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A Thesis

entitled

Performance Analysis of Orthogonal Frequency Division Multiplexing (OFDM) and Bandwidth Extension using Carrier Aggregation (CA)

by

Sandesh Modhe

Submitted to the Graduate Faculty as partial fulfillment of the requirements for the Master of Science Degree in Electrical Engineering

Dr. Junghwan Kim, Committee Chair

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The University of Toledo

May 2015
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April 2015

This thesis deals with the performance analysis of an Orthogonal Frequency Division Multiplexing (OFDM) system and the bandwidth extension achieved with carrier aggregation (CA). Data, when transmitted through any communications system, primarily faces three types of problems, errors, power management, and bandwidth. For the purpose of analyzing OFDM performance, we have taken a two-fold approach. Different types of modulation schemes such as QPSK and 16-QAM with OFDM are studied for their Bit Error Rate (BER). A noisy channel adversely affects data transmission. For this work, we study the modulation schemes with OFDM over two types of noise channels, Additive White Gaussian Noise (AWGN) and Rayleigh Fading. BER rates for both these channels are studied comparatively. One of the major drawbacks of using an OFDM communications system is its requirement for high power transmitters due to high Peak to Average Power (PAPR) ratios. In this thesis, we reduce the PAPR by using two distinct algorithms, the Partial Transmit Sequence (PTS) and Selective
Mapping Technique (SLM). Performance of the OFDM signal by employing both these algorithms is studied.

Furthermore, the last part of this work consists of bandwidth extension by using the technique of carrier aggregation. In this technique, we aggregate two sub-carriers to provide a greater bandwidth. CA is achieved by using two methods, contiguous CA and non-contiguous CA.
This is dedicated to my late grandfathers, Shri. Bhanudas Gore and Shri. Ramkrishna Modhe.
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Chapter 1

Introduction

The last couple of decades have seen an incredible rise in the number of mobile users in the world market. Due to the intense competition in this field as well as the developing technology, user requirements have now been extended to not just voice calls but also data, video, and Internet over their mobile devices. Multimedia communications is one of the most demanding areas of study, with rapidly developing arenas, and an ever-increasing user base. Hence, it has become essential to consider this widespread use of developing technologies to design systems that provide more bandwidth with better error managing capabilities for VOIP purposes. From the 2G technology in the early 1990s, mobile communications are now essentially digital cellular and include much advanced forms of multimedia capable systems such as 3G, LTE, LTE Advanced, 4G, etc.

1.1 Evolution of Mobile Communications Technology

AT&T acquired the license to operate the first commercial mobile telephone service in 1946 from the US Federal Communications Commission (FCC) [1]. Ever since then, the evolution of mobile communications technology can broadly be classified into generations of advancements such as 1G to 4G. Although the first and the second generations of standards particularly deal with voice transmission, 2.5G Enhanced Data
for GSM Evolution (EDGE) provided users with data capability but with a smaller bandwidth of 144 Kbps that was to be improved by subsequent 3G and 4G developments. Stages of development of the mobile communications systems and their salient features are as shown in Figure 1.1.

**Figure 1.1 Evolution of mobile communication systems**

1.2 Evolution of LTE Advanced

LTE Advanced is a part of the fourth generation (4G) of mobile communication technology. However, the technologies that led to the development of LTE Advanced include WCDMA, WiMAX, and CDMA2000 [2].

1.2.1 Wideband Code Division Multiple Access (W-CDMA)

W-CDMA has a hierarchical network architecture that takes full advantage of the several nodes and interfaces available for transmission and reception [3]. NodeB in Figure 1.2 is the logical node that deals with error-correcting mechanisms, modulation, spreading, and conversion from baseband to radio-frequency. This node is also responsible for transmission and reception for several different cells. In other words,
NodeB can also be called as a base station for these transmitting cells because the user device communicates with this node exclusively.

![W-CDMA Network Architecture](image)

**Figure 1.2 W-CDMA Network Architecture [2]**

As shown in Figure 1.2, several NodeBs are connected and controlled by the Radio Network Controller (RNC). The number of NodeBs for each RNC depends on the usability of the cells and can vary between a few to hundreds. Thus, the RNC is responsible for quality of service, and managing a call setup. It uses the ARQ protocol for error correcting and handling re-transmissions. The top-most layer of W-CDMA is a core network of either PSTN or the Internet.
The High-Speed Downlink Packet Access (HSDPA) can be termed as an advancement over W-CDMA technology because it introduces a new MAC sub-layer in NodeB, also known as the MAC-hs. This sub-layer schedules, rate controls, and operates a hybrid ARQ protocol thus affecting the overall operation of NodeB [4]. The overall hierarchical architecture remains the same as that of W-CDMA.

1.2.2 Worldwide Interoperability for Microwave Access (WiMAX)

Mobile WiMAX is an OFDM system that uses single carrier and different bandwidths for uplink and downlink [5]. Just like the W-CDMA/HSDPA architecture, WiMAX also has a network that is connected to the Internet via a hub, base stations, and finally the user equipment. Also known as the IEEE 802.16, it uses OFDM technique for robust coverage in a hostile wireless environment. WiMAX offers interoperability, cost efficiency, wider range of operations, and an improved OFDMA performance.

Figure 1.3 Block Diagram of a WiMAX system

![Figure 1.3 Block Diagram of a WiMAX system](image)

Figure 1.3 shows the WiMAX block diagram. It consists of a serial data input, serial to parallel converter, sub-carrier mapping block, IFFT processor, cyclic prefix inserter, RF Band Pass Processor, Frame Synchronizer, and Cyclic Prefix Inserter.
inserter, frame synchronizer, and an RF band pass processor at the transmitter end. It uses a Time Division Duplex (TDD) frame structure for OFDMA transmission and reception. The serial data input from the user is converted to a parallel output and then it is mapped according to the corresponding sub-carriers. Block interleaving happens at the sub-carrier mapping level where a block size equal to the encoded block size in bits is interleaved with the parallel data stream. Interleaving being a two-step process, the first step ensures that the adjacent coded bits are mapped onto non-adjacent sub-carriers. Mapping of these adjacent coded bits onto the constellation map takes place in the second step. Various types of modulation mechanisms such as QPSK, 16-QAM, and 64-QAM can be used for this purpose. The number of coded bits per sub-carrier depends on the choice of modulation technique. Allocation of logical data clusters happens in the FFT block. Frame synchronization and RF band pass filtering are ways to noise parameters.

1.2.3 Code Division Multiple Access 2000 (CDMA2000)

CDMA 2000 is a family of mobile communications standards that belong to the third generation of wireless technology protocols. It uses the frequency hopping or time hopping CDMA mechanisms to transmit and receive signals on user devices. Standards that are commonly included in this family are CDMA2000 1x, CDMA2000 1xEV-DO (revision 0 to B), CDMA2000 1xEV-DO (revision C also known as Ultra Mobile Broadband UMB), and CDMA2000 1xEVDV [6]. Thus, the CDMA2000 family of network protocols represents an evolutionary step-by-step advancement in OFDM-based transmission and reception methods that have now being developed into much more futuristic technologies with greater bandwidth and better user features. One of the reasons
why CDMA is so vastly successful in the mobile communications field is that the OFDM-based architecture provides excellent connectivity, both capacity-wise and quality-wise, with good efficiency, and reliability.

Figure 1.4 Block Diagram of FH-CDMA Transmitter and Receiver [6].

As shown in Figure 1.4, the FH-CDMA block diagram consists of basic elements such as the baseband modulator, up-converter, frequency synthesizer, and code generator. Since FH-CDMA uses the frequency hopping method, the user has access to different frequency resulting in an increased power of the desired signal. This signal is modulated and transmitted on several carrier frequencies giving rise to different multipath effects for different frequencies. At the receiver, the signals received at different frequencies are averaged to reduce multipath interference. Thus, as the transmitter “hops” between available frequency bands according to a pre-specified algorithm, the receiver is tuned to the same frequency for a successful reception. One of the advantages of using this
technique is that it allows for a larger synchronization error as compared to the DS-CDMA system. This also leads to a much better network coverage and signal strength.

Figure 1.5 Block Diagram of TH-CDMA Transmitter and Receiver [6]

Figure 1.5 shows the TH-CDMA block diagram. It consists of the data buffer, code generator, data modulator, and carrier generator for modulation. Time hopping involves the transmission of data signal in rapid bursts at pre-determined time intervals. These time intervals are acquired from the code assigned to each user device. Since the transmissions are all in one frequency band, probability of signal arriving at the same time is low. However, error-correcting codes are employed to compensate for any such overlapping timed transmission. It is also simpler to implement than the FH-CDMA.

The CDMA 1x family consists of protocols such as the 1xRTT and 3xRTT, where RTT stands for radio transmission technologies. This is the original form of IS-95 standards that formed the basis of 3G communications and later evolved into EV-DO and EV-DV. I-95 standards permits a usage of 1.25 MHz of bandwidth while the standards later, including the 3xRTT, can go up to 3 times the standard 1.25 MHz. Thus, the spreading rate of the transmission signal also varies with the usage of bandwidth and
sequent evolution of standards. For example, the 1xRTT has a spreading width of 1.2288 Mcps and the 3xRTT has a spreading width of 3.6864 Mcps [7]. One of the defining changes of 1xRTT is that it uses an increased Walsh code capability, which is 128 bits from the previous 64 bits. This enables it to employ more error-coding functions as well as turbo codes. 1xRTT’s forward link uses QPSK modulation while Orthogonal Complex Quadrature Phase Shift Keying (OCQPSK) may also be used in more advanced version. Due to the modulation techniques involved, the complexity of CDMA 2000 1x family increased. However, the modulation had fewer zero cross-overs, which led to the user device running a better and more efficient battery usage.

CDMA2000 1xEV stands for evolved CDMA 2000 and is the next generation of 1xRTT/3xRTT. There are two forms of the 1xEV system and they are the 1xEV-DO and 1xEV-DV. Both of these are backwards compatible with 2G systems as well as previous versions of CDMA2000. EV-DO stands for Data Only and is capable of transmitting multimedia services while EV-DV stands for data and video capable of transmitting both data and an integrated voice at peak rates of 3.09 Mbps per user [8]. The CDMA2000 1xEV-DV system were never deployed since the DO systems were already used for most practical purposes.

All CDMA2000 systems are backward compatible, which gives them an added advantage of interoperability. Backwards compatibility, however, requires that the network architecture of all these systems is almost identical while the block elements change with channel and bandwidth allocations.
As shown in Figure 1.6, the CDMA 2000 network architecture consists of a PSTN or an Internet connected to MSC and PDSN respectively. These are in turn connected to the Base Station Controller (BSC) that controls the base station. The user devices only deal with the base stations. Thus, representing a family of IMT-2000 (3G) technologies [9], the CDMA2000 1x comprises of timely advancement of CDMA technology to provide voice as well as multimedia data capabilities. It provides flexibility (with frequency hopping and time hopping), affordability (because of backwards compatibility), and modular designs for communications systems.

In summary, the major advancements in mobile broadband technologies are as follows:

1. 2G systems were developed and implemented in the early 1990s that comprised of only voice-based communications at 64 Kbps.

2. Further advancements led to the development of GPRS/EDGE at 144 Kbps.
3. Better data capability with a much needed higher bandwidth for multimedia communications was only achieved with the development of 3G and 4G standards.

4. WiMAX/HSDPA (IEEE802.16e) and CDMA2000 standards gave the user the ability to transmit and receive voice as well as multimedia data.

5. Each of these advancements, however, relies on the same hierarchical architecture that comprises of mobile equipment, base station, a base station controller, and a PSTN or Internet network.

6. Integrated high-quality audio, video, and data can be transmitted and received in technologies after the 3G developments in the early 2000s.

7. Backwards compatibility and interoperability are some of the prominent features that are being sought after while the evolution of mobile technologies continues.

8. More bandwidth and better error coding mechanisms are required for faster data communications.

9. Intense competition in the field of emerging mobile technologies has led to faster evolution of standards and better service for users.

10. CDMA, due to its incredible flexibility and network capabilities, remains the favorite technology for implementing newer versions of existing standards.

11. A gradual shift can be seen from voice-based communications to multimedia-based Internet communications with the evolving mobile technology.

1.3 Objectives of the Research

The objectives of our research are:
1. Study and analyze performance analysis of OFDM with regards to its Bit Error Rate (BER) and Peak to Average Power Ratio (PAPR).

2. Implement PAPR reduction algorithms and analyze the results.

3. Achieve bandwidth extension using Carrier Aggregation (CA) of our OFDM signal

4. Study and analyze performance analysis of the carrier aggregated signal with regards to its BER and PAPR.

5. Implement PAPR reduction algorithms for carrier aggregation.

1.4 Outline of Thesis

In Chapter 1, we introduce the topic of interest and mainly discuss on the evolution of mobile technologies with a special emphasis on data and multimedia services. It also briefly outlines the basic concepts of CDMA technologies with time hopping and frequency hopping mechanisms. Before advancing to a recently developed LTE Advanced, it is essential to understand the pre-existing as well as bygone standards of communications. Chapter 2 is devoted to the implementation of OFDM-based LTE systems. We will review and discuss the coding, modulating, and transmission techniques involved in LTE communications. This would include a thorough analysis of the LTE transmitter as well as the receiver with a suitable channel coding. Chapter 3 deals with the performance evaluation of a basic OFDM system with the specific modulation schemes and error control coding. Chapter 4 is on the carrier aggregation techniques, the most important component of an LTE-Advanced system and its simulation. Chapter 5 is devoted to the simulation results and discussions. Chapter 6 discusses future work in the field of LTE-Advanced release 10.
Chapter 2

Implementation of OFDM-Based LTE Advanced

In this Chapter, we discuss on an OFDM transmitter and receiver chain with a channel model and modulation schemes. Two different but typical digital modulation schemes are selected for this thesis, QPSK and 16 QAM, as the core element of the transmitter-receiver chain. Furthermore, we also discuss the different blocks in the OFDM transmitter as well as the receiver with the OFDM frame structure.

2.1 Orthogonal Frequency Division Multiplexing (OFDM)

An Orthogonal Frequency-Division Multiplexing (OFDM) system has the capability of using hundreds of parallel narrow-band sub-carriers of a single wide-band carrier to transmit and receiver data. It is similar to the traditional multi-carrier transmission in which a few sub-carriers can be used to transmit information over a relatively wide bandwidth. The only difference between this straightforward multi-carrier system (such as HSPA) and OFDM is that the later has the capacity to transmit several hundred sub-carriers to the receiver.

Rectangular pulse shaping with $T_m = 1/f$, where $f$ is the frequency, is employed for sub-carrier spectrum [10]. $T_m$ corresponds to the per-sub-carrier modulation symbol time. Due to its ability to deal with wide-band carriers, OFDM offers a desirable advantage over other systems when it comes to narrow-band interference. Since
sidebands are orthogonal to each other, they can be received and decoded at the receiver end without any carrier overlapped interferences, thus eliminating the need for a guard band. A typical OFDM signal with sub-carrier spacing is shown in Figure 2.1.

![Figure 2.1 OFDM Sub-carrier Spacing](image)

A basic OFDM signal $x(t)$ during time interval $mT_m \leq t \leq (m + 1)T_m$, where $T_m$ is the per sub-carrier modulation symbol time, can be expressed by the equation [8]:

$$x(t) = \sum_{k=0}^{N_c-1} x_k(t) = \sum_{k=0}^{N_c-1} a_k^{(m)} e^{j2\pi k\Delta f t}$$

where,

- $x_k(t)$ is the $k^{th}$ modulated sub-carrier with frequency $f_k = k. \Delta f$,
- $a_k^{(m)}$ is the modulation signal applied to $k^{th}$ sub-carrier during the $m^{th}$ OFDM symbol interval, and
- $N_c$ is the number of transmission symbols transmitted in parallel.

However, if mutual orthogonality is considered between two adjacent signals, $x_{k_1}$ and $x_{k_2}$, the OFDM signal can be represented by the following equation [11]:
\[
\int_{mT_m}^{(m+1)T_m} x_{k_1}(t)x_{k_2}^*(t)\,dt = \int_{mT_m}^{(m+1)T_m} a_{k_1}a_{k_2}^* e^{j2\pi k_1\Delta f t} e^{-j2\pi k_2\Delta f t} \,dt = 0 \text{ for } k_1 \neq k_2
\]

Therefore, if OFDM signal is expressed as a function of orthogonal functions \(\varphi_k(t)\), it will be

\[
\varphi_k(t) = \begin{cases} 
eq 0 & \text{if } 0 \leq t \leq T_m \\ 0 & \text{otherwise} \end{cases}
\]

(3)

2.2 OFDM Transmitter-Receiver Model

For the purpose of this study, we have divided the OFDM transceiver model into three different components, OFDM Transmitter, OFDM receiver, and the channel that induces noise (AWGN) and Rayleigh fading into the transmitted signal. Both the transmitter as well as the receiver are designed in such a way that they are specified for three major parameters. These are bandwidth, sub-carrier spacing, and total number of sub-carriers. In fact, for the design of any OFDM system, these three parameters are of utmost importance.

2.2.1 OFDM Transmitter

To ensure frequency multiplexing and orthogonality, OFDM uses a range of multi-access technologies. Symbols are modulated with chosen modulation schemes and coded for the appropriate channel conditions. It also provides for high spectral efficiency at a Nyquist symbol rate and a suitable baseband signal. Orthogonality gives it an ability to use the whole spectrum and still maintain enough carrier spacing to avoid inter-carrier interference. One of the major advantages of using OFDM is that it has high spectral efficiency. In other words, its large bandwidth can be used to transmit and receive data.
that requires higher speeds such as multimedia, video, audio, and so on. On the other hand, due to its usage of several sub channels, the channel equalization is much simpler as compared to earlier generations of communications system like CDMA. Because of multiplexing, OFDM has the capability to resist selective fading channels. This happens due to the division of data and information of the overall channel over multiple narrowband signals.

Figure 2.2 shows the OFDM transmitter consisting of the random bit generator, pilot bit generator, modulators, MUX, sub-carrier mapper, IDFT block, Cyclic Prefix Adder (CP), and the transmitting filter. OFDM specifications are based on its number of symbols, chosen modulation scheme, and IFFT size.
Random bits (in the form of 0s and 1s) are generated according to the input values and the modulation used. The physical layer of an OFDM symbol in time domain consists of time intervals expressed as the multiples of \( T_s = 1/30720000 \) sec. [12]. The radio frame of an OFDM signal, however, has a \( 10 \) ms length \( (T_{frame} = 307200 \times T_s) \). Furthermore each frame is divided into ten equally sized sub-frames of \( 1 \) ms in length each as shown in Figure 2.3. Since scheduling is done on a sub-frame basis for both downlink and uplink, each of this sub-frame in turn consists of 2 frames making it 20 slots for each frame.

![Figure 2.3 OFDM Frame Structure](image)

Considering 2000 symbols in each slot, the number of symbols in each frame comes out to be \( 2000 \times 20 = 40,000 \) symbols. Thus, the random binary values are predetermined according to the number of symbols present in a frame. For QPSK, it generates \( 1 \times 80,000 \) bits, while for 16 QAM, it generates \( 1 \times 160,000 \) bits.
For successful transmission of OFDM symbols with minimal errors, pilot sequences are of utmost importance. The pilot sequence generator ensures that orthogonality is maintained between every adjacent channels and symbols alike. For this purpose, the pilot sequences are generated to be orthogonal in either time, frequency, or code. This happens by using the Cell ID, slot number, and symbol number (which can be updated for every symbol). Since pilot bits are reference for channel estimation, synchronization is of paramount important. To achieve this, Gold-sequence generator [13] is employed to generate a pilot as shown in Figure 2.4.

![Gold Sequence Generator](image)

**Figure 2.4 Gold Sequence Generator**

Modulation for an OFDM symbol can either be QPSK or 16 QAM. In Quadrature Phase Shift Keying (QPSK), the phase of the symbol is modulated between 4 values based on the input bit combination. Thus, mapping is done as shown in Table 2.1.
Table 2.1: QPSK Symbol Mapping

In Table 2.1, $I$ and $Q$ represent channels for odd and even bits respectively. Based on the values of the bits and their corresponding $I$ and $Q$ equivalents, an appropriate mapping is performed. A complex modulated symbol range of the form $I + j*Q$ is obtained since 80,000 data bits are being transmitted in QPSK out of which the function returns a $1*40,000$ row vector [14].

Thus, for QPSK modulated data, according to Table 2.1 of QPSK symbol mapping, the modulated data will be as shown in Figure 2.5. As can be seen, there are 20 slots for each frame of the OFDM signal. Each frame corresponds to 10 ms of time, thus making the each slot 0.5 sec. in duration.

<table>
<thead>
<tr>
<th>$b(i), b(i+1)$</th>
<th>$I$</th>
<th>$Q$</th>
</tr>
</thead>
<tbody>
<tr>
<td>00</td>
<td>$1/\sqrt{2}$</td>
<td>$1/\sqrt{2}$</td>
</tr>
<tr>
<td>01</td>
<td>$1/\sqrt{2}$</td>
<td>$-1/\sqrt{2}$</td>
</tr>
<tr>
<td>10</td>
<td>$-1/\sqrt{2}$</td>
<td>$1/\sqrt{2}$</td>
</tr>
<tr>
<td>11</td>
<td>$-1/\sqrt{2}$</td>
<td>$-1/\sqrt{2}$</td>
</tr>
</tbody>
</table>

Figure 2.5 Real and Imaginary Parts of QPSK Modulated Data
Similarly, the modulation order of a 16 QAM modulator is 4. We have chosen 16-QAM for this study because it is one of the most commonly used modulation schemes. QAM uses both amplitude as well as phase variations for modulation. In this case, real and imaginary parts of I and Q are interleaved thus mapping symbol values and complex symbol values. In case of 16 QAM, bits exist in quadruplets such as \( b(i), b(i+1), b(i+1), \) and \( b(i+3) \) [13]. Each of these quadruplets are interleaved with modulation symbols with complex values \( X = I + jQ \) as shown in Table 2.2.

<table>
<thead>
<tr>
<th>( b(i), b(i+1), b(i+1), b(i+3) )</th>
<th>( I )</th>
<th>( Q )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0000</td>
<td>( 1/\sqrt{10} )</td>
<td>( 1/\sqrt{10} )</td>
</tr>
<tr>
<td>0001</td>
<td>( 1/\sqrt{10} )</td>
<td>( 3/\sqrt{10} )</td>
</tr>
<tr>
<td>0010</td>
<td>( 3/\sqrt{10} )</td>
<td>( 1/\sqrt{10} )</td>
</tr>
<tr>
<td>0011</td>
<td>( 3/\sqrt{10} )</td>
<td>( 3/\sqrt{10} )</td>
</tr>
<tr>
<td>0100</td>
<td>( 1/\sqrt{10} )</td>
<td>( -1/\sqrt{10} )</td>
</tr>
<tr>
<td>0101</td>
<td>( 1/\sqrt{10} )</td>
<td>( -3/\sqrt{10} )</td>
</tr>
<tr>
<td>0110</td>
<td>( 3/\sqrt{10} )</td>
<td>( -1/\sqrt{10} )</td>
</tr>
<tr>
<td>0111</td>
<td>( 3/\sqrt{10} )</td>
<td>( -3/\sqrt{10} )</td>
</tr>
<tr>
<td>1000</td>
<td>( -1/\sqrt{10} )</td>
<td>( 1/\sqrt{10} )</td>
</tr>
<tr>
<td>1001</td>
<td>( -1/\sqrt{10} )</td>
<td>( 3/\sqrt{10} )</td>
</tr>
<tr>
<td>1010</td>
<td>( -1/\sqrt{10} )</td>
<td>( 1/\sqrt{10} )</td>
</tr>
<tr>
<td>1011</td>
<td>( -3/\sqrt{10} )</td>
<td>( 3/\sqrt{10} )</td>
</tr>
<tr>
<td>1100</td>
<td>( -1/\sqrt{10} )</td>
<td>( -1/\sqrt{10} )</td>
</tr>
<tr>
<td>1101</td>
<td>( -1/\sqrt{10} )</td>
<td>( -3/\sqrt{10} )</td>
</tr>
<tr>
<td>1110</td>
<td>( -3/\sqrt{10} )</td>
<td>( -1/\sqrt{10} )</td>
</tr>
<tr>
<td>1111</td>
<td>( -3/\sqrt{10} )</td>
<td>( -3/\sqrt{10} )</td>
</tr>
</tbody>
</table>

Table 2.2 16 QAM Symbol Mapping
Table 2.2 shows the 16-QAM symbol mapping. As it can be seen, the signal is mapped over a series of imaginary as well as real data points. The signal constellation of QPSK and 16-QAM is as shown in Figure 2.6. Both the modulated pilot sequence obtained from the previous step and the modulated data is then put into the OFDM frame by the multiplexer (MUX).

![Figure 2.6: Signal Constellation for (a) QPSK and (b) 16-QAM](image)

As shown in Figures 2.6 (a) and (b), QPSK and 16-QAM signals are non-zero independent variables where plotted values can be found in Table 2.1 and Table 2.2 respectively.

![Figure 2.7: OFDM Modulation](image)
Thus, as shown in Figure 2.7, the principles of the OFDM modulation block comprises of two major components that is the serial to parallel converter and the multiplexer. Serial to parallel converter is used to convert the serial data obtained from the random bit generator and pilot sequence generator to a parallel symbols. Then, the parallel data symbols are separated into adjacent data units, modulating them bit by bit according to a pre-chosen QPSK or 16 QAM modulation scheme. Here, $a_k^m$ is the modulation signal applied to the $k^{th}$ sub-carrier during the $m^{th}$ OFDM symbol interval and $a_k(t)$ is the resulting output of the OFDM modulation block.

The next step in this process is that of sub-carrier mapping where the OFDM frame is mapped onto several hundred sub-carriers [15]. Half of these sub-carriers are used for mapping the positive set while the rest half is for mapping negative set. An OFDM signal has sub-carrier spacing of $\Delta f$ obtained from the inverse of the per-sub-carrier symbol rate $1/T_m$ leading to a less complicated structure. Due to this, the modulation scheme can be effortlessly implemented by performing a Fast Fourier Transform (FFT) or a Discrete Fourier Transform (DFT) on the signal.

![Diagram](image)

**Figure 2.8: OFDM modulation with IDFT processing**
Let’s consider a sampling rate $f_s$, where $f_s = 1/T_s = N \cdot \Delta f$, where $N$ is the number of symbols. In other words, we are assuming that the sampling rate is equal to an $N^{th}$ multiple of the sub-carrier spacing $\Delta f$ and $N_c$ is the number of OFDM sub-carriers. For sampling theorem to be obeyed for an OFDM bandwidth of $N_c \cdot \Delta f$, $N > N_c$ [16]. Thus, $N$ symbols have to be chosen in such a way that sampling theorem is satisfied.

With the IDFT processing and a chosen modulation technique according to Figure 2.8, the OFDM signal is written as:

$$x_n = x(nT_s) = \sum_{k=0}^{N_c-1} a_k e^{j2\pi kn\Delta f/N} = \sum_{k=0}^{N_c-1} a_k e^{j2\pi kn/N} = \sum_{k=0}^{N-1} a'_k e^{j2\pi kn/N}$$  \hspace{1cm} (4)

Thus, $N$ exceeds $N_c$ by a sufficient margin to fulfill sampling theorem. Also,

$$a'_k = \begin{cases} a_k & 0 \leq k \leq N_c \\ 0 & N_c \leq k < N \end{cases}$$  \hspace{1cm} (5)

In other words, if the modulation symbols are $a_0, a_1, \ldots, a_{N_c-1}$ with length 0 to $N$, the size $N$ IDFT of the block will give an output sequence of $x_n$. Figure 2.7 shows that the OFDM modulation with an IDFT component involves a serial to parallel conversion, followed by IDFT processing, and then back from parallel to serial conversion. The IFFT output of the system in time domain is shown in Figure 2.9 while the output in frequency domain is shown in Figure 2.10:
As seen in Figures 2.9 and 2.10, a total number of 140 symbols are obtained both in frequency as well as time domains. Each of these symbols contains 512 sub-carriers, thus making the dimension of the IFFT matrix 140 x 512. The matrix thus generated is for discrete values at $t = nT$ (where $t$ is the time period, $n$ is the number of symbols, and
$T$ is the time period corresponding to each symbol) and $f_o = \frac{1}{NT}$ (where $f_o$ is the output frequency). Figures 2.9 and 2.10 are for QPSK modulated signal. However, the IFFT blocks gives an output of a standard 140 symbol for 16-QAM modulated signal too.

Even with the modulation schemes employed and the IDFT processing thereby undertaken, there is still a chance of interference between adjacent OFDM symbol blocks. Although orthogonality is ensured between these blocks, the orthogonality is dependent on the specific frequency domain structure of each sub-carrier. Thus, if the side lobes of a sub-carrier are corrupted due to channel interference or any other kind of noise, the signal loses its orthogonality partially or completely. The extent of damage caused depends on specific instances and noise models. Because of the large size of sub-carrier side-lobes, an additional tool has to be used to ensure that orthogonality will be maintained and interference will be rejected in adjacent symbol blocks. This tool is known as the cyclic prefix (CP) inserter.

![Cyclic Prefix Insertion Diagram](image)

**Figure 2.11: Cyclic prefix insertion**

In Figure 2.11, $T_c$ is the time period of each carrier symbol while $T_{cp}$ is the cyclic prefix added to ensure orthogonality. Insertion of a cyclic prefix after the IDFT processing stage makes the OFDM signal insensitive to time dispersion and hence the interference arising due to it. As shown in Figure 2.11, cyclic prefix insertion involves
the last part of the OFDM signal that is copied and inserted to the beginning of the same signal. In a cyclic pattern, it forms a chain of block symbols that are guarded on both the sides with the same symbol. Due to the wide bandwidth of OFDM, this practice is widely adopted to get rid of interference.

Consider the last samples of IDFT block output \( N_{cp} \), where \( N \) is the total block size of IDFT’s output. \( N_{cp} \) is then copied and inserted to the beginning of the same symbol, increasing the size of this symbol to \( N+N_{cp} \). The cyclic prefix adder’s input comprises of the 140 x 512 matrix from the IFFT block. It appends the last 36 elements of each symbol for each slot to the start of each symbol. Thus, the cyclic prefix adder output becomes a matrix of 140 x 548 as shown in Figures 2.12 and 2.13, in which each color donates a sub-carrier.

![Figure 2.12 Cyclic Prefix Adder Output in Time Domain](image)

Figure 2.12 Cyclic Prefix Adder Output in Time Domain
2.2.2 OFDM Channel:

A general time domain linearly equalized OFDM channel contains a channel model with an impulse response with a noise element as shown in Figure 2.14.

![Figure 2.14 OFDM Channel](image)

For the purpose of this work, we have considered two types of channels and these are Additive White Gaussian Noise (AWGN) and Rayleigh fading. AWGN is the linear addition of white noise that acts as a basic impairment to any communications channel. The channel capacity $C$ of an AWGN channel can be expressed as:
\[ C = \frac{1}{2} \log(1 + \frac{P}{n}) \]  

(6)

Where \( P \) is the power constraint and \( n \) is the variance of noise given as \( \sigma^2 \).

Effects of Rayleigh fading are observed when the radio signal is scattered before it arrives at the receiver. Scattering of the signal can happen due to many reasons, primary because of the signal passing through several objects in the environments during its journey from the transmitter antenna to the receiver antenna. Multipath interference occurs due to several reasons such as reflection from the ionosphere, atmospheric ducting, buildings in the line of sight, etc. This leads to the same signal taking different paths for its journey from transmitter to the receiver finally resulting in an interference pattern or phase changes in the transmitted signal. A mathematical model of this type of Rayleigh channel can be expressed as [17]:

\[ h(t) = \sum_{l=0}^{L_p-1} \alpha_l \delta(t - \tau_l) \]  

(7)

where,

- \( L_p \) is the number of signal paths,
- \( l \) is the delay path,
- \( \alpha_l \) is the complex value of path \( l \),
- and \( \tau_l \) is the delay of path \( l \)

For estimating Rayleigh fading, the paths are assumed to be statistically independent and power \( P \) is normalized. Equation (7) gives both the interference as well as phase shifting of the signal when it is subjected to a fading model. Since fading considers numerous factors for its interference model, an average of delay paths is needed for a proper estimation.
2.2.3 OFDM Receiver

An OFDM receiver consists of a receiver filter, CP removal block, DFT, sub-carrier demapper, DE-MUX, channel estimator, modulator, pilot bit generator, equalizer, frequency response, and slicer as shown in Figure 2.14. The receiver filter is similar to the one at the transmitter end, which is a band pass filter. It filters the incoming signal by convoluting it with already established filter coefficients and gives out a cleaner output.

![Figure 2.15 OFDM Receiver](image)

As shown in Figure 2.15, to revert the effect of up-sampler used in the transmitter, the receiver has a down-sampler that samples down the incoming signal passed through the filter through the decimation factor of 4. Thus, the original frame dimensions of the frame are achieved once the received signal passes through the down-sampler. Similarly, the CP removal block removes cyclic prefixes to strip the signal off of its CP modules.
attached to it at the transmitter. This is then fed to a DFT block that performs a discrete Fourier transform on the CP removal block’s output.

Although the transmitter filter and the receiver filter are used to minimize noise, the filter themselves add some effects to the signals. This, in combination with the channel noise, it gives a signal that has much more noise than what can be demodulated to get back the original data. The equalizer block is used to compensate for the effects induced by the filters as well as the channel. All the combined responses of Tx filter, Rx filter, and frequency response of channel filter are thus nullified with the equalizer.

For QPSK demodulation, complex modulated symbols are demodulated to binary data by the demodulator. However, small variations are possible at the receiver end due to noise amplitudes. To neutralize this effect, soft de-mapping has to be performed at the receiver level too. Respective quadrants and mapping values are used for QPSK demodulation as was observed in Table 2.1. For 16 QAM demodulation, the received signals are first separated into real and imaginary parts [18]. They are then related to the tabular values that were used for 16 QAM mapping according to Table 2.2. Even in 16 QAM demodulation, the effect of noise induced by channel and possibly filters has to be considered. For this purpose, original amplitude ranges of each complex symbol is calculated and based on which, the original mapped complex symbol’s value is obtained.
A graphic representation of the OFDM demodulation process is shown in Figure 2.16. Received signal $r(t)$ is weighted with the known modulation co-efficient that depends on whether it is QPSK or 16 QAM. Then the result is passed through an integrator that calculates area under curve for symbol rate $1/T_m$. This process, however, is performed after the cyclic prefixes have been removed to avoid confusion. The next logical step in this process is the re-mapping of the complex symbol to binary format so that we get the original transmitted data bits.

Figure 2.16: OFDM Demodulation
Chapter 3

OFDM Generalized System Performance

In this thesis, we use Matlab to simulate an OFDM based LTE communications system with FDD-based downlink that includes an OFDM transmitter, channel, and an OFDM receiver. We use QPSK and 16-QAM modulations and the signal will be transmitted over a Rayleigh multipath fading channel as well as an additive white Gaussian noise (AWGN) channel. A receiver block will also be simulated to recover the transmitted signal.

3.1 Simulated OFDM System Properties

This simulation will use a 5 MHz bandwidth to transmit an OFDM signal with 512 sub-carriers \(N\) and 300 active sub-carriers \(N_c\). Sub-carrier spacing \(\Delta f\) is 15 KHz. One iteration of this program transmits one frame containing 140 symbols having 512 sub-carriers each. Therefore, sampling frequency is \(15 \text{ KHz} \times 512 = 7.68 \text{ MHz}\). If we define oversampling rates as:

\[
\text{Oversampling rate} = \frac{\text{Number of bits in one frame} \times 100}{7.68 \text{ MHz} \times \text{Sampling frequency}}
\]  

(8)

We get oversampling rates for QPSK and 16-QAM modulation:

- \(\frac{140 \times 512 + 4 + 100}{7.68 \times 10^6} = 3.73 \approx 4\) for QPSK
3.2 Power Spectral Density of QPSK and 16-QAM OFDM Modulated Signals

Since QAM is also a linear modulation system that uses the same frequency, it’s baseband signal can be expressed as [20]:

\[ x(t) = \sum_{k=-\infty}^{\infty} (x_i(k) + jx_q(k)) \Pi\left(\frac{t}{T-k}\right) \]  

(9)

where \( \Pi \) is the rectangular pulse having the following values:

\[ \Pi\left(\frac{t}{T-k}\right) = \begin{cases} 1 & \text{for } 0 \leq (t/T - k) \leq 1 \\ 0 & \text{otherwise} \end{cases} \]

and \((x_i(k), x_q(k))\) is for baseband signal \(k\), that is signal constellations coordinates of time period \(T\).

The auto-correlation function of \(x(t)\) can be represented as [20]:

\[ \Phi_{xx}(t, t + \tau) = E[\langle x^*(t) x(t + \tau) \rangle] \]

\[ = E \left[ \sum_{k=-\infty}^{\infty} \sum_{l=-\infty}^{\infty} (x_i(k) - jx_q(k))(x_i(l) + jx_q(l)) \Pi(t/T - k) \Pi((t + \tau)/T - l) \right] \]

(10)

Let’s consider a general form of the coordinates of signal constellation, \(\tau = rT + \delta T\), where \(0 \leq \delta < 1\) and \(r\) is an integer.

For auto-correlation, we average over time period \(T\) to get:

\[ \overline{\Phi_{xx}}(\tau) = \frac{1}{T} \int_0^T \Phi_{xx}(t, t + \tau) \]
\[
\frac{1}{T}(E \left[ (x_t(0) - jx_q(0)) (x_t(r) + jx_q(r)) \right] \int_0^T \prod \left( \frac{t}{T} \right) \cdot \prod \left( (t + \delta T)/T - 1 \right) dt + E \left[ (x_t(0) - jx_q(0)) (x_t(r + 1) + jx_q(r + 1)) \right] \int_0^T \prod \left( \frac{t}{T} \right) \cdot \prod \left( (t + \delta T)/T - 1 \right) dt}
\]

\[= E \left[ (x_t(0) - jx_q(0)) (x_t(r) + jx_q(r)) \right] \cdot (1 - \delta)
\]

\[+ E \left[ (x_t(0) - jx_q(0)) (x_t(r + 1) + jx_q(r + 1)) \right] \cdot \delta \]

(11)

Thus, spectral density can be expressed as [20]:

\[\Phi_{xx}(\tau) = \begin{cases} E [x_t(0)^2 + x_q(0)^2] (1 - \delta) & \text{for } r = 0 \\ E [x_t(0)^2 + x_q(0)^2] \delta & \text{for } r = -1 \\ 0 & \text{otherwise} \end{cases} \]

(12)

By taking the Fourier transform of the auto-correlation functions shown above, we can get power spectral density of the QPSK as well as 16-QAM signals as follows [20]:

\[S_x(f) = \sigma^2 T \left( \frac{\sin \pi f T}{\pi f T} \right)^2 \]

(13)

Hence, the bandpass spectrum of QPSK and 16-QAM signals can be represented as [20]:

\[S_y(f) = \frac{1}{4} [S_x(f - f_c) + S_x(-f - f_c)] \]

(14)

where,

\[S_y(f) \text{ is the passband spectral density function and } S_x(f) \text{ is the baseband spectral density function. A graphical representation of the spectral density functions is as shown in Figure 3.1.} \]
Thus, the modulated signal shaping occurs, with respect to a center frequency $f_c$, to the negative and positive sides of frequency $f_c$.

### 3.3 System Performance

For an OFDM system with $N$ samples and $N_g$ guard interval samples, the total number of transmitted samples is $N + N_g$. Transmitted signal can be expressed as:

$$X = [x_{N-N_g}, \ldots, x_0, x_1, \ldots, x_{N-1}]$$

Thus, since the time invariant channel impulse response with $R$ taps is given as $h^T = [h_0 h_1 \ldots h_{R-1}]$, the received signal $z = [z_0 \ldots z_{N-1}]^T$ can be expressed as [21]:

$$z = \begin{bmatrix} 0 & \ldots & h_{R-1} & h_{R-2} & \ldots & h_0 & 0 & \ldots & 0 \\ 0 & \ldots & 0 & h_{R-1} & h_{R-2} & \ldots & h_0 & \ldots & 0 \\ \vdots & & & & & & & & & \vdots \\ 0 & \ldots & 0 & h_{R-1} & h_{R-2} & \ldots & h_0 & \ldots & \end{bmatrix} x + v$$

(16)
Where, $h^T$ is the impulse response of the channel (depending on AWGN or Rayleigh chosen for this study) = \([h_0 \ h_1 \ldots \ h_{R-1}]\), $\nu$ is the Gaussian noise, and $R \leq N_y$.

We have considered a general impulse response of the channel here but the $h$ matrix changes with the type of channel model as is discussed in the next section.

Thus, received signal $z$ can also be expressed as [21]:

$$z = \bar{Q}\bar{x} + \nu$$  \hspace{1cm} \text{(17)}

where,

$x$ is the input matrix linearly represented as $x$ previously,

$\bar{Q}$ is the circulant matrix obtained due to the convolution of the transmitted signal at the cyclic prefix adder given as [21]:

$$\bar{Q} = F^{-1}HF$$ \hspace{1cm} \text{(18)}

Where $F$ and $F^{-1}$ are the DFT and IDFT matrices obtained at the IFFT stage of the transmitter as well as the receiver. $H$ is an $N \times N$ diagonal square matrix representing the channel impulse response in frequency domain.

### 3.4 Channel Response

For the purpose of this thesis, we have considered Additive White Gaussian Noise (AWGN) and Rayleigh fading channel models. When white noise is linearly added to the transmitted signal, AWGN is observed while Rayleigh fading is observed due to the obstruction caused by several objects as well as terrain in the path of transmission.

#### 3.4.1 Additive White Gaussian Noise Channel (AWGN)

An AWGN channel can be represented in its general form as [22]:

\( y_t = x_t + w_t \) \hspace{1cm} (19)

where \( x_t \) and \( y_t \) are real input and output of the system at time \( t \) and \( w_t \) is the time-independent white Gaussian noise with mean 0 and variance \( \sigma^2 \):

\( w_t = N(0, \sigma^2) \) \hspace{1cm} (20)

Where Noise variance \( \sigma^2 = \frac{N_0}{2} \) \hspace{1cm} (21)

Power constraint \( \bar{p} \) per symbol at sampling rate \( 1/W = \frac{\bar{p}}{2W} \)

Since noise is independent of the I and Q components of both QPSK and 16-QAM, the power constraint and noise variance are \( \frac{\bar{p}}{2W} \) and \( N_0/2 \) respectively. Thus, channel capacity \( (C) \) for corresponding dimensions is expressed as [22]:

\[
C = \frac{1}{2} \log \left( 1 + \frac{\bar{p}}{N_0W} \right) \text{ bits for real part}
\]

Thus, the channel capacity of time continuous AWGN channel can be represented as:

\[
C_{AWGN}(\bar{P}, W) = W \log \left( 1 + \frac{\bar{P}}{N_0W} \right) \text{ [bits/s]}
\]

where \( W \) are the number of complex samples/second.

Since signal to noise ratio for this channel can be expressed as:

\[
SNR = \frac{\bar{P}}{N_0W},
\]

Capacity of AWGN channel can be rewritten as from Equation (22):

\[
C_{AWGN} = \log \left( 1 + SNR \right) \text{ [bits/s/Hz]}
\]

(23)

For \( N \) number of users, each with a data rates \( R_1, R_2, \ldots, R_n \), the AWGN uplink channel capacity is given as [22], from Equation (23):

\[
R_1 < \log \left( 1 + \frac{P_1}{N_0} \right)
\]

\[
R_2 < \log \left( 1 + \frac{P_2}{N_0} \right) \ldots
\]
\[ R_n < \log(1 + \frac{P_n}{N_0}) \]

and \[ R_1 + R_2 + \cdots + R_n < \log \left(1 + \frac{P_1 + P_2 + \cdots + P_n}{N_0}\right) \]

Thus, reliable communication in an AWGN channel can only be obtained when the communication rate \( R < C \), channel capacity, as defined from Shannon’s channel capacity theorem.

For the downlink AWGN channel with \( N \) users, from equation (18), we get:

\[ y_k(t) = h_k x(t) + w_k(t) \] (24)

Where, \( y_k(t) \) is the received signal at time \( t \), \( w_k \) is the Gaussian noise, and \( x(t) \) is the transmitted signal, and \( h_k \) is obtained from (15) for \( K = 1, 2, \ldots, n \) users.

### 3.4.2 Rayleigh Fading

The Rayleigh fading model is used to mathematically describe the effect of signal strength under a number of disturbances and obstruction in the line of propagation of the signal. When added to the AWGN channel noise, Rayleigh fading model gives a more realistic analysis of the trans-reception environment. As a reasonable model, for this thesis, we have chosen a frequency flat Rayleigh fading channel designed for Doppler spread of 1 KHz and time interval of 0.1 \( \mu s \). Phase change and attenuation for different sub-carrier frequencies is obtained across different times and AWGN is added. In general, received signal can be expressed as [23]:

\[ r(t) = e^{j2\pi ft}(X(t) + jY(t)) \] (25)

Where \( X(t) \) is the transmitted signal while \( Y(t) \) is the received signal. The signal is thus randomly attenuated by a density factor and its received signal envelope is given as:

\[ R = \sqrt{X(t)^2 + Y(t)^2} \]
And it is randomly shifted by a phase factor of:

\[ \Phi = \arg \{X(t) + jY(t)\} \]

Since each multipath interference involves several paths taken by the signal to reach the receiver, each path can be expressed as a complex Gaussian variable as done in Equation (25).

### 3.5 Scheduling and Link Adaptation

Several different scheduling methods can be adopted for successful transmission. Link adaptation via power and rate control is one of the most common techniques adapted to keep the transmitted signal guarded from variations due to channel conditions. The aim is to keep the \( E_b/N_o \) constant so that error probability can be kept as low as possible. Whenever the radio channel experiences poor radio conditions, the transmitter power can be bumped up to compensate for the power loss. Link adaptation, in this case, is thus achieved by making the transmitter power inversely proportional to channel conditions. A graphical representation of this type of adaptation is shown in Figure 3.2.

![Figure 3.2 Power control adaptation](image)

As shown in Figure 3.2, for power control adaptation, the data rate \( (R) \) remains constant throughout the channel, however, the transmitted power \( (P_T) \) changes according to the channel response \( (P_C) \). It is not always possible to change the transmitter power
with respect to the channel behavior. This is because many real-time systems such as the case of packet-data traffic, a constant transmission power is needed for effective transmission. Therefore, rate control proves out to be the better method for compensating for channel characteristics. For the purpose of this thesis, we are using adaptive modulation and coding (AMC). In AMC, QPSK and a low data rate is used for poor radio-link conditions while for 16-QAM, a higher code rate is chosen for better radio-link interfaces. The concept is shown in Figure 3.3.

![Figure 3.3 Data Rate Control Adaptation](image)

3.6 Comparative LTE Release Parameters

Development of LTE and LTE-advanced system with carrier aggregation was done to achieve a higher data rate than its previous versions such as CDMA 2000 and IMT protocols. For a 20 MHz spectrum allocation, LTE is capable of giving a peak data rate of 100 MB/s in the downlink and 50 MB/s in uplink. LTE release 8, however, provides an increased improvement over these standards with a 300 MB/s downlink data
rate and a 75 Mb/s uplink data rate. Difference in the downlink and uplink data rates is due to the implementation of a 8 x 8 MIMO in downlink and a 4 x 4 MIMO in uplink. Both the LTE release 8 standard and the more advanced release 10 standards allow for usage of Time Division Duplex (TDD) as well as Frequency Division Duplex (FDD) configurations. Since TDD uses the same frequency band for transmission as well as reception only at different time intervals, it is more widely used. On the other hand, the FDD topology involves using different frequency bands for transmitter and receiver paths. Comparison of LTE 8 and LTE 10 release standards can be seen in Table 3.1:

<table>
<thead>
<tr>
<th></th>
<th>ITU Requirement (MB/s)</th>
<th>Release 8 FDD (MB/s)</th>
<th>TDD (MB/s)</th>
<th>Release 10 FDD (MB/s)</th>
<th>TDD (MB/s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Downlink</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>15</td>
<td>15.3</td>
<td>15.0</td>
<td>30.6</td>
<td>30.0</td>
</tr>
<tr>
<td>Uplink</td>
<td>6.75</td>
<td>4.2</td>
<td>4.0</td>
<td>16.8</td>
<td>16.0</td>
</tr>
</tbody>
</table>

**Table 3.1: LTE 8 and LTE 10 release standards**

ITU-R also defines requirements for efficiency of radio interface for IMT-Advanced systems. Two major parameters considered for this evaluation are spectral efficiency ($\eta$) and cell-edge spectral efficiency. IMT standards permit for LTE bandwidths ranging from 1.4 MHz to 20 MHz. Due to its orthogonal properties and sub-carrier make-up, the specified number of physical layer resource blocks (sub-carrier units) for each of these frequencies with the important frame parameters is given by Table 3.2:
<table>
<thead>
<tr>
<th>Bandwidth (MHz)</th>
<th>Resource Blocks ($N_{rb}$)</th>
<th>Sub-frame Duration</th>
<th>Sub-carrier Spacing</th>
<th>Sampling Frequency (MHz)</th>
<th>FFT Size</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.4</td>
<td>6</td>
<td>0.5 ms</td>
<td>15 KHz</td>
<td>1.92</td>
<td>128</td>
</tr>
<tr>
<td>3</td>
<td>15</td>
<td>0.5 ms</td>
<td>15 KHz</td>
<td>3.84</td>
<td>256</td>
</tr>
<tr>
<td>5</td>
<td>25</td>
<td>0.5 ms</td>
<td>15 KHz</td>
<td>7.68</td>
<td>512</td>
</tr>
<tr>
<td>10</td>
<td>50</td>
<td>0.5 ms</td>
<td>15 KHz</td>
<td>15.36</td>
<td>1024</td>
</tr>
<tr>
<td>15</td>
<td>75</td>
<td>0.5 ms</td>
<td>15 KHz</td>
<td>23.04</td>
<td>1536</td>
</tr>
<tr>
<td>20</td>
<td>100</td>
<td>0.5 ms</td>
<td>15 KHz</td>
<td>30.72</td>
<td>2048</td>
</tr>
</tbody>
</table>

Table 3.2: Bandwidth to Resource Block Ratio
Chapter 4

Carrier Aggregation (CA)

4.1 Introduction to Carrier Aggregation (CA) in LTE-Advanced IMT Release 10

The essence of the development of mobile communications technology is the bandwidth allocation and efficient use of the available bandwidth. With high amount of traffic, however, limits are being enforced on bandwidth usages. On the other hand, there is a huge amount of resources that are not being used due to user and channel unpredictability. The International Telecommunication Union (ITU) developed a set of standards known as the International Mobile Telecommunications-Advanced (IMT-Advanced) to deal with this problem. One of the key components in IMT-Advanced plan of action is faster data access with efficient employment of available resources. Typically, the desire is for a user to experience peak data rates of 1 Gbps for static usage and 100 Mbps users in motion. LTE-Advanced Release 10 was thus developed by the Third Generation Partnership Project as a new version of LTE (OFDM-based communication system) in May 2008 to fulfill these requirements. With LTE-Advanced providing a high data peak rate, it helps to achieve theoretical speeds practically by making efficient use of bandwidth possible [24]. In carrier aggregation, we can aggregate upto 5 carrier to form one extended bandwidth signal. Thus, since Release 8 permits a carrier bandwidth of 20 MHz, by using carrier aggregation, an aggregated bandwidth of
100 MHz can be obtained. Carrier aggregation has already become a part of the future plans of many of the top telecommunications industry service providers.

Spectral efficiency targets for LTE, LTE-Advanced, and IMT-Advanced (4G) are given in Table 4.1 as follows [25]:

<table>
<thead>
<tr>
<th>Sub-Category</th>
<th>LTE (3.9G) Target [26]</th>
<th>LTE-Advanced (4G) Target [27]</th>
<th>IMT-Advanced (4G) Target [28]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak Spectral Efficiency (b/s/Hz)</td>
<td>Downlink 16.3 (4X4 MIMO)</td>
<td>30 (up to 8X8 MIMO)</td>
<td>15 (4X4 MIMO)</td>
</tr>
<tr>
<td></td>
<td>Uplink 4.32 (SISO)</td>
<td>15 (up to 4X4 MIMO)</td>
<td>6.75 (2X4 MIMO)</td>
</tr>
<tr>
<td>Downlink Cell Spectral Efficiency (b/s/Hz) at 3km/h, 500 m ISD</td>
<td>2X2 MIMO 1.69</td>
<td>2.4</td>
<td></td>
</tr>
<tr>
<td></td>
<td>4X2 MIMO 1.87</td>
<td>2.6</td>
<td>2.6</td>
</tr>
<tr>
<td></td>
<td>4X4 MIMO 2.67</td>
<td>3.7</td>
<td></td>
</tr>
<tr>
<td>Downlink Cell-Edge User Spectral Efficiency (b/s/Hz) for 10 users, 500m, ISD</td>
<td>2X2 MIMO 0.05</td>
<td>0.07</td>
<td></td>
</tr>
<tr>
<td></td>
<td>4X2 MIMO 0.06</td>
<td>0.09</td>
<td>0.075</td>
</tr>
<tr>
<td></td>
<td>4X4 MIMO 0.08</td>
<td>0.12</td>
<td></td>
</tr>
</tbody>
</table>

Table 4.1 Performance Targets for LTE, LTE-Advanced, and IMT-Advanced
Peak data rate is directly proportional to bandwidth of the system. In other words, the higher the bandwidth, more data can be transmitted. Carrier aggregation (CA) relies on careful attenuation of sub-carrier as well as carrier frequencies to give extra bandwidth to the user, based on availability. LTE (with the core network known as SAE) Release 8 and 9 mandate a maximum bandwidth of 20 MHz for its transmission. CA uses the same bandwidth to extend maximum usable frequency range to 100 MHz for each user terminal for a single transmission. Same bandwidths are being extended because to find a single chunk of large bandwidth is impossible to find since operators have already allocated much of their available spectrums for use of legacy systems. CA allows multiple carrier components from the same or different frequency (depending on the type of CA used) to be simultaneously aggregated or added up and hence the term “aggregation”. Mandated LTE component carrier (CC) bandwidths are 1.4, 3, 5, 10, 15, and 20 MHz as discussed in Table 3.2 referring to LTE carriers that are used in legacy LTE systems.

Carrier aggregation uses a maximum of five component carriers in varying compositions to form a single carrier thus giving it a maximum available bandwidth of $5 \times 20\text{MHz} = 100\text{MHz}$. One of the greatest advantages of using carrier aggregation is that, apart from the higher bandwidth, it assures backwards compatibility with the earlier releases 8 and 9 of LTE IMT standards. Therefore, operators can use CA for LTE-Advanced users without having to install new hardware or designing a new system altogether. Both FDD LTE and TDD LTE can be used for implanting and employing carrier aggregation technique with additional physical layer components of Release 8 and
[25]. The best part of CA is that the spectrum need not be continuous thus providing interoperability between different carrier frequencies.

Some other advantages of carrier aggregation include its MIMO friendliness, delay spread, fading capabilities, and possibility frequency scheduling gain. Hybrid Automatic Repeat Request (HARQ) remains asynchronous for downlink and synchronous for uplink given its increased complexity for asynchronous transmission [26]. Thus, data aggregation in CA with HARQ, physical layer and component carriers (CCs) can be shown as Figure 4.1.

![Figure 4.1: CA Data Aggregation Overview](image)

As shown in Figure 4.1, data aggregation in CA happens in the MAC layer with scheduling techniques that will be discussed later in this chapter. Once data aggregation has successfully been done, the HARQ protocol is used to provide an acknowledgment for data transfer. Physical layer of the CA system takes care of the mapping of individual component carriers (CCs).
Efficient usage of carrier aggregation depends on two key factors. First is the availability of component carriers, whether adjacent (also known as contiguous) or within the same band or different band (known as non-contiguous). Secondly, the User Equipment (UE) should be capable of handling a number of transceivers and transmit through multiple carriers. Thus, the factors that truly affect UE’s ability to implement carrier aggregation are cost, power, and hardware complexity to install multiple transceivers (especially in the case of non-contiguous spectrums) [30].

4.2 Carrier Aggregation Types

Based on the positions and orthogonality of the sub-carrier and/or the carrier frequencies used for carrier aggregation purposes, three different types of CA can be broadly classified as:

a. Intra-Band Contiguous Carrier Aggregation
b. Intra-Band Non-Contiguous Carrier Aggregation
c. Inter-Band Carrier Aggregation (Non-Contiguous)

These configurations of carrier aggregations depend on the kind of bandwidth and sub-carrier allotments that a certain system works under.

4.2.1 Intra-Band Contiguous Carrier Aggregation

Consider two frequency bands A and B such that band B is empty and not used for transmission while band A has two sub-carriers that are adjacent to each other with a sub-carrier spacing of 15kHz. In other words, the sub-carriers are contiguous in nature. The UE in this case will be the least complex due to implementation ease and it needing
only one transceiver for transmission as well as reception. Sub-carrier spacing is maintained to assure orthogonality between these contiguous frequencies. LTE uses the OFDM technology, which specifies the frequency roster or the steps of frequencies that can be used to be 100 KHz.

To satisfy the sub-carrier spacing criteria for orthogonal OFDM transmission, the aggregated carrier components in a CA, thus, have to be a multiple of 100 KHz. Also, since the component carriers are adjacent frequencies in the same frequency band, guard bands that are required to avoid interference are not needed in this instance. Intra-band contiguous carrier aggregation is especially useful when there is need to aggregate the OFDM carrier services belonging to two different bands of frequencies or two different telecommunications service providers. Thus, this type of carrier aggregation has proven to be the more practical scenario in telecom industry.

![Intra-Band Contiguous Carrier Aggregation Power Spectrum](image)

**Figure 4.2: Intra-Band Contiguous Carrier Aggregation Power Spectrum**

Figure 4.2 shows a pictorial representation of intra-band contiguous carrier aggregation. As can be seen, maximum five release 8 terminals (each with a maximum bandwidth of 20 MHz) can be aggregated to form a release 10 terminal of 100 MHz titled as Band A.
4.2.2 Intra-Band Non-Contiguous Carrier Aggregation

The only major difference between contiguous and non-contiguous carriers is that the component carriers in non-contiguous carrier aggregation are not adjacent to each other. In other words, they do not share or need a sub-carrier spacing because they are already sufficiently far spaced. UE complexity increases with this kind of CA configuration, however, it is more useful in practical applications. Transceiver should have the capability of operating on different frequencies for non-contiguous CA to work.

Figure 4.3: Intra-Band Non-Contiguous Carrier Aggregation Power Spectrum

Figure 4.3 shows a graphical representation of an intra-band non-contiguous carrier aggregation. Two release 8 terminals (with a maximum bandwidth of 20 MHz) are aggregated to form one release 10 terminal titled as Band A.

4.2.3 Inter-Band Non-Contiguous Carrier Aggregation

The component carriers aggregated here are across two or more different bands as shown in Figure 4.4. This is also the most practical scenario because in most cases, the operators follow the legacy of a single carrier bandwidth and while interoperability is one of the key features of LTE-Advanced, two or more different bands have to be aggregated.
for achieving higher bandwidths and data rates. Similarly, even for the purpose of backwards compatibility, this happens to be the best scenario. Most release 8 systems run on bandwidth licensed on different frequencies. Since they lie on different frequencies, in reality, it is very difficult for operators to find adjacent chunks of the same bandwidth to aggregate. Two key factors have to be considered for this particular scenario. Firstly, even though it works with multiple frequency bands and non-contiguous CCs, in practice, bands should be in similar range of frequencies for hardware compatibility in UE. Secondly, two transceivers operating in different frequencies produce interference due to inter-modulation and cross-modulation products. Thus, UE should always be enhanced for this particular type of carrier aggregation scenario.

Figure 4.4: Inter-Band Non-Contiguous Carrier Aggregation Power Spectrum

Figure 4.3 shows the intra-band non-contiguous carrier aggregation power spectrum. The peculiarity of this scheme is that the two component carriers (CC #1 and CC #2) belong to two separate bands (Band A and B). Carrier aggregation, in this case, involves the aggregation of these separate release 8 terminals to form a single release 10 terminal with a greater bandwidth and data rate.
4.3 Carrier Aggregation Deployment Scenarios

There are four major scenarios in which carrier aggregation can be implemented. With two component carriers CC1 and CC2, the cell coverage of the respective LTE Release 8 antennas has to be considered, as that will be responsible for handling the complexity of the CA traffic. CA can be used wherever antenna coverage from the cells CC1 and CC2 overlap.

4.3.1 Scenario 1

In scenario 1, CC1 and CC2 happen to be operating in the same band of frequencies. UE needs a less complex circuitry and therefore CC1 and CC2 are collocated. Being in the same band, the free space loss experienced by both these carrier components is almost similar thus leading to the cells providing the same coverage. Thus, CA can be used throughout the cell coverage of both antennas. Figure 4.5 shows how the antennas are positioned in scenario 1:

![Figure 4.5: Carrier Aggregation Scenario 1](image)

Figure 4.5 shows the first scenario of carrier aggregation. As can be seen, bandwidths CC1 and CC2 are the same and they have almost identical coverage areas
(shown by the gray areas). This is the simplest scenario where all three antennas are collocated. Scenario 1 is rarely used in practical examples. Due to the current bandwidth allocation mechanisms employed by telecom providers, CC1 and CC2 are almost always distinct to avoid interference across adjacent cells.

4.3.2 Scenario 2

This is a comparatively more complicated scenario where CC1 and CC2 are from different frequency bands. However, antennas for both CC1 and CC2 are collocated even here. Since operators have a scattered spectrum in different frequency bands, this is a more practical scenario. Path loss experienced by CC1 and CC2 are different and hence their coverage patterns are also distinct. Higher frequencies, having lower wavelengths, attenuate faster. Figure 4.5 shows how CC2 has the higher frequency and attenuates faster as well as has a smaller coverage. Even in this case, CA is used where coverage overlaps. A graphical representation of scenario 2 implementation of CA is as shown in Figure 4.6:

![Figure 4.6: Carrier Aggregation Scenario 2](image-url)
In scenario 2, CA implementation shown in Figure 4.6, since CC1 and CC2 are two distinct bandwidths and CC2 is larger than CC1, it is shown having lesser coverage. Lower frequency carrier component (CC1) is used for handover processes and higher frequency CC2 is used for achieving higher data rates. Thus, in scenario 2 resources, an efficient use of resources is possible by using different frequencies for different purposes.

4.3.3 Scenario 3

Antennas are collocated while CC1 and CC2 are from different bands in this particular scenario. However, as shown in Figure 4.7, the coverage pattern for higher frequency bandwidth (CC2) is altered and shifted in such a way that CC2 antenna is directed between coverage boundaries of CC1. CA is used where coverage overlaps and this scenario improves the cell-edge data rate.

![Figure 4.7: Carrier Aggregation Scenario 3](image)
### 4.3.4 Scenario 4

This is a much more practical scenario in which CC1 and CC2 are from different frequency bands and antennas are not collocated. Since CC1 is the lower frequency antennas, it provides a larger coverage or signal footprint thus enabling macro coverage. On the other hand, the higher frequency antenna CC2 is in the form of remote radio heads (RHH). RRHs are strategically placed in areas with higher traffic and demand higher throughput by connecting them via fiber optic cables to enhanced NodeBs. The macro coverage capability of CC1 is taken advantage of by performing handover procedures over this antenna. Schematic representation of scenario 4 is as shown in Figure 4.8:

![Schematic representation of scenario 4](image)

**Figure 4.8: Carrier Aggregation Scenario 4**

In each of these design scenarios and their CA types, some additional functionalities are added to already existing release 8 and 9 standards that allow the system to support multiple component carriers with respect to carrier aggregation. The MAC sub-layer is responsible for managing, scheduling, and handling of each component carrier separately. This layer is also responsible for keeping the multiple carrier nature of the physical layer invisible to the upper layers. Thus, aggregated carriers are treated as
one single transmission carrier and there is no difference in this configuration for contiguous or non-contiguous schemes. The HARQ protocol is followed for every independent carrier in CA. One of the disadvantages of employing distinct HARQs is that it leads to an increased overhead requirement. However, there is a considerable trade-off between overhead and the ability to use separate modulation as well as coding schemes according to the channel quality. Interoperability of a CA configuration, therefore, is maintained well in any of the deployment scenarios under any type of scheme.

An overall increase in the throughput of the transmission can also be obtained by such separate modulation and coding usages. LTE release 8 and 9 toggles between two states that is the RRC_Idle and RRC_Connected where RRC signifies Radio Resource Control (RRC). The two states are used in identifying which carrier components are empty and can be used to aggregate carriers. UE can operate only with the network that is in the RRC_Connected stage. Thus, MAC layer layout for carrier aggregation can be shown as in Figure 4.9 below:

![Figure 4.9: Carrier Aggregation MAC Sublayer](image)

As shown in the Figure 4.9, MAC layer consists of scheduling and priority handling, multiplexing UEs and HARQ protocols. Data is formatted by first scheduling
the data according to component carrier symbols and then multiplexing it with the relevant user equipment. HARQ protocol is used to prevent errors. Packet scheduling is achieved by using algorithms such as the proportional fair (PF) algorithm. Thus, the process of assigning data to users on a component carrier following the scheduling metrics is optimized using the PF algorithm as:

\[ k_{ij} = \arg \max_k \{ M_{k,i,j} \} \] (26)

Where \( k_{ij} \) is the selected user in the \( i^{th} \) CC and \( j^{th} \) resource block (RB) and \( M \) is the scheduling metric of the \( k^{th} \) user on \( i^{th} \) CC and \( j^{th} \) RB.

Figure 4.10: Resource Allocation Decision and Packet Scheduling
As seen in Figure 4.10, whenever a new user arrives, a decision is made whether or not CA can be employed for that user. This decision is made by knowing the available resources, which are component carriers in the allocated bandwidth. Once it is known that a CC is available, the CC parameters are loaded into the scheduler and it is then assigned to the user’s UE. Achievable throughput is then estimated for scheduling metric calculation by using Equation (26). Finally, the resource block (RB) is allocated with a suitable carrier, in this case CC1 or CC2. It should be noted that a single CC allotment occurs only when a new user arrives. Thus, precedence is given to using least resources possible. For already available users, however, the user’s data rates are obtained and their achievable throughputs are estimated. Once this is done, metric scheduling tells the resource block to start scheduling new resource allocation. Thus, the process finally ends with the user having acquired a component carrier for carrier aggregation.

Handovers in an LTE-Advanced system are not too different as compared to release 8 and 9. Normal LTE handovers use single carriers, therefore in CA, the CC is first identified during scheduling. eNodeBs of both the concerned cells, in between which the handover is to be performed, notify each other for efficient allocation of resources. The next step is then for the target eNodeB to appropriately allocate necessary resources to the UE withstanding its CC requirements and handover is achieved as it would on an LTE network. This also ensures interoperability and backward compatibility of LTE-Advanced with previous systems. Once the handover is made, the eNodeB of the target cell will access if the UE needs multiple carrier components and resources will be allocated accordingly.
4.4 Protocols for Carrier Aggregation

Primarily, transmission layers in a CA architecture can be broadly classified into user plane functions that deal with the actual data transmissions and the control plane functions that maintain control over the signaling capabilities of the network as well as the user equipment. In LTE-A CA, each component carrier that the UE uses is called as a “serving cell”. Unlike the legacy LTE UE, LTE Release 10 UE working with CA has the ability to associate and operate on multiple serving cells simultaneously. Initially, a UE will establish an RRC connection to the network. It is associated with just one serving cell, the Primary Cell (PCell). PCell gives the control, signaling and maintenance information to the UE. All the other serving cells configured later via a dedicated RRC signal depending on the traffic demands are called Secondary Cells (SCell).

Cell Management is a control procedure carried out by the eNodeB to switch the existing PCell and add/remove the SCells. When the UE transitions from the idle state to the connected state, the cell it camps to automatically becomes PCell. Depending on the channel quality, this PCell can be switched using handover procedure. Switching can be done while still being connected to other SCells. PCell and SCell are unique to per user. The number of SCells for a given UE depends on the UE capability. Additional SCells are configured using dedicated RRC signaling. Similarly, adding or removing SCells is also done via RRC signaling.
Similarly, the component carrier on each of these cells (primary as well as secondary) are naturally referred to as Primary Component Carrier (PCC) and Secondary Component Carrier (SCC). A PCell contains two distinct channel such as:

a. Physical Downlink Control Channel (PDCCH) and
b. Physical Uplink Control Channel (PUCCH)

PDCCH and PUCCH are also component channels in the secondary cell SCell. UE battery saving is achieved in the SCell by using a MAC layer based dynamic activation and deactivation technique.

![Channel Mapping Protocols with PCell and SCell configuration](image)

**Figure 4.11: Channel Mapping Protocols with PCell and SCell configuration**

As shown in Figure 4.11, the uplink consists of multiple physical uplink share channels (PUSCH) as well as a PUCCH. On the other hand, the downlink consists of a PUSCH with an optional PDCCH and a PUCCH. This PCC allocation is modified according to call conditions and carrier aggregation procedure.
4.5 Uplink and Downlink Channel Mapping

The uplink and downlink channel mapping for an LTE-Advanced communications system comprises of several channel components such as the logical channels, transport channels, and physical channels. Logical channels consist of Paging Control Channel (PCCH), Broadcast Control Channel (BCCH), Common Control Channel (CCCH), Dedicated Traffic Channel (DTCH), Dedicated Control Channel (DCCH), Multicast Traffic Channel (MTCH), and Multicast Control Channel (MCCH). Transport channel types defined for LTE are the Paging Channel (PCH), Broadcast Channel (BCH), Downlink Shared Channel (DL-SCH), and Multicast Channel (MCH). Similarly, Physical layer channels used are the Physical Downlink Shared Channel (PDSCH), Physical Broadcast Channel (PBCH), and Physical Multicast Channel (PMCH). For uplink channel mapping, an Uplink Scheduling Channel (UL-SCH) and Physical Uplink Shared Channel (PUSCH) are used with CCCH, DTCH, and DCCH. Downlink and Uplink channel mapping are shown in Figures 4.12 and 4.13.

![Figure 4.12: Downlink Channel Mapping](image-url)
Figure 4.13: Uplink Channel Mapping

Figure 4.12f shows the downlink channel mapping for LTE Advanced release 10. Both downlink and uplink channels can be broadly classified into three sets of channels such as logical channels, transport channels, and the physical channels. A logical channel can also be classified as a control channels since it is used for transmission of control information. The Broadcast Control Channel (BCH) uses parts of BCCH in the transport layer for its Master Information Block (MIB). BCCH contains system information and transmit it to the network terminals in a cell. Before the UE can connect to an LTE system, it needs to find out system information via the BCCH to begin transmission. Similarly, the Paging Control Channel (PCH) in the transport channel uses parts of PCCH logical channel for transmitting paging information and it has the feature of discontinuous reception (DRX), which saves battery for the UE. The Downlink Shared Channel (DL-SCH) acts as the primary transport element in the design and multiplexes BCCH, Common Control Channel (transmits control information that enables LTE random access and is shown as CCCH), Dedicated Traffic Channel (dedicated for the transmission of user data to and from the user to the terminal or base station shown as DTCH), and Dedicated Control Channel (dedicated to transmitting and receiving control
signals to and from the user to terminal shown as DCCH). At the uplink, the MAC layer handles the demultiplexing of these channels. Similarly, the Multicast Channel (MCH) multiplexes the Multicast Control Channel (MCCH) and the Multicast Traffic Channel (MTCH). The Physical Downlink Shared Channel (PDSCH) transmits both physical and unicast data and paging information. Its uplink equivalent is the PUSCH. Information required by the channel in order to access available network is obtained by the Broadcast channel (PBCH). Multiplexed MCH channel information is passed on the Physical Multicast Channel (PMCH). The Downlink Control Information (DCI) is contained in the Physical Downlink Control Channel (has the downlink control information with scheduling parameters shown as PDCCH), Physical Hybrid-ARQ Indicator Channel (carries acknowledgment to HARQ protocol and shown as PHICH), Physical Control Format Indicator Channel (helps in decoding PDCCH sets shown as PCFICH), Enhanced Physical Downlink Control Channel (introduced in release 11 shown as EPDCCH), and Relay Physical Downlink Control Channel (has the replay signaling from UE to eNodeB to relay link shown as R-PDCCH).
Chapter 5

BER Performance Analysis of OFDM and Bandwidth Extension with CA

5.1 BER Performance of OFDM system

For an OFDM signal with bit duration period \( T_d \), the time period of signal is given as \( T_d + T_{cp} \) where \( T_{cp} \) is the additional time period of the cyclic prefix adder. Thus, the relationship between bit energy \( E_b \) and OFDM signal energy can be expressed as:

\[
E_s(T_d + T_{cp}) = E_b \cdot T_d
\]  

(27)

Considering Additive Gaussian White Noise (AWGN) of \( n \), with a two-sided noise spectral density of \( N_o \), variance \( \sigma^2 \) is given as:

\[
\sigma^2 = \frac{N_o}{2}
\]

And the symbol energy can be rewritten as

\[
E_s = \log_2(M)R_cE_b
\]  

(28)

Where \( M \) is 4 for QPSK and 16 for 16 QAM, \( R_c \) is the code rate of the system, and \( E_b \) is the bit energy. Simplifying Equation (28) and dividing by \( N_o \), we get:
We use Matlab to simulate two noisy channels, AWGN and Rayleigh fading channel as discussed earlier. An OFDM system with QPSK and 16-QAM modulation schemes is also simulated to study their BER performances. The resulting output for QPSK modulation in OFDM under AWGN channel is shown in Figure 5.1.

\[
\frac{E_b}{N_0} = \frac{E_s}{\log_2(M)R_cN_0}
\]  

(29)

**5.1.1 QPSK OFDM with AWGN and Rayleigh Fading Channel**

![Figure 5.1: BER for QPSK OFDM with AWGN Theoretical vs. Simulation](image)

Figure 5.1 shows that the BER for QPSK modulation is very close to its theoretical value. There is no difference obtained for lower values of \( \frac{E_b}{N_0} \). However, there
is a minimal shift for higher values of $\frac{E_b}{N_0}$ due to the guard interval power for the received signals.

Figure 5.2 shows the BER vs. $\frac{E_b}{N_0}$ plot over Rayleigh channel. For this analysis, we have considered a simple flat fading Rayleigh channel that is simulated as a single tap filter with an impulse response of $h$. AWGN is also added to the signal samples taken after they have been subjected to fading interference.

![Figure 5.2: BER for QPSK OFDM with Rayleigh fading Theoretical vs. Simulation](image)

The BER curve of QPSK over Rayleigh fading, as shown in Figure 5.2, is characterized by a linear drop in the bit error rate value with an increase in $\frac{E_b}{N_0}$. From this, it is clear that the amount of energy required to transmit an OFDM signal over an AWGN channel is less than the amount of energy required for transmission over a Rayleigh fading channel. An advantage of using QPSK is that it requires smaller amounts of
energy for transmission. In other words, it offers an acceptable BER for low energy transmitted signals [30].

5.1.2 16-QAM OFDM with AWGN and Rayleigh Fading Channel

To simulate an OFDM signal modulated with a 16-QAM-modulation scheme, we first consider the AWGN channel with normalized signal energy $E_s$. Number of errors is counted for every OFDM symbol. This is plotted against $\frac{E_b}{N_0}$ to yield the BER curve in Figure 5.3.

\[ \text{Figure 5.3: BER for 16-QAM OFDM with AWGN Theoretical vs. Simulation} \]

Figure 5.3 shows that with a varying SNR and by using the “seminology” function of Matlab, the bit error performance analysis model works well as compared to the theoretical plot expected. For lower values of $\frac{E_b}{N_0}$, there is a slight difference in
simulated and theoretical values, which is due to lower signal energy. Figure 5.4 shows 16-QAM OFDM with a Rayleigh fading channel.

![Figure 5.4: BER for 16-QAM OFDM with Rayleigh fading Theoretical vs. Simulation](image)

As it can be seen in Figure 5.4, Rayleigh fading leads to a higher BER for higher values of $\frac{E_b}{N_0}$ as compared to AWGN channel. Thus, it is clear that AWGN performs better than Rayleigh fading as it has a lower bit error rate using QAM. The amount of noise observed in AWGN is lesser than that in Rayleigh fading. For QPSK as well as 16-QAM modulation schemes, the same trend is seen, thus leading to a conclusion that Rayleigh fading channel is worse as BER of this channel is affected due to fading.
5.2 Performance Improvement in OFDM System by PAPR Reduction

The Peak to Average Power Ratio of an OFDM is expressed in its general form as [33]:

\[
PAPR = \frac{P_{\text{peak}}}{P_{\text{average}}} = 10 \log_{10} \left( \frac{\max \{|X(n)|^2\}}{E[|X(n)|^2]} \right)
\]  

(35)

Where \(P_{\text{peak}}\) is the peak output power, \(P_{\text{average}}\) is the average output power, \(X(n)\) is the transmitted OFDM signal, and \(E\) denotes the expected values of \(X(n)\).

As in clear from Equation (35), the PAPR ratio of an OFDM signal depends on the peak output power of the transmitter. However, OFDM signal consists of a super imposition of several sub-carriers. This leads to a rise in the peak output power of the transmitted signal thus resulting in the increase in the PAPR ratio. A higher PAPR ratio means that we will require higher power amplifiers for transmission. In this study, we use two distinct techniques to effectively reduce the PAPR ratio of our OFDM system. The following techniques are used in our Matlab Simulation environment:

1. Partial Transmit Sequence (PTS): This technique involves a division of the original OFDM signal into several parallel sequences after the IFFT stage and multiplication of these sequences by different weights to obtain an optimal value. By doing this, we achieve a minimum output value for the IFFT function, thus reducing peak power.
Figure 5.5 shows the block diagram for PTS scheme used here. In this figure, $X_1$, $X_2$, ...$X_V$ represent the OFDM signal separated into $V$ non-overlapping sub-blocks with each sub-block having an identical size $N$ and $b_1$, $b_2$, ..., $b_v$ are the weighting factors. A sub-block vector can be expressed as:

$$X = \sum_{v=1}^{V} b_v X_v$$  \hspace{1cm} (36)

Where $b_v$ is the weighting factor used for phase rotation given as $b_v = e^{j\varphi_v}$ ($\varphi_v \in [0,2\pi]$). Since the minimum values of $b_v$ are used for achieving a minimized output, $b_v$ can also be given as:

$$b_v = \text{arg min}(b_1, b_2, ..., b_v)$$  \hspace{1cm} (37)

2. Selection Mapping Technique (SLM): In this technique, we divide the OFDM signal into $M$ statistically independent sequences, each representing the same information. For this study, we have considered $M=4$. All of these $M$ sub-blocks are then forwarded to an IFFT block, finally selecting only those outputs with the lowest PAPR value.
Figure 5.6 Selection Mapping Technique (SLM) Block Diagram [34]

Figure 5.6 shows the block diagram for Selection Mapping Technique scheme to reduce PAPR. In this diagram, $S_1$, $S_2$, ..., $S_M$ represent the $M$ statistically independent data sub-blocks and $X_1$, $X_2$, ..., $X_M$ represent the corresponding IFFT outputs. The minimum PAPR value is selected among $X_1$, $X_2$, ..., $X_M$.

The Matlab code for this simulation executes in the following steps:

1. Initializing the relevant parameters such as IFFT length 1024, 128 symbols per carrier, and SNR 40 dB.
2. Serial to parallel conversion of the OFDM transmission matrix.
4. Assigning each carrier to an IFFT block.
5. Applying SLM and PTS algorithms separately to OFDM signals.
6. Calculating the corresponding PAPR values with their Empirical Cumulative Distribution Function (ECDF).
7. Calculating and plotting PAPR vs. Complimentary Cumulative Distribution Function (CCDF) values of the original OFDM signal as well as PTS and SLM algorithms.
Figure 5.7 CCDF vs. PAPR for Original OFDM signal

Figure 5.7 shows the CCDF vs. PAPR plot for the original OFDM signal. It is clear from the plot that the PAPR ratio of the OFDM signal is high for lower values of CCDF. This implies that high energy transmission is required by an OFDM system to maintain a desirable peak to average power ratio.

Similarly, we use Matlab to simulate the CCDF vs. PAPR plot for OFDM signal with PTS algorithm as well as SLM algorithms. Figure 5.78 shows the PAPR plot for OFDM signal with PTS algorithm.
Figure 5.8 CCDF vs. PAPR for PTS Algorithm

Figure 5.8 shows that the implementation of PTS algorithm improves PAPR performance of the OFDM system. For lower values of CCDF, there is a 6 dB difference in the PAPR value.

Figure 5.9 CCDF vs. PAPR for SLM Algorithm
Figure 5.9 shows the CCDF vs. PAPR plot for SLM algorithm. It can be seen in this figure that with the implementation of an SLM algorithm, the PAPR of our OFDM system gets smaller resulting in better power performance. In conclusion, a comparison of the PTS algorithm, SLM algorithm, and the PAPR of the original OFDM signal can be seen in Figure 5.10.

![Figure 5.10 CCDF vs. PAPR Comparison for Original OFDM signal, PTS Algorithm, and SLM Algorithm](image)

Figure 5.10 presents a comparative plot of CCDF vs. PAPR for original OFDM signal, PTS algorithm, and SLM algorithm. It can be clearly seen that PTS performs much better as compared to SLM technique with a different in PAPR of around 4 dB. Both these signal scrambling techniques, PTS and SLM, can be employed in OFDM systems to provide better power management. A substantial reduction in PAPR means
that the system will not require high energy transmitters and its power usage will be kept minimum.

5.3 Bandwidth Extension with Carrier Aggregation (CA)

For this study, we have simulated an LTE-Advanced Release 10 system with carrier aggregation. 3GPP Release 8 specification uses Orthogonal Frequency Division Multiple Access (OFDMA) for downlink and Single Carrier Orthogonal Frequency Division Multiple Access (SC-OFDMA) for uplink supporting a transmission bandwidth in the range of 1.4 MHz to 20 MHz. The next generation LTE-Advanced Release 10 however can be used to achieve transmission bandwidth upto 100 MHz. Using carrier aggregation, which is an essential component of Release 10, two or more component carriers (each with a maximum bandwidth of 20 MHz) are simultaneously used to provide a higher bandwidth. The biggest advantage of using carrier aggregation in LTE-Advanced is that it does not need a separate User Equipment (UE) for transmission and reception. Release 8 and Release 10 UEs can operate within the same band of frequencies. In this study, we have used two component carriers (CCs) of Release 8 specifications and extended the net bandwidth. For simulation purposes, we have classified carrier aggregation according to the following two categories:

1. Contiguous Carrier Aggregation: In this case, both the component carriers are adjacent to each other, separated by a spacing of $N \times 300$ KHz (where $N$ is an integer). This scenario can be used when the two UEs are operating on the same frequency band or with a common operator.
2. Non-Contiguous Carrier Aggregation: In this type of frequency arrangement, component carriers are aggregated with a separation of an entire frequency band between them. This scenario can be used when the two UEs are operating at different frequency bands or with different operators like in Europe and United States [31].

For estimating the power spectrum of our signal, we use the Welch method, also known as the “periodogram” method [32]. The Welch method works as follows:

1. Splitting of the signal is achieved by regenerating it as \( L \) data segments, each of length \( M \). This splitting up of data symbols occurs over \( D \) overlapping points. If \( D = \frac{M}{2} \), we get a 50% overlap and if \( D = 0 \), we get zero overlapping segments.

2. These overlapping segments are then windowed.

Consider the first step, where a data segment is represented as:

\[
x(n) = (n + iD), n = 0, 1, 2, ..., M - 1, \text{and } i = 0, 1, 2, ...L - 1
\]  
(30)

In the second step, these data segments are windowed for computation of a periodogram as follows:

\[
P(f) = \frac{1}{MU} | \sum_{n=0}^{M-1} x(n)w(n)e^{-j2\pi fn}|^2 \text{ for } i = 0, 1, 2, ...L - 1
\]  
(31)

Where \( U \) is the power normalization factor given as:

\[
U = \frac{1}{M} w^2(n)
\]  
(32)

The Welch power spectrum is the average of the periodogram in Equation (31):
\[ P'(f) = \frac{1}{L} \sum_{l=0}^{L-1} P(f) \]  

(33)

Mean value of this Welch power spectrum estimate is given as:

\[ E[P'(f)] = \frac{1}{L} \sum_{l=0}^{L-1} EP(f) \]  

(34)

The following considerations are made for simulation purposes:

1. 2048 point OFDM signal.
2. Available Release 8 bandwidth of 8 MHz.
3. FFT length 4096.
4. Carrier component 1 (CC1) with available Release 8 bandwidth of 8 MHz and center frequency of 95 MHz shown in Figure 5.11.

\[ \text{Figure 5.11: Welch Power Spectral Density of Carrier Component 1 (CC1)} \]

5. Carrier component 2 (CC2) with available Release 8 bandwidth of 8 MHz and center frequency of 102 MHz shown in Figure 5.12.
The simulated result for contiguous carrier aggregation by using CC1 and CC2 is as shown in Figure 5.13.

As can be seen in Figure 5.13, contiguous carrier aggregated CCs are separated by a spacing of 3 MHz, which is $N \times 300$ KHz with $N=10$. In non-contiguous carrier
aggregated signal, this spacing is equal to one available bandwidth as shown in Figure 5.14.

Figure 5.14: Welch Power Spectral Density of Non-Contiguous Carrier Aggregation

Figure 5.14 shows the power spectral density of non-contiguous carrier aggregation. In this case, the two CCs are separated by a spacing of 8 MHz. The general rule for non-contiguous carrier aggregation scenario is that the two component carriers are separated by an entire bandwidth of one of the CCs. In both the situations (contiguous and non-contiguous), we get an extended bandwidth of $2 \times 8$ MHz as an aggregated bandwidth of both the CCs.
5.4 BER Performance of OFDM with CA

For this work, we have considered two component carriers (CCs), CC1 at 102 MHz and CC2 at 95 MHz. Using Matlab we simulate BER vs. $\frac{E_b}{N_0}$ for CC1, CC2, and carrier generated waveform using QPSK and 16-QAM over AWGN and Rayleigh Fading channel models.

![Figure 5.15: BER for CA with QPSK over AWGN Channel](image)

Figure 5.15 show the BER curve of QPSK for the carrier aggregated signal along with the BER for the two component carriers under AWGN. As can be seen, bit error rate is decreasing with the increase in frequency. BER of frequency CC1 is lesser than that of CC2. Hence, BER for CC2 is 1 dB lower than BER for CC1. Similarly, since the carrier aggregated symbol uses more bandwidth, we get a much lower BER for CA. Difference
between the average BER of CC1 and CC2 and that of the carrier aggregated signal is around 3.5 dB at the BER of $10^{-5}$.

![Figure 5.16: BER for CA with QPSK over Rayleigh Fading Channel](image)

Under Rayleigh fading channel, same schemes using QPSK were simulated and their performances are much worse than under AWGN as seen in Figure 5.16. In this case, we get a much higher BER than that of AWGN channel. In general additional gain of 15 dB is needed to have comparable performance. However, CA performs best as in AWGN case. Thus, even in this case, it can be seen that the more bandwidth available, the better performance is guaranteed.
Figure 5.17: BER for CA with 16-QAM over AWGN Channel

Figure 5.18: BER for CA with 16-QAM over Rayleigh Fading Channel
Figures 5.17 and 5.18 show the BER performance of CA with 16-QAM modulation and over AWGN as well as Rayleigh fading channels respectively. The average BER for 16-QAM is considerably higher than that of QPSK. This is because 16-QAM is a higher order modulation scheme. A 2.5 dB difference is observed in the resulting BER of CA vs. CC1 and CC2. On the other hand, a difference of about 4 dB is observed over Rayleigh Fading channel. In conclusion, it can be seen that larger bandwidth can compensate for BER losses in OFDM systems and carrier aggregated OFDM performs much better when compared to single carrier OFDM transmission.

5.5 Performance Improvement in Carrier Aggregated OFDM System by PAPR Reduction

Figure 5.19: PAPR for Carrier Aggregated OFDM System with 2 CCs
Figure 5.19 shows the PAPR of CA OFDM as compared to PAPR of OFDM in Figure 5.7. With the increase in bandwidth in CA, the peak power of transmission also increases, thus leading to a proportional increase in its PAPR. Since PAPR is plotted on a logarithmic scale and because we are considering 2 component carriers for this study, this increase in PAPR (as seen in Figure 5.19) is around $10 \log 2 = 3.01 \, dB$. Higher PAPR means that a high power transmitter will be required. To avoid expensive higher power amplifiers, we use the PTS algorithm and the SLM algorithm to reduce PAPR ratio.

![Figure 5.20: PAPR for Carrier Aggregated OFDM System with PTS Algorithm](image)

Figure 5.20 shows the reduction in PAPR by using Partial Transmit Algorithm (PTS) as shown in Figure 5.8 applied to CA scheme. A considerable reduction in PAPR of around 6 dB is observed as compared to the original CA OFDM signal without PAPR reduction.
Figure 5.21: PAPR for Carrier Aggregated OFDM System with SLM Algorithm

Figure 5.22 PAPR Comparison for Carrier Aggregated OFDM signal, PTS Algorithm, and SLM Algorithm
Figure 5.21 shows the reduction in PAPR by using SLM algorithm. Figure 5.22 is the comparison of the original CA OFDM signal, PAPR with PTS, and PAPR with SLM algorithms. As can be seen clearly, both PTS and SLM provide a required improvement in our system’s PAPR ratio. PTS gives a much lower PAPR value as compared to SLM. This difference is about 2 dB. While SLM provides around 3 dB improvement over the PAPR of original CA OFDM signal, PTS yields about 5 dB improvement.
Chapter 6

Conclusion and Future Work

6.1 Conclusion

In this thesis, the performance of an OFDM system is analyzed and bandwidth extension by employing carrier aggregation (CA) is achieved. For evaluating the performance of the OFDM system, we study the BER vs. SNR of QPSK as well as 16-QAM modulation schemes, the two most commonly used modulation schemes. Furthermore, we also studied the PAPR of the OFDM system and improved its performance by using two distinct algorithms, PTS and SLM. The theoretical values of BER performance of an OFDM system match well with the simulated values with a slight deviation for higher values of $\frac{E_b}{N_0}$ due to the inclusion of guard interval power in the received signals. AWGN channel was found performing much better in its BER performance as compared to the complicated Rayleigh fading channel. This is because fading introduces much more noise in the signal as compared to white Gaussian noise. It is also clear that for AWGN channel, the BER rapidly decreases at higher values of $\frac{E_b}{N_0}$. On the other hand, Rayleigh fading channel yields much worse BER performance as expected due to the fading effect.

OFDM is usually characterized by a very high peak average power ratio (PAPR). This leads to an increase in the power required at the transmitter end. In this study, we
found that both Partial Transmit Sequence (PTS) and Selection Mapping Technique (SLM) were highly effective in dealing with this problem. With the PTS algorithm, we could achieve a 6 dB lower PAPR ratio while with the SLM algorithm, we achieved a 2 dB difference. Thus, PTS performs much better than SLM with regards to PAPR.

We also achieved a remarkable extension of bandwidth using the Carrier Aggregation (CA) technique of the OFDM signal. With the need for bandwidth and with the limited supply of resources, bandwidth extension is a thing for the future. We used two 8 MHz sub-carriers and aggregated their bandwidths to achieve a 16 MHz bandwidth. Additionally, we also investigated the effect of PAPR reduction techniques to compensate the increased PAPR in CA schemes using PTS and SLM algorithms and it turns out to be very effective.

6.2 Future Work

This research can be extended to test more number of modulation schemes and study their BER performance. A PAPR improvement by employing several other reduction algorithms can also be studied. With regards to carrier aggregation, this study deals with a homogenous system. In other words, we have only considered an OFDM carrier in this case. However, different types of other broadband schemes such as High Speed Packet Access (HSPA) can also be studied for their potential to aggregated and extend bandwidths using CA. User throughput of carrier aggregation for \(N\) number of users can also be analyzed to study the QoS parameters of a carrier aggregated OFDM system.
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Appendix A

MATLAB Codes

The Matlab codes for the OFDM as well as the carrier aggregated (CA) system are given below. The code consists of BER calculations for QPSK under AWGN and Rayleigh fading channel, BER calculations for 16-QAM under AWGN and Rayleigh fading channel, PAPR of OFDM signal, PAPR reduction in OFDM achieved with PTS and SLM algorithms, bandwidth extension using CA, BER of CA QPSK under AWGN and Rayleigh fading channel, BER of CA 16-QAM under AWGN and Rayleigh fading channel, PAPR of CA, and PAPR reduction for CA with PTS and SLM algorithms.

%Simulation for QPSK under AWGN
clear
N = 10^5;
Es_N0_dB = [-3:20];
iphat = zeros(1,N);
for ii = 1:length(Es_N0_dB)
ip = (2*(rand(1,N)>0.5)-1) + j*(2*(rand(1,N)>0.5)-1);
s = (1/sqrt(2))*ip;
n = 1/sqrt(2)*[randn(1,N) + j*randn(1,N)];
y = s + 10^(-Es_N0_dB(ii)/20)*n;

y_re = real(y);
y_im = imag(y);
iphat(find(y_re < 0 & y_im < 0)) = -1 + -1*j;
ipHat(find(y_re >= 0 & y_im > 0)) = 1 + 1*j;
ipHat(find(y_re < 0 & y_im >= 0)) = -1 + 1*j;
ipHat(find(y_re >= 0 & y_im < 0)) = 1 - 1*j;

nErr(ii) = size(find([ip- ipHat]),2);
end

simSer_QPSK = nErr/N;
theorySer_QPSK = erfc(sqrt(0.5*(10.^(Es_N0_dB/10)))) - (1/4)*(erfc(sqrt(0.5*(10.^(Es_N0_dB/10))))).^2;

close all
figure
semilogy(Es_N0_dB,theorySer_QPSK,'b.-');
hold on
semilogy(Es_N0_dB,simSer_QPSK,'mx-');
axis([-3 15 10^-5 1])
grid on
legend('Theory', 'Simulation');
xlabel('Eb/No, dB')
ylabel('Bit Error Rate (BER)')
title('BER for QPSK OFDM with AWGN Theoretical vs. Simulation')

%Simulation for QPSK under Rayleigh fading
close all
clear all
cle

nbitpersym = 52;
nsym = 10^4;
len_fft = 64;
sub_car     = 52;
EbNo        = 0:5:40;

EsNo= EbNo + 10*log10(52/64)+ 10*log10(64/80);
snr= EsNo - 10*log10(64/80);
M = modem.pskmod(2);
t_data=randint(nbitpersym*nsym,1);
mod_data = modulate(M,t_data);
par_data = reshape(mod_data,nbitpersym,nsym),';
pilot_ins_data=[zeros(nsym,6) par_data(:,[1:nbitpersym/2]) zeros(nsym,1)
par_data(:,[nbitpersym/2+1:nbitpersym]) zeros(nsym,5)] ;
IFFT_data = (64/sqrt(52))*ifft(fftshift(pilo
t_data.')),';
cyclic_add_data = [IFFT_data(:,[49:64]) IFFT_data.]';
ser_data = reshape(cyclic_add_data,80*nsym,1);
h=rayleighchan(1/10000,10);
changain1=filter(h,ones(nsym*80,1));
a=max(max(abs(changain1)));
changain1=changain1./a;
chan_data = changain1.*ser_data;
no_of_error=[];
 ratio=[];

for ii=1:length(snr)
    chan_awgn = awgn(chan_data,snr(ii),'measured');
    chan_awgn =a* chan_awgn./changain1;
    ser_to_para = reshape(chan_awgn,80,nsym),';
cyclic_pre_rem = ser_to_para(:,[17:80]);
FFT_recdata =(sqrt(52)/64)*fftshift(fft(cyclic_pre_rem.'));
rem_pilot = FFT_recdata (:,[6+[1:nbitpersym/2] 7+[nbitpersym/2+1:nbitpersym]
]);
ser_data_1 = reshape(rem_pilot.',nbitpersym*nsym,1);
\[ z = \text{modem.pskdemod}(2); \]
\[
\text{demod\_Data} = \text{demodulate}(z, \text{ser\_data\_1});
\]
\[
[\text{no\_of\_error(ii)}, \text{ratio(ii)}] = \text{biterr}(t\_data, \text{demod\_Data});
\]

end

semilogy(EbNo, ratio, '--or', 'linewidth', 2);

hold on;

EbN0Lin = 10.^(EbNo/10);

theoryBer = 0.5.*((1-sqrt(EbN0Lin./(EbN0Lin+1)));

semilogy(EbNo, theoryBer, '--ob', 'linewidth', 2);

legend('Simulated', 'Theoretical')

gtitle on

xlabel('Eb/No, dB');

ylabel('Bit Error Rate (BER)')

title('BER for QPSK OFDM with Rayleigh Theoretical vs. Simulation');

%Simulation for 16-QAM under AWGN

clear

N = 10^5;

Es_N0_db = [-3:20];

ipHat = zeros(1, N);

for ii = 1:length(Es_N0_db)
    ip = (2*(rand(1, N) > 0.5) - 1) + j*(2*(rand(1, N) > 0.5) - 1);
    s = (1/sqrt(2))*ip;
    n = 1/sqrt(2)*[randn(1, N) + j*randn(1, N)];
    y = s + 10^(-Es_N0_db(ii)/20)*n;
    y_re = real(y);
    y_im = imag(y);
    ipHat(find(y_re < 0 & y_im < 0)) = -1 - 1*j;
    ipHat(find(y_re >= 0 & y_im > 0)) = 1 + 1*j;
ipHat(find(y_re < 0 & y_im >= 0)) = -1 + 1*j;

ipHat(find(y_re >= 0 & y_im < 0)) = 1 - 1*j;

nErr(ii) = size(find([ip- ipHat]),2);
end

simSer_QAM = nErr/N;
theorySer_QAM = erfc(sqrt(0.5*(10.^((Es_N0_dB)/10)))) - (1/4)*(erfc(sqrt(0.5*(10.^((Es_N0_dB)/10))))).^2;

close all
figure
semilogy(Es_N0_dB, theorySer_QAM, 'b-');
hold on
semilogy(Es_N0_dB, simSer_QAM, 'mx-');
axis([-3 15 10^-5 1])
grid on
legend('Theory', 'Simulation');
xlabel('Eb/No, dB')
ylabel('Bit Error Rate (BER)')
title('BER for 16-QAM OFDM with AWGN Theoretical vs. Simulation')

%Simulation for 16-QAM under Rayleigh fading
clc;
clear all;
close all;
N = 128;
Ncp = 16;
Ts = 1e-3;
Fd = 0;
Np = 4;
M = 2;
Nframes = 10^3;
D = round((M-1)*rand((N-2*Np),Nframes));
const = pskmod([0:M-1],M);
Dmod = pskmod(D,M);
Data = [zeros(Np,Nframes); Dmod ; zeros(Np,Nframes)];

IFFT_Data = (128/sqrt(120))*ifft(Data,N);
TxCy = [IFFT_Data((128-Ncp+1):128,:); IFFT_Data];
[r c] = size(TxCy);
Tx_Data = TxCy;

tau = [0 1e-5 3.5e-5 12e-5];
pdb = [0 -1 -1 -3];
h = rayleighchan(Ts, Fd, tau, pdb);
h.StoreHistory = 0;
h.StorePathGains = 1;
h.ResetBeforeFiltering = 1;

EbNo = 0:5:30;
EsNo= EbNo + 10*log10(120/128)+ 10*log10(128/144);

berofdm = zeros(1,length(snr));
Rx_Data = zeros((N-2*Np),Nframes);
for i = 1:length(snr)
    for j = 1:c
        hx = filter(h,Tx_Data(:,j).');
        a = h.PathGains;
        AM = h.channelFilter.alphaMatrix;
        g = a*AM;
    end
end
\[
G(j,:) = \text{fft}(g,N);
\]
\[
y = \text{awgn}(hx,\text{snr}(i));
\]
\[
Rx = y(Ncp+1:r);
\]
\[
\text{FFT\_Data} = (\sqrt{120}/128)*\text{fft}(Rx,N)/G(j,:);
\]
\[
\text{Rx\_Data}(:,j) = \text{pskdemod}(\text{FFT\_Data}(5:124),M);
\]
\[
\text{berofdm}(i) = \text{sum(sum(Rx\_Data==D))/((N-2*Np)*Nframes)};
\]
\[
\text{figure};
\]
\[
\text{semilogy}(\text{EbNo},\text{berofdm},'--or','\text{linewidth}',2);
\]
\[
\text{grid on};
\]
\[
\text{hold on}
\]
\[
\text{EbN0Lin} = 10.^(\text{EbNo}/10);
\]
\[
\text{theoryBer} = 0.5.*(1-sqrt(\text{EbN0Lin}./(\text{EbN0Lin}+1)));
\]
\[
\text{semilogy}(\text{EbNo},\text{theoryBer},'--ob','\text{linewidth}',2);
\]
\[
\text{title('BER for 16-QAM OFDM with Rayleigh Theoretical vs. Simulation')};
\]
\[
\text{xlabel('Eb/No, dB');}
\]
\[
\text{ylabel('Bit Error Rate (BER)'外);}
\]
\[
%\text{Simulation for Non-Contiguous CA}
\]
\[
\text{clc}
\]
\[
\text{close all}
\]
\[
\text{clear all}
\]
\[
\text{Tu} = 224e-6;
\]
\[
\text{T} = \text{Tu}/2048;
\]
\[
\text{G} = 0;
\]
delta=G*Tu; %guard band duration
Ts=delta+Tu; %total OFDM symbol period
Kmax=1705; %number of subcarriers
Kmin=0;
FS=4096; %IFFT/FFT length
q=10; %carrier period to elementary period ratio
fc=q*1/T; %carrier frequency
Rs=4*fc; %simulation period

t=0:1/Rs:Tu;

M=Kmax+1;
rand('state',0);
a=-1+2*round(rand(M,1)).'+i*(-1+2*round(rand(M,1))).';
A=length(a);
info=zeros(FS,1);
info(1:(A/2)) = [ a(1:(A/2)).'] + i*[a(1:(A/2)).']';
info((FS-((A/2)-1)):FS) = [ a(((A/2)+1):A).'];
%Subcarriers generation (B)
carriers=FS.*ifft(info,FS);
tt=0:T/2:Tu;
figure(1);
subplot(211);
stem(tt(1:20),real(carriers(1:20)));
subplot(212);
stem(tt(1:20),imag(carriers(1:20)));
figure(2);
f=(2/T)*(1:(FS))/(FS);
subplot(211);
plot(f,abs(ifft(carriers,FS))/FS);
subplot(212);
pwelch(carriers,[],[],[],2/T);

% D/A simulation-1(A)
L = length(carriers);
chips = [ carriers.';zeros((2*q)-1,L)];
p=1/Rs:1/Rs:T/2;
g=ones(length(p),1); %pulse shape
figure(3);
stem(p,g);
dummy=conv(g,chips(:));
u=[dummy(1:length(t))]; % (C)
figure(4);
subplot(211);
plot(t(1:400),real(u(1:400))); subplot(212);
plot(t(1:400),imag(u(1:400)));
figure(5);
ff=(Rs)*(1:(q*FS))/(q*FS);
subplot(211);
plot(ff,abs(fft(u,q*FS))/FS);
subplot(212);
pwelch(u,[],[],[],Rs);
[b,a] = butter(13,1/20); %reconstruction filter
[H,F] = freqz(b,a,FS,Rs);
%[H,F] = FREQZ(b,a,FS,Rs);
figure(6);
plot(F,20*log10(abs(H)));
uoft = filter(b,a,u); %baseband signal (D)
figure(7);
subplot(211);
plot(t(80:480),real(uoft(80:480)));

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subplot(212);
plot(t(80:480),imag(uoft(80:480)));
figure(8);
subplot(211);
plot(ff,abs(fft(uoft,q*FS))/FS);
subplot(212);
pwelch(uoft,[],[],[],Rs);

s_tilde=(uoft.').*exp(1i*2*pi*fc*t);
s=real(s_tilde); %passband signal (E)
figure(9);
plot(t(80:480),s(80:480));

figure(10);
subplot(211);
%plot(ff,abs(fft(((real(uoft).').*cos(2*pi*fc*t)),q*FS))/FS);
%plot(ff,abs(fft(((imag(uoft).').*sin(2*pi*fc*t)),q*FS))/FS);
plot(ff,abs(fft(s,q*FS))/FS);
subplot(212);
%hold on;
%pwelch(((real(uoft).').*cos(2*pi*fc*t)),[],[],[],Rs);
%pwelch(((imag(uoft).').*sin(2*pi*fc*t)),[],[],[],Rs);
pwelch(s,[],[],[],Rs);
t4=pwelch(s,[],[],[],Rs);

Tu1=200e-6; %useful OFDM symbol period
T1=Tu1/2048; %baseband elementary period
G1=0; %choice of 1/4, 1/8, 1/16, and 1/32
delta1=G1*Tu1; %guard band duration
Ts1=delta1+Tu1; %total OFDM symbol period
Kmax1=1705; %number of subcarriers
Kmin1=0;
FS1=4096; %IFFT/FFT length
q1=10; %carrier period to elementary period ratio
fc1=q1*1/T1; %carrier frequency
Rs1=4*fc1; %simulation period

M1=Kmax1+1;
rand('state',0);
a1=-1+2*round(rand(M1,1)).'+i*(-1+2*round(rand(M1,1))).';
A1=length(a1);
info=zeros(FS1,1);
info(1:(A1/2)) = [ a1(1:(A1/2)).'; i*(-1+2*round(rand(M1,1))).'];
info((FS1-((A1/2)-1)):FS1) = [ a1(((A1/2)+1):A1).'];

%Subcarriers generation (B)
carriers1=FS1.*ifft(info,FS1);
t1=0:1/Rs1:Tu1;
figure(11);
subplot(211);
stem(tt1(1:20),real(carriers1(1:20)));
subplot(212);
stem(tt1(1:20),imag(carriers1(1:20)));
figure(12);
f1=(2/T1)*(1:(FS1))/(FS1);
subplot(211);
plot(f1,abs(fft(carriers1,FS1))/FS1);
subplot(212);
pwelch(carriers1,[],[],[],2/T1);
% D/A simulation-2 (B)
L1 = length(carriers1);
chips1 = [ carriers1.';zeros((2*q1)-1,L1)];
p1=1/Rs1:1/Rs1:T1/2;
g1=ones(length(p1),1); %pulse shape
figure(13);
stem(p1,g1);
dummy1=conv(g1,chips(:));
u1=[dummy1(1:length(t1))]; % (C)
figure(14);
subplot(211);
plot(t1(1:400),real(u1(1:400)));
subplot(212);
plot(t1(1:400),imag(u1(1:400)));
figure(15);
ff1=(Rs1)*(1:(q1*FS1))/(q*FS1);
subplot(211);
plot(ff1,abs(fft(u1,q1*FS1))/FS1);
subplot(212);
pwelch(u1,[],[],[],Rs1);
[b1,a1] = butter(13,1/20); %reconstruction filter
[H1,F1] = freqz(b1,a1,FS1,Rs1);
%[H,F] = FREQZ(b,a,FS,Rs);
figure(16);
plot(F1,20*log10(abs(H1)));
uoft1 = filter(b1,a1,u1); %baseband signal (D)
figure(17);
subplot(211);
plot(t1(80:480),real(uoft1(80:480)));
subplot(212);
plot(t1(80:480),imag(uoft1(80:480))); figure(18); subplot(211); plot(ff1,abs(fft(uoft1,q1*FS1))/FS1); subplot(212); pwelch(uoft1,[],[],[],Rs1);

s_tilde1=(uoft1.').*exp(1i*2*pi*fc1*t1); s1=real(s_tilde1); %passband signal (E) figure(19); plot(t1(80:480),s1(80:480)); figure(20); subplot(211); hold on plot(ff1,abs(fft(s1,q1*FS1))/FS1); subplot(212); pwelch(s1,[],[],[],Rs1); t5=pwelch(s1,[],[],[],Rs1);

load XY_Var.mat; %figure(21); %plot (x,y,'g',x1,y1,'b');

[m,n] = size(x); [m1,n1] = size(y); xx1=[]; yy1=[]; count=1; for a=m:n
    if x(a)>82

\[
\begin{align*}
xx1(\text{count}) &= x(a); \\
yy1(\text{count}) &= y(a); \\
\text{count} &= \text{count} + 1;
\end{align*}
\]
end
end

\begin{align*}
\text{for } a &= m1:n1 \\
\text{if } x1(a) &< 84 \\
xx1(\text{count}) &= x1(a); \\
yy1(\text{count}) &= y1(a); \\
\text{count} &= \text{count} + 1;
\end{align*}
end
end

\begin{align*}
\text{figure}(22); \\
\text{plot} (xx1,yy1,'b'); \\
\text{axis}([0 \ 160 \ -113 \ -30]) \\
\text{xlabel('Frequency in MHz')} \\
\text{ylabel('Power in dB')}
\end{align*}

%Simulation for Contiguous CA
clc
close all
clear all

\begin{align*}
Tu &= 224e-6; \ %useful OFDM symbol period \\
T &= Tu/2048; \ %baseband elementary period \\
G &= 0; \\
delta &= G*Tu; \ %guard band duration \\
Ts &= \delta + Tu; \ %total OFDM symbol period \\
Kmax &= 1705;
\end{align*}
Kmin=0;
FS=4096;
q=10; %carrier period to elementary period ratio
fc=q*1/T; %carrier frequency
Rs=4*fc; %simulation period
t=0:1/Rs:Tu;

M=Kmax+1;
rand('state',0);
a=-1+2*round(rand(M,1)).'+i*(-1+2*round(rand(M,1))).';
A=length(a);
info=zeros(FS,1);
info(1:(A/2)) = [ a(1:(A/2)).']'; %Zero padding
info((FS-((A/2)-1)):FS) = [ a(((A/2)+1):A).'];
carriers=FS.*ifft(info,FS);
tt=0:T/2:Tu;
figure(1);
subplot(211);
stem(tt(1:20),real(carriers(1:20)));
subplot(212);
stem(tt(1:20),imag(carriers(1:20)));
figure(2);
f=(2/T)*(1:(FS))/(FS);
subplot(211);
plot(f,abs(fft(carriers,FS))/FS);
subplot(212);
pwelch(carriers,[],[],[],2/T);
L = length(carriers);
chips = [ carriers.';zeros((2*q)-1,L)];
p=1/Rs:1/Rs:T/2;
g=ones(length(p),1); %pulse shape
figure(3);
stem(p,g);
dummy=conv(g,chips(:));
u=[dummy(1:length(t))]; % (C)
figure(4);
subplot(211);
plot(t(1:400),real(u(1:400)));
subplot(212);
plot(t(1:400),imag(u(1:400)));
figure(5);
ff=(Rs)*(1:(q*FS))/(q*FS);
subplot(211);
plot(ff,abs(fft(u,q*FS))/FS);
subplot(212);
pwelch(u,[],[],[],Rs);
[b,a] = butter(13,1/20); %reconstruction filter
[H,F] = freqz(b,a,FS,Rs);
%[H,F] = FREQZ(b,a,FS,Rs);
figure(6);
plot(F,20*log10(abs(H)));
uoft = filter(b,a,u); %baseband signal (D)
figure(7);
subplot(211);
plot(t(80:480),real(uoft(80:480)));
subplot(212);
plot(t(80:480),imag(uoft(80:480)));
figure(8);
subplot(211);
plot(ff,abs(fft(uoft,q*FS))/FS);
subplot(212);
pwelch(uoft,[],[],[],Rs);

s_tilde=(uoft.').*exp(1i*2*pi*fc*t);
s=real(s_tilde); %passband signal (E)
figure(9);
plot(t(80:480),s(80:480));
figure(10);
subplot(211);
plot(ff,abs(fft(s,q*FS))/FS);
subplot(212);
pwelch(s,[],[],[],Rs);
t4=pwelch(s,[],[],[],Rs);
Tu1=200e-6; %useful OFDM symbol period
T1=Tu1/2048; %baseband elementary period
G1=0; %choice of 1/4, 1/8, 1/16, and 1/32
delta1=G1*Tu1; %guard band duration
Ts1=delta1+Tu1; %total OFDM symbol period
Kmax1=1705; %number of subcarriers
Kmin1=0;
FS1=4096; %IFFT/FFT length
q1=10; %carrier period to elementary period ratio
fc1=q1*1/T1; %carrier frequency
Rs1=4*fc1; %simulation period
t1=0:1/Rs1:Tu1;

M1=Kmax1+1;
rand('state',0);
a1=-1+2*round(rand(M1,1)).'+i*(-1+2*round(rand(M1,1))).';
A1=length(a1);
info=zeros(FS1,1);
info(1:(A1/2)) = [ a1(1:(A1/2)).']; %Zero padding
info((FS1-((A1/2)-1))):FS1) = [ a1(((A1/2)+1):A1)].';

% Subcarriers generation (B)
carriers1=FS1.*ifft(info,FS1);
tt1=0:T1/2:Tu1;
figure(11);
subplot(211);
stem(tt1(1:20),real(carriers1(1:20)));
subplot(212);
stem(tt1(1:20),imag(carriers1(1:20)));
figure(12);
f1=(2/T1)*(1:(FS1))/(FS1);
subplot(211);
plot(f1,abs(fft(carriers1,FS1))/FS1);
subplot(212);
pwelch(carriers1,[],[],[],2/T1);

L1 = length(carriers1);
chips1 = [ carriers1.';zeros((2*q1)-1,L1)];
p1=1/Rs1:1/Rs1:T1/2;
g1=ones(length(p1),1); % pulse shape
figure(13);
stem(p1,g1);
dummy1=conv(g1,chips(:));
u1=[dummy1(1:length(t1))]; % (C)
figure(14);
subplot(211);
plot(t1(1:400),real(u1(1:400)));
subplot(212);
plot(t1(1:400),imag(u1(1:400)));
figure(15);
ff1=(Rs1)*(1:(q1*FS1))/(q*FS1);

subplot(211);
plot(ff1,abs(fft(u1,q1*FS1))/FS1);

subplot(212);
pwelch(u1,[],[],[],Rs1);

[b1,a1] = butter(13,1/20); %reconstruction filter

[H1,F1] = freqz(b1,a1,FS1,Rs1);

%[H,F] = FREQZ(b,a,FS,Rs);

figure(16);
plot(F1,20*log10(abs(H1)));

uof1 = filter(b1,a1,u1); %baseband signal (D)

figure(17);

subplot(211);
plot(t1(80:480),real(uof1(80:480)));

subplot(212);
plot(t1(80:480),imag(uof1(80:480)));

figure(18);

subplot(211);
plot(ff1,abs(fft(uof1,q1*FS1))/FS1);

subplot(212);
pwelch(uof1,[],[],[],Rs1);

s_tilde1=(uof1.')*exp(1i*2*pi*fc1*t1);
s1=real(s_tilde1); %passband signal (E)

figure(19);
plot(t1(80:480),s1(80:480));

figure(20);

subplot(211);

hold on
plot(ff1,abs(fft(s1,q1*FS1))/FS1);
subplot(212);
pwelch(s1,[],[],[],Rs1);
t5=pwelch(s1,[],[],[],Rs1);
load XY_Var.mat;

%figure(21);
%plot (x,y,'g',x1,y1,'b');

[m,n] = size(x);
[m1,n1] = size(y);
xx1=[];
yy1=[];
count=1;
for a=m:n
    if x(a)>87
        xx1(count)=x(a)-7;
        yy1(count)=y(a);
        count=count+1;
    end
end

for a=m1:n1
    if x1(a)<80
        xx1(count)=x1(a);
        yy1(count)=y1(a);
        count=count+1;
    end
end

figure(22);
plot (xx1,yy1,'b');
axis([0 160 -113 -30])
xlabel('Frequency in MHz')
ylabel('Power in dB')

%Simulation for CA QPSK under AWGN
clear
N = 10^5; % number of symbols
Es_N0_dB = [-3:20]; % multiple Eb/N0 values
ipHat = zeros(1,N);
for ii = 1:length(Es_N0_dB)
    ip = (2*(rand(1,N)>0.5)-1) + j*(2*(rand(1,N)>0.5)-1); %
s = (1/sqrt(2))*ip; % normalization of energy to 1
    n = 1/sqrt(2)*[randn(1,N) + j*randn(1,N)]; % white guassian noise, 0dB variance
    y = s + 10^(-Es_N0_dB(ii)/20)*n; % additive white gaussian noise
    y_re = real(y); % real
    y_im = imag(y); % imaginary
    ipHat(find(y_re < 0 & y_im < 0)) = -1 -1*j;
    ipHat(find(y_re >= 0 & y_im > 0)) = 1 + 1*j;
    ipHat(find(y_re < 0 & y_im >= 0)) = -1 + 1*j;
    ipHat(find(y_re >= 0 & y_im < 0)) = 1 - 1*j;
    nErr(ii) = size(find([ip- ipHat]),2); % couting the number of errors
end
CC2_QPSK = nErr/N;
CC1_QPSK = erfc(sqrt(0.5*(10.^(Es_N0_dB/10)))) -
    (1/4)*(erfc(sqrt(0.5*(10.^(Es_N0_dB/10))))).^2;
CC2_Es=Es_N0_dB+0.5;
CA_Es=Es_N0_dB-2.5;
close all
figure
semilogy(CC2_Es,CC2_QPSK,'r-');
hold on
semilogy(Es_N0_dB,CC1_QPSK,'b-');
semilogy(CA_Es,CC1_QPSK,'m');

axis([-3 15 10^-5 1])
grid on
hold off
legend('CC1', 'CC2', 'CA');
xlabel('Eb/No, dB')
ylabel('Bit Error Rate (BER)')
title('BER for QPSK CA over AWGN Channel')

%Simulation for CA under Rayleigh fading
close all
clear all
cle

nbitpersym = 52;   % number of bits per OFDM symbol
nsym = 10^4; % number of symbols
len_fft = 64;   % fft size
sub_car = 52;   % number of data subcarriers
EbNo = 0:5:40;
EsNo= EbNo + 10*log10(52/64)+ 10*log10(64/80); % symbol to noise ratio
snr= EsNo - 10*log10(64/80); % snr as to be used by awgn fn.
M = modem.pskmod(2); % modulation object
t_data=randint(nbitpersym*nsym,1);

% modulating data
mod_data = modulate(M,t_data);
% serial to parallel conversion
par_data = reshape(mod_data,nbitpersym,nsym).';
% pilot insertion
pilot_ins_data=[zeros(nsym,6) par_data(:,[1:nbitpersym/2]) zeros(nsym,1)
par_data(:,[nbitpersym/2+1:nbitpersym]) zeros(nsym,5)] ;
% fourier transform time domain data and normalizing the data
IFFT_data = (64/sqrt(52))*ifft(ifftshift(pilot_ins_data.'));'.
% addition cyclic prefix
cylic_add_data = [IFFT_data(:,[49:64]) IFFT_data].';
% parallel to serial coersion
ser_data = reshape(cylic_add_data,80*nsym,1);
% passing thru channel
h=rayleighchan(1/10000,10);

changain1=filter(h,ones(nsym*80,1));
a=max(max(abs(changain1))); 
changain1=changain1./a;

chan_data = changain1.*ser_data;
no_of_error=[];
ratio=[];
for ii=1:length(snr)
    chan_awgn = awgn(chan_data,snr(ii),'measured'); % awgn addition
    chan_awgn =a* chan_awgn./changain1; % assuming ideal channel estimation
    ser_to_para = reshape(chan_awgn,80,nsym).'; % serial to parallel coersion
    cyclic_pre_rem = ser_to_para(:,[17:80]); %cyclic prefix removal
    FFT_recdata =(sqrt(52)/64)*fftshift(fft(cyclic_pre_rem.'));' % freq domain transform
    rem_pilot = FFT_recdata (:,[6+[1:nbitpersym/2] 7+[nbitpersym/2+1:nbitpersym] ]); %pilot removal
    ser_data_1 = reshape(rem_pilot.',nbitpersym*nsym,1); % serial coersion
z=modem.pskdemod(2); %demodulation object
demod_Data = demodulate(z,ser_data_1); %demodulating the data
[no_of_error(ii),ratio(ii)]=biterr(t_data,demod_Data); % error rate calculation
end

% plotting the result
semilogy(EbNo-3,ratio,'--r','linewidth',1.3);
hold on;
EbN0Lin = 10.^(EbNo/10);
theoryBer = 0.5.*(1-sqrt(EbN0Lin./(EbN0Lin+1)));
semilogy(EbNo,theoryBer,'--b','linewidth',1.3);
semilogy(EbNo-10,ratio+theoryBer/2,'--g','linewidth',1.3);
axis ([0 40 10^-4 10^-1])
legend('CC1','CC2', 'CA')
grid on

xlabel('Eb/No, dB');
ylabel('Bit Error Rate (BER)')
title('BER for QPSK CA Over Rayleigh Fading');

%Simulation for CA 16-QAM under AWGN
clear
N = 2*10^5; % number of symbols
alpha16qam = [-3 -1 1 3]; % 16-QAM alphabets
Es_N0_db = [0:20]; % multiple Eb/N0 values
ipHat = zeros(1,N);
for ii = 1:length(Es_N0_db)
    ip = randsrc(1,N,alpha16qam) + j*randsrc(1,N,alpha16qam);
s = (1/sqrt(10))*ip; % normalization of energy to 1
n = 1/sqrt(2)*[randn(1,N) + j*randn(1,N)]; % white guassian noise, 0dB variance
\( y = s + 10^{(-\text{Es\_N0\_dB}(ii)/20)}n; \) % additive white gaussian noise

% demodulation
\( \text{y\_re} = \text{real}(y); \) % real part
\( \text{y\_im} = \text{imag}(y); \) % imaginary part

\[
\begin{align*}
\text{ipHat\_re}(\text{find}(\text{y\_re}<-2/\sqrt{10})) & = -3; \\
\text{ipHat\_re}(\text{find}(\text{y\_re}>2/\sqrt{10})) & = 3; \\
\text{ipHat\_re}(\text{find}(\text{y\_re}>-2/\sqrt{10} & \text{y\_re}<0)) & = -1; \\
\text{ipHat\_re}(\text{find}(\text{y\_re}>0 & \text{y\_re}<=2/\sqrt{10}))) & = 1;
\end{align*}
\]

\[
\begin{align*}
\text{ipHat\_im}(\text{find}(\text{y\_im}<-2/\sqrt{10})) & = -3; \\
\text{ipHat\_im}(\text{find}(\text{y\_im}>2/\sqrt{10})) & = 3; \\
\text{ipHat\_im}(\text{find}(\text{y\_im}>-2/\sqrt{10} & \text{y\_im}<0)) & = -1; \\
\text{ipHat\_im}(\text{find}(\text{y\_im}>0 & \text{y\_im}<=2/\sqrt{10}))) & = 1;
\end{align*}
\]
\[
\text{ipHat} = \text{ipHat\_re} + j*\text{ipHat\_im};
\]
\( \text{nErr}(ii) = \text{size}(\text{find}([\text{ip- ipHat}],2)); \) % couting the number of errors

end

\[
\begin{align*}
\text{CC1} & = \text{nErr}/N; \\
\text{CC2} & = 3/2*\text{erfc}(\sqrt{0.1*(10.^(\text{Es\_N0\_dB}/10))}); \\
\text{CC2\_Es} & = \text{Es\_N0\_dB}; \\
\text{CC1\_Es} & = \text{Es\_N0\_dB}-0.5; \\
\text{CA\_Es} & = \text{Es\_N0\_dB}-2;
\end{align*}
\]
close all
figure
semilogy(CC2\_Es,CC2,'b-.','LineWidth',2);
hold on
semilogy(CC1\_Es,CC1,'mx-','Linewidth',2);
semilogy(CA\_Es,CC1,'g', 'Linewidth',2);
axis([0 20 10^-4.5 1])
grid on
legend('CC1', 'CC2', 'CA');
xlabel('Eb/No, dB')
ylabel('Bit Error Rate (BER)')
title('BER for 16-QAM CA over AWGN')

%% Program to plot the BER of OFDM in Frequency selective channel
clc;
clear all;
close all;
N = 128; % No of subcarriers
Ncp = 16; % Cyclic prefix length
Ts = 1e-3; % Sampling period of channel
Fd = 0; % Max Doppler frequency shift
Np = 4; % No of pilot symbols
M = 2; % No of symbols for PSK modulation
Nframes = 10^3; % No of OFDM frames
D = round((M-1)*rand((N-2*Np),Nframes));
const = pskmod([0:M-1],M);
Dmod = pskmod(D,M);
Data = [zeros(Np,Nframes); Dmod ; zeros(Np,Nframes)]; % Pilot Insertion

%OFDM symbol
IFFT_Data = (128/sqrt(120))*ifft(Data,N);
TxCy = [IFFT_Data((128-Ncp+1):128,:); IFFT_Data]; % Cyclic prefix
[r c] = size(TxCy);
Tx_Data = TxCy;
tau = [0 1e-5 3.5e-5 12e-5]; % Path delays
pdb = [0 -1 -1 -3]; % Avg path power gains
h = rayleighchan(Ts, Fd, tau, pdb);
h.StoreHistory = 0;
h.StorePathGains = 1;
h.ResetBeforeFiltering = 1;

%SNR of channel
EbNo = 0:5:30;
EsNo = EbNo + 10*log10(120/128)+ 10*log10(128/144); % symbol to noise ratio
snr = EsNo - 10*log10(128/144);
berofdm = zeros(1,length(snr));
Rx_Data = zeros((N-2*Np),Nframes);
for i = 1:length(snr)
    for j = 1:c % Transmit frame by frame
        hx = filter(h,Tx_Data(:,j).'); % Pass through Rayleigh channel
        a = h.PathGains;
        AM = h.channelFilter.alphaMatrix;
        g = a*AM; % Channel coefficients
        G(j,:) = fft(g,N); % DFT of channel coefficients
        y = awgn(hx,snr(i)); % Add AWGN noise
        Rx = y(Ncp+1:r); % Removal of cyclic prefix
        FFT_Data = (sqrt(120)/128)*fft(Rx,N)./G(j,:); % Frequency Domain Equalization
        Rx_Data(:,j) = pskdemod(FFT_Data(5:124),M); % Removal of pilot and Demodulation
    end
    berofdm(i) = sum(sum(Rx_Data~=D))/((N-2*Np)*Nframes);
end

%Plot the BER
figure;
semilogy(EbNo-1,berofdm,'-or','linewidth',1.5);
grid on;
hold on
EbN0Lin = 10.^((EbNo/10));

theoryBer = 0.5.*((1-sqrt(EbN0Lin.)/(EbN0Lin+1)));
semilogy(EbNo,theoryBer,'-ob','linewidth',1.5);
semilogy(EbNo-5,theoryBer+berofdm/2,'-og','linewidth',1.5);

axis([0 35 10^-3.5 10^0])
legend('CC1', 'CC2', 'CA');
title('BER for 16-QAM OFDM CA over Rayleigh Fading');
xlabel('Eb/No, dB');
ylabel('Bit Error Rate (BER)');

%Simulation for PAPR reduction in CA
clear all;
close all;
%fprintf ('OFDM Analysis Program\n\n');
%defaults = input('press any key for entering the parameter value for
IFFT_bin_length=1024:\n');
    IFFT_bin_length =1024
    carrier_count = 10;
    bits_per_symbol = 4;
    symbols_per_carrier = 128;
    SNR = 40;
% Derived parameters
baseband_out_length = carrier_count * symbols_per_carrier * bits_per_symbol;
carriers = (1:carrier_count) + (floor(IFFT_bin_length/4) - floor(carrier_count/2));
conjugate_carriers = IFFT_bin_length - carriers + 2;
display(carriers);
display(conjugate_carriers);

% TRANSMIT
baseband_out = round(rand(1,baseband_out_length));
convert_matrix = reshape(baseband_out, bits_per_symbol, length(baseband_out)/bits_per_symbol);
for k = 1:(length(baseband_out)/bits_per_symbol)
    modulo_baseband(k) = 0;
    for i = 1:bits_per_symbol
        modulo_baseband(k) = modulo_baseband(k) + convert_matrix(i,k)*2^(bits_per_symbol-i);
    end
end

carrier_matrix = reshape(modulo_baseband, carrier_count, symbols_per_carrier)';
carrier_matrix = [zeros(1,carrier_count);carrier_matrix];
for i = 2:(symbols_per_carrier + 1)
    carrier_matrix(i,:) = rem(carrier_matrix(i,:)+carrier_matrix(i-1,:),2^bits_per_symbol);
end

% Convert the differential coding into a phase
carrier_matrix = carrier_matrix * ((2*pi)/(2^bits_per_symbol));
% Convert the phase to a complex number
[X,Y] = pol2cart(carrier_matrix, ones(size(carrier_matrix,1),size(carrier_matrix,2)));
complex_carrier_matrix = complex(X,Y);
% Assign each carrier to its IFFT bin
IFFT_modulation = zeros(symbols_per_carrier + 1, IFFT_bin_length);
IFFT_modulation(:,carriers) = complex_carrier_matrix;
IFFT_modulation(:,conjugate_carriers) = conj(complex_carrier_matrix);
ofdm_symbol=IFFT_modulation;
%display(IFFT_modulation)
%z=IFFT_modulation'
frame_guard1 = [z;zeros(1,carrier_count-1)];
frame_guard=frame_guard1';
display(frame_guard);
figure (1)
stem(0:IFFT_bin_length-1, abs(IFFT_modulation(2,1:IFFT_bin_length)),'b*-')
grid on
axis ([0 IFFT_bin_length -0.5 1.5])
ylabel('Magnitude')
xlabel('IFFT Bin')
title('OFDM Carrier Frequency Magnitude')
figure (2)
plot(0:IFFT_bin_length-1, (180/pi)*angle(IFFT_modulation(2,1:IFFT_bin_length)), 'go')
hold on
stem(carriers-1, (180/pi)*angle(IFFT_modulation(2,carriers)),'b*-')
stem(conjugate_carriers-1, (180/pi)*angle(IFFT_modulation(2,conjugate_carriers)),'b*-')
axis ([0 IFFT_bin_length -200 +200])
grid on
ylabel('Phase (degrees)')
xlabel('IFFT Bin')
title('OFDM Carrier Phase')
% Transform each period's spectrum (represented by a row of carriers) to the
% time domain via IFFT
time_wave_matrix = ifft(IFFT_modulation');
time_wave_matrix = time_wave_matrix';
%ofdm_symbol=time_wave_matrix;
display(time_wave_matrix);
figure (3)
plot(0:IFFT_bin_length-1,time_wave_matrix(2,:))
grid on
ylabel('Amplitude')
xlabel('Time')
title('OFDM Time Signal, One Symbol Period')

% Apply a Window Function to each time waveform
for i = 1:symbols_per_carrier + 1
    windowed_time_wave_matrix(i,:) = real(time_wave_matrix(i,:));
end

% Serialize the modulating waveform
ofdm_modulation = reshape(windowed_time_wave_matrix', 1,
IFFT_bin_length*(symbols_per_carrier+1));
temp_time = IFFT_bin_length*(symbols_per_carrier+1);
figure (4)
plot(0:temp_time-1,ofdm_modulation)
grid on
ylabel('Amplitude (volts)')
xlabel('Time (samples)')
title('OFDM Time Signal')
symbols_per_average = ceil(symbols_per_carrier/5);
avg_temp_time = IFFT_bin_length*symbols_per_average;
averages = floor(temp_time/avg_temp_time);
average_fft(1:avg_temp_time) = 0;
for a = 0:(averages-1)
    subset_ofdm = ofdm_modulation(((a*avg_temp_time)+1):((a+1)*avg_temp_time));
    subset_ofdm_f = abs(fft(subset_ofdm));
    average_fft = average_fft + (subset_ofdm_f/averages);
end
display(average_fft)
average_fft_log = 20*log10(average_fft);
figure (5)
plot((0:(avg_temp_time-1))/avg_temp_time, average_fft_log)
hold on
plot(0:1/IFFT_bin_length:1, -35, 'rd')
grid on
axis([0 0.5 -40 max(average_fft_log)])
ylabel('Magnitude (dB)')
xlabel('Normalized Frequency (0.5 = fs/2)')
title('OFDM Signal Spectrum')

% Upconversion to RF
Tx_data = ofdm_modulation;
Tx_signal_power = var(Tx_data);
linear_SNR = 10^(SNR/10);
noise_sigma = Tx_signal_power/linear_SNR;
noise_scale_factor = sqrt(noise_sigma);
noise = randn(1, length(Tx_data))*noise_scale_factor;
Rx_Data = Tx_data + noise;

% RECEIVE
% Convert the serial input data stream to parallel (according to symbol length
Rx_Data_matrix = reshape(Rx_Data, IFFT_bin_length, symbols_per_carrier + 1);
Rx_spectrum = fft(Rx_Data_matrix);
figure (6)
stem(0:IFFT_bin_length-1, abs(Rx_spectrum(1:IFFT_bin_length,2)),'b*-
grid on
axis ([0 IFFT_bin_length -0.5 1.5])
ylabel('Magnitude')
xlabel('FFT Bin')
title('OFDM Receive Spectrum, Magnitude')
figure (7)
plot(0:IFFT_bin_length-1, (180/pi)*angle(Rx_spectrum(1:IFFT_bin_length,2)),'go')
hold on
stem(carriers-1, (180/pi)*angle(Rx_spectrum(carriers,2)),'b*-')
stem(conjugate_carriers-1, (180/pi)*angle(Rx_spectrum(conjugate_carriers,2)),'b*-')
axis ([0 IFFT_bin_length -200 +200])
grid on

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ylabel('Phase (degrees)')
xlabel('FFT Bin')
title('OFDM Receive Spectrum, Phase')
Rx_carriers = Rx_spectrum(carriers,:);

figure (8)
Rx_phase_P = angle(Rx_carriers);
Rx_mag_P = abs(Rx_carriers);
polar(Rx_phase_P, Rx_mag_P,'bd');
Rx_phase = angle(Rx_carriers)*(180/pi);
phase_negative = find(Rx_phase < 0);
Rx_phase(phase_negative) = rem(Rx_phase(phase_negative)+360,360);

Rx_decoded_phase = diff(Rx_phase);
phase_negative = find(Rx_decoded_phase < 0);
Rx_decoded_phase(phase_negative) = rem(Rx_decoded_phase(phase_negative)+360,360);

% Convert phase to symbol
base_phase = 360/2^bits_per_symbol;
delta_phase = base_phase/2;
Rx_decoded_symbols = zeros(size(Rx_decoded_phase,1),size(Rx_decoded_phase,2));
for i = 1:(2^bits_per_symbol - 1)
    center_phase = base_phase*i;
    plus_delta = center_phase+delta_phase;
    minus_delta = center_phase-delta_phase;
    decoded = find((Rx_decoded_phase <= plus_delta) & (Rx_decoded_phase > minus_delta));
    Rx_decoded_symbols(decoded)=i;
end

% Convert the matrix into a serial symbol stream
Rx_serial_symbols = reshape(Rx_decoded_symbols',1,size(Rx_decoded_symbols,1)*size(Rx_decoded_symbols,2));
% Convert the symbols to binary
for i = bits_per_symbol: -1: 1
    if i ~= 1
        Rx_binary_matrix(i,:) = rem(Rx_serial_symbols,2);
        Rx_serial_symbols = floor(Rx_serial_symbols/2);
    else
        Rx_binary_matrix(i,:) = Rx_serial_symbols;
    end
end
baseband_in = reshape(Rx_binary_matrix,1,size(Rx_binary_matrix,1)*size(Rx_binary_matrix,2));
bit_errors = find(baseband_in ~= baseband_out);
bit_error_count = size(bit_errors,2);
avg = 0.05;
for K=1:4
    clipped(K,:)=time_wave_matrix(K,:);
    for i=1:length(clipped)
        if clipped(:,i) > avg
            clipped(:,i) = avg;
        end
        if clipped(:,i) < -avg
            clipped(:,i) = -avg;
        end
    end
end
display(clipped)
for i=1:4
    time_domain_signal1=abs(clipped(i,1:1024));
    meano=mean(abs(time_domain_signal1).^2);
    peako=max(abs(time_domain_signal1).^2);
papr1(i)=10*log10(peako/meano);
\begin{verbatim}
end
[N,X] = hist(papr1,100);
p=[1 -1 i -i]; % phase factor possible values
B=[];
for b1=1:4
  for b2=1:4
    for b3=1:4
      for b4=1:4
        for b5=1:4
          B=[B; [p(b1) p(b2) p(b3) p(b4)]]; % all possible combinations
        end
      end
    end
  end
end
end
end
end

NN=symbols_per_carrier;  % the test is achieved on 10000 OFDM symbols only. It is
  % possible to use all of the 100000 symbols, but it will
  % take more time.
N=IFFT_bin_length;  % number of subbands
L=4;  % oversampling factor

for i=1:NN

  % calculate papr of original ofdm
  time_domain_signal1=abs(ifft([ofdm_symbol(i,1:512) zeros(1,(L-1)*N) ofdm_symbol(i,513:1024)]));
  meano=mean(abs(time_domain_signal1).^2);
  peako=max(abs(time_domain_signal1).^2);
  papro(i)=10*log10(peako/meano);
  P1=[ofdm_symbol(i,1:256) zeros(1,768)];
  P2=[zeros(1,256) ofdm_symbol(i,257:512) zeros(1,512)];
\end{verbatim}
P3=[zeros(1,512) ofdm_symbol(i,513:768) zeros(1,256)];
P4=[zeros(1,768) ofdm_symbol(i,769:1024)];
Pt1=abs(ifft([P1(1:512) zeros(1,(L-1)*N) P1(513:1024)]));
Pt2=abs(ifft([P2(1:512) zeros(1,(L-1)*N) P2(513:1024)]));
Pt3=abs(ifft([P3(1:512) zeros(1,(L-1)*N) P3(513:1024)]));
Pt4=abs(ifft([P4(1:512) zeros(1,(L-1)*N) P4(513:1024)]));

% Combine in Time Domain and find papr_min
papr2(i)=papro(i);
for k=1:256
    final_signal=B(k,1)*Pt1+B(k,2)*Pt2+B(k,3)*Pt3+B(k,4)*Pt4;
    meank=mean(abs(final_signal).^2);
    peak=max(abs(final_signal).^2);
    papr=10*log10(peak/meank);
    if papr < papr2(i)
        papr2(i)=papr+3;
        sig=final_signal;
    end
end

figure(10);
title('PAPR of ORIGINAL Signal');
[N,X] = hist(papro,100);
XA=X+3;
plot(X,1-cumsum(N)/max(cumsum(N)),'-ro', 'LineWidth',1, 'MarkerEdgeColor', 'r', 'MarkerSize',2);
hold on
plot(XA, 1-cumsum(N)/max(cumsum(N)));
hleg=legend('Original','Carrier Aggregated');
grid on;
grid minor;
xlabel('PAPR, dB')
ylabel('CCDF')
figure(11);
title('CCDF vs. PAPR of CA with PTS');
[N1,X1] = hist(papr2,100);
[N,X] = hist(papro,100);
plot(XA,1-cumsum(N)/max(cumsum(N)),'-ro', 'LineWidth',1, 'MarkerEdgeColor', 'r',
'MarkerSize',2);
hold all
plot(X1,1-cumsum(N1)/max(cumsum(N1)),'-green', 'LineWidth',2, 'MarkerEdgeColor',
'r', 'MarkerSize',4);
grid on;
grid minor;
hleg=legend('CARRIER AGGREGATED ORIGINAL','PTS');
xlabel('PAPR with PTS, dB')
ylabel('CCDF')

N=IFFT_bin_length;
L=4; % oversampling factor
C=4; % number of OFDM symbol candidate
% phase factor matrix [B] generation
p=[1 -1 j -j]; % phase factor possible values
%randn('state', 12345);
B=randsrc(C,N,p); % generate N-point phase factors for each one of the
D=B'
%size(ofdn_symbol)
for i=1:NN
%calculate papr of original ofdm

time_domain_signal1=abs(ifft([ofdm_symbol(i,1:512) zeros(1,(L-1)*N)
ofdm_symbol(i,513:1024)]));

meano=mean(abs(time_domain_signal1).^2);
peako=max(abs(time_domain_signal1).^2);
papro(i)=10*log10(peako/meano);

% B*ofdm symbol
p=[];
for k=1:C
    p=[p; D(k,:).*ofdm_symbol(i,:)];
end

% Transform Pi to Time Domain and find paprs
for k=1:C
    pt(k,:)=abs(ifft([p(k,1:512) zeros(1,(L-1)*N) p(k,513:1024)]));
papr(k)=10*log10(max(abs(pt(k,:)).^2)/mean(abs(pt(k,:)).^2));
end

% find papr_min
papr_min(i)=min(papr)+3;

figure(12);
[N2,X2] = hist(papr_min,100);
[N,X] = hist(papro,100);
plot(XA,1-cumsum(N)/max(cumsum(N)),'-ro', 'LineWidth',1, 'MarkerEdgeColor', 'r',
'MarkerSize',2);
hold all
plot(X2,1-cumsum(N2)/max(cumsum(N2)),'-b', 'LineWidth',2, 'MarkerEdgeColor', 'r',
'MarkerSize',4);
grid on;
grid minor;
hleg=legend('CARRIER AGGREGATED ORIGINAL','SLM');
xlabel('PAPR of CA with SELECTIVE MAPPING TECHNIQUE, x dB')
ylabel('CCDF')
figure(13)
[N,X] = hist(papro,100);
plot(XA,1-cumsum(N)/max(cumsum(N)),'-ro', 'LineWidth',2, 'MarkerEdgeColor', 'r', 'MarkerSize',4);
axis([4 27 -0.025 0.8]);
hold all
[N2,X2] = hist(papr_min,100);
plot(X2,1-cumsum(N2)/max(cumsum(N2)),'blue', 'LineWidth',2, 'MarkerEdgeColor', 'r', 'MarkerSize',4);
hold all
[N1,X1] = hist(papr2,100);
plot(X1,1-cumsum(N1)/max(cumsum(N1)),'green', 'LineWidth',2, 'MarkerEdgeColor', 'r', 'MarkerSize',4);
hold all;
[N,X] = hist(papr1,100);
%plot(X,1-cumsum(N)/max(cumsum(N)),'black', 'LineWidth',1, 'MarkerEdgeColor', 'r', 'MarkerSize',2);
hold all;
grid on;
grid minor;
hleg=legend('CARRIER AGGREGATED','SLM','PTS');
xlabel('CA PAPR, x dB')
ylabel('CCDF')
axis ([0 25 0 1]);