|^2 + \gamma_{in}}} - 1. \]  

(3.29)

We see that the output SNR is determined by the channel frequency response \(H[k]\), or equivalently, the channel impulse response \(h[i]\). Assuming QPSK modulation for data symbols and making a Gaussian distribution approximation for ISI, the bit error probability of the equalizer for a realization of the channel impulse response can be evaluated as follows,

\[ \frac{1}{N} \sum_{n=0}^{N-1} Q(\sqrt{\gamma[n]}) \]  

(3.30)

where the Q-function is defined as,

\[ Q(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} e^{-\frac{t^2}{2}} dt. \]  

(3.31)

Also, assuming that the channel impulse response has \(L\) independent paths, each of which is modelled as an independent complex Gaussian process. The average BER for such frequency-selective fading channel can be evaluated as,

\[ P_e(M, L) = E_h \left\{ \frac{1}{N} \sum_{n=0}^{N-1} Q(\sqrt{\gamma[n]}) \right\} \]  

(3.32)

where \(E\{\cdot\}\) denotes the ensemble averaging over all possible \(h[i]\). It can be seen that Equation 3.32 is a function of the data group size \(M\) and the multipath length \(L\).
3.2.2 BER Lower Bounds and Application

To show the relationship between the system performance and the group size as well as the relationship between the system performance and the channel diversity order, two sets of BER lower bounds using the MMSE equalization are worked out. The first set represents the best possible performance for a given block size $M$. It is assumed that the channel provides a full multipath diversity, that is $L \gg M$, so that $H[lN+n]$ at different $ls$ become independent complex Gaussian variables with unit variance and are alternatively denoted as $\alpha_1$ for convenience. Then the average BER can be alternatively evaluated as

$$P_e(M, \infty) = E_\alpha \left\{ Q \left( \sqrt{\frac{1}{M} \sum_{l=0}^{M-1} \frac{1}{\gamma_{in}[\alpha_l]^2+1}} - 1 \right) \right\}$$

(3.33)

where $E\{\cdot\}$ denotes the ensemble average over $\alpha_0, \alpha_1, \ldots, \alpha_{M-1}$. In Equation 3.33, $\gamma_{in}$ can be expressed as $2\frac{E_b}{N_0}$ for QPSK, where $E_b$ is the signal energy per bit and the $N_0$ is the noise power spectral density.

The second set of lower bounds indicates the best performance for a given number of channel multipath $L$ (referred to as multipath diversity order) with sufficiently large data block size $M$. Let $M \to \infty$, the average BER can be evaluated as

$$P_e(\infty, L) = E_h \left\{ Q \left( \sqrt{\frac{1}{2\pi} \int_0^{2\pi} \frac{1}{\gamma_{in}[H(e^{j\omega})]^2+1} d\omega - 1} \right) \right\}$$

(3.34)

where $H(e^{j\omega})$ is the fourier transform of $h[i]$.

This section has shown that the performance of the pre-coded OFDM systems using MMSE equalization is determined by the precoding data group size and the diversity order that the multipath channel can provide.

When you compare the two above equations for $P_e(1, \infty) = P_e(\infty, 1)$, which means there is no diversity gain or no pre-coding ($M = 1$), the best performance
a conventional OFDM system can achieve is the same as the performance with
diversity order one ($L = 1$). This is consistent with what is known about
c conventional OFDM.

When $1 < M = L < \infty$, we have $P_e(M, \infty) > P_e(\infty, L)$. This means that
for a given data group size $M$ the pre-coded OFDM system can not achieve
the performance which the system could potentially offer with diversity order
$L = M$. However, as $M$ become larger, the performance gradually approaches
the best performance that the system provides for a given diversity order. This
is studied in later chapters through simulation results using different sized $M$.

From $P_e(\infty, L)$ we can predict what is the best performance the system could
potentially offer once the multipath diversity is given. From $P_e(M, \infty)$ we
can compare system performance under different pre-coding sizes and decide
a proper $M$ subject to some complexity constraint.

### 3.3 Examples of Similar Contributions

In [75] the authors discuss an adaptive modulation scheme to increase the trans-
mission rate by changing the channel modulation scheme according to the esti-
mated channel state information as discussed earlier in this chapter.

Since its implementation depends on the channel environment of the system and
control period by using feedback information, in [75] they present an evaluation
for the effects of various modulation scheme combinations, target BER, Doppler
frequency, and various adaptation intervals as control period on the performance
of adaptive OFDM.

The authors also propose a predicted feedback information scheme which in-
creases the adaptation interval using the predicted power estimation, in order
to reduce the transmission time of feedback information from receiver to trans-
mitter.

Their computer simulation results show that the case with BPSK, QPSK and
$16QAM$ modulation combination at target BER $10^{-2}$ achieves $2Mbit/s$ im-
Block Spread Orthogonal Frequency Division Multiplexing

Improvement over other combination cases in high Doppler frequency. On the other hand, at target BER $10^{-3}$, the case with BPSK, QPSK, $8PSK$ and $16QAM$ modulation combination achieves 3Mbit/s improvement compared to the case of target BER $10^{-2}$. It is also shown that the predicted feedback information scheme effectively reduces the transmission time of feedback information from the receiver to transmitter.

The same problem which was mentioned earlier also applies to a scheme such as this. There is an increase of overhead since this needs to have a feedback system to check the condition of the channel to retransmit the data with a higher order modulation, in the case of QPSK, BPSK etc to 16, 64- $QAM$.

Using Pre-coded OFDM will reduce the time for the transmission, at the same time ensuring that the overall bit performance is improved drastically.

In [76] the authors provide link adaptation techniques, where the modulation, coding rate, and/or other signal transmission parameters are dynamically adapted to the changing channel conditions - which have emerged as powerful tools for increasing the data rate and spectral efficiency of wireless data-centric networks.

While there has been significant progress on understanding the theoretical aspects of time adaptation in LA protocols, new challenges surface when dynamic transmission techniques are employed in broadband wireless networks with multiple signaling dimensions. Those additional dimensions are mainly frequency, especially in multicarrier systems, and space in multiple-antenna systems, particularly multi-array multiple-input multiple-output communication systems.

The authors provide an overview of the challenges and promises of link adaptation in future broadband wireless networks. They suggest guidelines to help in the design of robust, complexity/cost-effective algorithms for these future wireless networks. This would seem as an interesting development, but a bit old, again as above the spreading matrices have a better use in this and using transmission diversity will allow the system to improve the performance.
In [77] the authors investigates different diversity concepts for OFDM systems which include,

1. Space frequency block codes (SFBCs) and
2. Code division multiplexing (CDM).

The performances of both, and of combinations of both, are compared for coded OFDM systems in a Rayleigh fading channel.

The applied SFBCs and CDM have in common that they have no rate loss and thus do not decrease the bandwidth efficiency of the system. While the SFBCs exploit space and frequency diversity, CDM exploits frequency and time diversity in OFDM systems. Additionally, the effects due to diversity gains by channel coding are taken into account in the investigations.

These are interesting papers that could be useful when discussing the time delayed-BSOFDM. This is where we use the time delay of the blocks so it would encounter an uncorrelated channel and therefore increasing the BER performance of the system. This is experimentally studied in frequency selective channels.

In [78] the authors present an improved maximum likelihood estimation (MLE) algorithm for preamble-based OFDM frequency synchronization. This proposed algorithm employs time diversity schemes to reduce the effects of fading channels. Two combining methods include,

1. Selection diversity (SD).

These are investigated and analyzed. Simulation results demonstrate that the proposed algorithm significantly improves the performance of frequency synchronization in multipath fading channels. This is useful in having another discussion of time diversity and using different diversity types.
In [79] the symmetric conjugate property of the inter-carrier interference (ICI) weighting function is investigated. Based on that, the ICI self-cancellation of symmetric data-conjugate method is studied to suppress phase noise (PHN). A new ICI cancelling demodulation (ICD) scheme with frequency diversity combining is proposed.

Under flat fading channel, the carrier-to-interference ratio (CIR) improvement from this proposed method is larger than 22 dB, and can tolerate a phase noise with standard deviation of $7^{\text{deg}}$ when BER is $10^{-3}$.

In [80] exploiting the effect of frequency diversity based on the frequency-domain channel correlation of OFDM systems is presented. They propose an improved space-time block coding with frequency diversity for OFDM systems in frequency selective fading channels. With only two transmit antennas, the proposed scheme can achieve a fourth-order diversity gain at low complexity. Simulation results for the proposed scheme demonstrate that this joint exploitation of spatial diversity and frequency diversity leads to an improvement in BER performance.

In this thesis the same concept is employed except we do this for BSOFDM. We use spatial diversity with the number of transmit antennas plus the frequency diversity with the spreading matrices to improve the BER for OFDM.

In [81] an OFDM system with adaptive frequency diversity and decision combining is presented and analyzed. The system involves the transmission of duplicate data over a subset of the total number of subcarriers available to the entire OFDM system. The decisions from these subsets are optimally combined at the receiver to arrive at an overall decision about the transmitted data.

The optimal combination of these subsets is based on weighting factors, which are adaptively chosen based on fading estimates over each subcarrier. The channel is assumed to be nonselective Rayleigh with additive, white Gaussian noise (AWGN).

Computer simulations are presented for a variety of systems and compared with
the performances of several nonadaptive OFDM systems. It is shown that the adaptive frequency diversity with decision combining in OFDM systems can provide enhanced bit error rate performance at the cost of data throughput of the system. The last point is the weakness of such a system.

In [82] the authors propose adaptive subcarrier selection (ACS), as a technique for improving the error performance of grouped linear constellation pre-coding (GLCP) OFDM in the low-to-medium SNR range (0–20 dB). Recently, GLCP-OFDM has been proposed to achieve frequency diversity (multipath diversity) in OFDM overcoming the problem of channel nulls (deep fades in frequency-domain).

In GLCP-OFDM, symbol pre-coding is performed on subgroups of OFDM subcarriers rather than on the total number of OFDM subcarriers which is large in practice. This reduces the receiver decoding complexity while maintaining the high SNR frequency diversity benefit - though the error performance in the low SNR regime is worse than a non-precoded system.

In contrast, the proposed GLCP-ACS-OFDM improves the error performance in the low-to-medium SNR range without increasing the receiver complexity. Simulation results demonstrate the superior performance of GLCP-ACS-OFDM over GLCP-OFDM for a practical range of SNR as applicable to mobile wireless communications.

This is BSOFDM but called pre-coding. This does pre-coding only on a small number of subcarriers not all. The authors says that this will improve the SNR in mid-range.

### 3.4 Conclusion

This chapter focuses on a particular area of improvement to OFDM called Block Spread OFDM (BSOFDM) or pre-coded OFDM. This chapter also briefly highlights current work with different types of diversity which are related to the contributions of this thesis.
Chapter 4

Rotation Spreading Matrix for Block Spread OFDM

4.1 Introduction

Frequency diversity as discussed in the literature chapter has been realized as a method to improve the bit error rate of an OFDM system especially to negate the effects of wireless communication channels. One particular method employs the spreading matrices for this task. But as discussed in the previous chapter, not all spreading matrices are the real contribution factors to this improvement but rather the size of the blocks in regards to Block Spread OFDM. In particular, spreading matrices which are considered to be of the unitary class.

This chapter introduces a new spreading matrix for BSOFDM called Rotation Spreading matrix which does not belong to the unitary class of matrices. This work has been published in a number of international conferences and where applicable these references will be given. This chapter also presents different studies and applications for this new spreading matrix.

Primarily this new spreading matrix is used in what has been described as Block Spread OFDM (BSOFDM). This is when the full set of subcarriers are divided into smaller blocks and using spreading matrices to spread the data across these blocks so to achieve multipath diversity across each block at the receiver [10], [11].
The block spreading matrices are used to introduce dependence among the subcarriers. $N$ subcarriers are split into $\frac{N}{M}$ of blocks of size $M$, where $M = 2$ is used for this example. Then each of the blocks are multiplied by a $2 \times 2$ spreading matrix $U_2$. The length two output vectors are interleaved using general block interleaving to ensure the symbols are statistically independent so as to encounter independent fading channels. This will ensure in a dispersive frequency selective channel the data is statistically less likely to become corrupted and studies and simulations have shown this to be correct.

The transmitter’s IFFT has the interleaved data passed through it and this data is sent across the frequency selective channel. The data is passed through an FFT processor at the receiver and deinterleaved before using block by block processing.

### 4.2 New Rotation Spreading Matrix for Block Spread OFDM

This new Spreading matrix called the Rotation Spreading matrix is presented in [9]. This new spreading matrix is applicable to systems such as the Block Spread OFDM, where it is used to introduce frequency diversity to achieve a higher gain of BER. The Rotation Spreading matrix is presented in Equation 4.1 below

$$U = \begin{bmatrix} 1 & \tan(\alpha) \\ \tan(\alpha) & -1 \end{bmatrix}$$  \hspace{1cm} (4.1)$$

where the angle $\alpha$ is the rotation angle and is introduced by the user.

The most important aspect for any spreading matrix is that it maintains orthogonality for systems which are used for multi-user communications, especially the OFDM system. So the first test required for such a spreading matrix is the orthogonal test which means that if the Rotation Spreading matrix $U$ is multiplied by it conjugate, then the result is an identity matrix $I$. This is expressed in the
following equation

\[ UU' = U'U = AI \] (4.2)

where \( A \) is the scaling factor and is equal to \( A = 1 + \tan^2(\alpha) \). The method by which the scaling factor is calculated is the Rotation Spreading matrix is multiplied by its conjugate which yields the following

\[ UU' = \begin{bmatrix} 1 + \tan^2(\alpha) & 0 \\ 0 & 1 + \tan^2(\alpha) \end{bmatrix}. \] (4.3)

The equation below then proves that the calculated scaling factor \( A \) is correct

\[ \frac{1}{\sqrt{A}} U' \times \frac{1}{\sqrt{A}} U = I. \] (4.4)

The result of the equation given above when the Rotation Spreading matrix is used is that it confirms that the matrix maintains its orthogonal properties. This leads to the next important test for any matrix used for frequency diversity - does it increase the correlation between the transmitted symbols and if so what is the significance of such an increase?

The answer is simple. The higher the correlation between the transmitted symbols, the better the performance of such a system. This is due to the physical qualities of the wireless channel that was discussed in Chapter 2. There will be weak transmission strength which translates to a loss of information carrying symbols. So what a higher correlation between symbols basically means is there are more copies of the same bit of information and the likelihood of these being lost is minimized. This process is achieved through the multiplication of the blocks \( q \) by the \( M \times M \)-sized matrix and is described as
\[ P = U \times q \]  \hspace{1cm} (4.5)

where \( U \) is the spreading matrix and \( q \) is a vector of \( M \) sized blocks. This translates directly into better overall systems performance in terms of BER versus SNR.

In the Literature Review chapter other popular spreading matrices have been discussed and their results analyzed. One of the most popular and often most used is the Hadamard matrix and its derivative the Rotated Hadamard matrix. This thesis will set out to do a comparative study of the existing matrices with the Rotation Spreading matrix. This is done through experimental simulations.

The first notable advantage that the new Rotation matrix has over the other two is its flexibility in developing different types of constellations for the transmissions.

Depending on the choice of the angle \( \alpha \), different combinations for the modulation schemes are possible. If one was to use the example of QPSK modulation at the transmission, a higher order modulation is developed after the multiplication of the blocks by the spreading matrices. When compared with the other two spreading matrix, which can be seen in the literature review, the combination for the constellation is the same all the time, regardless of the situation of the channel or the system. For the case of the Hadamard matrix it only produced a nine point constellation, which meant that not many copies of the constellation symbols were developed. This directly translated into a very poor performance for the Hadamard spreading matrix in a BSOFDM system.

The following figures depict the varying constellation points which are achievable using the new Rotation Spreading matrix for a block size of \( M = 2 \) and the Rotation Spreading matrix of size \( U_{2 \times 2} \). Figure 4.1 depicts the constellation scatter plot of QPSK which will be used for this study.

Figures 4.2, 4.5, 4.6 and 4.7 depict the different constellation points which are achievable using the new matrix with angles \( \alpha = \frac{\pi}{3}, \frac{\pi}{6}, \frac{\pi}{9} \) and \( \frac{\pi}{4} \) respectively.
Figure 4.1 The QPSK constellation points.

Figure 4.2 The Rotation Spreading matrix for block spread OFDM with rotation $\frac{\pi}{3}$. 
Figure 4.3 The Rotation Spreading matrix for block spread OFDM with rotation $\frac{\pi}{2}$.

Figure 4.4 The Rotation Spreading matrix for block spread OFDM with rotation $\pi$. 
Rotation Spreading Matrix for Block Spread OFDM

Figure 4.5 The Rotation Spreading matrix for block spread OFDM with rotation $\frac{\pi}{6}$.

Figure 4.6 The Rotation Spreading matrix for block spread OFDM with rotation $\frac{\pi}{9}$. 
As can be seen from Figure 4.7, since $\tan(\pi/4) = 1$, the end result is the same as the Hadamard matrix.

A study, through simulation, is carried out in [14] which produced a study of different angles for the Rotation Spreading matrix. It is important to know which angle produces a better result in different environments.

The results of this study can be seen depicted from Figure 4.8 through to Figure 4.12. The channel model used for this study is the UWB described in the literature review chapter. This simulation set out to study the angle which achieves the best result in UWB channel models within the BSOFDM system. In Figure 4.8, using the UWB channel model $CM1$, it can be seen that the angle $\pi/3$ achieves the best performance in terms of BER in a BSOFDM system. This angle achieved a BER gain of approximately over $6dB$ over normal OFDM.

This is also true for packet error rate (PER) for the same angles shown in the Figure 4.8.

When the angle $\alpha = \pi/3$ is compared in the same environment to other angles such as $\pi/9$, $\pi$, $\pi/4$, it also outperforms these.
Figure 4.8 The BER using Rotation Spreading matrix with angles $\frac{\pi}{6}$, $\frac{\pi}{3}$ and $\frac{\pi}{7}$.

Figure 4.9 The Rotation Spreading matrix with angles $\frac{\pi}{6}$, $\frac{\pi}{3}$, $\frac{\pi}{2}$, $\frac{\pi}{7}$, $\frac{\pi}{9}$ and $\frac{\pi}{4}$ using CM1 channel.
It is interesting to note that the angle $\frac{\pi}{4}$ has a similar performance to that of the Hadamard matrix since the $\tan\left(\frac{\pi}{4}\right) = 1$ and substituting this value into the new matrix will result in a Hadamard matrix. The two angles $\frac{\pi}{2}$ and $\pi$ will yield the same result as the QPSK modulation. The reason for this is that the two mentioned angles simply rotate the QPSK modulation back onto itself. This can be seen from the constellation points presented above in Figures 4.3 and 4.4.

For the UWB channel model CM2 shown in Figure 4.10, the angle $\frac{\pi}{7}$ outperforms all the other angles. This achieved a bit error rate gain of approximately 16dB over OFDM.

For both the UWB channel models CM3 and CM4 the angles $\frac{\pi}{3}$ and $\frac{\pi}{7}$ again outperform the other angles shown, which can be seen in Figures 4.11 and 4.12. It can be seen that the Hadamard matrix (also the angle $\frac{\pi}{4}$) are outperformed by a significant BER gain. It can be seen that the angle $\frac{\pi}{6}$ has a similar performance to that of the angles $\frac{\pi}{3}$ and $\frac{\pi}{7}$. The angle $\frac{\pi}{9}$ does not perform as well, although it does out perform the angle $\frac{\pi}{4}$ which represents the Hadamard matrix.

Ultimately the spreading matrix which achieves equal distance between each
The final study that is required to be carried out is a BER versus SNR plot to compare the three spreading matrices in a multipath environment.

As can be seen from the results depicted in Figure 4.13, the Rotation Spreading matrix outperforms the Hadamard and Rotated Hadamard in a BSOFDM system across the UWB channel CM1. The Rotation Spreading matrix achieves a gain of well over 15dB over the Hadamard matrix and over 2.5dB gain over the Rotated Hadamard matrix. These results are achieved at reasonable SNR. The worst performance, as expected, is the OFDM system without any pre-coding or frequency diversity employed. This is followed by the pre-coded OFDM when using the Hadamard matrix. This only achieved a 2dB gain over the uncoded
OFDM system at a very large SNR. The Rotated Hadamard achieved a much better result than the Hadamard in this environment. This achieved approximately 10dB gain over the OFDM system and approximately 8dB gain over the pre-coded OFDM system when using the Hadamard matrix.

As discussed and can be seen from Figure 4.13 the new Rotation Spreading matrix has outperformed other spreading matrices which are employed in the OFDM system for frequency diversity in the UWB channel for direct line of sight at a very close distance - CM1. These experimental results use the optimal decoder maximum likelihood (ML) with the rotation angle $\alpha = \frac{\pi}{3}$ for the new Rotation Spreading matrix. These results use 10000 OFDM packets for transmission with QPSK modulation employed at the transmission.

Table 4.2 summaries the SNR needed to achieve the BER of $10^{-3}$ when comparing OFDM without diversity, diversity with Hadamard, Rotated Hadamard and the Rotation Spreading matrices. As can be seen the Rotation Spreading matrix outperforms the others in CM1 UWB channel.

This type of improvement shown through dB gain is also seen when the new Rotation Spreading matrix faced other types of channels for indoor propagation
Figure 4.13 The new Rotation Spreading matrix shown outperforming Rotated Hadamard and Hadamard matrices in UWB CM1.

Table 4.1
Summary of diversity gain in UWB channel CM1 comparing the Hadamard, Rotated Hadamard and the Rotation Spreading matrix.
Figure 4.14 The new Rotation Spreading matrix shown outperforming Rotated Hadamard and Hadamard matrices in UWB CM2 using the ML decoder.

In Figure 4.14, the new Rotation Spreading matrix is compared with the other two spreading matrices in CM2 UWB channel. This channel is non-line of sight for a very short distance. Again, the results show that the new Rotation Spreading matrix outperforming the others. The same system conditions discussed above are used in this experiment. The Rotation Spreading matrix gained an extra 15dB over the Hadamard matrix. It also achieves a gain of approximately 1.5dB over the Rotated Hadamard. These gains are achieved at reasonable SNR.

Table 4.2 summaries the diversity gain in channel CM2. In Figure 4.15, the new Rotation Spreading matrix again is shown outperforming the other two. This UWB channel (CM3) models longer distances with no line of sight. It is worth while mentioning that the Hadamard matrix shows no improvements over the un-coded OFDM which means that in this environment the Hadamard matrix is just increasing the system complexity with no benefit. The Rotated Hadamard does achieve good result as can be seen. The Rotation Spreading
Table 4.2
Summary of diversity gain in UWB channel CM2 comparing the Hadamard, Rotated Hadamard and the Rotation Spreading matrix.

<table>
<thead>
<tr>
<th>Without Diversity</th>
<th>$BER$</th>
<th>$SNR(dB)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Hadamard</td>
<td>$10^{-3}$</td>
<td>22</td>
</tr>
<tr>
<td>Rotated Hadamard</td>
<td>$10^{-3}$</td>
<td>13</td>
</tr>
<tr>
<td>Rotation Spreading</td>
<td>$10^{-3}$</td>
<td>13</td>
</tr>
<tr>
<td>Rotated Hadamard</td>
<td>$10^{-5}$</td>
<td>17.1</td>
</tr>
<tr>
<td>Rotation Spreading</td>
<td>$10^{-5}$</td>
<td>16.5</td>
</tr>
</tbody>
</table>

**Figure 4.15** The new Rotation Spreading matrix shown outperforming Rotated Hadamard and Hadamard matrices in UWB CM3 using the ML decoder.

matrix does however achieve a gain of approximately $2dB$ over the Rotated Hadamard.

In slow fading channels, it is interesting to note that the Rotated Hadamard achieves the same result as the Hadamard matrix. This would indicate that these type of matrices are only useful in a frequency selective channel such as the UWB channels. The new Rotation Spreading matrix outperforms the two spreading matrices by approximately $2dB$. This indicates that this new spreading matrix is useful in any environment that the system may experience during transmission. This is depicted in Figure 4.18.
Table 4.3
Summary of diversity gain in UWB channel CM3 comparing the Hadamard, Rotated Hadamard and the Rotation Spreading matrix.

<table>
<thead>
<tr>
<th>Method</th>
<th>BER</th>
<th>SNR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Without Diversity</td>
<td>$10^{-3}$</td>
<td>$&gt;&gt; 25$</td>
</tr>
<tr>
<td>Hadamard</td>
<td>$10^{-3}$</td>
<td>$&gt;&gt; 22$</td>
</tr>
<tr>
<td>Rotated Hadamard</td>
<td>$10^{-3}$</td>
<td>14</td>
</tr>
<tr>
<td>Rotation Spreading</td>
<td>$10^{-5}$</td>
<td>14</td>
</tr>
<tr>
<td>Rotated Hadamard</td>
<td>$10^{-5}$</td>
<td>19.5</td>
</tr>
<tr>
<td>Rotation Spreading</td>
<td>$10^{-5}$</td>
<td>18.5</td>
</tr>
</tbody>
</table>

Figure 4.16 The Hadamard matrix versus un-coded OFDM in two ray slow fading channel.
Figure 4.17 depicts the pre-coded OFDM when using the rotated Hadamard matrix. This also shows no improvement. The experimental results used Zero forcing decoder.

The final area of study would be to see what type of decoder can and should be used with this new Rotation Spreading matrix when used for frequency diversity in OFDM.

### 4.3 A Study of Different Decoders for the New Rotation Spreading Matrix for Block Spread OFDM in UWB Channels

In [16] a study of the different decoders at the receiver for BSOFDM is presented. If there are $k$ subcarriers in a BSOFDM system and the channel described as $h(k)$, the received signal is described as $R(k)$ and the equalizer to be used to be described as $g(k)$, then for a one tap equalizer at the receiver the expression can be presented as follows
Rotation Spreading Matrix for Block Spread OFDM

For de-spreading using \( M = 2 \) blocks as an example, the following expression can be used to describe the process at the receiver

\[
Output = U^{-1} \times \begin{bmatrix}
g(k_1) \times R(k_1) \\
g(k_2) \times R(k_2)
\end{bmatrix} = \begin{bmatrix}
\hat{a}_1 \\
\hat{b}_2
\end{bmatrix}
\]

(4.7)

where the \( U^{-1} \) is the inverse of the unitary matrix used at the transmitter for spreading the blocks of \( M \) size and \( \hat{a}_1 \) and \( \hat{b}_2 \) are the de-spread and decoded symbols at the receiver. Then the study of different decoders can be carried out, but first a brief description is given for each decoder.

4.3.1 Maximum Likelihood (ML) Decoder

The Maximum Likelihood (ML) Decoder is known to have the best performance of all the decoders presently available, as it calculates all possible combinations at the receiver before making a decision. This is also known to be the most complex and most system implementations avoid using this decoder due to this

\[
C_k = g(k) \times R(k).
\]

(4.6)
issue. The following is the mathematical expression used for the ML decoder [83], it can be shown that the maximum likelihood criterion for the receiver is

\[ g(k) = \min_z \left| \frac{\hat{H}Z - X}{\sigma} \right|^2 \]  

(4.8)

where \( Z \) is the transmitted data and \( \hat{H} \) is the channel estimate, \( X \) is the received data with the following format,

\[ X = HZ + n \]  

(4.9)

where \( n \) is the Gaussian noise with variance \( \sigma^2 \).

### 4.3.2 Minimum Mean Square Error (MMSE) Decoding

The MMSE is a useful alternative to that of the ML decoder described above as it can achieve good results in terms of Bit Error Rate (BER) at low Signal to Noise Ratio (SNR). This is less complex than the ML. The down side of this decoder is at high SNR, the performance is similar to that of the Zero Forcing (ZF) decoder but unlike the ZF decoder, as seen from Equation 4.12, the interferences \( a \) and \( b \) are not forced to zero and as such the noise is not amplified.

The MMSE can be described as follows

\[ g_k = \frac{h_k^*}{|h_k|^2 + \frac{1}{SNR_c}} \]  

(4.10)

where \( h_k \) is assumed to be the known channel. At very high SNR, the equation above can be shown to be as follows

\[ g_k = \frac{1}{h_k} \]  

(4.11)
which says that it is a ZF decoder. This can be seen in the results at high SNR. So the MMSE is a good solution for BSOOFDM at low SNR because it allows a good compromise between noise and the ISI minimization. It is robust (avoids problems with channel zeros) and widely used in practice.

### 4.3.3 Zero Forcing (ZF) Decoding

The Least Square (LS) or Zero forcing (ZF) decoder is the simplest method used for decoding. The complexity is also the least. It is described in Equation 4.11 and as discussed earlier at high SNR has the same performance as that of the MMSE. With the ZF decoder, the interference is forced to zero, or assumed to be zero, which in turn amplifies the noise and can be seen from the following mathematical expression for BSOOFDM,

$$\hat{a} = xa + yb + n$$  \hspace{1cm} (4.12)

$$\hat{b} = xb + ya + n$$  \hspace{1cm} (4.13)

If the first received symbol is expressed as $\hat{a}$ and the second is expressed as $\hat{b}$, then the first equation shows that for the received symbol $a$, the interference of received symbol $b$ is also available. So if one is to use the zero forcing of Equation 4.11, then the noise $n$ is amplified.

### 4.3.4 Maximal Ratio Combining (MRC) Decoding

The MRC is a decoder which is described as follows

$$g_k = h_k^*.$$  \hspace{1cm} (4.14)

After the channel estimation has taken place, the conjugate of the estimated channel is calculated. This has the worst performance for BSOOFDM between all the decoders discussed in this chapter and can be seen in the results section. This is due to its inability to compensate for the spreading and de-spreading which takes place in BSOOFDM.
This had a poor performance, but is better suited when applied with code gain and will be discussed in the final chapter.

### 4.3.5 Equal Gain Combining (EGC) Decoding

The Equal Gain Combining decoder does not attempt to equalize the effect of the channel distortion in any way, but it is desirable for its simplicity. The EGC decoder can be expressed as

\[
g_k = \frac{h_k^*}{|h_k|}.
\]  

(4.15)

This has a similar expression to the MMSE, but unlike the MMSE, regardless of the SNR, the same format remains. The result shows that this does not have the same performance as MMSE, but has a better performance than the MRC in this system.

For the simulation results the transmission is carried over the UWB channels and that the cyclic prefix is of sufficient length (longer than the maximum path delay). The assumption is that the channel is known at the receiver. The modulation scheme at the transmitter and receiver used is QPSK. The number of subcarriers used range from \( N = 16 \) to \( N = 128 \) with the number of packets simulated ranging from \( a = 10000 \) to \( a = 100000 \).

As can be seen from Figures 4.19 to 4.25 the ML decoder performance the best in terms of BER as expected. This comes with the increased complexity as ML tries all different combinations at the receiver. The MMSE outperforms the ZF at smaller SNR and is less complex than the ML but at high SNR the performance is the same as the ZF decoder. These same figures also show that the MRC decoder performs the worst out of the five decoders studied. The EGC decoder, while outperforming the MRC decoder, does not show any improvement over the normal QPSK modulation although block spreading is used. So the advantage that the Rotation Spreading matrix offers across UWB channels is not shown when using the MRC and the EGC decoding.
Rotation Spreading Matrix for Block Spread OFDM

Figure 4.19 BER between MRC, EGC, MMSE, ZF and ML decoders Rotation Spreading matrix N=16 for BSOFDM in UWB CM1.

Figure 4.20 BER between MRC, EGC, MMSE, ZF and ML decoders Rotation spreading matrix N=16 for BSOFDM in UWB CM2.
Figures 4.21, 4.22 and 4.23 shows simulation results using the new Rotation Spreading matrix with angle $\alpha = \frac{\pi}{3}$ using $N = 32, 64, 128$ subcarriers across the UWB channel $CM_1$. These figures shows that the ML outperforms all the other decoders studied. The MMSE shows a small improvement at lower SNR over the ZF decoder. The ZF decoder also shows good performance. The MRC again is shown to be the poorest decoder used, while the EGC does not show any improvement over QPSK modulation used although using block spreading.

The following simulation results shown below use 10000 BSOFDM packets, using for the spreading matrix the Rotation Spreading matrix with rotation angle $\alpha = \frac{\pi}{3}$, subcarriers $N = 64, 128, 512$ and two ray slow fading channel.

As can be seen from the results in Figures 4.26, 4.27 and 4.28, as the block sizes increased the system performance in terms of BER improved [21]. As an example of better results using larger blocks for the block size $M = 16$ using $N = 64$ subcarriers yields BER of zero, which is also true for $N = 128$ and $N = 512$ subcarriers. Please note $h2$ channel represents a two ray slow fading channel.
Rotation Spreading Matrix for Block Spread OFDM

Figure 4.22 BER between MRC, EGC, MMSE, ZF and ML decoders Rotation Spreading matrix N=64 for BSOFDM in UWB CM1.

Figure 4.23 BER between MRC, EGC, MMSE, ZF and ML decoders Rotation Spreading matrix N=128 for BSOFDM in UWB CM1.
Figure 4.24 BER between MRC, EGC, MMSE, ZF and ML decoders Rotation Spreading matrix N=16 for BSOFDM in UWB CM3.

Figure 4.25 BER between MRC, EGC, MMSE, ZF and ML decoders Rotation Spreading matrix N=16 for BSOFDM in UWB CM4.
Figure 4.26 Simulation results comparing different block sizes, using Rotation Spreading Matrix with $\alpha = \frac{\pi}{3}$ $N = 64$, two ray slow fading channel

Figure 4.27 Simulation results comparing different block sizes, using Rotation Spreading Matrix with $\alpha = \frac{\pi}{3}$ $N = 128$, two ray slow fading channel
Figure 4.28 Simulation results comparing different block sizes, using Rotation Spreading Matrix with $\alpha = \frac{\pi}{3}$, $N = 512$, two ray slow fading channel

### 4.4 Conclusion

This chapter has proposed a new solution to increasing the correlation between transmitted symbols in OFDM system by employing frequency diversity through the development of a new spreading matrix called the Rotation Spreading matrix first presented in [9]. This is done by rotating the coordinates of the modulation symbols at the transmitter and then taking the inverse at the receiver. This allows the user to increase the bandwidth efficiency without necessarily having to trade off the BER performance of the system.

A study for varying angles for the Rotation Spreading matrix is presented [14]. This study is carried in a BSOFDM system over the four models proposed for the IEEE.15.3a task group for UWB channels. This study allowed further analysis of this new spreading matrix and under certain conditions which angles were more efficient than others.

The advantages that can be noted is its flexibility in determining different structures of matrices and the simple angle rotation allows an improvement to take place over more traditional spreading matrices. At the same time, this matrix can reproduce existing matrices depending on the angle used, for example $\frac{\pi}{4}$.
would result in the Hadamard matrix.

It can be stated at the end of this study, that the angles which perform the best across UWB channel for matrix size of \( U_{2 \times 2} \) are the angles \( \frac{\pi}{3}, \frac{\pi}{6}, \text{ and } \frac{\pi}{7} \). Angles \( \frac{\pi}{4}, \frac{\pi}{2}, \) and \( \pi \) do not achieve the same performance as those listed above, since the rotation of the modulation resulted in the same constellation.

A study of different decoders that can be used with the new Rotation Spreading matrix for BSOFDM system over the same UWB channels which is presented in [16]. The ML showed that it outperforms all the other decoders, as expected, due to it taking all possible combinations at the receiver and calculating the required constellation points. This comes at the expense of increased complexity as the size of the \( N \) subcarriers increased. The MMSE showed good performance at low SNR and is less complex than the ML. This has a good compromise between noise and ISI minimization and is widely used in practice due to its robustness. The disadvantage of such a decoder is at high SNR it has the same performance as the ZF decoder. Out of all the practical decoders the ZF is the least complex and at high SNR has the same performance as the MMSE. For BSOFDM, in frequency selective channels such as the UWB, it continues to display the advantage of frequency diversity that BSOFDM shows.

It can be concluded that for the new Rotation Spreading matrix for BSOFDM in frequency selective channels, such as UWB, that the MMSE be used as it offers the most practical solution in terms of complexity and performance. Today there are a number of alternative methods used with the MMSE to further help improve the performance while maintaining the robustness and the complex less structure.

Finally, this new spreading matrix is another method to employ frequency diversity for OFDM systems. The studies and analysis showed this is a very good candidate for such a task.
Chapter 5

Higher Order Rotation Spreading Matrix

5.1 Introduction

In the previous chapter a new spreading matrix called the Rotation Spreading matrix was proposed as a method to employ frequency diversity to improve the overall performance of OFDM. As discussed in the literature review, the performance of BSOFDM is, in one way or another, reliant on the size of the block which is being produced for BSOFDM, $M$. So this requires a method by which the Rotation Spreading matrix is made scalable. That is to obtain higher order Rotation Spreading matrices.

This chapter presents and discusses two methods to achieve higher order Rotation Spreading matrix which are based on recursive methods for the Hadamard and the Scalable Complementary Sequences. This work was also published in a book chapter [22].

5.2 Higher Order Rotation Matrix Based on the Recursive Method

In [15] the first method is presented to produce higher order Rotation Spreading matrices. This method is similar to the one used for achieving higher order
Hadamard matrices. If it can be assumed that the Rotation Spreading matrix is a square matrix $U_N$ of dimensions $N \times N$ with the three kinds of elements $\tan(\alpha)$, 1 and $-1$, which satisfies

$$U_N U_N^T = U_N^T U_N$$
$$= AI_N$$

(5.1)

where $U_N^T$ stands for transpose of $U_N$ and $I_N$ is the identity matrix of order $N$ and $A$ is a scaling factor of $1 + \tan^2(\alpha)$. Equation 5.1 shows that any two sequences given by rows or columns of $U_N$ are orthogonal. Then the Rotation Spreading matrix of $N = 2^n$ ($n \geq 0$) can be generated using the simple recursive procedure presented below

$$U_{2N} = \begin{bmatrix} U_N & U_N \\ U_N & -U_N \end{bmatrix}.$$  (5.2)

Using the method described in Equation 5.2 for the Rotation Spread matrix, the $U_{4\times4}$ would have the following structure

$$U_4 = \begin{bmatrix} 1 & \tan(\alpha) & 1 & \tan(\alpha) \\ \tan(\alpha) & -1 & \tan(\alpha) & -1 \\ 1 & \tan(\alpha) & -1 & -\tan(\alpha) \\ \tan(\alpha) & -1 & -\tan(\alpha) & 1 \end{bmatrix}$$  (5.3)

where $\alpha$ is the rotation angle. Then the higher order matrix $U_{8\times8}$ would have the following structure
Higher Order Rotation Spreading Matrix

$$U_8 = \begin{bmatrix}
1 & \tan(\alpha) & 1 & \tan(\alpha) & 1 & \tan(\alpha) & 1 & \tan(\alpha) \\
\tan(\alpha) & -1 & \tan(\alpha) & -1 & \tan(\alpha) & -1 & \tan(\alpha) & -1 \\
1 & \tan(\alpha) & -1 & -\tan(\alpha) & 1 & \tan(\alpha) & -1 & -\tan(\alpha) \\
\tan(\alpha) & -1 & -\tan(\alpha) & 1 & \tan(\alpha) & -1 & -\tan(\alpha) & 1 \\
1 & \tan(\alpha) & 1 & \tan(\alpha) & -1 & -\tan(\alpha) & -1 & -\tan(\alpha) \\
\tan(\alpha) & -1 & \tan(\alpha) & -1 & -\tan(\alpha) & -1 & \tan(\alpha) & -1 \\
\end{bmatrix}$$

(5.4)

where $\alpha$ is the rotation angle.

The Rotation Spreading matrix has been proven to outperform other spreading matrices in UWB channels and were presented and discussed in the previous chapter.

Based on this method to achieve higher order Rotation Spreading matrix, it can be shown that these new higher order structures maintained the important condition seen in Equation 5.1. Also, as will be shown in this chapter through experimental results, the higher order Rotation Spreading matrix achieves a better system performance when compared with the Hadamard and Rotated Hadamard matrices when used for pre-coding for OFDM.

Figure 5.1 depicts the result of the BER versus SNR of the new Rotation Spreading matrix using $M = 16$ sized blocks with the Rotation Spreading matrix $U_{16 \times 16}$ compared to the Hadamard matrix in a two ray slow fading channel. The Rotation Spreading matrix outperforms the Hadamard by approximately over 3 dB. The number of subcarriers is 32. Please note where $H2$ is used in the headings of the graphs this represents a two ray slow fading channel.

Figure 5.2 depicts the Rotation Spreading matrix using $M = 8$ sized blocks comparing the Hadamard $H_{8 \times 8}$ matrix. The Rotation Spreading matrix outperforms the Hadamard by approximately 2 dB.

Figure 5.3 depicts the Rotation Spreading matrix with the block size of $M = 4$ versus the Hadamard $H_{4 \times 4}$ matrix. As can be seen from this figure the Rotation Spreading matrix outperforms the Hadamard by approximately 2 dB.
Figure 5.1 BER M=16 Rotation Spreading matrix versus Hadamard in a two ray model channel N=32 subcarriers.

Figure 5.2 BER M=8 Rotation Spreading matrix versus Hadamard in a two ray model channel N=32 subcarriers.
5.3 Higher Order Rotation Spreading Matrix Based on the Complete Complementary Sets of Sequences (CCSS) Method

In [18] another method for achieving higher order Rotation Spreading matrix is proposed which was based on the Complete Complementary method proposed in [84] for achieving higher order Hadamard matrix.

In [84], the authors showed a new method to expand the Hadamard matrix into higher order. This method was employed to expand the Rotation Spreading matrix and can be described as follows

\[ U_{2N} = \begin{bmatrix} U_N & \tilde{U}_N \\ U_N & -\tilde{U}_N \end{bmatrix} \]  

(5.5)

where \( \tilde{U}_N = P_N U_N Q_N \), which denotes an equivalent rotation matrix obtained by permuting the rows and columns of \( U_N \). \( P_N \) and \( Q_N \) are arbitrary monomial permutation matrices which have exactly one non-zero entry in every row and
column and therefore satisfying the conditions

\[
P_N P_N^T = P_N^T P_N = Q_N Q_N^T = Q_N^T Q_N = I_N. \tag{5.6}
\]

where \( P_N \) is

\[
P_N = \begin{bmatrix}
0 & I_{\frac{N}{2}} \\
I_{\frac{N}{2}} & 0
\end{bmatrix} \tag{5.7}
\]

and \( Q_N = I_N \). \( \tilde{U} \) is obtained by exchanging the upper and lower half of \( U_N \) and it was shown in [84] to have good complementary properties.

When this method is applied to the Rotation Spreading matrix, the following structure for a four by four Rotation Spreading matrix \( U \) is achieved

\[
U_4 = \begin{bmatrix}
1 & \tan(\alpha) & \tan(\alpha) & -1 \\
\tan(\alpha) & -1 & 1 & \tan(\alpha) \\
1 & \tan(\alpha) & -\tan(\alpha) & 1 \\
\tan(\alpha) & -1 & -1 & -\tan(\alpha)
\end{bmatrix}. \tag{5.8}
\]

The eight by eight \( U \) Rotation Spreading matrix based on the CCSS expansion can be shown as follows

\[
U_8 = \begin{bmatrix}
1 & \tan(\alpha) & \tan(\alpha) & -1 & 1 & \tan(\alpha) & -\tan(\alpha) & 1 \\
\tan(\alpha) & -1 & 1 & \tan(\alpha) & \tan(\alpha) & -1 & -1 & -\tan(\alpha) \\
1 & \tan(\alpha) & -\tan(\alpha) & 1 & 1 & \tan(\alpha) & \tan(\alpha) & -1 \\
\tan(\alpha) & -1 & -1 & -\tan(\alpha) & \tan(\alpha) & -1 & 1 & \tan(\alpha) \\
1 & \tan(\alpha) & \tan(\alpha) & -1 & -1 & -\tan(\alpha) & \tan(\alpha) & -1 \\
\tan(\alpha) & -1 & 1 & \tan(\alpha) & -\tan(\alpha) & 1 & 1 & \tan(\alpha) \\
1 & \tan(\alpha) & -\tan(\alpha) & 1 & -1 & \tan(\alpha) & -\tan(\alpha) & 1 \\
\tan(\alpha) & -1 & -1 & -\tan(\alpha) & -\tan(\alpha) & 1 & -1 & -\tan(\alpha)
\end{bmatrix} \tag{5.9}
\]
Figure 5.4 Comparing Rotation Spreading matrix \((\frac{\pi}{3})\) higher order using CCSS with Rotated Hadamard and Hadamard in a two ray channel \(N = 16, M = 4\).

where \(\alpha\) is the rotation angle. Again, the condition presented in Equation 5.1 is maintained.

The following experimental results show that the expansion based on CCSS method also outperforms the Hadamard and Rotated Hadamard using block sizes of \(M = 16, M = 8\) and \(M = 4\) across frequency selective channels and slow fading channels. As can be seen from Figures 5.4, 5.5 and 5.6 the Rotation Spreading matrices outperforms the Rotated Hadamard and Hadamard matrices in BSOFDM across fading channels by more than 2dB. The rotation angle used for these figures was \(\alpha = \frac{\pi}{3}\), with the subcarriers ranging from \(N = 16\) to \(N = 64\).

This allows the comparison of the two expansion methods for higher order in UWB channels for the Rotation Spreading matrix. For block size of \(M = 4\), when comparing the two methods in \(CM1\) (line of site in short distances) there is no difference in terms of BER gain. This is depicted in Figure 5.7.

In Figure 5.8 it can be seen when using block size of \(M = 4\) with the number of subcarriers \(N = 64\) in UWB channel \(CM2\), the CCSS expansion method for the Rotation Spreading matrix outperforms the recursive expansion method by
Figure 5.5 Comparing Rotation Spreading matrix ($\frac{\pi}{3}$) higher order using CCSS with Rotated Hadamard and Hadamard in a two ray channel $N = 16$ $M = 8$.

Figure 5.6 Comparing Rotation matrix ($\frac{\pi}{3}$) higher order using CCSS with Rotated Hadamard and Hadamard in a two ray channel $N = 64$ $M = 16$. 
Higher Order Rotation Spreading Matrix

Figure 5.7 Comparing Recursive method with CCSS method expansion $M = 4$ $N = 64$ in UWB CM1.

a gain of 0.5dB. This is a very small gain.

In Figure 5.9, there is a further improvement of the CCSS expansion method over the recursive expansion method when the two are compared for block size of $M = 4$ using $N = 64$ subcarriers across the UWB channel CM3 of approximately 1.5dB.

Figures 5.10 and 5.11 depict similar results for block size $M = 8$ for the UWB channels CM1 to CM4.

When the CCSS expansion method is compared with the recursive expansion method for higher order Rotation Spreading matrix in slow fading channel, seen in Figure 5.12, the two methods have a very similar result and no advantage can be observed over the other in terms of BER gain. Both expansion methods still outperformed the Rotated Hadamard by approximately 2dB.

In the previous chapter, a study was carried out for the different angles for the Rotation Spreading matrix. It mentioned that there are some angles which cannot be used since the rotation achieved the same modulation scheme as the modulation at the transmission. These angles included $\alpha = \frac{\pi}{4}$, $\alpha = \pi$ and $\alpha = \frac{\pi}{2}$ when using the QPSK modulation. These angles appeared not to be useful. As
Figure 5.8 Comparing Recursive method with CCSS method expansion $M = 4$ $N = 64$ in UWB CM2.

Figure 5.9 Comparing Recursive method with CCSS method expansion $M = 4$ $N = 64$ in UWB CM3.
Figure 5.10 Comparing Recursive method with CCSS method expansion $M = 8\  N = 64$ in UWB $CM1$.

Figure 5.11 Comparing Recursive method with CCSS method expansion $M = 8\  N = 64$ in UWB $CM2$. 
 Higher Order Rotation Spreading Matrix

BER comparison of two expansion methods Vs Rotated Had BSOFDM QPSK M=16 N=64 H2 channel

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure5_12.png}
\caption{Comparing recursive method with CCSS method expansion $M = 16$ $N = 64$ in a two ray channel.}
\end{figure}

As can be seen in the next section, when the Rotation Spreading matrix applies the two methods discussed above for higher order sized matrices, this limitation is no longer applicable.

\section{A Study of Different Angles for Higher Order Rotation Spreading Matrix for BSOFDM in UWB Channels}

This section presents a study into different angles for higher order Rotation Spreading matrix developed for BSOFDM which is presented in [17]. It was shown previously that for block size $M = 2$ the angle $\alpha = \frac{\pi}{3}$ achieved the best result in terms of BER in UWB channels. It was discovered that this is no longer the case when the higher order Rotation Spreading matrix is used for larger $M$ sized blocks and that other angles produced better results. This showed that the Rotation Spreading matrix advantage over existing spreading matrices such as the Hadamard is its flexibility to be adapted to different communication.
systems and channels due to it maintaining its orthogonality for higher order matrices.

Figures 5.13-5.18 depict the different constellation points which are achievable when using the rotation angles $\alpha = \frac{\pi}{3}, \frac{\pi}{6}, \frac{\pi}{5}, \frac{\pi}{7}, \frac{\pi}{9}$ and $\frac{3\pi}{4}$ respectively for block size $M = 4$ with the number of subcarriers of $N = 512$.

An interesting feature of this higher order Rotation Spreading matrix is for angles such as $\alpha = \frac{\pi}{4}$, the modulation constellation no longer rotates on it self as it does with the case of block size $M = 2$ in [14]. As can be seen from Figure 5.19, it still increased the correlation between the transmitted symbols. This was also true for the angles $\frac{\pi}{2}$ and $\pi$ which for $M = 2$ do nothing to improve the correlation of the symbols transmitted. These can be seen in Figures 5.20 and 5.21.

It can be shown that for higher order Rotation Spreading matrices that different angles outperform, in terms of BER, the angle shown in [14]. It can also be shown that where previously angles such as $\alpha = \frac{\pi}{4}, \alpha = \frac{\pi}{2}$ and $\alpha = \pi$ could not be used for a block size of $M = 2$, however for higher order Rotation Spreading matrix these show an increase in the correlation between the transmitted
Figure 5.14 Higher Order Rotation Spreading matrix with rotation $\alpha = \frac{\pi}{6}$, $M = 4$, $N = 512$.

Figure 5.15 Higher Order Rotation Spreading matrix with rotation $\alpha = \frac{\pi}{5}$, $M = 4$, $N = 512$. 
Figure 5.16 Higher Order Rotation Spreading matrix with rotation $\alpha = \frac{\pi}{7}$, $M = 4$, $N = 512$.

Figure 5.17 Higher Order Rotation Spreading matrix with rotation $\alpha = \frac{\pi}{9}$, $M = 4$, $N = 512$. 
Figure 5.18 Higher Order Rotation Spreading matrix with rotation $\alpha = \frac{3\pi}{4}$ $M = 4$ $N = 512$.

Figure 5.19 Higher Order Rotation Spreading matrix with rotation $\alpha = \frac{\pi}{4}$ $M = 4$ $N = 512$. 
Figure 5.20 Higher Order Rotation Spreading matrix with rotation $\alpha = \frac{\pi}{2}$ $M = 4$ $N = 512$.

Figure 5.21 Higher Order Rotation Spreading matrix with rotation $\alpha = \pi$ $M = 4$ $N = 512$. 
The angle $\alpha = \frac{\pi}{4}$ outperforms the other angles such as $\frac{\pi}{3}$ in UWB channels and can be seen in Figure 5.22.

For angle $\alpha = \frac{3\pi}{4}$ the Rotation Spreading matrix for block size $M = 4$ in $CM3$ achieves the best result when compared with the angles $\frac{\pi}{3}$, $\frac{\pi}{6}$, $\frac{\pi}{7}$ and $\frac{\pi}{5}$, which can be seen in Figure 5.24. Also angle $\alpha = \frac{\pi}{5}$ produces good results and can be seen in Figure 5.25 across UWB channel $CM3$ and in Figure 5.26 across UWB channel $CM1$.

The advantage of this Rotation Spreading matrix in higher order block sizes is its flexibility in adapting to the required communication environment. When the block size $M = 2$ is applied in [14] it is shown that some angles do not improve the performance of the system. For higher order matrices the very same angles show that they can be used and a better BER is achievable when compared with other angles. The channels used for this study are the UWB channels $CM1$ to $CM4$. The subcarriers modelled vary from $N = 64$ to $N = 512$. The block size is $M = 4$ and the decoder at the receiver used is the Zero Forcing assuming that the channel is known. Not all results are shown due to space limits.
Figure 5.23 Comparing all angles in UWB CM3, it can be seen that $\frac{\pi}{2}$ although for $M = 4$ increases correlation between symbols does not achieve better results.

Figure 5.24 Comparing all angles in UWB CM3, it can be seen that $\frac{3\pi}{4}$ has better results for $M = 4$ $N = 64$. 
Higher Order Rotation Spreading Matrix

Figure 5.25 Comparing all angles in UWB CM3, it can be seen that $\pi/5$ has better results, $M = 4$ $N = 64$.

Figure 5.26 Comparing all angles in UWB CM1, it can be seen that $\pi/5$ has better results for $M = 4$ $N = 512$. 
5.5 Conclusion

Two methods are presented to allow the expansion or make scalable the new Rotation Spreading matrix for higher order block sizes for the BSOFDM or precoded OFDM system. The first method is based on the recursive that is used for the Hadamard matrix and is presented in [15]. The second method presented a way to increase the Rotation Spreading matrix to higher order based on a method proposed in [84], which is presented in [18].

These two methods proposed to achieve higher order matrices for the Rotation Spreading matrix showed that this matrix still outperforms the other spreading matrices in slow fading and selective frequency channels. This is due to its ability to maintain its orthogonality for higher order block sizes.

A study of varying angles for higher order Rotation Spreading matrix presented in [17] for BSOFDM system over the four UWB channel models is presented. Although in [14] it was shown that for block size of $M = 2$ the angles $\alpha = \frac{\pi}{3}$ produced the best result in terms of BER across UWB channels, that is no longer the case for higher order Rotation Spreading matrix. It is also shown that although some angles such as $\alpha = \frac{\pi}{4}, \frac{\pi}{2}$ and $\pi$ for $M = 2$ were useless due to them simply rotating the modulation scheme back onto itself in the case of QPSK modulation and did not increase the correlation between the transmitted signals. These same angles for higher order Rotation Spreading matrix are shown to not only increase the correlation but in some cases such as the angle $\alpha = \frac{\pi}{4}$ outperform other angles in UWB channels.

It can be stated that the previous limitation of the angles mentioned in the previous chapter no longer hold for larger block sizes of $M$ for higher order Rotation Spreading matrix. It can be confirmed that the angles that achieve the greatest $dB$ gain across UWB channel models for higher order block sizes are the angles $\alpha = \frac{\pi}{5}, \frac{\pi}{6}, \frac{3\pi}{4}$ and $\frac{\pi}{4}$. 
Chapter 6

Delayed Block Spread OFDM

6.1 Introduction

Time diversity is discussed in the literature review as a method which can be employed to improve the overall OFDM systems performance. There are a number of ways to implement such a diversity scheme, which are also discussed in the literature chapter. This chapter presents a method to apply time diversity to BSOFDM to improve the overall system performance.

Delayed Block Spread OFDM delays the transmitted blocks by a time interval, $\tau$. The purpose of such a method is to ensure that each block, for this example the $M = 2$ will be used, encounters uncorrelated channels. As discussed in the previous chapter, spreading matrices such as the Hadamard and the Rotated Hadamard do not perform well in slow fading channels. In fact, these two spreading matrices do not show any improvement over an un-coded OFDM system. This delayed block spread OFDM method allows such spreading matrices systems to achieve better BER gains in these types of environments.

6.2 Description of Delayed Block Spread OFDM System

The idea behind the delayed Block Spread OFDM (d-BSOFDM), which is presented in [10] and [12], is to exploit time diversity to further improve BSOFDM.
The functionality is similar to BSOFDM discussed in the two previous chapters except the $q$ blocks are spread across a number of OFDM symbols rather than just the one OFDM symbol. As shown in Figure 6.1, the odd blocks are spread across the first OFDM symbol and the even blocks are spread across the second OFDM symbol introducing time diversity among the blocks.

The motivation behind this is the fact that although conventional BSOFDM improves on the system performance of the conventional OFDM in frequency selective channels, our studies have shown that it does not do so under flat fading conditions which do occur during transmission. An example of this is shown in Figures 6.2 and 6.3 for block size of $M = 2$, using BPSK modulation at the transmission. When comparing the BER versus SNR for OFDM and BSOFDM systems, the same performance is achieved and no improvement is seen when applying frequency diversity (this is important since applying frequency diversity increases the complexity of the system).

The unitary spreading matrix applied to achieve frequency diversity for these simulation results is the Rotated Hadamard matrix, which has the following
Figure 6.2 Bit Error Rate versus SNR comparing the BSOFDM and OFDM in Flat Fading environment.

Figure 6.3 Packet Error Rate versus SNR comparing the BSOFDM and OFDM in Flat Fading environment.
structure

\[ U = \frac{1}{\sqrt{2}} \times H_{M \times M} \times \text{diag}(\exp(\frac{j\pi m}{W})) \]  

(6.1)

where the \( H_{M \times M} \) is the Hadamard matrix, \( m = [0 < m < M - 1] \) and \( M \) is the size of the block. The \( W \) is chosen so that \( \frac{2\pi}{\theta} \) is the smallest angle which the constellation rotates back onto itself and for MPSK this equals two and for QAM this equals four. Chapter 3 discusses that the Hadamard and Rotated Hadamard do not improve the OFDM system across slow fading channels. When these unitary matrices are used with the d-BSOFDM system, these matrices do show improvement.

At the receiver, the received data is de-multiplexed and recombined in the same order as the blocks are transmitted.

To choose the delay for each block a look at the the Rayleigh channel model is required. The Rayleigh channel can be modelled by using the Bessel Function of the first order. To exploit time diversity between the blocks, the time delay \( \tau \) is chosen by taking the point on the time axis \( \beta \) to be approximately 2.5, this can be seen in Figure 6.4.

If the the maximum doppler shift \( f_d \) is chosen at 100Hz, then the delay between the blocks to be calculated to achieve the time diversity is

\[ \tau = \frac{\beta}{2 \times \pi \times f_d} \]

(6.2)

\[ \tau = \frac{2.5}{2 \times \pi \times 100} = 3.9 \times 10^{-3}. \]

If \( \tau \) is calculated incorrectly or not accurately enough, while the blocks are not fully correlated, the blocks are not independent of each other. If the correlation is as high as 0.7, the system will potentially lose a 0.5dB in performance. The studies have shown improvements on the conventional BSOIFDM.
In [12] d-BSOFDM is applied to a multiple path environment and the same improvement in flat fading is also achieved in this environment.

### 6.3 Results

Table 6.1 lists the parameters used for the simulation results in this section. The results show that employing this type of time diversity to pre-coded OFDM or BSOFDM the BER gain will improve.

As can be seen from Figures 6.2 and 6.3, which compare the conventional BSOFDM and OFDM across flat fading channels, no improvement is seen. As time diversity is introduced the results show improvement. These results can
be seen in Figures 6.5 and 6.6 for block size of $M = 2$, when using BPSK at the transmitter. There is an approximate gain of $0.7dB$. The decoder used at the receiver is the ZF. The same gain in dB is also achieved for packet error rate versus SNR. The number of subcarriers simulated is $N = 64$. Figures 6.7 and 6.9 depict the BER and PER versus SNR when using $N = 128$ subcarriers. In this case a gain of $0.7dB$ is achieved. Again using the BPSK at the transmission and ZF decoder at the receiver. While the results show improvement, it is expected further improvement is possible when uncorrelated and independent blocks are achieved.

The time diversity improvement discussed above has also, as expected, been translated from the flat fading channels to the frequency selective environment using both the BPSK and QPSK modulation. Figures 6.9 and 6.10 shows time diversity introduced into BSOFDM while in frequency selective channel.

The number of subcarriers used in these simulation results is $N = 128$ with 40000 OFDM packets simulated. QPSK modulation is used at the transmission and at the receiver the decoder used is the ZF. The dB gain achieved is approximately $0.7dB$. 

Figure 6.5 Bit Error Rate versus SNR comparing the delayed BSOFDM and BSOFDM in Flat Fading environment.
Figure 6.6 Packet Error Rate versus SNR comparing the delayed BSOFDM and BSOFDM in Flat Fading environment.

Figure 6.7 BER of conventional BSOFDM versus time diversity BSOFDM using BPSK modulation.
Figure 6.8 PER of conventional BSOFDM versus time diversity BSOFDM using BPSK modulation.

Figure 6.9 BER of conventional BSOFDM versus time diversity BSOFDM using QPSK modulation.
Delayed Block Spread OFDM

Figure 6.10 PER of conventional BSOFDM versus time diversity BSOFDM using QPSK modulation.

6.4 Conclusion

This chapter presents another method to improve the overall BER system performance of BSOFDM system based on time diversity. It is shown in the two previous chapters that BSOFDM systems does not perform well under flat fading conditions when employing spreading matrices such as the Hadamard or the Rotated Hadamard. In fact, no system improvement is shown.

Delayed Block Spread OFDM shows that by using time diversity the system does improve in terms of BER gain. It shows that the Hadamard matrix can still be used in such an environment. The d-BSOFDM, delays the block into another OFDM packet and this will allow the $M$ blocks to encounter different uncorrelated channels therefore improving the overall system performance.

If the blocks do not achieve a correlation of zero and are closer to a correlation of one, then the performance degraded and the system potentially loses up to 0.5dB of performance.
Chapter 7

A New Approach to BSOFDM - Parallel Concatenated Spreading Matrices OFDM

7.1 Introduction

Coding gain has been previously used in error correction codes such as Turbo codes and Low Density Parity checks (LDPC) to achieve better performance. This type of gain is applied to BSOFDM in a new system described as Parallel Concatenated Spreading Matrices (PCSM) OFDM which is presented in [13] and [23]. This new method spreads the same data $n$ times. While BSOFDM improved the overall BER performance on OFDM in frequency selective channels, this new approach further improves the BER of BSOFDM by over $3dB$ gain in frequency selective channels.

7.2 System Description of PCSM-OFDM

Figure 7.1 depicts the block diagram of the new system described as Parallel Concatenated Spreading Matrices OFDM or PCSM-OFDM. The same data is copied into two streams of block size $\frac{N}{M}$, $d_1$ and $d_2$, where $N$ is the number of subcarriers and $M$ is the block size. The same data is used for each stream. The block size $M = 2$ is used for this example, and each stream is spread using
A New Approach to BSOFDM - Parallel Concatenated Spreading Matrices

![Diagram of BSOFDM system](image)

**Figure 7.1** The new approach to BSOFDM, Parallel Concatenated Spreading Matrices.

A unitary spreading matrix $U$ of size $M \times M$. The streams of $M$ sized blocks can be described as $d_1 U_1$ and $d_2 U_2$. The streams are multiplexed and the same process which is applied to BSOFDM is applied to PCSM-OFDM after this point.

Figure 7.2 depicts another two configurations for the PCSM-OFDM. These include the use of a random interleaver.

At the receiver, the same process applied to BSOFDM is applied to PCSM-OFDM except the de-multiplexing is used to separate the two streams apart. To make full use of the coding gain the two data streams are combined before de-spreading takes place and the combining is done by using Maximum Ratio Combining (MRC) or Equal Gain combining (EGC) and can be represented by the following equation

$$ R_1 = \frac{(\alpha_1^* r_1 + \alpha_2^* r_2)}{|\alpha_1|^2 + |\alpha_2|^2} \quad (7.1) $$
A New Approach to BSOFDM - Parallel Concatenated Spreading Matrices
OFDM

Figure 7.2 (a) PCSM-OFDM with the combining after the de-spreading and (b) an interleaver included before the second spreading matrix.
where $\alpha_1$ and $\alpha_2$ are the estimated channel weights and $r_1$ and $r_2$ are the useful data from each stream. The system can also apply what is known as Equal Gain Combining (EGC) where the channel weights $\alpha_1$ and $\alpha_2$ are given equal priority of 1, and Equation 7.1 becomes

$$R_1 = \frac{\alpha_1^*}{|\alpha_1|} r_1 + \frac{\alpha_2^*}{|\alpha_2|} r_2.$$  

(7.2)

The results are described below with the decoder used for these simulation results being the Zero Forcing decoder and the channels used for these simulations are slow fading and frequency selective UWB channels. The combining unless otherwise specified is the EGC and the number of packets simulated is 10000 packets. The spreading matrices used in these studies are the Hadamard, Rotated Hadamard and the Rotation Spreading matrices.

Figure 7.3 depicts the BER versus SNR in a slow fading channel when using the Hadamard matrix. The number of subcarriers used is $N = 128$. The BER gain of PCSM-OFDM over BSOFDM is greater than 3$dB$. This showed that employing this new structure greatly improved the performance of the OFDM
system. The same performance is also seen when used with $N = 64$ subcarriers in the same environment. As expected the same performance is also seen when using the new Rotation Spreading matrix and Rotated Hadamard in the same environment. These results are presented in Figure 7.5 and 7.6 when using $N = 16$ subcarriers.

Figure 7.7 compares the PCSM-OFDM using $N = 16$ subcarriers with classical BSOFDM $N = 32$ subcarriers to ensure that there is a real improvement with this new approach and not just due to an increase in samples across the channel. The $N = 32$ for BSOFDM is used to compare the same sample number across the channel, and as can be seen the improvement is still evident from the simulation results. It can be concluded that the improvement is not due to the increase in samples. In the slow fading channel there is a greater than $4dB$ improvement in terms of BER.

Across frequency selective channels such as UWB, this new system also showed superior performance to the classical BSOFDM. UWB channels $CM3$ and $CM4$ (which represent medium to long distances of non-line of sight) are used for the purpose of studying the PCSM-OFDM system. Figures 7.8 and 7.9 depicts
Figure 7.5 PCSM-OFDM compared to BSOFDM using the Rotation Spreading matrix N=16.

Figure 7.6 PCSM-OFDM compared to BSOFDM using the Rotated Hadamard matrix N=16.
the simulation results of BER versus SNR of the Hadamard matrix in $N = 64$ and $N = 128$ subcarriers respectively. Again PCSM-OFDM outperforms the classical BSOFDM, with a gain in dB of over $2dB$. It is important to note that these improvements are achieved at low SNR not at high SNR.

In Figure 7.2, two different configuration of structures for the PCSM-OFDM are shown. The simulation results do not show any difference in dB gain between the two structures. These results are depicted in Figures 7.10 and 7.11. The only real advantage of one over the other is the combination of the two streams is done before the de-spreading will ensure that only one inverse unitary matrix is used at the receiver. This reduces the overall complexity of the system.

### 7.3 Higher Order Parallel Concatenated Spreading Matrices OFDM

This section studies higher order Parallel Concatenated Spreading Matrix OFDM (PCSM-OFDM) which is presented in [20]. PCSM-OFDM is shown to greatly improve on the classic Block Spread OFDM, this continued the work on PCSM-
A New Approach to BSOFDM - Parallel Concatenated Spreading Matrices

Figure 7.8 PCSM-OFDM $N = 64$ in UWB channel $CM3$.

Figure 7.9 PCSM-OFDM $N = 128$ in UWB channel $CM4$ compares with BSOFDM.
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**Figure 7.10** PCSM-OFDM $N = 128$ in UWB channel $CM4$ compares interleaver and non-interleaver using first combination.

**Figure 7.11** PCSM-OFDM $N = 128$ in UWB channel $CM4$ compares interleaver with combination before and after de-spreading.
OFDM and studies higher order in terms of an increased number of streams. It is shown as the number of streams increased so does the BER gain.

### 7.3.1 System Description of Higher Order PCSM-OFDM

A very important question in any system is scalability. This question is applied to PCSM-OFDM and is depicted in Figure 7.12, where \( n \) is the number of streams required.

If the system is to use the higher order PCSM-OFDM, then at the receiver Equation 7.1 is transformed into

\[
R_1 = \frac{\alpha_1^* r_1 + \alpha_2^* r_2 + \alpha_3^* r_3 + \cdots + \alpha_n^* r_n}{|\alpha_1|^2 + |\alpha_2|^2 + |\alpha_3|^2 + \cdots + |\alpha_n|^2} \tag{7.3}
\]

and if the system required the use of EGC, then Equation 7.3 would be transformed into

\[
R_1 = \frac{\alpha_1^*}{|\alpha_1|} r_1 + \frac{\alpha_2^*}{|\alpha_2|} r_2 + \cdots + \frac{\alpha_n^*}{|\alpha_n|} r_n \tag{7.4}
\]
where $\alpha_n$ are the estimated channel weights and $r_n$ representing the useful channel information.

The PCSM-OFDM scalability improved the system performance with the increase of streams in terms of gain in $dB$.

Figure 7.13 compares four, three and two streams for the PCSM-OFDM. The simulation result used $N = 64$ subcarriers with 10000 OFDM packets simulated. As can be seen from the results, as the number of streams increased the BER performance and the $dB$ gain improves. With each stream increase a gain of 1$dB$ is achieved.

Figures 7.14, 7.15, 7.16 and 7.17 compares the same conditions mentioned above but with the number of subcarriers increased from $N = 16$ through to $N = 128$. All these experimental results through simulations are across frequency selective channels UWB and slow fading channels. The decoder at the receiver used is the zero forcing with the unitary matrix used for spreading being the Hadamard matrix.
Figure 7.14 PCSM-OFDM higher order 4, 3 and 2 compared a slow fading channel using $N=128$ subcarriers.

Figure 7.15 PCSM-OFDM compared higher order 3 against 4 $N=128$. 
Figure 7.16 PCSM-OFDM compared higher order 3 against 4 \( N = 64 \).

Figure 7.17 PCSM-OFDM compared higher order 3 against 4 \( N = 32 \).
7.4 Conclusion

This chapter presented a new approach to Block spread OFDM called Parallel Concatenated Spreading Matrices OFDM (PCSM-OFDM) which is presented in [13]. The data is copied into two streams (which become parallel to each other) of M sized blocks and both these streams are spread using the same unitary spreading matrix. Then the two streams are multiplexed and the normal procedure is carried out in BSOFDM and OFDM systems. At the receiver after channel equalization the signal is de-multiplexed and combination using MRC or EGC is done before the de-spreading using the inverse of the unitary matrix. From the simulation results presented, this new approach to OFDM called PCSM-OFDM outperforms the BSOFDM by greater than $4dB$ in slow fading channels and over $3dB$ in frequency selective channels such as UWB channels. Other combinations of PCSM-OFDM included an interleaver between the two streams at the transmission side and carrying out the de-spreading before the two streams at the receiver are combined. The same performance is achieved using both combinations as the original configuration of PCSM-OFDM, but the de-spreading after the combining would reduced complexity as the system is only required to use one unitary matrix inverse. Overall, this system is recommended for wireless communication systems were OFDM and BSOFDM are applied. Higher order PCSM-OFDM is then presented in this chapter [20]. This system showed that with the increase of the number of $n$ streams the system gained an extra $1dB$ gain.
Chapter 8

Conclusion

As a direct result of the contributions presented in this thesis 15 internationally peer-reviewed publications have been realized. This includes two book chapters. These contributions are listed in Chapter one.

The literature review is presented in Chapter two. This chapter concentrates on wireless communications in particular solutions to OFDM when used in indoor environments. This begins with presenting some of the fundamental concepts of wireless communications. It discusses some of the affects in the physical layer regarding channel properties in particular concentrating on indoor propagation.

In Chapter Three the main area of study, primarily OFDM and current contributions to improve aspects of this multi-carrier scheme, are presented and discussed. A particular area of improvement to OFDM called Block Spread OFDM (BSOFDM) or pre-coded OFDM, which utilizes frequency diversity to improve system performance, is presented and discussed. This is concluded by briefly highlighting current work with different types of diversity which are related to the contributions of this thesis.

In Chapter Four, a new spreading matrix called the Rotation Spreading matrix is presented. This matrix is used to apply frequency diversity to OFDM. The advantages for this new spreading matrix over more traditional spreading matrix such as the Hadamard and the Rotated Hadamard become obvious in frequency selective channels such as the UWB channels.
It is proven that the Rotation Spreading matrix outperforms the other spreading matrices due to its flexibility in achieving varying combination for modulation after the spreading is carried out depending on the spreading angle chosen. The new solution is used to increase the correlation between transmitted symbols in OFDM system by employing frequency diversity through. This is first presented in [9]. This is done by rotating the coordinates of the modulation symbols at the transmitter and then taking the inverse at the receiver. This allows the user to increase the bandwidth efficiency without necessarily having to trade off the BER performance of the system. In frequency selective channels when compared with the Hadamard and Rotated Hadamard, approximately 12dB gain and 2dB gain was achieved over the two respectively.

This chapter then presents a study for varying angles which can be used for the new spreading matrix, which is first presented in [14] for BSOFDM system across the UWB channels. This study allowed further analysis of this new spreading matrix and under certain conditions which angles are more efficient than others.

The advantages can be noted is its flexibility in determining different structures of matrices and the angle rotation allows an improvement to take place over more traditional spreading matrices. At the same time, this matrix can reproduce existing matrices depending on the angle used, for example $\frac{\pi}{4}$ would result in the Hadamard matrix for block size of $M = 2$.

It can be stated at the end of this study, that the angles which perform the best over UWB channels are the angles $\alpha = \frac{\pi}{3}, \frac{\pi}{6}$ and $\frac{\pi}{7}$. Angles $\frac{\pi}{4}, \frac{\pi}{2}$ and $\pi$ do not achieve the same performance as those listed above, since the rotation of the modulation resulted in the same constellation for a block size of $M = 2$.

The final study carried out for this matrix is of different decoders that can be used with the new Rotation Spreading matrix for BSOFDM system at the receiver. This is first presented in [16]. The ML decoder showed that it outperforms all the other decoders, as expected, due to it taking all possible combinations at the receiver and calculating the required constellation points. This
comes at the expense of increased complexity as the size of the N subcarri-
er increased. The MMSE showed good performance at low SNR and was less
complex than the ML. This has a good compromise between noise and ISI mini-
mization and is widely used in practice due to its robustness. The disadvantage
of such a decoder is at high SNR it has the same performance as the ZF. Out
of all the practical and useful decoders the ZF is the least complex and at high
SNR has the same performance as the MMSE. For BSOFDM, in frequency
selective channels such as the UWB, it continues to display the advantage of
frequency diversity that BSOFDM shows.

It can be concluded that for the new Rotation Spreading matrix for BSOFDM
in frequency selective channels, that the MMSE be used as it offers the most
practical solution in terms of complexity and performance. Today there are a
number of alternative methods used with the MMSE to further help improve the
performance while maintaining the robustness and the complex less structure.
An analytical study of the Rotation Spreading matrix is also presented (first
presented in [21]).

Chapter Five proposes two methods which expand or make scalable the new
Rotation Spreading matrix for higher order block sizes in a BSOFDM system.
The first method is based on the recursive procedure which is also applied to
the Hadamard matrix and is first presented in [15]. The second employed a
method to increase the Rotation Spreading matrix to higher order based on
CCSS proposed in [84], which is first studied and presented in [18].

These two methods proposed for achieving higher order matrices for the Ro-
tation Spreading matrix showed that this matrix still outperformed the other
spreading matrices in slow fading and selective frequency channels. This is due
to its ability to maintain its orthogonality.

Then a study of varying angles for higher order Rotation Spreading matrix is
presented and analyzed. This work is first presented in [17]. Although in [14] it
is shown that for block size of $M = 2$ the angles $\alpha = \frac{\pi}{3}$ produced the best result
in terms of BER over UWB channels, that is no longer the case for higher order
Conclusion

Rotation Spreading matrix. It is also shown that although some angles such as \( \alpha = \frac{\pi}{4}, \frac{\pi}{2} \) and \( \pi \) for block size of \( M = 2 \) were useless due to them simply rotating the modulation scheme back onto itself in the case of QPSK modulation and did not increase the correlation between the transmitted signals, these same angles for higher order Rotation Spreading matrix were shown to not only increase the correlation but in some cases such as the angle \( \alpha = \frac{\pi}{4} \) outperformed other angles across UWB channels.

It can be said that the previous limitation of some angles no longer hold for larger block sizes of \( M \) for higher order Rotation Spreading matrix. It can be stated that the advantages of the Rotation Spreading matrix is its flexibility in determining different structures of matrices and the angle rotation allows an improvement to take place over more traditional spreading matrices.

It can be stated that the angles which perform the best across UWB channel models for block size \( M = 4 \) to \( M = 16 \) are the angles \( \frac{\pi}{5}, \frac{\pi}{6}, \frac{3\pi}{4} \) and \( \frac{\pi}{4} \).

Time diversity is proposed as another method to improve the system performance of BSOFDM in Chapter Six. This chapter proposes a new system called delayed Block Spread OFDM (d-BSOFDM) to improve on BSOFDM systems not performing well under flat fading conditions when employing spreading matrices such as the Hadamard or the Rotated Hadamard. In fact, no system improvement is shown.

Delayed BSOFDM showed that by using time diversity the system did improve in terms of BER gain. It showed that the Hadamard matrix can still be used in such an environment. The Delayed BSOFDM, delays the block into another OFDM packet and this intern will allow the \( M \) blocks to encounter different uncorrelated channels therefore improving the overall system performance. This new system is first proposed in [10] and [12].

Chapter Seven presents a new approach to Block Spread OFDM called Parallel Concatenated Spreading Matrices OFDM (PCSM-OFDM) which employs coding gain. This is first presented in [13] and in the book chapter [23]. The data is split into two or more streams of \( M \) sized blocks at the transmitter side of the
system and these streams are spread using the same unitary spreading matrix discussed above. Then these streams are multiplexed and the normal procedure is carried out in BSOFDM and OFDM systems. At the receiver after channel equalization the signal is de-multiplexed and combination using Maximum Ratio Combining (MRC) or Equal Gain Combining (EGC). The de-spreading using the inverse of the unitary matrix is then carried out.

From the simulation results presented, the PCSM-OFDM outperforms the BSOFDM by greater than $4dB$ in slow fading channels and over $3dB$ in frequency selective channels such as UWB channels. Other combinations of PCSM-OFDM included an interleaver between the two streams at the transmitter and carrying out the de-spreading before the two streams at the receiver are combined. Both showed no improvement on the current setup, but the de-spreading after the combining would reduce complexity as you are only required to use one unitary matrix inverse. Overall, this system is recommended for wireless communication systems were OFDM and BSOFDM are used.

Then it presents a continuation study of the new approach Parallel Concatenated Spreading Matrices OFDM (PCSM-OFDM) first presented in [20]. At the transmitter the data is split into $n$ streams of M sized blocks and these $n$ streams are spread using the same unitary spreading matrix such as the Hadamard or the Rotation Spreading matrix. Then each of the $n$ streams are multiplexed and the normal procedure is carried out in BSOFDM and OFDM systems. At the receiver after channel equalization the signal is de-multiplexed and combination using Maximum Ratio Combining (MRC) or Equal Gain Combining (EGC) is done before the de-spreading using the inverse of the unitary matrix.

From the simulation results presented, the higher order PCSM-OFDM system shows that with the increase of $n$ streams the system gained an extra $1dB$. This then is up to the designer of the system to choose if extra complexity in the system is required for these improvements. Overall, this system is recommended for wireless communication systems were OFDM and BSOFDM are used.
Bibliography


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