OFDM PAPR REDUCTION WITH LINEAR CODING

Thesis submitted in the partial fulfillment of requirement for the award of degree of

Master of Engineering
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Submitted by

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DECLARATION

I, Sumeet Singh, hereby certify that the work which is being presented in this thesis entitled "OFDM PAPR REDUCTION WITH LINEAR CODING" by me in partial fulfillment of the requirements for the award of degree of Master of Engineering in Electronics and Communication Engineering from Thapar University (Deemed University), Patiala, is an authentic record of my own work carried out under the supervision of Mr. Karamjeet Sandha.

The matter presented in this thesis has not been submitted in any other University/Institute for the award of any other degree.

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It is certified that the above statement made by the student is correct to the best of my knowledge and belief.

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ABSTRACT

Wireless communications have been developed widely and rapidly in the modern world especially during the last decade. Recent advances in wireless communication systems have increased the throughput over wireless channels. The reliability of wireless communication has also been increased. But still the bandwidth and spectral availability demands are endless. The need to achieve reliable wireless systems with high spectral efficiency, low complexity and good error performance results in continued research in this field

In this thesis, reduction of the Peak-to-Average Power Ratio (PAPR) of Orthogonal Frequency Division Multiplexing (OFDM) is studied. A PAPR reduction method is proposed that is based on block coding the input data and modifying the codeword. The method makes use of the error correction capability of the block code employed. The performance of the algorithm has been investigated through theoretical models and computer simulations. The algorithm performance is examined through computer simulations and it is found that power reductions are obtained.

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CHAPTER ONE

INTRODUCTION

Multicarrier communications is a technique that has recently seen rising popularity in wireless and wireline applications [1], [2], [3]. In the last years wireless communications have experienced a fast growth due to the high mobility that they allow. However, wireless channels have some disadvantages, like multipath fading, that make them difficult to deal with. A modulation that efficiently deals with selective fading channels is orthogonal frequency division multiplexing (OFDM). The increasing interest in this technique can be ascribed to the advancing capabilities of digital signal processors. International standards making use of OFDM for wireless LAN's are currently being established by IEEE 802.11 and ETSI BRAN committees. For wireless applications, OFDM-based systems can be of interest because they can provide a greater immunity to impulse noise and fast fades and eliminate the need for equalizers, while efficient hardware implementations for small numbers of carriers can be realized using fast Fourier transform (FFT) techniques.

OFDM (orthogonal frequency division multiplexing) have been proposed for many different types of systems from DAB (digital audio broadcasting) to radio LANs (local area networks). Orthogonal frequency-division multiplexing (OFDM) is a method of transmitting data simultaneously over multiple equally spaced carrier frequencies, using Fourier transform processing for modulation and demodulation. The method has been proposed or adopted for many types of radio systems such as wireless local-area networks [4] and digital audio and digital video broadcasting [5]. OFDM offers many well-documented advantages for multicarrier transmission at high data rates, particularly in mobile applications. Specifically, it has inherent resistance to dispersion in the propagation channel. Furthermore, when coding is added it is possible to exploit frequency diversity in frequency-selective fading channels to obtain excellent performance under low signal-to-noise conditions. For these reasons OFDM is often preferable to constant envelope modulation with adaptive equalization (and indeed is arguably less complex to implement).

The advantage of such schemes is that unlimited transmission rates are theoretically possible in highly time dispersive channels. Also, by introducing a ‘guard-period’ and using differential encoding, reliable transmission over spectrally shaped channels is possible without any equalisation.

The disadvantage is that multicarrier signals exhibit a high peak-to-average power ratio (PAPR). If nonlinearities are overloaded by large signal peaks, intermodulation among subcarriers and undesired out-of-band radiation is caused. Hence, amplifiers must operate with large power back-offs to keep out-of-band power below specified limits. Linear and consequently inefficient amplifiers are required for the amplification of these signals to avoid distortion and spectral spreading. If the peak transmit power is limited, either by regulatory or application constraints, this has the effect of reducing the average power allowed under OFDM relative to that under constant power modulation techniques. This in turn reduces the range of OFDM transmissions. Moreover, to prevent spectral growth of the OFDM signal in the form of intermodulation among subcarriers and out-of-band radiation, the transmit amplifier must be operated in its linear region (i.e., with a large input backoff), where the conversion from dc to RF power is inefficient. This may have a deleterious effect on battery lifetime in mobile applications. In many low-cost applications the drawbacks of high PAPR outweigh all the potential benefits of OFDM systems. In addition to this, some hand regulations specify a PEP (peak envelope power) limit for a given band, and consequently the choice of a multicarrier transmission scheme, for a system, does not make the most of this type of power limit. There is clearly a necessity to limit the PAPR of multicarrier signals.

PAPR reduction techniques have been proposed to reduce the PAPR problem in OFDM transmitter. Some techniques use coding, in which data sequence is embedded in a larger sequence and only a subset of all the possible sequences are used to exclude patterns with high PAPR. While the coding technique reduces PAPR, the also reduces transmission rate, significantly so for a large number of subcarriers. Multiple signal representation techniques have been proposed. These include such as amplitude clipping, partial transmit sequence (PTS) [6], selected mapping (SLM) [7], [8], and block coding [9]. The block coding scheme is attractive due to its inherent error control capability, tightly controlled PAPR, and efficient encoding and decoding schemes. The basic principle of block coding in phase shift keying (PSK) modulation is to use the Golay complementary sequences as the transmitted OFDM sequences. These techniques require side information to be transmitted from the transmitter to the receiver to recover the original data block from the received signal. The OFDM sequences constructed from Golay sequences not only enjoy the low PAPR values but also have better error-correcting capabilities in PSK constellations.

In this thesis a method based on coding is proposed to reduce the PAPR value of an OFDM signal. The method uses Golay Complementary coding and the error correction capability of the code. In the literature there are a number of similar methods proposed to reduce the PAPR value. Their performance is evaluated according to the final PAPR value achieved only. However, some of the methods result in an increase in Bit Error Rate (BER) and some of the methods require transmission of side information on a secure channel for the correct reception of the transmitted data. These effects need to be considered for accurate performance calculations. Hence, there is a deficiency in the performance evaluation criterion. This thesis provides a means of performance evaluation of a PAPR reducing algorithm.

This thesis is organized as follows. Chapter 2 is a rather detailed overview of OFDM. Main equations are derived and main techniques of OFDM such as windowing, guard Interval & cyclic prefixing, synchronization, bit loading and PAPR are explained. Afterwards, applications of OFDM are discussed.

Chapter 3 provides a deep insight into the PAPR problem by discussing the methods used in the literature. Block diagram of each method is provided.

In Chapter 4, the PAPR reduction method suggested in the thesis is explained. Detailed Encoding and Decoding Methods are discussed to clarify how the algorithm works. It also contains the graphs explaining system performance. Additionally, this chapter provides the numerical results obtained by algorithm implementation.

Chapter 5 provides the Literature review. In this chapter, various literature works are discussed.

Chapter 6 gives the Gaps in study and Thesis objective.

Finally, Chapter 7 includes some concluding remarks.

CHAPTER TWO

OVERVIEW OF OFDM

In this chapter, first the detail overview of OFDM will be mentioned including history and general structure of an OFDM system. The second part will be about main techniques of OFDM. Finally, applications of OFDM will be examined.

2.1 History of OFDM

OFDM is a special case of multicarrier transmission, where a single data stream is transmitted over a number of lower-rate subcarriers (SCs). It is worth mentioning here that OFDM can be seen as either a modulation technique or a multiplexing technique. One of the main reasons to use OFDM is to increase robustness against frequency-selective fading or narrowband interference. In a single-carrier system, a single fade or interferer can cause the entire link to fail, but in a multicarrier system, only a small percentage of the SCs will be affected. Error-correction coding can then be used to correct for the few erroneous SCs. The concept of using parallel-data transmission and frequency-division multiplexing (FDM) was developed in the mid-1960s [21]. Some early development is traced back to the 1950s. A U.S. patent was filed and issued in January 1970.

In a classical parallel-data system, the total signal frequency band is divided into N nonoverlapping frequency subchannels. Each subchannel is modulated with a separate symbol, and then the N subchannels are frequency multiplexed. It seems good to avoid spectral overlap of channels to eliminate interchannel interference. However, this leads to inefficient use of the available spectrum. To cope with the inefficiency, the ideas proposed in the mid-1960s were to use parallel data and FDM with overlapping subchannels, in which each, carrying a signaling rate b, is spaced b apart in frequency to avoid the use of high-speed equalization and to combat impulsive noise and multipath distortion, as well as to use the available bandwidth fully.

Figure 2.1 illustrates the difference between the conventional nonoverlapping multicarrier technique and the overlapping multicarrier modulation technique. By using the overlapping multicarrier modulation technique, we save almost 50% of bandwidth. To realize this technique, however, we need to reduce cross talk between SCs, which means that we want orthogonality between the different modulated carriers.

The word “orthogonal” indicates that there is a precise mathematical relationship between the frequencies of the carriers in the system. In a normal FDM system, many carriers are spaced apart in such a way that the signals can be received using conventional filters and demodulators. In such receivers, guard bands are introduced between the different carriers and in the frequency domain, which results in a lowering of spectrum efficiency.

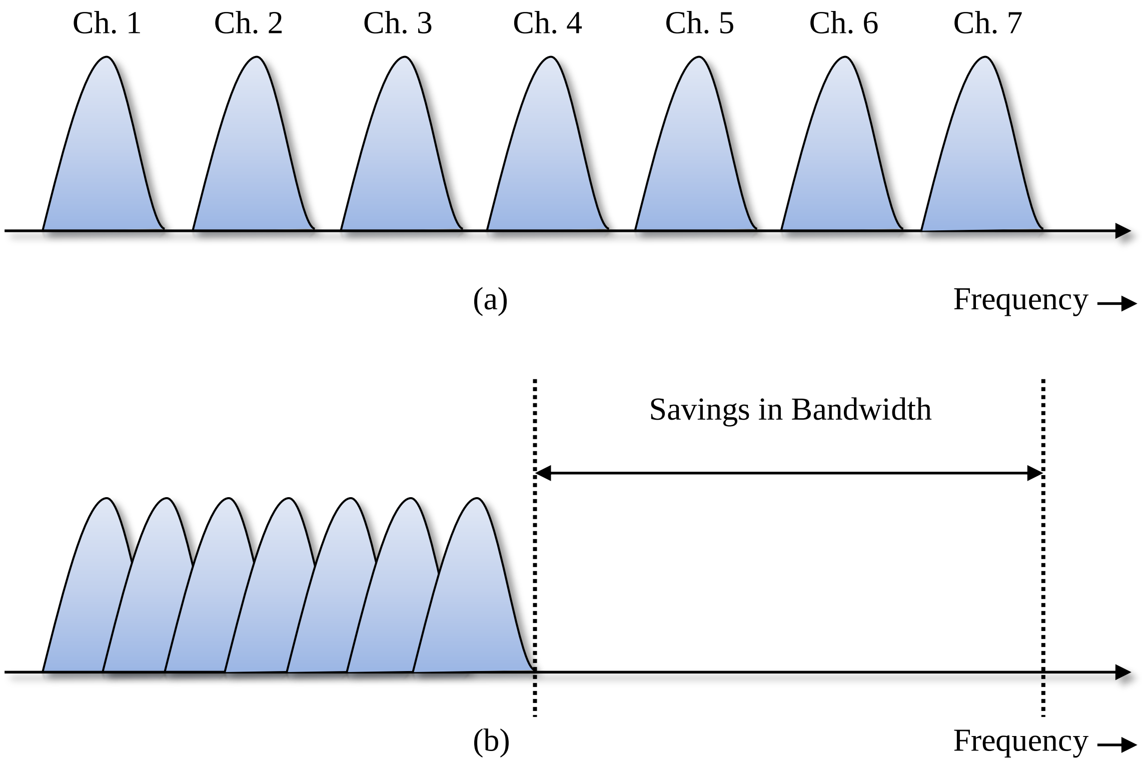
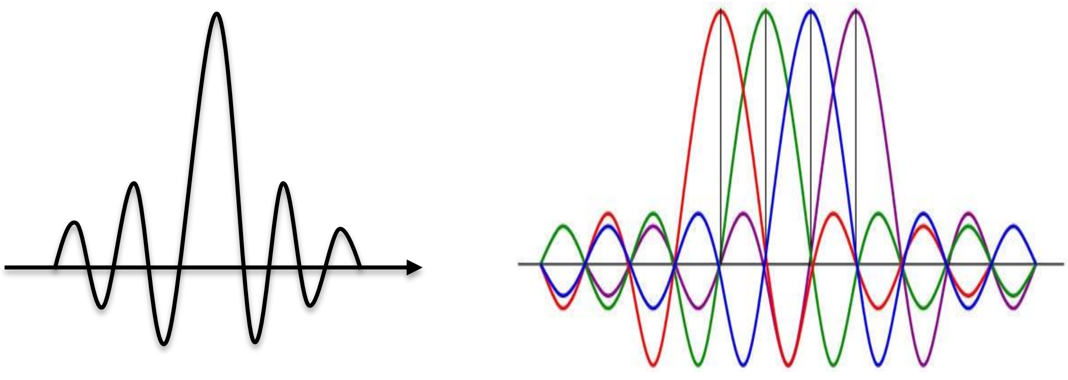


Figure 2.1 Concept of the OFDM signal: (a) conventional multicarrier technique, and (b) orthogonal multicarrier modulation technique.

It is possible, however, to arrange the carriers in an OFDM signal so that the sidebands of the individual carriers overlap and the signals are still received without adjacent carrier interference. To do this the carriers must be mathematically orthogonal. The receiver acts as a bank of demodulators, translating each carrier down to dc, with the resulting signal integrated over a symbol period to recover the raw data. If the other carriers all beat down the frequencies that, in the time domain, have a whole number of cycles in the symbol period T, then the integration process results in zero contribution from all of these other carriers. Thus, the carriers are linearly independent (i.e., orthogonal) if the carrier spacing is a multiple of 1/T.

Much of the research focuses on the highly efficient multicarrier transmission scheme based on “orthogonal frequency” carriers. In 1971, Weinstein and Ebert applied the discrete Fourier transform (DFT) to parallel-data-transmission systems as part of the modulation and demodulation process. Figure 2.2(a) shows the spectrum of the individual data of the subchannel. The OFDM signal, multiplexed in the individual spectra with a frequency spacing b equal to the transmission speed of each SC, is shown in Figure 2.2(b). Figure 2.2 shows that at the center frequency of each SC, there is no cross talk from other channels. Therefore, if we use DFT at the receiver and calculate correlation values with the center of frequency of each SC, we recover the transmitted data with no cross talk.



(a) (b)

Figure 2.2 Spectra of (a) an OFDM subchannel and (b) an OFDM signal.

Moreover, to eliminate the banks of SC oscillators and coherent demodulators required by FDM, completely digital implementations could be built around special-purpose hardware performing the fast Fourier transform (FFT), which is an efficient implementation of the DFT. Recent advances in very-large-scale integration (VLSI) technology make high-speed, large-size FFT chips commercially affordable. Using this method, both transmitter and receiver are implemented using efficient FFT techniques that reduce the number of operations from N^2 in DFT to N logN.

In the 1960s, the OFDM technique was used in several high-frequency military systems such as KINEPLEX, ANDEFT, and KATHRYN. For example, the variable-rate data modem in KATHRYN was built for the high-frequency band. It used up to 34 parallel low-rate phase-modulated channels with a spacing of 82 Hz.

In the 1980s, OFDM was studied for high-speed modems, digital mobile communications, and high-density recording. One of the systems realized the OFDM techniques for multiplexed quadrature amplitude modulation (QAM) using DFT [10]; also, by using pilot tone, stabilizing carrier and clock frequency control and trellis coding could also be implemented. Moreover, various-speed modems were developed for telephone networks.

In the 1990s, OFDM was exploited for wideband data communications over mobile radio FM channels, high-bit-rate digital subscriber lines (HDSL; 1.6 Mbps), asymmetric digital subscriber lines (ADSL; up to 6 Mbps), very-high-speed digital subscriber lines (VDSL; 100 Mbps), digital audio broadcasting (DAB), and high definition television (HDTV) terrestrial broadcasting [11–13].

2.2 OFDM Introduction and System Model

OFDM is a parallel transmission scheme, where a high-rate serial data stream is split up into a set of low-rate substreams, each of which is modulated on a separate subcarrier (FDM). Thereby, the bandwidth of the subcarriers becomes small compared with the coherence bandwidth of the channel; that is, the individual subcarriers experience flat fading, which allows for simple equalization. This implies that the symbol period of the substreams is made long compared to the delay spread of the time-dispersive radio channel.

OFDM transmits data by using a large number of narrow-band subcarriers. These subcarriers are regularly spaced in frequency, forming a block of spectrum. The frequency spacing and time synchronization of the subcarriers is chosen in such a way that the subcarriers are orthogonal, meaning that they do not cause interference to one another. This is despite the subcarriers overlapping each other in the frequency domain. By selecting a special set of (orthogonal) carrier frequencies, high spectral efficiency is obtained because the spectra of the SCs overlap, while mutual influence among the SCs can be avoided. The derivation of the system model shows that by introducing a cyclic prefix, the orthogonality can be maintained over a dispersive channel. In OFDM, subcarrier spacing is kept at minimum, while still preserving the time domain orthogonality between subcarriers, even though the individual frequency spectrum may overlap. The minimum subcarrier spacing should equal to 1/T, where T is the symbol period.

Hereafter are summarized the main advantages and drawbacks of OFDM systems.

First the main advantages :

- high spectral efficiency. The signal is built in frequency domain so it can be shaped in order to use the available bandwidth as efficiently as possible.
- efficient in multipath environments. OFDM increases the symbol duration by N, being N the number of subcarriers and this together with the cyclic prefix can completely eliminate ISI. Because the bandwidth of each subcarrier is narrow compared to the coherence bandwidth of the channel, fading can be considered as flat which reduces the complexity of the receiver.
- simple digital realization by using the FFT operation. Thanks to the advances in digital signal processing and VLSI, the realization of the FFT operations is now simple to implement.
- low complex receivers due to avoidance of ICI and ISI. Each sub-channel can be considered separately since ICI does not affect the signal and each channel suffers flat fading so complex equalizers can be avoided.

- different modulation schemes can be used on individual sub-carriers. This way more robust modulations can be used at the subcarriers that suffer more fading, if channel estimation is used.

And the main drawbacks are:

- accurate frequency and time synchronization is required.
- more sensitive to Doppler spreads than single-carrier schemes.
- sensitive to frequency offset and phase noise caused by imperfections in the transmitter and the receiver oscillators.
- guard interval causes loss in spectral efficiency.
- high PAPR.

2.2.1 OFDM System Model

Mathematically, the OFDM signal is expressed as a sum of the prototype pulses shifted in the time and frequency directions and multiplied by the data symbols. An OFDM signal in baseband is defined as:

$$x(t) = \sum_{n=0}^{N-1} \left(a_n \, e^{j2\pi f_n t} \, w(t)\right) \qquad 0 \leq t \leq T, \qquad (2.1)$$

where, $a_n$denotes the complex symbol modulating the n-th carrier, w(t) is the time window function defined in the interval [0,T], N is the number of subcarriers, and T is the duration of an OFDM symbol. Subcarriers are spaced Δf=1/T apart. Each subcarrier is located at:

$$f = \frac{n}{T}\,, \qquad 0 < n < N-1$$

In order to maintain the orthogonality between the OFDM symbols, the symbol duration and subchannel space must meet the condition TΔf = 1. In (2.1) time-limited complex exponential signals $\left\{ \left(e^{j2\pi f_n t}\right)_{n\,=\,0}^{N\,-\,1} \right.$ which represent the different subcarriers at $f = \frac{n}{T}$

in the OFDM signal. These signals are defined to be orthogonal if the integral of the products for their common (fundamental) period is zero, that is :

$$\rho_{ki} = \left(\frac{1}{T}\right)\int_0^T e^{j2\pi f_k t}\, e^{-j2\pi f_i t}\, dt$$

$$\rho_{ki} = \left(\frac{1}{T}\right)\int_0^T \exp(\mathrm{j}\frac{2\pi kt}{T}) \exp(-\mathrm{j}\frac{2\pi it}{T})\, dt$$

$$\rho_{ki} = \left(\frac{1}{T}\right)\int_0^T \exp(\mathrm{j}\frac{2\pi(k-i)t}{T})\, dt \tag{2.2}$$

As can be seen from (2.2),

$$\rho_{ki} = \begin{cases} 1, & i = k \\ 0, & i \neq k \end{cases} \tag{2.3}$$

Therefore, OFDM signal of the form (2.1) satisfies the condition of mutual orthogonality between subcarriers in the symbol interval. The k-th subcarrier OFDM symbol should be down converted with a frequency of k/T for data modulation, and then integrated over the symbol period [14]. We can assume that w(t) is a rectangular window defined in [0,T] for simplicity. Then above operation may be shown as:

Resultant signal,

$$S_k = \left(\frac{1}{T}\right)\int_0^T \exp(-\,\mathrm{j}\frac{2\pi kt}{T}) \sum_{n=0}^{N-1}\left(a_n \exp(\mathrm{j}\frac{2\pi nt}{T})\right)\, dt$$

$$S_k = \sum_{n=0}^{N-1}(a_n)\left(\frac{1}{T}\right)\int_0^T \exp(-\,\mathrm{j}\frac{2\pi kt}{T})\, \exp(\mathrm{j}\frac{2\pi nt}{T})\, dt \tag{2.4}$$

From (2.1) to (2.4), the complex equivalent lowpass signal transmitted can be directly given. The complex envelope of the OFDM signal is written as:

$$x(t) = \sum_{i} x(t - iT)$$

$$x(t) = \sum_{i} \sum_{n=0}^{N-1} \left(a_{n,i} \exp(\mathrm{j} \frac{2\pi n(t - iT)}{T}) \right) w(t - iT) \tag{2.5}$$

Where, $a_{n,i}$ represents the data constellations the n-th carrier of the i-th OFDM symbol.

In a digital system, this modulated waveform can be generated by an IDFT or by its computationally efficient implementation, the IFFT. The data constellations $a_{n,i}$ are the input to this IFFT, the TD OFDM symbol is its output.

The influence of the time-variant, multipath fading radio channel is expressed by its (lowpass equivalent) IR $h(\tau,t)$, plus additive white Gaussian noise (AWGN) $n(t)$:

$$r(t) = h(\tau, t) * x(t) + n(t) = \int_{0}^{\tau_{max}} h(\tau, t) x(t - \tau) d\tau + n(t) \tag{2.6}$$

Two assumptions are made to simplify the derivation of the received signal. The channel is considered quasistatic during the transmission of the kth OFDM symbol; thus, $h(\tau,t)$ simplifies to $h_k(\tau)$. Furthermore, the maximum excess delay $\tau_{max} < T_{guard}$. Therefore, there is no interference of one OFDM symbol on the effective period of the consecutive one; that is, ISI is suppressed in case of sufficiently accurate time synchronization.

Assuming knowledge of the exact time instants kT at which the OFDM symbols start, we try to extract the transmitted signal constellations $a_{n,i}$ from the received signal $r(t)$.The received signal constellations are denoted $b_{n,i}$.

$$b_{n,i} = \frac{1}{T_0} \int\limits_{t=kT}^{kT+T_0} r(t) e^{-j2\pi i(t-kT)/T_0} \, dt \tag{2.7}$$

$$b_{n,i} = \frac{1}{T_0} \int\limits_{t=kT}^{kT+T_0} \left[\int_0^{\tau_{max}} h(\tau,t) x(t-\tau) d\tau + n(t) \right] e^{-j2\pi i(t-kT)/T_0} \, dt \tag{2.8}$$

$$b_{n,i} = \frac{1}{T_0} \int\limits_{t=kT}^{kT+T_0} \left[\int_0^{\tau_{max}} h(\tau,t) \sum_{n=0}^{N-1} \left(a_{n,i} e^{\frac{j2\pi n(t-kT-\tau)}{T_0}} \right) d\tau + n(t) \right] e^{-j2\pi i(t-kT)/T_0} \, dt$$

We finally obtain,

$$b_{n,i} = a_{n,i} h_{n,i} + n_{n,i}$$

Channel estimation is required in order to retrieve the data contained in these signal constellations because the receiver must have a phase (and amplitude) reference to detect the transmitted symbol correctly.

2.2.2 OFDM Block Diagram and Transmission

To overcome the frequency selectivity of the wideband channel experienced by single-carrier transmission, multiple carriers (OFDM) can be used for high rate data transmission. Figure 2.3(a) shows the basic structure and concept of a OFDM transmission system [5]. Here, a wideband signal is into several narrowband signals at the transmitter and is at the receiver so that the frequency-selective wideband channel can be approximated by multiple frequency-flat narrowband channels as depicted in Figure 2.3(b). Note that the frequency non-selectivity of narrowband channels reduces the complexity of the equalizer for each subchannel. As long as the orthogonality among the subchannels is maintained, the ICI (inter-carrier interference) can be suppressed, leading to distortionless transmission [5].

Orthogonal frequency division multiplexing (OFDM) transmission scheme is another type of a multichannel system, which is similar to the FMT transmission scheme in the sense that it employs multiple subcarriers. It does not use individual bandlimited filters and oscillators for each subchannel and furthermore, the spectra of subcarriers are overlapped for bandwidth efficiency, unlike the FMT scheme where the wideband is fully divided into N orthogonal narrowband subchannels. The multiple orthogonal subcarrier signals, which are overlapped in spectrum, can be produced by generalizing the single-carrier Nyquist criterion into the multi-carrier criterion. In practice, discrete Fourier transform (DFT) and inverse DFT (IDFT) processes are useful for implementing these orthogonal signals. Note that DFT and IDFT can be implemented efficiently by using fast Fourier transform (FFT) and inverse fast Fourier transform (IFFT), respectively.

In the OFDM transmission system, N-point IFFT is taken for the transmitted symbols $\left\{[X(k)]_{k=0}^{N-1}\right.$, so as to generate $\left\{[x(n)]_{n=0}^{N-1}\right.$, the samples for the sum of N orthogonal subcarrier signals. Let y[n] denote the received sample that corresponds to x[n] with the additive noise w[n] (i.e., y[n] = x[n] + w[n]). Taking the N-point FFT of the received samples, $\left\{[y(n)]_{n=0}^{N-1}\right.$, the noisy version of transmitted symbols $\left\{[Y(k)]_{k=0}^{N-1}\right.$, can be obtained in the receiver. The inherent advantages of the OFDM transmission will be detailed later in this chapter. As all subcarriers are of the finite duration T, the spectrum of the OFDM signal can be considered as the sum of the frequency shifted sinc functions in the frequency domain as illustrated in Figure 2.4, where the overlapped neighbouring sinc functions are spaced by 1/T. The discrete multi-tone (DMT) scheme used in ADSL (Asymmetric Digital Subscriber Line) and Zipper-based VDSL (Very high-rate Data digital Subscriber Line) also has the same structure as OFDM.

Since each subcarrier signal is time-limited for each symbol (i.e., not band-limited), an OFDM signal may incur out-of-band radiation, which causes non-negligible adjacent channel interference (ACI). It is clearly seen from Figure 2.5 that the first side lobe is not so small as compared to the main lobe in the spectra. Therefore, OFDM scheme places a guard band at outer subcarriers, called virtual carriers (VCs), around the frequency band to reduce the out-of-band radiation. The OFDM scheme also inserts a guard interval in the time domain, called cyclic prefix (CP), which mitigates the inter-symbol interference (ISI) between OFDM symbols.

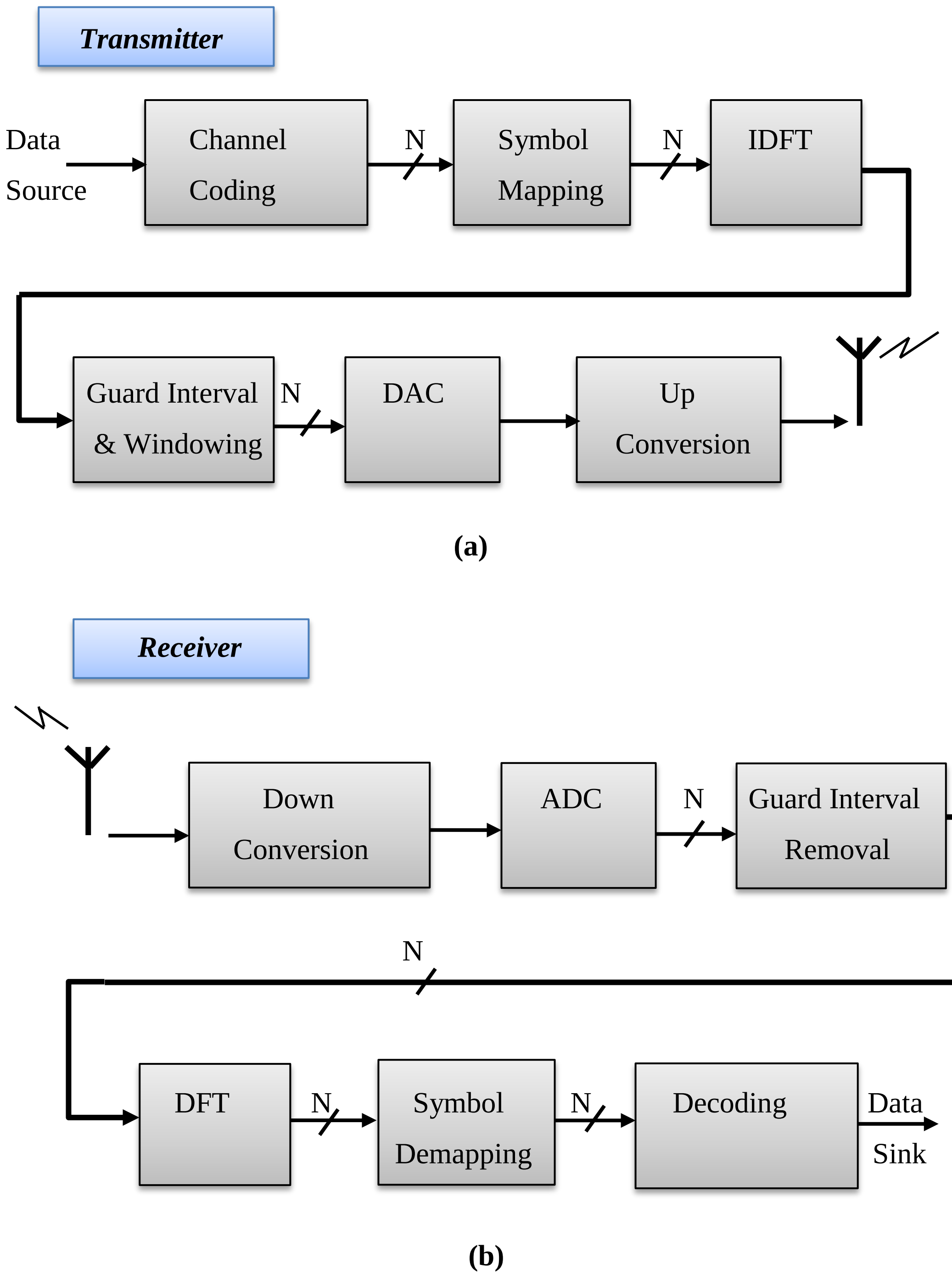


Figure 2.3 Block Diagram of an OFDM Transmitter & Receiver.

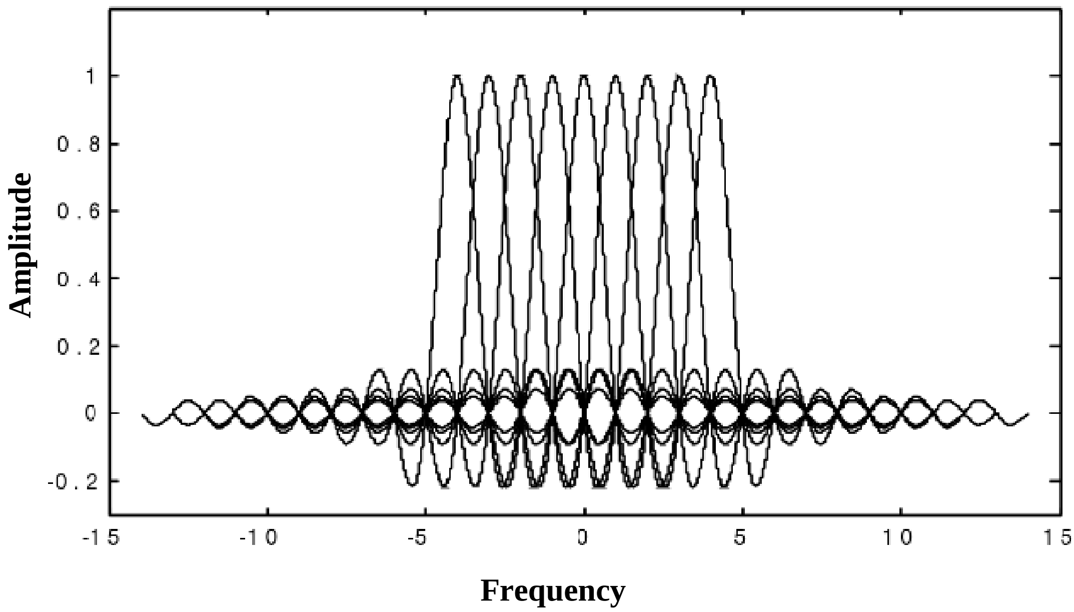


Figure 2.4 The spectrum of OFDM signal.

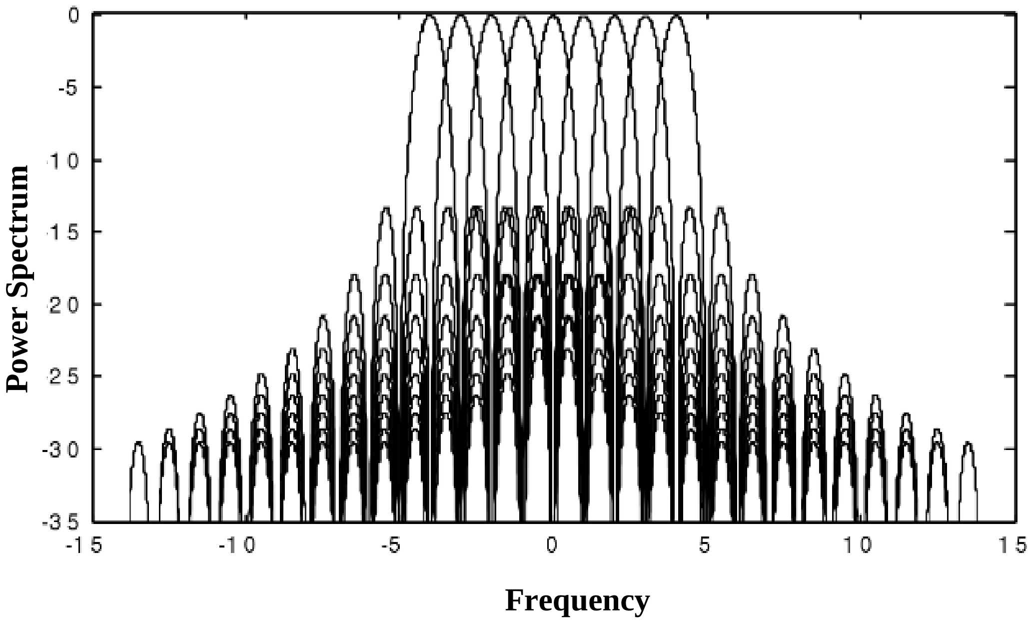


Figure 2.5 Power spectrum of OFDM signal (dB).

2.2.3 Design of OFDM Signal

The proposal of a realistic OFDM-based communications system was one of the goals of this research project. Therefore, we elaborate here on some hardware related design considerations, which are often neglected in theoretical studies. Elements of the transmission chain that have impact on the design of the transmitted OFDM signal include the following:

- ❖ The time-dispersive nature of the mobile channel. The transmission scheme must be able to cope with this.
- ❖ The bandwidth limitation of the channel. The signal should occupy as little bandwidth as possible and introduce a minimum amount of interference to systems on adjacent channels.
- ❖ The TF of the transmitter/receiver hardware. This TF reduces the useable bandwidth compared to the theoretical one given by the sampling theorem. That is, some oversampling is required.
- ❖ Phase jitter and frequency offsets of the up- and down-converters, and Doppler spreading of the channel.

2.2.3.1 Windowing

A rectangular pulse has a very large bandwidth due to the sidelobes of its FT being a sinc function. Windowing is a well-known technique to reduce the level of these sidelobes and thereby reduce the signal power transmitted out of band. In an OFDM system, the applied window must not influence the signal during its effective period. Therefore, cyclically extended parts of the symbol are pulse shaped as depicted in Figure 2.6 [15]. Note that this additional cyclic prefix extends the GI to some extent; that is, the delay-spread robustness is slightly enhanced. On the other hand, the efficiency is further reduced, as the window part is also discarded by the receiver. The orthogonality of the SCs of the OFDM signal is restored by the rectangular receiver filter implemented by the DFT, requiring the correct estimation of the DFT start time kT, where T is the OFDM symbol period. The symbol periods in Figure 2.6 are given as times. Since the implementation is usually done on digital hardware, those periods are also often defined in terms of samples. N, N_{guard}, and N_{win} then define the number of samples in the effective part, guard, and windowing interval, respectively. The effective part is also referred to as the FFT part because this part of the OFDM symbol is applied to the FFT to recover the data at the receiver.

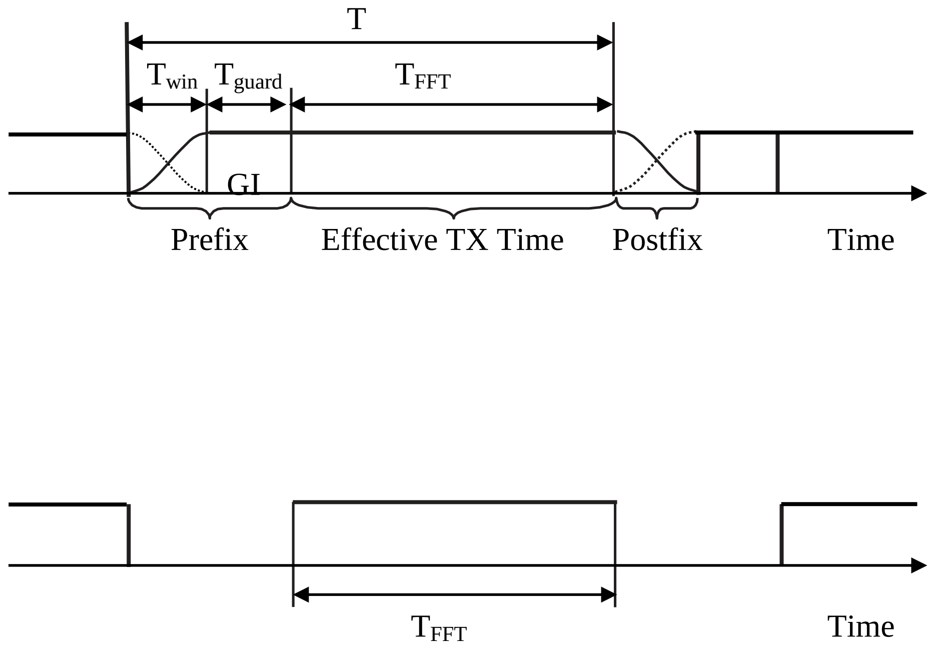


Figure 2.6 Cyclic extension and windowing of the OFDM symbol.

2.2.3.2 Guard Interval and Cyclic Prefix

We may insert zero into the guard interval. This particular approach is adopted by multiband-OFDM (MB-OFDM) in an UltraWide-band (UWB) system. Figure 2.7 show OFDM symbols with GI. Even with the length of ZP longer than the maximum delay of the multipath channel, a small STO causes the OFDM symbol of an effective duration to have a discontinuity within the FFT window and therefore, the guard interval part of the next OFDM symbol is copied and added into the head part of the current symbol to prevent ICI.

Since the GI is filled with zeros, the actual length of an OFDM symbol containing GI is shorter than that of an OFDM symbol containing CP and accordingly, the length of a rectangular window for transmission is also shorter, so that the corresponding sinc-type spectrum may be wider. This implies that compared with an OFDM symbol containing CP, an OFDM symbol containing GI has PSD (Power Spectral Density) with the smaller inband ripple and the larger out-of-band power, allowing more power to be used for transmission with the peak transmission power fixed.

Note that the data rate of the OFDM symbol is reduced by $T / T_{sym} = T / (T + T_{guard})$ times due to the guard interval.

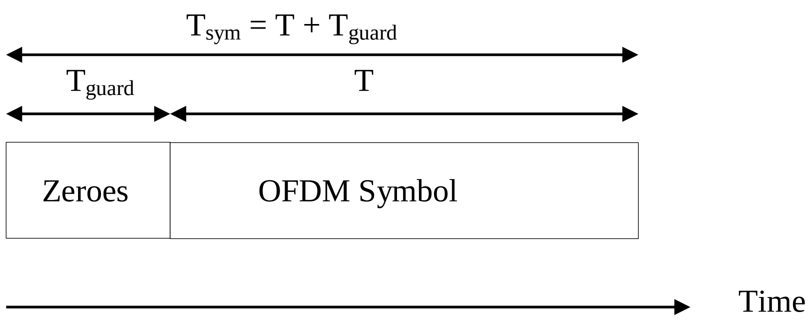


Figure 2.7 Insertion of Guard Interval.

The OFDM guard interval can be inserted in two different ways. One is the zero padding (ZP) that pads the guard interval with zeros. The other is the cyclic extension of the OFDM symbol (for some continuity) with CP (cyclic prefix. CP is to extend the OFDM symbol by copying the last samples of the OFDM symbol into its front. Let T_{guard} denote the length of CP in terms of samples. Then, the extended OFDM symbols now have the duration of $T_{sym} = T + T_{guard}$. Figure 2.8 shows OFDM symbol, which has the CP of length T_{guard}, while illustrating the OFDM symbol of length $T_{sym} = T + T_{guard}$. It can be seen from this figure that if the length of the guard interval (CP) is set longer than or equal to the maximum delay of a multipath channel, the ISI effect of an OFDM symbol on the next symbol is confined within the guard interval so that it may not affect the FFT of the next OFDM symbol, taken for the duration of T. This implies that the guard interval longer than the maximum delay of the multipath channel allows for maintaining the orthogonality among the subcarriers. As the continuity of each delayed subcarrier has been warranted by the CP, its orthogonality with all other subcarriers is maintained over T, such that

$$\frac{1}{T}\int_{0}^{T} e^{j2\pi f_k(t-t_0)}\, e^{-j2\pi f_i(t-t_0)}\, dt = 0\,, \qquad k \neq i$$

for the first OFDM signal that arrives with a delay of t_0, and

$$\frac{1}{T}\int_{0}^{T} e^{j2\pi f_k(t-t_0)}\, e^{-j2\pi f_i(t-t_0-T_s)}\, dt = 0\,, \qquad k \neq i$$

for the second OFDM signal that arrives with a delay of $t_0 + T_s$.

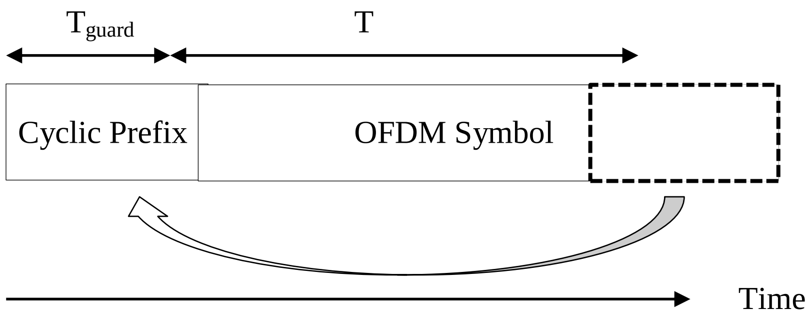


Figure 2.8 Insertion of Cyclic Prefix.

2.2.4 Synchronization

OFDM, like any other digital communication system, requires synchronization. However, OFDM as a multicarrier system has a different structure than a single-carrier system and so has different requirements and different resources. For example in OFDM, one can tolerate larger errors in estimating the start of a symbol than in a single-carrier system. This is due to OFDM's longer symbol period and its cyclic prefix. On the other hand, frequency synchronization in OFDM must be tighter than that in single-carrier systems, due to the narrowness of the OFDM subcarriers. In terms of resources, OFDM has a structure that is not available in single-carrier systems that is useful for synchronization. For example, most OFDM systems have a cyclic prefix that, as we will see, can be used for synchronization. The cyclic prefix can act as pilot data. Often, an OFDM symbol itself is used as pilot data. In this case, the structure of the OFDM symbol can be exploited for time and frequency offset estimation. The choice of pilots versus no-pilots depends on many parameters: the operating SNR, the size of the cyclic prefix, coherent versus differential modulation. Whether to insert pilot data or use OFDM symbols as pilot data often depends on how much overhead the system can tolerate.

Frequency offsets are typically introduced by a (small) frequency mismatch in the local oscillators of the transmitter and the receiver. Doppler shifts can be neglected in indoor environments. The impact of a frequency error can be seen as an error in the frequency instants, where the received signal is sampled during demodulation by the FFT. Figure 2.9 depicts this twofold effect. The amplitude of the desired SC is reduced, and ICI arises from the adjacent SCs.

Mathematically, a carrier offset can be accounted for by a frequency shift δf and a phase offset θ in the lowpass equivalent received signal

$$r'(t) = r(t)e^{j(2\pi\delta f t+\theta)}$$

With (2.7) we obtain

$$b_{n,i} = \frac{1}{T_0}\int_{t=kT}^{kT+T_0} r(t)e^{j(2\pi\delta f t+\theta)}e^{-j2\pi i(t-kT)/T_0}\,dt \tag{2.9}$$

$$b_{n,i} = e^{j2\pi\theta}\frac{1}{T_0}\int_{t=kT}^{kT+T_0}\left[\int_0^{\tau_{max}} h(\tau,t)x(t-\tau)d\tau + n(t)\right]e^{j2\pi\delta f t}e^{-j2\pi i(t-kT)/T_0}\,dt \tag{2.10}$$

The ICI term can be seen as an additional noise term and can thus be represented as a degradation of SNR. et al. [16] have determined the amount of degradation for AWGN channels, and Moose [17] has done so for dispersive fading channels. Frequency offsets up to 2% of the SC spacing F are negligible, according to their results. Even 5% to 10% can be tolerated in many situations.

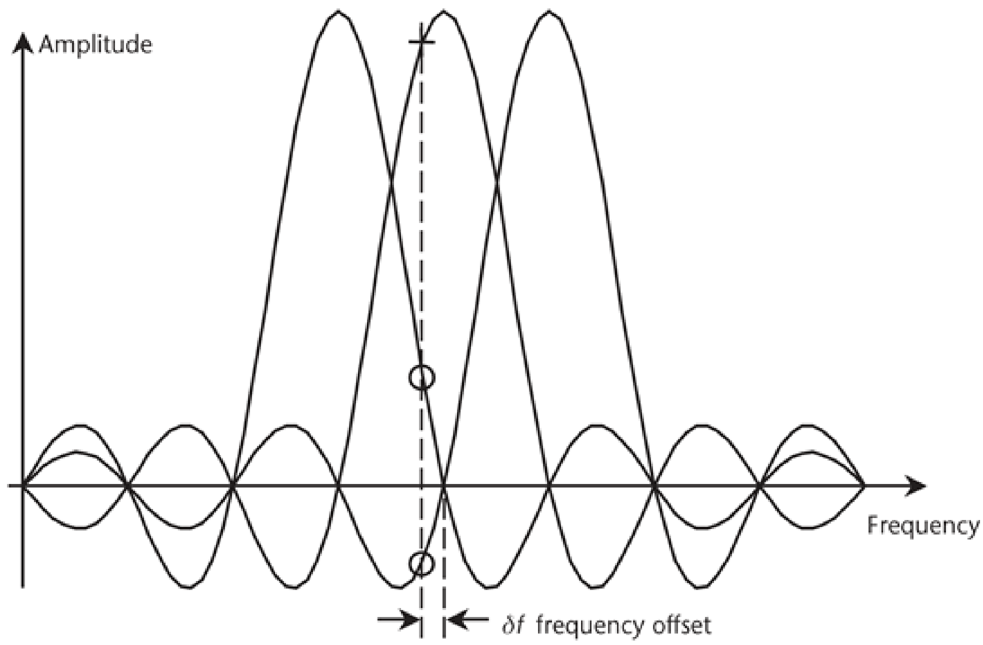


Figure 2.9 Inter-carrier interference (ICI) subject to carrier frequency offset.

2.2.4.1 Timing and Frame Synchronization

Initially a frame (packet or slot) synchronization circuit is required to detect the frame starting point. This is usually achieved by correlating the incoming signal with a known preamble. The same circuit is sometimes used to adjust for initial gain control. Therefore, the threshold of the detection circuit should be adjusted accordingly. A general block diagram is shown in Figure 2.10. Since timing offset does not violate orthogonality of the symbols it can be compensated after the receiver FFT.

The effect of the timing offset is a phase rotation which linearly increases with sub-carrier order. In order to estimate the timing offset, we should solve a regression problem with proper weighting to represent the effect of fading in the channel.

Frequency offset estimation in packet data communications requires fast acquisition. Hence, phase-lock techniques are not applicable for initial coarse correction. For this

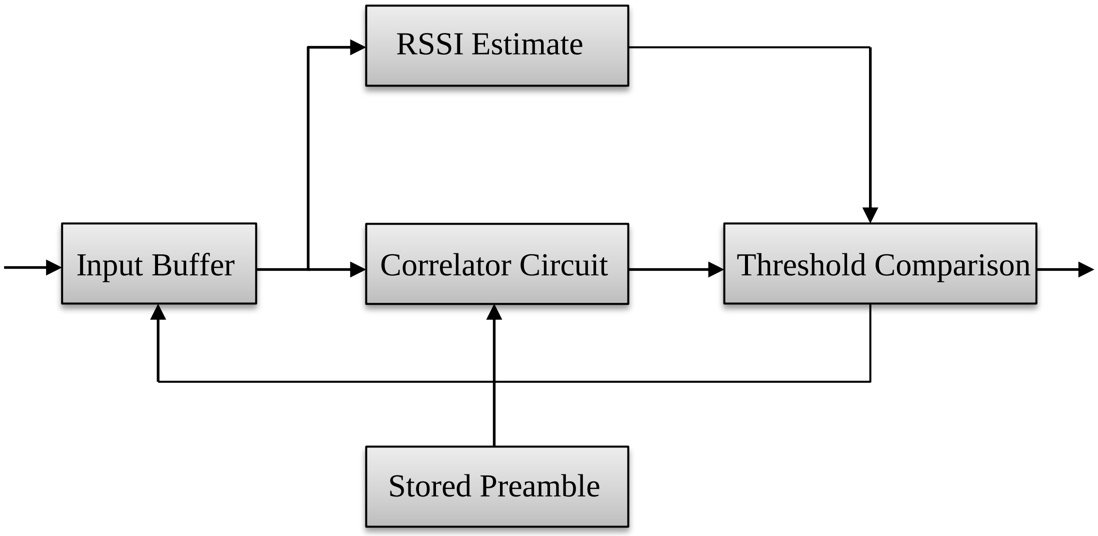


Figure 2.10 Frame synchronization.

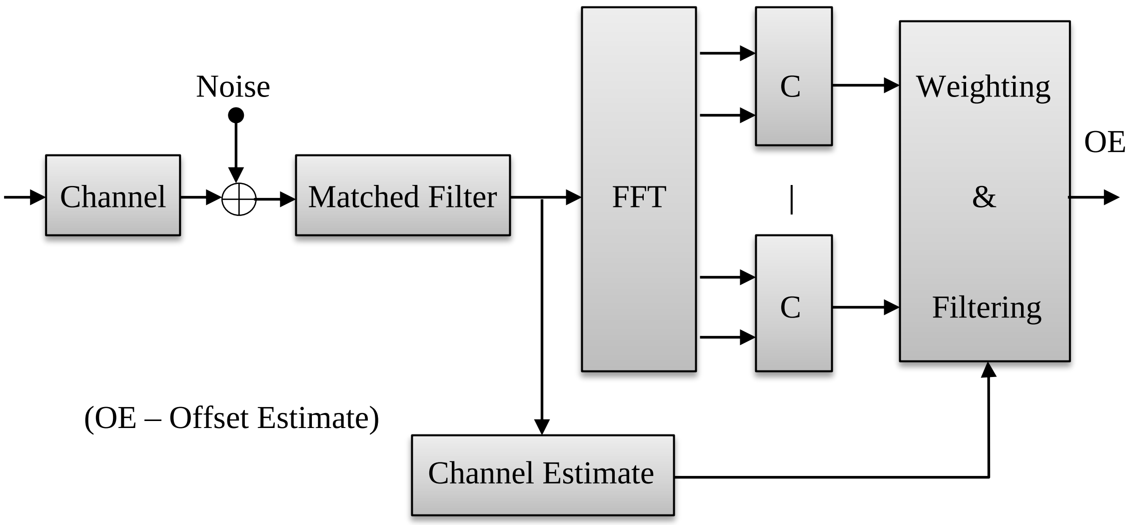


Figure 2.11 Timing offset estimate.

purpose, a pilot or training sequence should be transmitted. This can be achieved by sending a pilot in one or two bins and estimate the frequency offset by measuring the energy spread in other bins, or by using a training sequence or pilot in every bin and measuring the phase difference by repeating the same pattern. Usually, the second approach is more reliable, as additive noises in different bins are independent and the estimate convergence will be faster. In the following, we discuss both techniques and compare the result analytically and through simulation.

2.2.4.2 Frequency Synchronization

One of the main drawbacks of OFDM is its sensitivity to carrier frequency offset. The carrier frequency error results in SNR degradation, caused by two main phenomena:

- Reduction of amplitude of the desired subcarrier.
- ICI caused by neighboring carriers.

The amplitude loss occurs because the desired subcarrier is no longer sampled at the peak of the sinc function of DFT. The sinc function is defined as $sinc(x) = \frac{\sin(x)}{x}$. Adjacent carriers cause interference, because they are not sampled at the zero-crossings of they sinc functions. The overall effect on SNR is analyzed in [16], and for relatively small frequency errors, the degradation in dB was approximated by

$$SNR_{LOSS} = \frac{10}{3ln10}\,(\pi T f_e)^2 \frac{E_s}{N_0}\; dB$$

where f_e is the frequency error as a fraction of the subcarrier spacing and T is the sampling period. The performance effect varies strongly with the modulation used, naturally, constellations with fewer points can tolerate larger frequency errors than large constellations. Large constellations are exceedingly difficult to use with an OFDM system. However, keep in mind that a large constellation automatically implies higher operating SNR than with a small constellation. This directly improves the performance of the frequency error estimators.

In [18], the various algorithms that have been developed to estimate carrier frequency offsets in OFDM systems are divided into three types:

Type I	Data-aided algorithms; these methods are based on special training information embedded into the transmitted signal.

Type II	Non data-aided algorithms that analyze the received signal in frequency domain.
Type III	Cyclic prefix based algorithms that use the inherit structure of the OFDM signal provided by the cyclic prefix.

For WLAN applications, type 1 is the most important. The preamble allows the receiver to use efficient maximum likelihood algorithms to estimate and correct for the frequency offset, before the actual information portion of the packet starts. The algorithms belonging to types 2 and 3 are better suited for broadcast or continuous transmission type OFDM systems. Owing to the nature of this book, we will focus on data-aided frequency synchronization algorithms in this section.

2.2.4.3 Effect of Phase Noise

Intermediate stages of modulation, such as between RF and IF, can introduce degradation due not only to frequency error as described previously, but also to phase noise in the local oscillators involved in such modulation. Phase noise is the zero mean random process of the deviation of the oscillator's phase from a purely periodic function at the average frequency. The oscillator output may be considered to be an ideal sine wave phase modulated by the random process. The power spectral density is generally normalized to the power of the sine wave. Higher quality (and more costly) oscillators will typically have lower phase noise than cheaper ones. Therefore, one may expect greater phase noise at the mobile set in a wireless system, as opposed to the base station, and similarly in the customer terminal as opposed to the central office in a subscriber line transmission system.

2.2.5 Bit loading

Another important concept about OFDM is the bit loading. If the available transmit power and input data is equally divided among the subcarriers, the technique is named as *fixed* or *uniform loading*. The subcarriers at which the channel attenuation is maximum, mostly determine the performance of a fixed loading OFDM system under Rayleigh fading. Under frequency selective fading, such systems may perform poorly. An efficient method to cope with this problem is to employ *adaptive loading* of the available transmit power and input bits over the subcarriers according to channel state. By this way, subcarriers experiencing severe fading are loaded with a small number of bits, or may not be used at all. The advantage it provides in increasing the capacity can be computed for fading channels as described in [19].

2.3 Peak-to-Average Power Ratio (PAPR)

2.3.1 Introduction

An OFDM signal consists of a number of independently modulated SCs, which can give a large peak-to-average power ratio (PAPR) when added up coherently. When N signals are added with the same phase, they produce a peak power that is N times the average power. This effect is illustrated in Figure 2.12. For this example, the peak power is 16 times the average value. The peak power is defined as the power of a sine wave with an amplitude equal to the maximum envelope value. Hence, an unmodulated carrier has a PAPR of 0 dB. An alternative measure of the envelope variation of a signal is the Crest factor, which is defined as the maximum signal value divided by the RMS signal value. For an unmodulated carrier, the Crest factor is 3 dB. This 3-dB difference between the PAP ratio and Crest factor also holds for other signals, provided that the center frequency is large in comparison with the signal bandwidth.

A large PAPR brings disadvantages like an increased complexity of the analog-to-digital (A/D) and digital-to-analog (D/A) converters and a reduced efficiency of the RF power amplifier. To reduce the PAPR, several techniques have been proposed, which basically can be divided in three categories. First, there are signal distortion techniques, which reduce the peak amplitudes simply by nonlinearly distorting the OFDM signal at or around the peaks. Examples of distortion techniques are clipping, peak windowing, and peak cancellation. Second, there are coding techniques that use a special FEC code set that excludes OFDM symbols with a large PAPR. The third technique scrambles each OFDM symbol with different scrambling sequences and selecting the sequence that gives the smallest PAPR. This chapter discusses all of these techniques, but first analyzes the PAPR distribution function. This will give a better insight in the PAPR problem and will explain why PAPR reduction techniques can be quite effective.

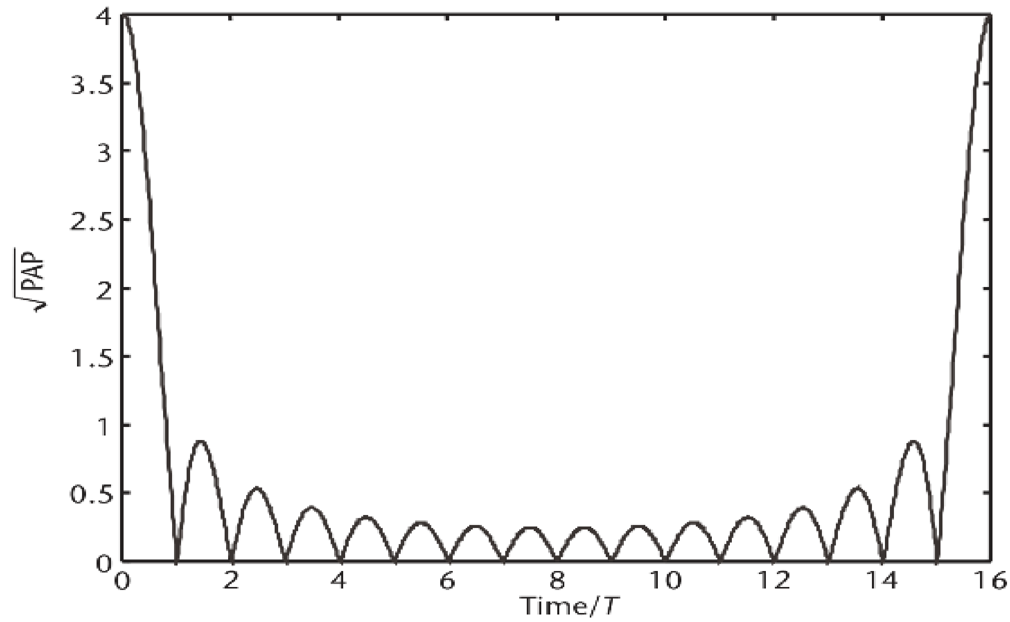


Figure 2.12 Square root of PAP ratio for a 16-channel OFDM signal.

2.3.2 Peak to Average Power Ratio(PAPR)

Presence of large number of independently modulated sub-carriers in an OFDM system the peak value of the system can be very high as compared to the average of the whole system. This ratio of the peak to average power value is termed as Peak-to-Average Power Ratio. Coherent addition of N signals of same phase produces a peak which is N times the average signal.

The major disadvantages of a high PAPR are :

1. Increased complexity in the analog to digital and digital to analog converter.
2. Reduction is efficiency of RF amplifiers.

2.3.3 PAPR of a Multicarrier Signal

Let the data block of length N be represented by a vector $X = [\, X_0, X_{1,\ldots,}\ X_{N-1}\,]^T$. Duration of any symbol X_k in the set $\boldsymbol{X}$ is T and represents one of the sub – carriers $\{\, f_{n,}\ n = 0,1,\ldots, N\text{-}1 \,\}$ set. As the N sub – carriers chosen to transmit the signal are orthogonal to each other, so we can have $f_n = n\Delta f$ where $n\Delta f = 1/NT$ and NT is the duration of the OFDM data block $\boldsymbol{X}$. The complex data block for the OFDM signal to be transmitted is given by

$$x(t) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} X_n e^{j2\pi n \Delta f t}, \qquad 0 \leq t \leq NT \tag{2.11}$$

The PAPR of the transmitted signal is defined as

$$\mathrm{PAPR} = \frac{\max\ |\mathrm{x(t)}|^2}{\frac{1}{\mathrm{NT}} \int_0^{NT} |\mathrm{x(t)}|^2\, dt}$$

Reducing the $max|x(t)|$ is the principle goal of PARP reduction techniques. Since, discrete- time signals are dealt with in most systems, many PAPR techniques are implemented to deal with amplitudes of various samples of $x(t)$. Due to symbol spaced output in the first equation we find some of the peaks missing which can be compensated by oversampling the equation by some factor to give the true PAPR value.

2.3.4 Cumulative Distribution Function

The Cumulative Distribution Function (CDF) is one of the most regularly used parameters, which is used to measure the efficiency of any PAPR technique. Normally, the Complementary CDF (CCDF) is used instead of CDF, which helps us to measure the probability that the PAPR of a certain data block exceeds the given threshold.

By implementing the Central Limit Theorem for a multi – carrier signal with a large number of sub-carriers, the real and imaginary part of the time – domain signals have a mean of zero and a variance of 0.5 and follow a Gaussian distribution. So Rayleigh distribution is followed for the amplitude of the multi – carrier signal, where as a central chi-square distribution with two degrees of freedom is followed for the power distribution of the system. The CDF of the amplitude of a signal sample is given by,

$$F(z) = 1 - \exp(z) \tag{2.12}$$

The CCDF of the PAPR of the data block is desired is our case to compare outputs of various reduction techniques. This is given by

$$\begin{aligned} P(PAPR > z) &= 1 - \mathrm{P}(PAPR \leq z) \\ &= 1 - \mathrm{F}(z)^N \\ P(PAPR > z) &= 1 - (1 - \exp(-z))^N \end{aligned} \tag{2.13}$$

2.4 Applications of OFDM

OFDM is used as the transmission technique in digital audio and television broadcasting applications, and in wireless LAN applications. This section describes the main applications of OFDM.

2.4.1 Digital Audio Broadcasting (DAB)

Current analog FM radio broadcasting system cannot satisfy the demands of the future, which are :

- Excellent sound quality.
- Large number of stations.
- Small portable receivers.
- No quality impairment due to multipath propagation or signal fading.

Current analog FM radio broadcasting systems have reached the limits of technical improvement. DAB is a digital technology offering considerable advantages over today's FM radio.

Eureka project EU 147 : DAB

- Launched at 1986.
- First phase: 4 year plan (1987-1991) of research and development.
- Participants from Germany, France, Netherlands and United Kingdom.

- Second phase (1992-1994, 170 man-years) : completion development of individual system specifications, development of ASICs, and considerations of additional services.

Key Features :

- Ability to deliver CD-quality stereo sound.
- Ease of use of DAB receivers.
- Switch between the eight or more stations carried by every single multiplex.
- No need for drivers to retune as they cross a country.
- Wider choice of programs.
- Each multiplex is able to carry up to six full-quality stereo programs.
- DAB can carry text and images as well as sound.
- All but the smallest will be able to display at least two 16-character lines of text.
- Selection by name or programme type.
- Enabling broadcasters to transmit programme-associated data (PAD) such as album title, song lyrics, or contact details.
- DAB can be transmitted at lower power than today's FM and AM services without loss of coverage.
- DAB combines two advanced digital technologies to achieve robust and spectrum-efficient transmission of high-quality audio and other data.
- DAB uses the MPEG Audio Layer II system to achieve a compression ratio of 7:1 without perceptible loss of quality.
- The signal is then encoded at a bit rate of 32 to 384 kbps, depending on the desired sound quality and the available bandwidth.
- Signal is individually error protected and labeled prior to multiplexing. Independent data services are similarly encoded.
- The coded orthogonal frequency division multiplex (COFDM) technology is used for transmission.
- 2.3 million bits of the multiplexed signal in time and across 1,536 distinct frequencies within the 1.5 MHz band.
- An conventional FM network must use different frequencies in each area. In a DAB network, all transmitters operate on a single frequency.

- ❖ Such a single frequency network (SFN) makes DAB's use of the radio spectrum over three times more efficient than conventional FM.
- ❖ DAB is designed for terrestrial, cable and for future satellite broadcasts.

2.4.2 HDTV - Digital Video Broadcasting (DVB)

- ❖ Audio and video-centric
- ❖ Very large files transmission.
- ❖ Quality of service issues.
    - o Guaranteed bandwidth
    - o Jitter
    - o Delay
- ❖ Large, scalable audience
- ❖ Broadband downstream, narrowband up
- ❖ Satellite: DVB-S
- ❖ Terrestrial: ATSC, DVB-T
- ❖ European standard for transmission of digital TV via satellite, cable or terrestrial :
    - o DVB-S (satellite)
        - ▪ QPSK – quadrature phase-shift keying.
    - o DVB-T (terrestrial)
        - ▪ COFDM – coded orthogonal frequency division multiplexing.
- ❖ MPEG-2 compression and transport stream.
- ❖ Support for multiple, encrypted program stream.

2.4.3 Magic WAND

The Magic WAND (Wireless ATM Network Demonstrator) project was a part of European ACTS (Advanced Communications Technology and Servers) program. The Magic WAND consortium members implemented a prototype wireless ATM network based on OFDM. This prototype had a large impact on standardization activities in the 5 GHz band. The wireless ATM application based on OFDM forms the basis for the standardization of the HYPERLAN type I physical layer.

The number of subcarriers is 16, with a guard time of 400 ns. With this guard time, rms

delay spreads up to 100 ns are completely tolerated without employing equalization. While this is sufficient for most of the office buildings, a realistic product would require more delay spread robustness in order to cover large office buildings and factory halls.
The subcarrier modulation is 8 PSK with a symbol rate of 13.3 M symbols/s, yielding a raw data rate of 40 Mbps. The rate 1/2 complementary coding reduces the net rate to 20 Mbps. The subcarrier spacing is 1.25 MHz, which gives a total bandwidth of 20 MHz. The packet preamble is 8.4 μs in duration and consists of one OFDM symbol, repeated 7 times. This preamble is used for packet detection, automatic gain control, frequency offset estimation, symbol timing and channel estimation. Transmission is done on half-slots, which consists of 9 OFDM symbols, carrying 27 bytes. An ATM cell is mapped on two consecutive half-slots, which form a full-slot. Of the 54 bytes carried on one full slot, 53 bytes are allocated to ATM cell traffic and remaining 1 byte is reserved for physical layer control signaling.

The WAND modem uses complementary codes for both forward-error correction and PAP reduction. Length 8 complementary codes are used, with 4 8-PSK symbols input, generating eight complex code outputs. By taking the IFFT of this complex coded output, an OFDM signal with a low PAP ratio is obtained. To encode 16 subcarriers, two length 8 codes are interleaved, which maximizes the benefits of frequency diversity. The minimum distance of the length 8 code is 4 symbols, and hence three arbitrary subcarriers can be erased without causing errors, provided that the data loaded onto the remaining subcarriers are not in error.

2.4.4 Wireless LAN Networks

1. IEEE 802.11 - The first international standard for WLAN, 1997.
    a. Infrared (IR) baseband PHY (1Mbps, 2Mbps).
    b. Frequency hopping spread spectrum (FHSS) radio in 2.4GHz band (1Mbps, 2Mbps).
    c. Direct sequence spread spectrum (DSSS) radio in the 2.4GHz band (1Mbps, 2Mbps).

2. IEEE 802.11a, 1999.
    a. 5GHz band.

   b. Orthogonal frequency division multiplexing (OFDM).
   c. 6Mbps to 54Mbps.

3. IEEE 802.11g.
   a. (802.11b + 80211a) operating at 2.4GHz band.
   b. ERP-DSS/CCK: IEEE 802.11b-1999.
   c. ERP-OFDM: IEEE 802.11a-1999.
   d. PBCC (optional).
   e. CCK-OFDM (optional).

4. IEEE 802.11h/D2.2, 2002.
   a. Radar detection in 5GHz band.
   b. Regulatory (ETSI EN 301 893 v.1.2.1).
   c. Power control.

2.4.5 IEEE 802.16 - Broadband Wireless Access System (BWAS)

BWAS is the broadband wireless technology used to deliver voice, data, Internet, and video service in the 25-GHz and higher spectrum.

Properties of 802.16 :

- ❖ Broad BW : up to 134 Mbps in 28 MHz wide channel (10-66 GHz).
- ❖ Support simultaneous multiple services with full QoS.
    - o IPv4, IPv6, ATM, Ethernet, etc.
- ❖ BW on demand (per frame).
- ❖ MAC designed for efficient spectrum use.
- ❖ Comprehensive, modern and extensible security.
- ❖ Support multiple frequency allocation from 2-66 GHz.
    - o OFDM & OFDMA for NLOS applications.

CHAPTER THREE

PAPR REDUCTION TECHNIQUES

3.1 Introduction

PAPR reduction techniques are classified into the different approaches: clipping technique, coding technique, probabilistic (scrambling) technique, adaptive predistortion technique, and DFT-spreading technique.

- The clipping technique employs clipping or nonlinear saturation around the peaks to reduce the PAPR. It is simple to implement, but it may cause in-band and out-of-band interferences while destroying the orthogonality among the subcarriers. This particular approach includes block-scaling technique, clipping and filtering technique, peak windowing technique, peak cancellation technique, Fourier projection technique, and decision-aided reconstruction technique [20,21].

- The coding technique is to select such codewords that minimize or reduce the PAPR. It causes no distortion and creates no out-of-band radiation, but it suffers from bandwidth efficiency as the code rate is reduced. It also suffers from complexity to find the best codes and to store large lookup tables for encoding and decoding, especially for a large number of subcarriers [22]. Golay complementary sequence, Reed Muller code, M-sequence, or Hadamard code can be used in this approach [23].

- The probabilistic (scrambling) technique is to scramble an input data block of the OFDM symbols and transmit one of them with the minimum PAPR so that the probability of incurring high PAPR can be reduced. While it does not suffer from the out-of-band power, the spectral efficiency decreases and the complexity increases as the number of subcarriersincreases. Furthermore, it cannot guarantee the PAPR below a specified level [9], [8]. This approach includes SLM (Selective

Mapping), PTS (Partial Transmit Sequence), TR (Tone Reservation), and TI (Tone Injection) techniques.

- ❖ The adaptive predistortion technique can compensate the nonlinear effect of a high power amplifier (HPA) in OFDM systems [24]. It can cope with time variations of nonlinear HPA by automatically modifying the input constellation with the least hardware requirement (RAM and memory lookup encoder). The convergence time and MSE of the adaptive predistorter can be reduced by using a broadcasting technique and by designing appropriate training signals.

- ❖ The DFT-spreading technique is to spread the input signal with DFT, which can be subsequently taken into IFFT. This can reduce the PAPR of OFDM signal to the level of single-carrier transmission. This technique is particularly useful for mobile terminals in uplink transmission. It is known as the Single Carrier-FDMA (SC-FDMA), which is adopted for uplink transmission in the 3GPP LTE standard [25].

3.2. PAPR Reduction by Clipping and Filtering

The clipping approach is the simplest PAPR reduction scheme, which limits the maximum of transmit signal to a pre-specified level. However, it has the following drawbacks :

- ❖ Clipping causes in-band signal distortion, resulting in BER performance degradation.
- ❖ Clipping also causes out-of-band radiation, which imposes out-of-band interference signals to adjacent channels. Although the out-of-band signals caused by clipping can be reduced by filtering, it may affect high-frequency components of in-band signal (aliasing) when the clipping is performed with the Nyquist sampling rate in the discrete-time domain. However, if clipping is performed for the sufficiently-oversampled OFDM signals (e.g., $L \geq 4$) in the discrete-time domain before a low-pass filter (LPF) and the signal passes through a band-pass filter (BPF), the BER performance will be less degraded.

❖ Filtering the clipped signal can reduce out-of-band radiation at the cost of peak regrowth. The signal after filtering operation may exceed the clipping level specified for the clipping operation [22].

Figure 3.1 shows a typical configuration. The conventional OFDM modulator is followed by a peak-clipping device, and a linear filter for removing the Out of Band (OOB) components. Finally, the up-converter translates the complex baseband signal into a real valued RF signal. The task of the linear filter is to reduce the unwanted OOB noise. Here, complete removal of the OOB components by an ideal low-pass filter is assumed. Of course, in practice, an ideal low-pass filter can only be approximated. Removal of the OOB noise by filtering leads to a regrowth of some of the signal peaks. The amount of CF regrowth may be determined by computer simulation.

While we have described the clipping and filtering approach for the time continuous baseband signal s(t) here, it might be realized in several variations. For example, clipping and filtering can be applied to the transmit signal at IF instead of baseband. In this case only a real-valued signal needs to be processed. Because digital filtering at IF has much higher computational complexity, other filter realizations, like surface acoustic wave (SAW) filters could be suitable solutions.

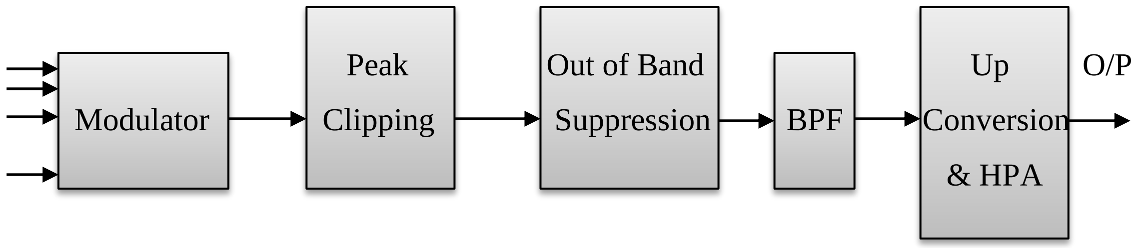


Figure 3.1 Block diagram of a PAPR reduction scheme using clipping and filtering.

Instead of the time-continuous signal (which, in practice, means a Sampled signal with a sufficiently high over-sampling rate), clipping and filtering could be applied to the equidistant samples s(kT/N). In this case the low-pass filtering can be considered part of the interpolation process in the digital to analog conversion.

Instead of clipping with subsequent filtering, the time domain signal s(t) is multiplied with some multiplication signal that drops down at time-instants where large peaks occur. Consequently, the peaks are reduced. The dips in the multiplication signal are smooth and chosen according to some window function that results in small spectral spreading only. In case of a few peaks only, this method proposes lower computational complexity since no additional filtering operation is required.

3.3 Peak Cancellation

The key element of all distortion techniques is to reduce the amplitude of samples whose power exceeds a certain threshold. In the case of clipping and Filtering, this was done by a nonlinear distortion of the OFDM signal, which resulted in a certain amount of out-of-band radiation. This undesirable effect can be avoided by performing a linear peak-cancellation technique, whereby a time-shifted and scaled-reference function is subtracted from the signal, such that each subtracted reference function reduces the peak power of at least one signal sample. By selecting an appropriate reference function with approximately the same bandwidth as the transmitted signal, it can be assured that the peak power reduction will not cause any out-of-band interference. One example of a suitable reference signal is a sinc function. A disadvantage of a sinc function is that it has infinite support. Hence, for practical use, it has to be time limited in some way. One way to do this without creating unnecessary out-of-band interference is to multiply it by a windowing function, for instance, a raised cosine window.

Peak cancellation can be done digitally after generation of the digital OFDM symbols. It involves a peak power detector, a comparator to see if the peak power exceeds the threshold and a scaling of the peak and surrounding samples. Figure 3.2 shows the block diagram of an OFDM transmitter with peak cancellation. Incoming data is first coded and converted from a serial bit stream to blocks of N complex signal samples. An IFFT operation is performed on each of these blocks. Then a cyclic prefix is added, extending the symbol size from N to N+Nc samples. After parallel-to-serial conversion, the peak cancellation procedure is applied to reduce PAPR. Except for the peak cancellation block, there is no difference with a standard OFDM transmitter. There is no difference for the receiver, so any standard OFDM receiver can be used.

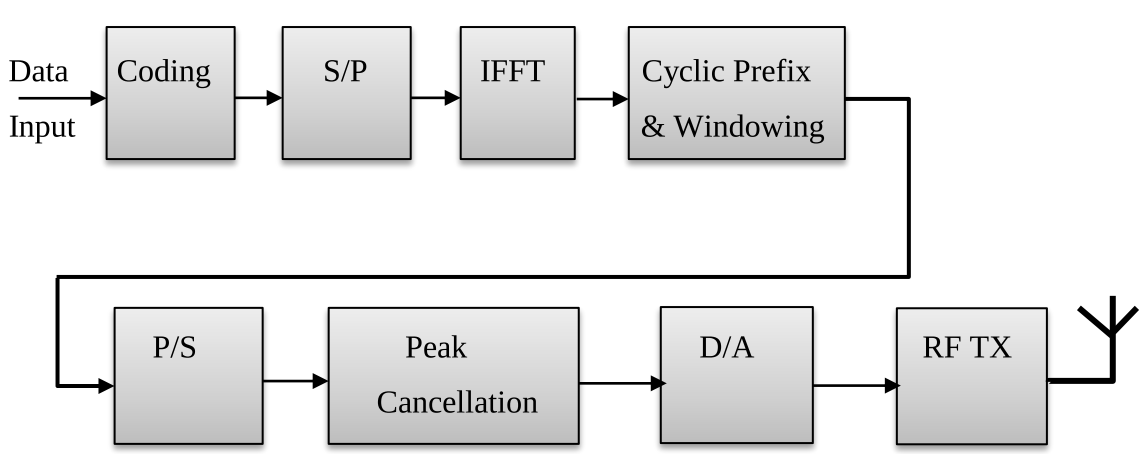


Figure 3.2 OFDM Transmitter with Peak Cancellation.

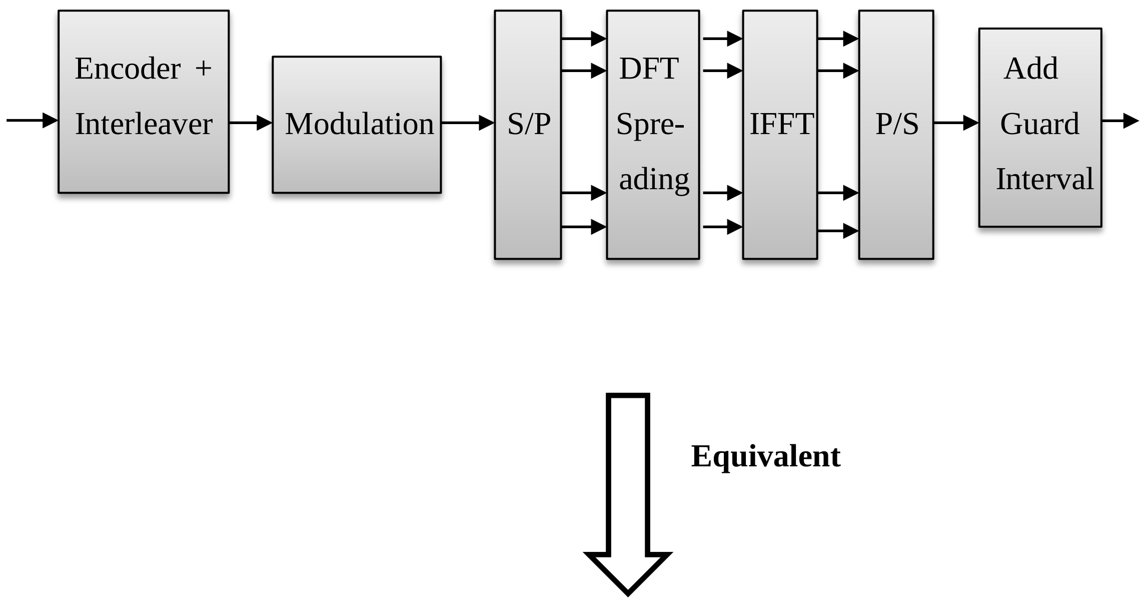
3.4 PAPR Reduction by Scrambling Techniques

The basic idea of symbol scrambling is that, for each OFDM symbol, the input sequence is scrambled by a certain number of scrambling sequences. The output signal with the smallest PAPR is transmitted. For uncorrelated scrambling sequences, the resulting OFDM signals and corresponding PAPR values will be uncorrelated, so if the PAPR for one OFDM symbol has a probability p of exceeding a certain level without scrambling, the probability is decreased to p^k by using k scrambling codes. Hence, symbol scrambling does not guarantee a PAPR below some low level; rather, it decreases the probability that high PAPR values occur. Initially proposed scrambling techniques were *Selected Mapping (SLM), TR (Tone Reservation), TI (Tone Injection) techniques* and *Partial Transmit Sequences (PTS)*. SLM and PTS improve the PAPR statistics by introducing little redundancy and also solve the problem signal distortion techniques which may cause in-band distortion and out-of-band noise.

3.5 DFT Spreading

Before discussing the DFT-spreading technique, let us consider OFDMA (Orthogonal Frequency-Division Multiple Access) system. As depicted in Figure 3.3, suppose that DFT of the same size as IFFT is used as a (spreading) code. Then, the OFDMA system becomes equivalent to the Single Carrier FDMA (SC-FDMA) system because the DFT and IDFT operations virtually cancel each other. In this case, the transmit signal will have the same PAPR as in a single-carrier system.

In OFDMA systems, subcarriers are partitioned and assigned to multiple mobile terminals (users). Unlike the downlink transmission, each terminal in uplink uses a subset of subcarriers to transmit its own data. The rest of the subcarriers, not used for its own data transmission, will be filled with zeros. Here, it will be assumed that the number of subcarriers allocated to each user is M. In the DFT-spreading technique, M-point DFT is used for spreading, and the output of DFT is assigned to the subcarriers of IFFT. The effect of PAPR reduction depends on the way of assigning the subcarriers to each terminal. There are two different approaches of assigning subcarriers among users: DFDMA (Distributed FDMA) and LFDMA (Localized FDMA). Here, DFDMA distributes M DFT outputs over the entire band (of total N subcarriers) with zeros filled in (N - M) unused subcarriers, whereas LFDMA allocates DFT outputs to M consecutive subcarriers in N subcarriers. When DFDMA distributes DFT outputs with equi-distance N / M = S, it is referred to as IFDMA (Interleaved FDMA) where S is called the bandwidth spreading factor.



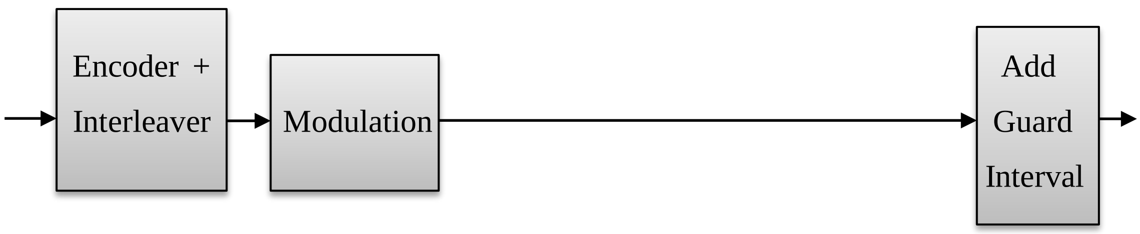


Figure 3.3 Equivalence of OFDMA system with DFT-spreading code to a single-carrier system.

3.6 PAPR Reduction by Coding

Due to bad PAPR of only small fraction of all generated OFDM symbols, another method for PAPR reduction, based on coding is suggested. The PAPR can be reduced by using a code that only produces OFDM symbols with PAPR below some desirable level. Of course, the smaller the desired PAPR level, the smaller the achievable code rate is. A PAPR of the maximum 3dB for the 8-carrier OFDM system can be achieved by 3/4-code rate block coding. Here, a 3-bit data word is mapped onto a 4-bit codeword. Then, the set of permissible code words with the lowest PAPRs in the time domain is chosen. The code rate must be reduced to decrease the desired level of PAPR. The block codes found through an exhaustive search are mostly based on Golay complementary sequence. Golay complementary sequence is defined as a pair of two sequences whose aperiodic autocorrelations sum to zero in all out-of-phase positions. Golay complementary sequences can be used for constructing OFDM signals with PAPR as low as 3dB. The possibility of using complementary codes for both PAPR reduction and forward error correction. Meanwhile, a large set of binary length 2^m Golay complementary pairs can be obtained from Reed-Muller codes. However, the usefulness of these coding techniques is limited to the multicarrier systems with a small number of subcarriers. In general, the exhaustive search of a good code for OFDM systems with a large number of subcarriers is intractable, which limits the actual benefits of coding for PAPR reduction in practical OFDM systems.

A sequence x of length N is said to be complementary to another sequence y if the following condition holds on the sum of both autocorrelation functions :

$$\sum_{k=0}^{N-1}(x_k\, x_{i+k} + y_k\, y_{i+k}) = \begin{cases} 2N, & i = 0 \\ 0, & i \neq 0 \end{cases} \tag{3.1}$$

By taking the FTs of both sides of (3.1), the above condition is translated into

$$|X(f)|^2 + |Y(f)|^2 = 2N \tag{3.2}$$

Here, $|X(f)|^2$ is the power spectrum of x, which is the FT of its autocorrelation function. From the spectral condition (3.2), it follows that the maximum value of the power spectrum is bounded by 2N :

$$|X(f)|^2 \leq 2N$$

Because the average power of $X(f)$ is equal to N, assuming that the power of the sequence x is equal to 1, the PAPR of $X(f)$ is bounded as

$$PAPR \leq \frac{2N}{N} = 2$$

In an OFDM transmission, normally the IFFT is applied to the input sequence x. However, because the IFFT is equal to the conjugated FFT scaled by 1/N, the conclusion that the PAPR is upper bounded by 2 is also valid when $X(f)$ is replaced by the inverse FT of the sequence x. Hence, by using a complementary code as input to generate an OFDM signal, the PAP ratio is not to exceed 3 dB.

CHAPTER FOUR

LITERATURE REVIEW

Richard G. Gibson and Jonathan Jedwab [32] explained the origin of all 4-phase Golay sequences and Golay sequence pairs of even length at most 26. The three-stage construction can also be used to derive minimum counts for 4-phase Golay sequences and sequence pairs of length greater than 26, although a more general result than Proposition 9 is needed for some lengths.

Frank Fiedler, Jonathan Jedwab and Matthew G. Parker [25] identified a framework of constructions from which all known Golay sequences and pairs of length 2^m over $Z_2{}^h$ can be obtained explicitly, and shown the key importance of Turyn's construction and its variations. They also considered the Golay sequences and pairs that can be obtained from an arbitrary initial Golay pair (a, b) by the iterative use of Budisin's construction, including the effect of (negative) reversal of intermediate sequences.

Richard G. Gibson and Jonathan Jedwab [33] presented a 4-phase Golay sequence pair of length s=5 (mod 8) is constructed from a Barker sequence of the same length whose even-indexed elements are prescribed. This explained the origin of the 4-phase Golay seed pairs of length 5 and 13. The construction cannot produce new 4-phase Golay sequence pairs, because there are no Barker sequences of odd length greater than 13.

Jonathan Jedwab and Matthew G. Parker [31] proposed the Constructions and nonexistence conditions for multi-dimensional Golay complementary array pairs. A construction for a d-dimensional Golay array pair from a (d+1)-dimensional Golay array pair is given. This used to explain and expand previously known constructive and nonexistence results in the binary case.

Moataz M. Salah and Ashraf A. Elrahman [14] proposed a modified OFDM scheme for image transmission. This new proposal was based on modification of the OFDM structure through using unequal power allocation for the successive OFDM symbols.

Unequal cyclic time guard was also applied with unequal power allocation. The proposed method was compared with the conventional OFDM.

Zizheng Cao, Jianjun Yu, Minmin Xia [20] theoretically investigated the signal degradation induced by intersubcarrier interference and frequency-selective fading in orthogonal frequency-division-multiplexing radio-over-fiber systems. Turbo codes and bit interleaver technologies were proposed to effectively mitigate unbalanced error distribution because of intersubcarrier interference and frequency-selective fading.

Seon Ae Kim, Dong Geon An [17] proposed a SNR estimation method which uses zero point auto-correlation of received signal per block and auto/cross- correlation of decision feedback signal in orthogonal frequency division multiplexing (OFDM) system. Proposed method can be studied into two types; Type 1 can estimate SNR by zero point auto-correlation of decision feedback signal based on the second moment property. Type 2 uses both zero point auto-correlation and cross-correlation based on the fourth moment property.

Lin Dong and Nam Yul Yu [30] presented a simple but novel technique to develop theoretical PAPR bounds of downlink OFDM system using Golay complementary sequences for spreading and coding. The developed PAPR bounds were independent of the spreading factor in uncoded OFDM. Furthermore, they have no dependency on the number of spreading processes as well as the spreading factor in coded OFDM.

Yen-Wen Huang Ying Li [28] presented a collection of 16-QAM Golay complementary sequences with new sequence lengths 7, 9, and 15. These Golay sequences can be recursively combined to form other 16-QAM Golay sequences with numerous new sequence lengths. The existence of fixed-alphabet Golay complementary sequences with these lengths was previously unknown.

Gaofei WU, Zilong WANG [35] enlarged the family size of near-complementary sequences by employing some new shortened and extended Golay complementary pairs as the seeds given by Yu and Gong. In this paper, some new shortened and extended Golay sequences are given.

John A. C. Bingham [2] discussed the following: the general technique of parallel transmission on many carriers; the performance that can be achieved on an undistorted channel; algorithms for achieving that performance; dealing with channel impairments;

improving the performance through coding; and methods of implementation. Also discuss duplex operation of MCM and the possible use of this on the GSTN.

J.A. Davis and J. Jedwab [27] proposed coding scheme for OFDM transmission, exploiting a previously unrecognised connection between pairs of Golay complementary sequences and second-order Reed-Muller codes. The scheme solved the notorious problem of power control in OFDM systems by maintaining a peak-to-mean envelope power ratio of at most 3dB while allowing simple encoding and decoding at high code rates for binary, quaternary or higher-phase signalling together with good error correction.

Kenneth G. Paterson [29] developed a powerful theory linking Golay complementary sets of polyphase sequences and Reed–Muller codes. The main result showed that any second-order coset of an M-ary generalization of the first order Reed–Muller code can be partitioned into Golay complementary sets whose size depends only on a single parameter that is easily computed from a graph associated with the coset.

Leonard J. Cimini [4] analysed and simulated of a technique for combating the effects of multipath propagation and co-channel interference on a narrow-band digital mobile channel. The system used the discrete Fourier transform to orthogonally frequency multiplex many narrow subchannels, each signaling at a very low rate, into one high-rate channel.

Beeta Tarokh and Hamid R. Sadjadpour [37] presented a technique to derive M-quadrature amplitude modulation (QAM) signals from quaternary phase-shift keying (QPSK) constellations when $M = 2^n$ and n is an even number. By utilizing QPSK Golay sequences, they constructed M-QAM sequences with low peak-to-mean envelope power ratios. Several upper bounds for these M-QAM sequences were derived.

Kenneth G. Paterson and Vahid Tarokh [36] established bounds on the region of achievable triples (R, d, PAPR(C)) where R is the code rate and d is the minimum Euclidean distance of the code. They prove a lower bound on PAPR in terms of R and d and show that there exist asymptotically good codes whose PAPR is at most 8 log n. They give explicit constructions of error-correcting codes with low PAPR by employing bounds for hybrid exponential sums over Galois fields and rings.

CHAPTER FIVE

GAPS IN STUDY & THESIS OBJECTIVE

5.1 Gaps in study

Kenneth G. Paterson, J.A. Davis and J. Jedwab and et. al. have done lots of work and researches on efficient OFDM transmission and PAPR reduction. They have proposed, some of them established various techniques to reduce the PAPR. Some of them worked on clipping technique, some of them worked on PTS , Scrambling techniques, DFT Spreading, coding etc. the most simple way is clipping but it introduces degradation BER and distort the signal. Other techniques are far better techniques than clipping but they require extra information, so the transmission rate decreases. While reducing PAPR, other factors like complexity, transmission rate, BER, error correction etc also is considered.

The most efficient and effective method to reduce the PAPR with better error correction, less complexity, higher code rate, reduced BER is through Coding techniques. Many codes are used to reduce the PAPR, but widely used codes are Golay Complementary codes. Various uses various schemes to generate efficient sequences to reduce the PAPR to minimum level with better error correction and reduced BER. J.A. Davis and J. Jedwab have done a great work in the field of generating efficient Golay complementary sequences. But none of the above has implemented the generated Golay sequences with MATLAB to show the actual performance of the OFDM system.

In this thesis, I have generated the sequences to provide the effective codeword for OFDM transmission by using the previous work done by various and used their work to reduce the PAPR to the minimum possible level at no cost of BER and system performance. Also the codes provide the good error correcting capability.

I have shown the simulation results of the scheme through MATLAB coding. The results show the significant reduction in PAPR with reduced BER and increased system performance.

5.2 Thesis Objective

Although OFDM has some advantages that make it suitable for fading channels, it presents a high peak-to-average power ratio (PAPR), which is one of the main drawbacks of OFDM systems. A simple technique used to reduce the PAPR of OFDM signals is to clip the signal to a maximum allowed value, at the cost of BER degradation and out-of-band radiation, other techniques (Partial Transmit Sequences, Selected Mapping, Peak Cancellation, DFT Spreading) reduce PAPR without degrading the BER but at the cost of adding extra information.

The goal of this thesis is to generate the complementary sequences from Golay codes using second-order Reed-Muller codes. The sequences solves the notorious problem of power control in OFDM systems by maintaining a peak-to-mean envelope power ratio at minimum possible level while allowing simple encoding and decoding at high code rates for binary, quaternary or higher-phase signalling together with good error correction.

The whole scheme is programmed with MATLAB to show the Simulation Results. The graph generated after simulation shows the reduction in Peak to Average Power Ratio (PAPR) and Bit error rate in comparison to SNR.

CHAPTER SIX

PAPR REDUCTION WITH GOLAY COMPLEMENTARY CODES & SYSTEM PERFORMANCE

6.1 PAPR Reduction with Golay complementary codes

6.1.1 Introduction

In this chapter, firstly the generation (Encoding) method is discussed for generating a set of complementary sequences, for symbol encoding to generate the effective codewords. Second section describes an optimal decoding technique for specific subsets of complementary codes, based on generalized Walsh-Hadamard encoding.

6.1.2 PAPR reduction : Algorithm Implementation

6.1.2.1 Encoding

The combination of the new results of [27,28,29,30,31] immediately suggests a practical OFDM coding scheme using 2^h -phase shift keying: allow as codewords only those Golay sequences described in [32,33] (for h = 1 and h > 1). This simultaneously confers tight envelope power control, by [32,33], and good error correction capability, by [34].

The Golay sequences in question occur as m!/2 cosets of RM_2^h (1,m) [35] and for convenience of implementation we use 2^w of these cosets, where 2^w is the largest integer power of 2 no greater than m!/2. Under this scheme we encode w + h(m + 1) information bits per OFDM symbol period. We use w bits to encode the choice of coset representative using a look-up table. The remaining h(m + 1) bits are converted to m + 1 information symbols $u_1, u_2, \ldots\ldots, u_m$, $u \in Z_2^h$ by taking each consecutive group of h bits to be the binary representation of an element of Z_2^h. The information symbols are then used to form the linear combination $\sum_{i=1}^{m} u_i x_i + u$, in which each symbol multiplies one row of the standard generator matrix for RM_2^h (1,m). This linear combination can be calculated in hardware in 2^m clock cycles using the encoding circuit for RM(1, m). The sum (over Z_2^h) of this linear combination with the selected coset representative is the OFDM codeword $(a_0, a_1, \ldots, a_{2^m-1})$, which is modulated prior to transmission . The code rate , namely the ratio of the number of information bits to the number of coded bits, is $(w + h(m + 1))=(2^m h)$, and we de ne the information rate to be h times the code rate. The information rate describes the increased rate at which information bits are encoded when we change the code from binary to quaternary, from quaternary to octary, and so on.

For example, consider the octary case with 16 carriers (h = 3, m = 4). The 12 coset representatives given by [32] are

$$(0\ 0\ 0\ 4\ 0\ 0\ 4\ 0\ 0\ 0\ 0\ 4\ 4\ 4\ 0\ 4) = 4 (x1\, x2 + x2\, x3 + x3\, x4);$$

$$(0\ 0\ 0\ 4\ 0\ 4\ 0\ 0\ 0\ 0\ 0\ 4\ 4\ 0\ 4\ 4) = 4 (x1\, x2 + x2\, x4 + x3\, x4);$$

$$(0\ 0\ 0\ 0\ 0\ 4\ 4\ 0\ 0\ 0\ 4\ 4\ 0\ 4\ 0\ 4) = 4 (x1\, x3 + x2\, x3 + x2\, x4);$$

$$(0\ 0\ 0\ 4\ 0\ 4\ 0\ 0\ 0\ 0\ 4\ 0\ 0\ 4\ 4\ 4) = 4 (x1\, x3 + x3\, x4 + x2\, x4);$$

$$(0\ 0\ 0\ 0\ 0\ 4\ 4\ 0\ 0\ 4\ 0\ 4\ 0\ 0\ 4\ 4) = 4 (x1\, x4 + x2\, x4 + x2\, x3);$$

$$(0\ 0\ 0\ 4\ 0\ 0\ 4\ 0\ 0\ 4\ 0\ 0\ 0\ 4\ 4\ 4) = 4 (x1\, x4 + x3\, x4 + x2\, x3);$$

$$(0\ 0\ 0\ 4\ 0\ 0\ 0\ 4\ 0\ 0\ 4\ 0\ 4\ 4\ 0\ 4) = 4 (x1\, x2 + x1\, x3 + x3\, x4);$$

$$(0\ 0\ 0\ 4\ 0\ 0\ 0\ 4\ 0\ 4\ 0\ 0\ 4\ 0\ 4\ 4) = 4 (x1\, x2 + x1\, x4 + x3\, x4);$$

$$(0\ 0\ 0\ 0\ 0\ 0\ 4\ 4\ 0\ 4\ 4\ 0\ 0\ 4\ 0\ 4) = 4 (x2\, x3 + x1\, x3 + x1\, x4);$$

$$(0\ 0\ 0\ 0\ 0\ 4\ 0\ 4\ 0\ 4\ 4\ 0\ 0\ 0\ 4\ 4) = 4 (x2\, x4 + x1\, x4 + x1\, x3);$$

$$(0\ 0\ 0\ 0\ 0\ 4\ 0\ 4\ 0\ 0\ 4\ 4\ 4\ 0\ 0\ 4) = 4 (x1\, x3 + x1\, x2 + x2\, x4);$$

$$(0\ 0\ 0\ 0\ 0\ 0\ 4\ 4\ 0\ 4\ 0\ 4\ 4\ 0\ 0\ 4) = 4 (x2\, x3 + x1\, x2 + x1\, x4);$$

of which we choose eight (say the first eight), so w = 3. The union of the eight cosets of RM_8 (1,4) having these coset representatives comprises the set of OFDM codewords, all of which have PMEPR of at most 2. The code forms a subcode of ZRM_8 (2,4) [36] and has minimum Hamming and Lee distance 4 and 8 respectively. An error of Hamming weight 1 can always be corrected, as can an error of Lee weight at most 3. The code rate is 3/8 and the information rate is 9/8 [37]. Given 18 information bits, three are used to select one of the eight coset representatives and the remaining 15 are regarded as the binary representation of five information symbols u1, *u2, u3, u4, u*. The linear combination *u1x1 + u2x2 + u3x3 + u4x4 + u* is calculated with reference to the generator matrix given by :

$$\begin{array}{c} \begin{bmatrix} 1&1&1&1&1&1&1&1&1&1&1&1&1&1&1&1\\ 0&0&0&0&0&0&0&0&1&1&1&1&1&1&1&1\\ 0&0&0&0&1&1&1&1&0&0&0&0&1&1&1&1\\ 0&0&1&1&0&0&1&1&0&0&1&1&0&0&1&1\\ 0&1&0&1&0&1&0&1&0&1&0&1&0&1&0&1 \end{bmatrix} \end{array} \begin{array}{c} 1\\ x_1\\ x_2\\ x_3\\ x_4 \end{array}$$

for RM_8 (1,4) and added to the selected coset representative. Suppose the 18 information bits are 011101111011110110. The rest three bits 011 select the coset representative (0004040000400444) (labelling the first eight coset representatives 000, 001,…., 111). The remaining 15 bits select the linear combination 5x1 + 7x2 + 3x3 + 6x4 + 6 = (6417530631642053), so the OFDM codeword is (6413570631242417).

The above coding scheme is restricted to the Golay sequences described in [27,30,31]. These sequences occur as m!/2 Golay cosets of RM_{2^h} (1,m) within a second-order linear code, where the second-order linear code is RM_2 (2,m) in the binary case h = 1 and is ZRM_{2^h} (2,m) in the non-binary cases h > 1.

We can increase the code rate, at the cost of progressively larger values of PMEPR, by including additional cosets of RM_{2^h} (1,m) within the same second-order code. These additional cosets do not necessarily comprise or even contain Golay sequences. Nonetheless we have found that partitioning the second-order code into cosets of RM_{2^h} (1,m) is an effective means of isolating codewords with large values of PMEPR. Alternatively we can increase the minimum Hamming distance, at the cost of a lower code rate, by choosing fewer than 2^w of the original m!/2 Golay cosets. In this way we can trade off code rate, PMEPR and error correction capability to provide a range of solutions to the envelope power problem. For implementation convenience we use $2^{w'}$ cosets of RM_2^h (1,m) for some integer w' to encode w' + h(m + 1) information bits, storing the coset representatives in a look-up table. We can determine the possible options for given h and m by arranging all the cosets of RM_2^h (1,m) (within the appropriate second-order code) in increasing order of their maximum PEP over the $2^{h(m+1)}$ codewords in the coset.

6.1.2.2 Decoding

An important attraction of the binary Golay code for applications purposes is that it is easy to decode. In particular, the first-order code RM_2 (1,m) can be decoded very efficiently by means of the fast Hadamard transform (FHT). In this section we give a fast decoding algorithm for RM_2^h (1,m) for any $h \geq 1$, requiring h FHTs and h encoding operations in RM_2^h (1,m). This algorithm acts as a decoder for RM_2^h (1,m) with respect to both Hamming and Lee distance: it always corrects errors of Hamming or Lee weight less than the limit $d/2 = 2^{m-2}$ guaranteed by the minimum Hamming or Lee distance $d = 2^{m-1}$ of the code [34]. In fact the class of errors which can always be corrected by the algorithm includes many whose Hamming or Lee weight greatly exceeds this limit. The algorithm can be used for soft-decision as well as hard-decision decoding. It is scalable in the sense that the decoder for RM_2^{h+1} (1,m) can be obtained directly from the decoder for RM_2^h (1,m) simply by including one additional iteration. We also extend the decoding algorithm, while maintaining its favourable properties, to deal with an arbitrary union of cosets of RM_2^h (1,m). We remark that [33] give an extension to non-binary cases of the standard FHT method for decoding RM_2 (1,m) but their extension applies to GRM(1,m) rather than to RM_2^h (1,m). We also note that [35] implicitly gives a hard-decision decoder for RM_2^h (1,3) with respect to Hamming distance but does not analyse which errors of Hamming weight greater than 1 can be corrected by this decoder and makes no mention of Lee weight.

We begin by summarising the standard FHT method for decoding RM_2 (1,m) as described in [38]. The Sylvester-Hadamard matrix $H_{2^m} = (H_{ij})$ of order 2^m is given by

$H_{ij} = (-1)^{\sum_{k=1}^{m} i_k j_k}$ for $i,j \in Z_2^m$, where (i_1, i_2,.., i_m) and (j_1 , j_2,…, j_m) are the binary representation of i and j respectively. The Hadamard transform of the row vector y = (y_0, y_1, …, y_{2m-1}) is $\tilde{y} = yH_2^m$. The Hadamard transform $\tilde{y}$ of a sequence y of length 2^m can be calculated rapidly by representing H_2^m as the product of m sparse matrices; we then call y the fast Hadamard transform (FHT) of y.

We now introduce the decoding algorithm by outlining the octary case h = 3. Suppose codeword $c \in RM_8$ (1,m) is received in error as r = (c + e) mod 8, where e is a sequence over Z_8. Write $= \sum_{i=1}^{m}(u_i x_i + u) \ mod\ 8$, where u_i , $u \in Z_8$. Let (v_{i2}, v_{i1}, v_{i0}) be the binary representation of u_i and let (v_2, v_1, v_0) be the binary representation of u, so that $u_i = 4v_{i2} + 2v_{i1} + v_{i0}$ and $u = 4v_2 + 2v_1 + v_0$. Then

$$c = (4f_2 + 2f_1 + f_0) \ mod\ 8, \qquad (6.1)$$

where

$$f_2 = (v_{i2}x_i + v_2) \ mod\ 2, \qquad (6.2)$$

$$f_1 = (v_{i1}x_i + v_1) \ mod\ 4, \qquad (6.3)$$

$$f_0 = (v_{i0}x_i + v_0) \ mod\ 8, \qquad (6.4)$$

Write the error e uniquely as e = $4e_2$ +$2e_1$ +e_0, where each e_k is a sequence over Z_2, so that

$$r = (4 (f_2 +e_2)+2 (f_1 +e_1)+(f_0 +e_0)) \ mod\ 8, \qquad (4.5)$$

Using the FHT, the decoding algorithm recovers the value f_0 by reducing modulo 2, then (assuming f_0 has been determined correctly) the value f_1 by reducing modulo 4, and finally (assuming f_0 and f_1 have been determined correctly) the value f_2, c is then recovered from (6.1).

Algorithm : (Decoding algorithm for RM_{2^h} (1,m))

1. Input the received codeword r as a sequence over Z_{2^h} of length 2^m. Set k = 0 and r_0 = r.

2. Define the sequence y by $(y)_i = 2^{k-1}\, wt_2^{\,k+1}\, ((rk)_i)$ for i = 0, 1,….., $2^m - 1$.

3. Let $\tilde{y}$ be the FHT of y and determine a value of $j \in Z_2^m$ for which $(\tilde{y})_j$ is an element of $\tilde{y}$ of largest magnitude. Let w be 0 or 1 according as $(\tilde{y})_j$ is positive or negative, and let (w_1 , w_2 ,...., w_m) be the binary representation of j. Set $f_k = \sum_{i=1}^{m}(w_i x_i + w) \ mod \ 2^{h-k}$.

4. If k = h - 1 then output the decoded codeword $(2^{h-1} f_{h-1} + 2^{h-2} f_{h-2} + ... + f_0) \ mod \ 2^h$. Else set $r_{k+1} = (r_k - 2^k f_k)$ mod 2^h, then increment k and go to Step 2.

6.2 System Performance

This chapter is devoted to the graphical and simulation results of the derivations of Section 6.1. First part will shows the graphs of the system performance of basic and modified OFDM system, second part will shows simulation graphs of PAPR and BER performances. Standard notations are used here and coding is done in MATLAB

6.2.1 System Behaviour

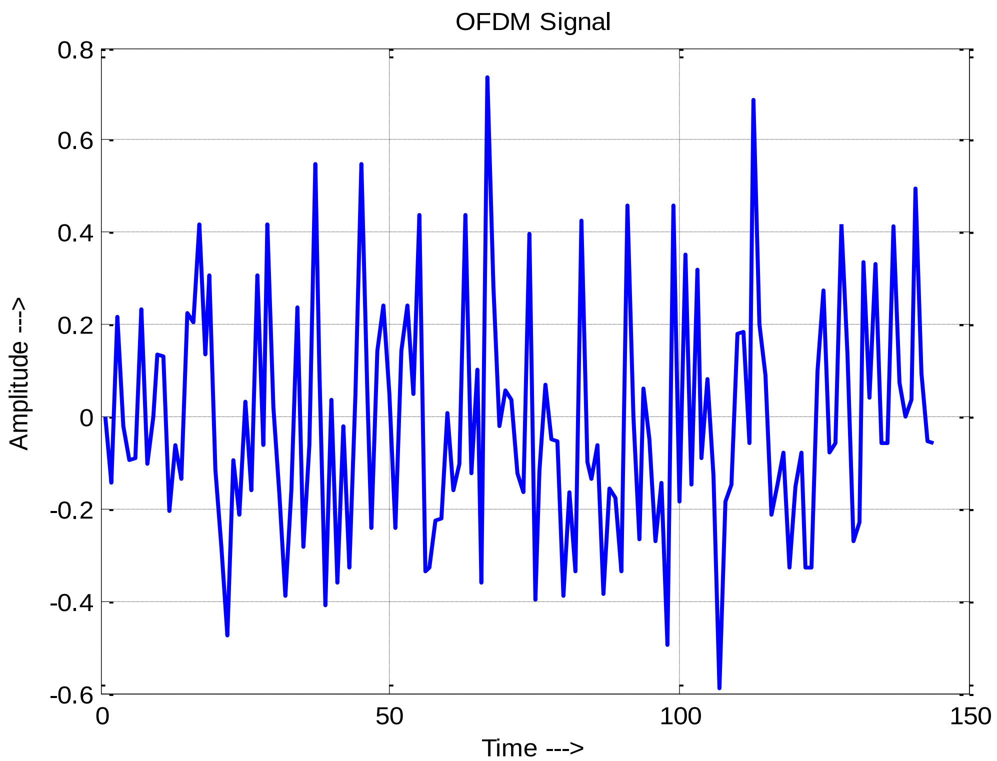


Figure 6.1 Signal Behaviour of Basic OFDM System.

Figure 6.1 shows the OFDM signal of basic OFDM system generated by taking basic parameters.

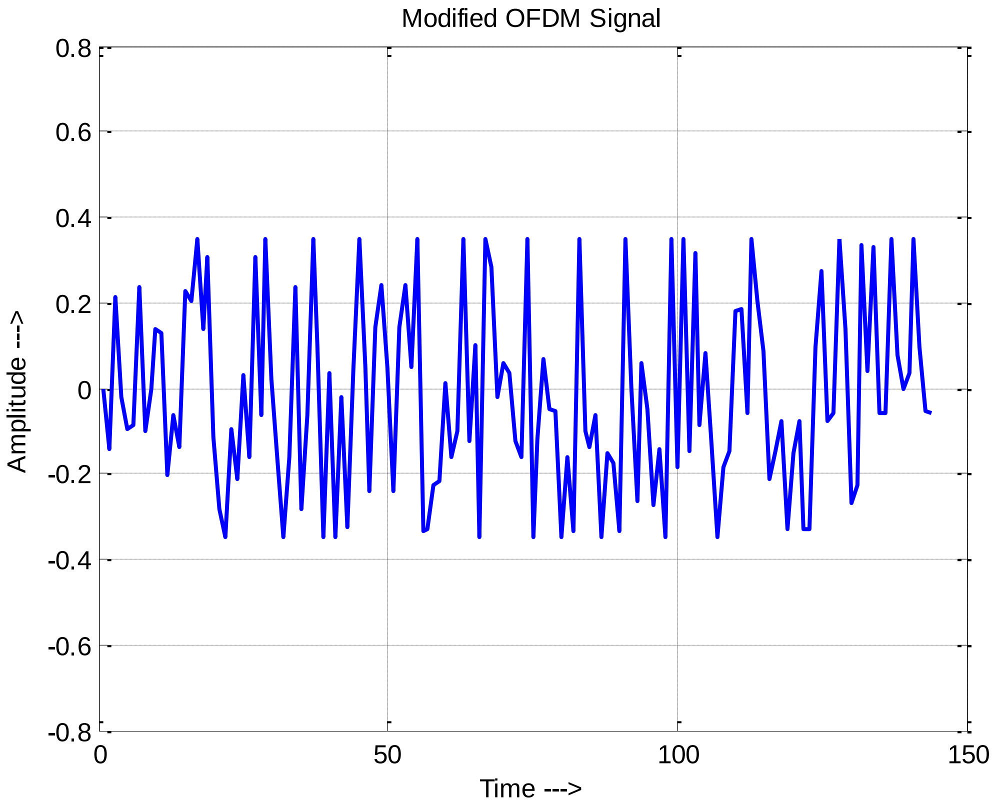


Figure 6.2 Signal Behaviour of coded OFDM System.

Figure 6.2 shows the reduced peaks after implementation of block coding on OFDM system. Graph clearly shows the significant reduction in PAPR.

6.2.2 Performance Graphs

Figure 6.3 demonstrates the BER performance comparison for two systems: Uncoded OFDM system and Coded OFDM system. It can be seen that coded OFDM system is superior to basic OFDM system.

Figure 6.4 demonstrates the PAPR performance comparison for two systems: Uncoded OFDM system and Coded OFDM system. It can be seen that coded OFDM system is superior to basic OFDM system.

As seen from Figure 6.3 and Figure 6.4, low BER and PAPR values are achieved, the system using PAPR reduction algorithm is successfully implemented.

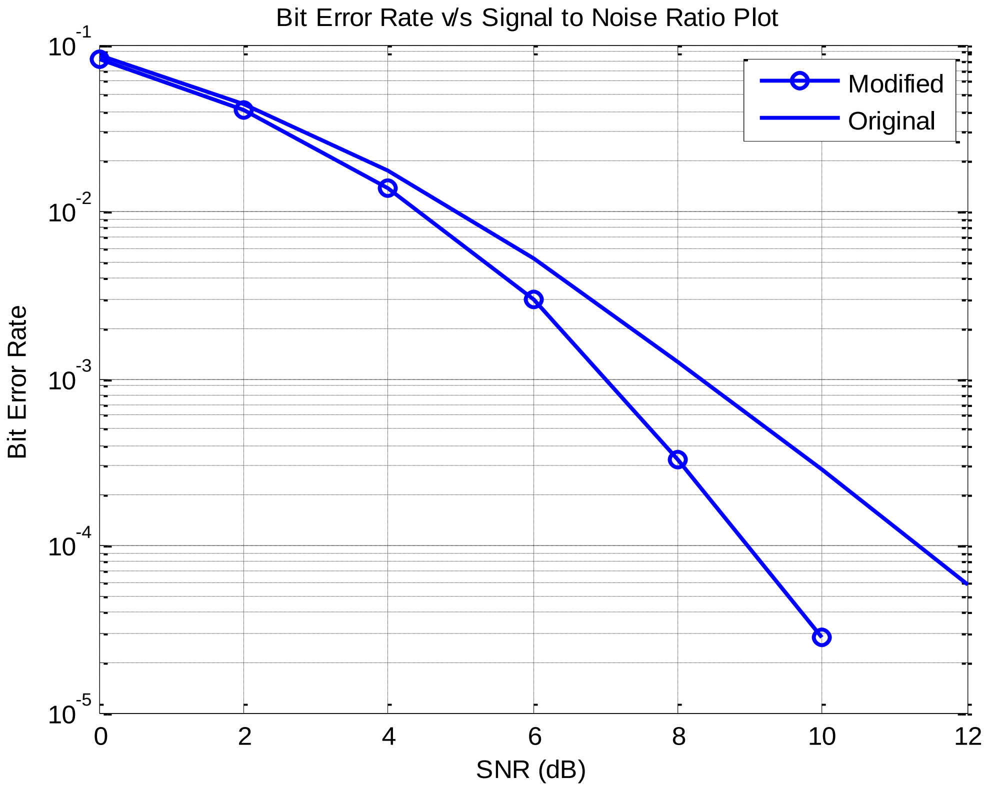


Figure 6.3 Bit Error Rate versus Signal-to-Noise Ratio.

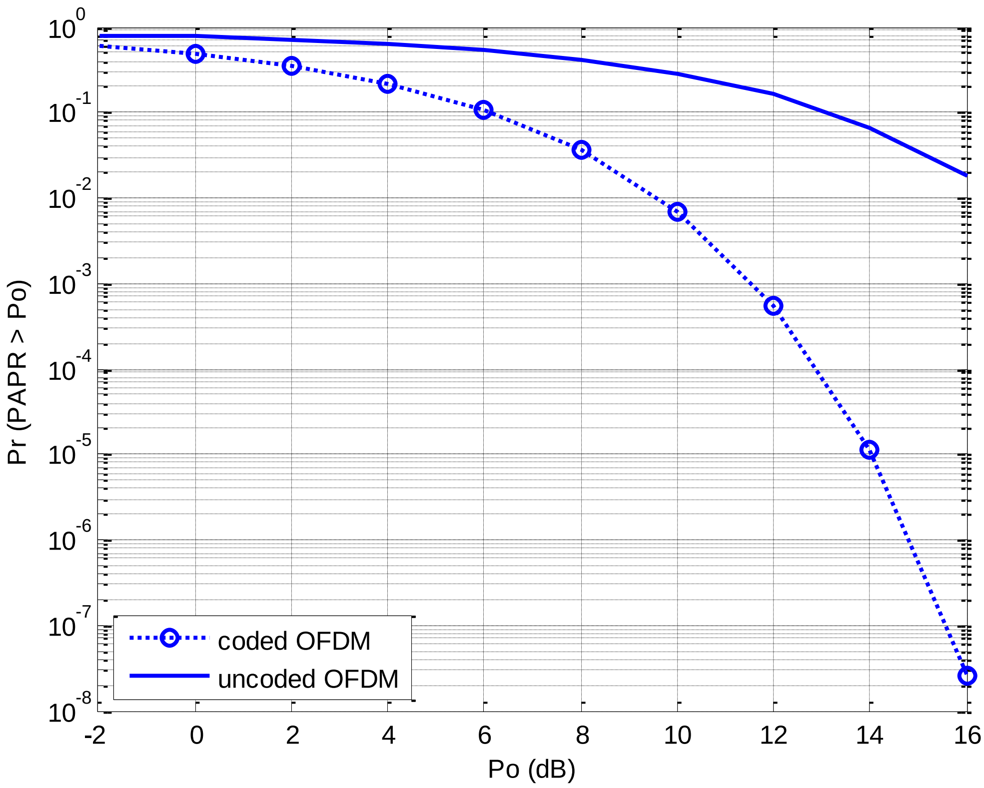


Figure 6.4 Effect of Linear Block Coding on PAPR.

CHAPTER SEVEN

CONCLUSION

7.1 Conclusion

In this thesis, a scheme to reduce PAPR of an OFDM signal is discussed. In the literature, the given methods are evaluated only by means of the PAPR reduction achieved by the technique. PAPR reduction is not a fair criterion to compare the performance of different systems. The performance of the suggested scheme is discussed by means of a new performance measure criterion that takes the basic system parameters like the peak transmit power, bit rate and noise level into account.

There are many different methods discussed in the literature for the purpose of reducing the PAPR of an OFDM signal. The methods can be grouped under three main categories as clipping, scrambling and coding. In this thesis, the method is offered for the PAPR reduction is based on block coding the input data. The modification of the input data does not result in a signal distortion since the employed block code has a sufficient error correction capability.

Each of the methods discussed in the literature has a drawback and these effects need to be considered for a real performance measurement. In this thesis, the performance is measured by the comparison of an Modified OFDM system applying the PAPR reduction scheme, with a simple OFDM system. Two systems are compared under the assumption that equal information bit error probabilities should be satisfied for both.

The simulation results indicate that, application of the scheme results in significant reduction in the PAPR values. As seen in Figure 6.4 the coded OFDM graph is below the uncoded OFDM graph which means with the implementation of the scheme PAPR values is reduced. Same criteria are followed in Figure 6.3 which also shows a decrease in BER. In order to examine the performance potential of the algorithm, extreme conditions were generated by injecting high PAPR producing symbols in two simulation cases. The calculations indicate that the transmit power of the system using the PAPR reduction algorithm is reduced.

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