Transmitter Linearisation using the LINC Technique and Digital Correction

Abdennour Azirar

Thesis Submitted to the University of Surrey for the Degree of Doctor of Philosophy

Unis

School of Electronics and Physical Sciences
University of Surrey
Guildford, Surrey GU2 7XH, UK

November 2005

© A. Azirar 2005
Abstract

Wireless high-speed digital communications systems are becoming increasingly commonplace for both commercial and military applications. One of the most effective techniques for combating multipath interference is the multicarrier OFDM (Orthogonal Frequency Division Multiplexing) scheme. However, the high PAPR (Peak-to-Average Power Ratio) of OFDM signals constitutes a problem when non-linear amplifiers are used. In this thesis the PAPR problem is introduced and the LINC (Linear Amplification with Nonlinear Components) technique is considered as a possible solution to counteract the PA (Power Amplifier) nonlinearity distortion. The proposed LINC technique can be implemented using DSP (Digital Signal Processing) techniques, has the potential of high IMD (Inter-modulation Distortion) suppression and it is unconditionally stable, which is a key advantage in broadband applications.

It's well known that the unwanted imperfections like the I/Q (In-phase/Quadrature) imbalances present in the analogue IQ modulators in a direct conversion OFDM system contribute to a loss of orthogonality and create ICI (Inter-carrier Interference). These impairments can also severely reduce the efficiency of the LINC technique itself. This thesis investigates a digital compensation mechanism to reduce the IQ imbalance errors in the direct conversion OFDM LINC transmitter.

The LINC technique is considered in the first part of this thesis as a possible linearisation technique to counteract the non-linear distortion caused by both the mixer and power amplifier in a single-carrier QPSK (Quaternary Phase Shift Keying) transmitter.

Prototype systems for single-carrier QPSK transmitter and multi-carrier OFDM transmitter have been constructed to demonstrate the proposed method’s capability.

Key words: Direct conversion, LINC, mixer Linearisation, nonlinear distortion, OFDM transmitter, power amplifier linearisation, PAPR, transmitter Linearisation, QPSK transmitter.
WWW: http://www.eim.surrey.ac.uk/
Acknowledgments

My sincerest thanks are due to all the people who have inspired and supported me during the course of this thesis. My gratitude goes to Prof Ian Robertson who gave me the opportunity to explore my own ideas, while tactfully ensuring that I follow a constructive path. I greatly value this freedom and especially appreciate his positive attitude. Beyond the call of duty, his advice has always been readily available, his enthusiasm a true source of inspiration and his dedicated friendship more than I could have asked for.

I owe my deepest thanks to my second supervisor Prof. Michael Kearney and Dr. Stepan Lucyszyn, who was my supervisor when the research project was first started.

I wish to express my sincere gratitude to Dr. Charles Free and Dr. Izzat Darwazeh who read through my thesis and gave their invaluable and constructive comments, questions and suggestions. I am also very grateful to all my colleagues at the MSRG. I extend my warmest thanks to Daniel Stephen and Imad Jallali.

I would like to acknowledge the financial support of the Engineering and Physical Science Research Council (EPSRC).

Last, but by no means least, I would like to thank all my family for bearing with me and all those papers everywhere, for unselfishly sharing all my ups and downs and for all the unconditional love, happiness and support they brought into my life. My special thanks go to my partner Lynn and my children Sophia, Tarik, Yasmin and Khalid.
Contents

Abstract ......................................................................................................................... ii

Acknowledgments ....................................................................................................... iv

Contents ......................................................................................................................... v

List of Figures .............................................................................................................. viii

List of Tables .............................................................................................................. xi

List of Abbreviations ................................................................................................. xii

Chapter 1 ....................................................................................................................... 1

Introduction .................................................................................................................. 1

1.1 Motivation .............................................................................................................. 1

1.2 Objectives of the Research .................................................................................. 4

1.3 Structure of the Thesis ......................................................................................... 5

1.4 Related Publications ............................................................................................ 6

Chapter 2 ....................................................................................................................... 7

Linearisation Techniques ............................................................................................. 7

2.1 Introduction .......................................................................................................... 7

2.2 Feedforward Concept ......................................................................................... 8

2.3 Predistortion Concept ......................................................................................... 9

2.4 Cartesian Loop Concept ..................................................................................... 12

2.5 Envelope Elimination and Restoration Concept .................................................. 13

2.6 Linear Amplification with Nonlinear Components Concept .................................. 15

2.7 Conclusion ........................................................................................................... 19
Chapter 3 ................................................................. 21

Single Carrier Transmitter Linearisation ......................... 21

3.1 Introduction .......................................................... 21
3.2 Single Carrier Transmitter Architecture ....................... 23
  3.2.1 Digital QPSK Modulator ......................................... 24
    3.2.1.1 Pulse Shaping Filters .............................................. 27
    3.2.1.2 QPSK Output Signal ............................................... 30
    3.2.1.3 Spectral Efficiency of QPSK Signals ......................... 32
    3.2.1.4 Measured Spectra of QPSK Modulator Output Signal .... 34
  3.2.2 RF Stage ............................................................ 36
    3.2.2.1 Nonlinearities ....................................................... 36
3.3 Single Carrier QPSK LINC Transmitter ....................... 38
  3.3.1 Experimental Results ............................................. 47
    3.3.1.1 Measurement Results of the Constant Envelope Signals 48
    3.3.1.2 Measurement Results for Mixer Linearisation ............... 50
    3.3.1.3 Measurement Results for QPSK Digital IF LINC Transmitter 55
3.4 Conclusion ............................................................ 58

Chapter 4 ........................................................................ 59

OFDM LINC Transmitter ................................................. 59

4.1 Introduction ............................................................ 59
4.2 OFDM Transmitter .................................................... 63
  4.2.1 Guard Interval and Coding ......................................... 66
  4.2.2 IFFT and Serial to Parallel Conversion ....................... 66
  4.2.3 RF Stage and Simulated OFDM Output Signal S(t) ......... 67
4.3 PAPR Reduction Techniques .......................................... 69
  4.3.1 Coding ................................................................. 69
    4.3.1.1 Complementary Golay Sequences ............................. 69
    4.3.1.2 Hadamard and Trellis Codes ................................. 70
    4.3.1.3 Trellis sequences .................................................. 70
  4.3.2 Orthogonal Pilot Sequences ....................................... 70
  4.3.3 Selective Mapping ................................................... 71
List of Figures

Figure 2-1: Block diagram of feedforward................................................................. 8
Figure 2-2: Block diagram of predistortion................................................................. 10
Figure 2-3: Block diagram of adaptive predistortion .................................................. 11
Figure 2-4: Block diagram of cartesian loop.............................................................. 12
Figure 2-5: Block diagram of EER ............................................................................ 14
Figure 2-6: Block diagram of a LINC architecture ..................................................... 16
Figure 2-7: Block diagram of CALLUM .................................................................... 17
Figure 3-1: Basic architecture of a conventional QPSK transmitter ....................... 24
Figure 3-2: Block diagram of a conventional digital QPSK modulator ..................... 25
Figure 3-3: Building QPSK symbols from binary data ............................................. 26
Figure 3-4: QPSK constellation diagram
   (a) without filtering............................................................................................. 29
   (b) with filtering, α= 0.75.....................................................................................
   (c) with filtering, α= 0.35.....................................................................................
Figure 3-5: Simulated time domain output signal of the QPSK modulator .............. 31
Figure 3-6: Simulated spectra
   (a) Inphase signal after being passed through the baseband filter
   (b) QPSK modulated signal..................................................................................
Figure 3-7: Measurement set-up for QPSK output modulator .................................. 34
Figure 3-8: Measured spectra of DSB QPSK output signal ...................................... 35
Figure 3-9: Block diagram of a conventional LINC transmitter .............................. 39
Figure 3-10: Block diagram of proposed digital IF transmitter ............................... 40
Figure 3-11. Vector diagram showing the relation between the input signal and the constant envelope signals .............................................................................. 42
Figure 3-12: Proposed digital IF LINC transmitter with digital signal components separator algorithm .................................................................................................... 46
Figure 3-13: Block diagram for measuring the constellation of the constant envelope $S_1(t)$.

Figure 3-14: Measurement set-up for constant envelope signal $S_1(t)$. 

Figure 3-15: Measured constellation of the constant envelope signal $S_1(t)$. 

Figure 3-16: Block diagram for frequency translation linearisation. 

Figure 3-17: Prototype system for frequency translation linearisation. 

Figure 3-18: Effectiveness of LINC technique in suppressing spectral regrowth, when the nonlinear mixer is used. 
   (a) Spectra without the LINC technique
   (b) Spectra with the LINC technique. 

Figure 3-19: Prototype system for QPSK digital IF LINC transmitter. 

Figure 3-20: Effectiveness of LINC technique in suppressing spectral regrowth, when both nonlinear mixer and power amplifier are used. 
   (a) Spectra without LINC technique
   (b) Spectra with LINC technique. 

Figure 4-1: PAPR of OFDM signal. 

Figure 4-2: Simplified block diagram of OFDM transmitter. 

Figure 4-3: OFDM output signal $S(t)$ 
   (a) Time domain 
   (b) Frequency domain. 

Figure 4-4: The proposed OFDM LINC transmitter. 

Figure 4-5: Analogue IQ modulator. 

Figure 4-6: OFDM LINC prototype system. 

Figure 4-7: Measurement results illustrating the output spectra of OFDM signal 
   (a) without LINC technique
   (b) with LINC technique. 

Figure 5-1: Analogue IQ modulator measurement set-up. 

Figure 5-2: HP-VEE program for measuring the analogue IQ modulator characteristics. 

Figure 5-3: Measured constellation of the analogue IQ modulator A. 

Figure 5-4: Direct conversion OFDM transmitter with digital IQ imbalance compensation algorithm. 
Figure 5-5: Prototype system of OFDM transmitter with digital IQ imbalance compensation .......... 96
Figure 5-6: Simulated spectra of an ideal direct conversion QPSK modulated OFDM signal .......... 97
Figure 5-7: Measurement results illustrating the output spectra for direct conversion OFDM signals
(a) OFDM spectra without IQ imbalance compensation ................................................................. 98
(b) OFDM spectra with IQ imbalance compensation ........................................................................ 98
Figure 5-8: Measured constellation of the analogue IQ modulator B .............................................. 100
Figure 5-9: Direct conversion OFDM LINC transmitter with digital IQ imbalance compensation
.................................................................................................................................................... 101
Figure 5-10: OFDM LINC transmitter with digital IQ imbalance prototype system ....................... 103
Figure 5-11: Measurement results illustrating the output spectra for OFDM signals
(a) with LINC technique but without IQ imbalance compensation
(b) with LINC technique and IQ imbalance compensation ............................................................. 104
List of Tables

Table 3-1: Characteristics of the power amplifier, model ZHL-42W........................................47

Table 3-2: Characteristics of the mixer, model ZEM-4300..........................................................48

Table 4-2: Main parameters for OFDM LINC prototype transmitter ...........................................80

Table 6-1: Thesis research contributions.........................................................................................109
### List of Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>ACI</td>
<td>Adjacent Channels Interference</td>
</tr>
<tr>
<td>ACPR</td>
<td>Adjacent Channel Power Ratio</td>
</tr>
<tr>
<td>AM</td>
<td>Amplitude Modulation</td>
</tr>
<tr>
<td>ASIC</td>
<td>Application Specific Integrated Circuit</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate</td>
</tr>
<tr>
<td>BPF</td>
<td>Bandpass Filter</td>
</tr>
<tr>
<td>BSK</td>
<td>Binary Phase Shift Keying</td>
</tr>
<tr>
<td>CALLUM</td>
<td>Combined Analogue Locked-loop Universal Modulator</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>COFDM</td>
<td>Coded Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>CP</td>
<td>Cyclic Prefix</td>
</tr>
<tr>
<td>CW</td>
<td>Continues Wave</td>
</tr>
<tr>
<td>DAB</td>
<td>Digital Audio Broadcasting</td>
</tr>
<tr>
<td>DAC</td>
<td>Digital to Analogue Converter</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DDS</td>
<td>Direct Digital Synthesiser</td>
</tr>
<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
</tr>
<tr>
<td>DSB</td>
<td>Double Sideband</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital Signal Processing</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Definition</td>
</tr>
<tr>
<td>--------------</td>
<td>----------------------------------</td>
</tr>
<tr>
<td>DVB</td>
<td>Digital Video Broadcasting</td>
</tr>
<tr>
<td>DVB-T</td>
<td>Terrestrial Digital Video Broadcasting</td>
</tr>
<tr>
<td>EER</td>
<td>Envelope Elimination and Restoration</td>
</tr>
<tr>
<td>ETSI</td>
<td>European Telecommunication Standards Institute</td>
</tr>
<tr>
<td>EVM</td>
<td>Error Vector Magnitude</td>
</tr>
<tr>
<td>EVP</td>
<td>Error Vector Phase</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transform</td>
</tr>
<tr>
<td>FIR</td>
<td>Finite Impulse Response</td>
</tr>
<tr>
<td>FPGA</td>
<td>Field Programmable Gate Array</td>
</tr>
<tr>
<td>FSK</td>
<td>Frequency Shift Keying</td>
</tr>
<tr>
<td>GPIB</td>
<td>General Purpose Interface Bus</td>
</tr>
<tr>
<td>HP</td>
<td>Hewlett Packard</td>
</tr>
<tr>
<td>HPA</td>
<td>High Power Amplifier</td>
</tr>
<tr>
<td>HP-VEE</td>
<td>HP-Visual Engineering Environment</td>
</tr>
<tr>
<td>ICI</td>
<td>Intercarrier Interference</td>
</tr>
<tr>
<td>IDFT</td>
<td>Inverse Discrete Fourier Transform</td>
</tr>
<tr>
<td>IEE</td>
<td>Institution of Electrical Engineers</td>
</tr>
<tr>
<td>IEEE</td>
<td>Institute of Electrical and Electronics Engineers</td>
</tr>
<tr>
<td>IEICE</td>
<td>Institute of Electronics, Information and Communication Engineers</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate Frequency</td>
</tr>
<tr>
<td>IFFT</td>
<td>Inverse Fast Fourier Transform</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
</tr>
<tr>
<td>-------------</td>
<td>--------------------------------------------</td>
</tr>
<tr>
<td>IM</td>
<td>Inter-modulation</td>
</tr>
<tr>
<td>IMD</td>
<td>Inter-modulation Distortion</td>
</tr>
<tr>
<td>IP3</td>
<td>Third Order Intercept Point</td>
</tr>
<tr>
<td>IRE</td>
<td>Institution of Radio Engineers</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter-symbol Interference</td>
</tr>
<tr>
<td>LINC</td>
<td>Linear Amplification with Nonlinear Components</td>
</tr>
<tr>
<td>LO</td>
<td>Local Oscillator</td>
</tr>
<tr>
<td>LPF</td>
<td>Low-pass Filter</td>
</tr>
<tr>
<td>LUT</td>
<td>Look-up Table</td>
</tr>
<tr>
<td>MCM</td>
<td>Multicarrier Modulation</td>
</tr>
<tr>
<td>MESFET</td>
<td>Metal Semiconductor Field Effect Transistor</td>
</tr>
<tr>
<td>MMW</td>
<td>Millimetre-Wave</td>
</tr>
<tr>
<td>m-PSK</td>
<td>Multi-level PSK</td>
</tr>
<tr>
<td>m-QAM</td>
<td>Multi-level QAM</td>
</tr>
<tr>
<td>MSRG</td>
<td>Microwave and Systems Research Group</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>PA</td>
<td>Power Amplifier</td>
</tr>
<tr>
<td>PAPR</td>
<td>Peak to Average Power Ratio</td>
</tr>
<tr>
<td>PSK</td>
<td>Phase Shift Keying</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quaternary Phase Shift Keying</td>
</tr>
<tr>
<td>Acronym</td>
<td>Definition</td>
</tr>
<tr>
<td>---------</td>
<td>------------</td>
</tr>
<tr>
<td>RC</td>
<td>Raised Cosine</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>VLSI</td>
<td>Very Large Scale Integration</td>
</tr>
<tr>
<td>WCDMA</td>
<td>Wideband Code Division Multiple Access</td>
</tr>
<tr>
<td>WLAN/MAN</td>
<td>Wireless Local/Metropolitan Area Networks</td>
</tr>
</tbody>
</table>
Chapter 1

Introduction

1.1 Motivation

Early wireless communication systems were based on constant envelope modulation schemes, most often FM (Frequency Modulation) and FSK (Frequency Shift Keying) [1,2]. The constant envelope feature of these schemes enables high efficiency amplification since the power amplifier can be operated in its most power efficient region near saturation without corrupting the information in signal transmission [3]. The drawback, however, is the inefficient use of the spectrum [4].

The emphasis on higher data rates and spectrum efficiency has driven the industry towards linear modulation techniques such as m-PSK (Multi-level Phase Shift Keying) and m-QAM (Multi-level Quadrature Amplitude Modulation). These modulation schemes have information in both phase and amplitude and require linear processing [5]. In practice it is typically the final power amplifier stage that constitutes the real challenges for the designer, because of the generation of IMD [6]. As a result of the nonlinear distortion, most of the IM (Inter-modulation) power appears as interference in adjacent channels making it difficult for receivers to correctly detect the information [7]. The simplest solution would be to operate the power amplifier in the linear region. However, limited battery capacity has imposed primary restrictions on the power consumption. Filtering the output of the transmitter using a BPF (Bandpass Filter) will help reduce this type of interference but usually will not eliminate the vast majority of IMD since some harmonics falls within the wanted frequency band. Linearisation of the PA is a desirable alternative.
Chapter 1. Introduction

The benefits of power amplifier linearisation are even more pronounced for multi-carrier applications such as OFDM scheme where the PAPR is large [8]. The OFDM scheme is becoming the chosen modulation technique for wireless communication systems [9], it supports high data rate wireless transmission using orthogonal frequency channels and does not require extensive equalisation [10], yet offer excellent immunity against fading and ISI (Inter-Symbol Interference) [11]. The major drawback of these systems is the large PAPR of the transmit signal, which renders a straightforward implementation a complicated and inefficient process [12].

Both linear and efficient power amplification can be accomplished using the outphasing concept first described in [13], and more commonly known as the LINC technique [14]. The LINC architecture of an RF (Radio Frequency) transmitter provides substantially linear amplification from two nonlinear amplifiers by decomposing the original amplitude and phase modulated signal into two constant amplitude envelope, phase varying signals, which, when combined, constructively and destructively interfere to reform the original signal. The output of the LINC amplifier, which is to be transmitted via an antenna, is an amplified form of the original signal. Because of the constant envelope, the information is not affected even when a grossly nonlinear device is used. This opens up the possibility of using power amplifiers driven deep into saturation.

The LINC technique was chosen in this work because it is suitable for digital implementation [15], especially with the emergence of high speed DSP as an enabling technology to implement the technique, and represents an important paradigm shift in PA design. The technique also has the potential of high IMD suppression [16] and completely avoids the nonlinear characteristic of the PAs [17] since the input signals are constant envelope signals. Moreover, the LINC technique is unconditionally stable since it’s based on open loop configuration [18], i.e. there is no delay as a consequence of the feedback loop, which is a key advantage in high speed OFDM communications systems.

Frequency translation devices are usually present in RF transmitters in order to upconvert the modulated baseband signal to the desired frequency, and that leads to another important source of interference. In a two step upconversion process, the modulated input signal is
generated at an IF (Intermediate Frequency), then frequency translated in one or more mixing stages to the final frequency.

A common problem with this arrangement is the requirement for image rejection filters. In addition, the mixing process will produce nonlinear distortion in the output signal, causing spectral regrowth [19]. Similar to the power amplifier, the nonlinear distortion can be reduced by ensuring that the input signal, in this case the IF signal, is operated well below compression. However, the mixers are required to operate at their maximum output in order to reduce the amount of gain required in the succeeding RF amplification stage. Spurious products that fall at offsets far from the wanted frequency can be suppressed through filtering, but those within the wanted frequency band will not be attenuated. The LINC technique is also considered in this work for mixers linearisation.

Another common way of mitigating those problems is to generate the wanted signal directly at the final frequency stage using an analogue IQ modulator. This arrangement is known as the direct conversion architecture. In this case, the I and Q baseband signals, or equivalently, the real and imaginary parts of the complex signal are used to directly modulate a carrier at the output frequency. Although spectral spreading of the signal into adjacent channels can still occur, the harmonic mixing effect is eliminated since there is only a single carrier component applied to the mixers.

A drawback with this arrangement is that the impairment caused by the IQ phase and gain errors of the analogue IQ modulators degrades the output signal of the high data rate multicarrier OFDM systems causing both out-of-band and in-band distortions [20-24]. The out-of-band interference is usually characterized by ACI (Adjacent Channel Interference), while the in-band distortion can be quantitatively described in terms of EVM (Error Vector Magnitude). These unwanted imperfections in the analogue IQ modulators are not only known to contribute to the loss of orthogonality and create intercarrier interference in direct conversion OFDM systems, but they are also known to reduce the performance of the LINC technique itself [25].

A further contribution of this thesis is the development of a digital IQ imbalance compensation technique that reduces the effect of the IQ imbalance caused by the analogue
IQ modulator in the direct conversion OFDM transmitter. Hence the ability of the LINC technique itself is extended to include this digital control mechanism to compensate for the imperfections of the two analogue IQ modulators. The mechanism stores the measured data of the analogue IQ modulators and adjusts the signal components of the LINC system in order to compensate for any differences in the characteristics of the separate signal paths, which would cause the combination not to accurately represent the original signal.

1.2 Objectives of the Research

The choice of modulation type involves many trade-offs to optimise efficiency in digital radio system design. Bandwidth efficiency is the extent to which a modulation format uses an allocated bandwidth and its ability to accommodate data within that limited bandwidth. Power efficiency describes the ability of the system to reliably send information at the lowest practical power level. In most systems, bandwidth efficiency is the highest priority. However, a change in one category has a direct effect on the other.

In the past, improvement in power efficiency could be achieved by sacrificing bandwidth efficiency. This is no longer acceptable, since the electromagnetic spectrum is already crowded and every communication system design must be as bandwidth efficient as it is power efficient. In this environment, modulation selection takes on increased importance. The new digital transmission systems, particularly those based on OFDM technology, feature good bandwidth efficiency as they exploit multilevel modulation schemes to transmit at high information rates with combination to a dense allocation of a large number of orthogonal subcarriers, making it suitable for many communications systems including digital broadcasting, wireless standards and seems to be an excellent candidate to fulfill on 4G (4th Generation) mobile radio systems. However, these modulation signals have a large PAPR, as a result, problems with nonlinear distortion become critical for system performance and, therefore, must be reduced to a minimum.

The objective of this research is to reduce the nonlinear distortion in the transmit system for both single carrier and multi-carrier OFDM transmitters to achieve excellent efficiencies and
the effect of the IQ imbalance using advanced digital signal processing techniques. The research efforts focuses on three areas:

- Mixer Linearisation
- Power amplifier Linearisation
- IQ imbalance compensation

1.3 Structure of the Thesis

The dissertation consists of six chapters, and it’s mainly concerned with the Linearisation of the power amplifier and improving the performance of the frequency translation in both single carrier and multi-carrier OFDM transmitters.

In Chapter 2, the fundamental concepts of existing Linearisation techniques are described. These techniques are feedforward, predistortion, cartesian loop, EER (Envelope Elimination and Restoration), adaptive baseband predistortion, LINC and CALLUM (Combined Analogue-Locked Loop Universal Modulator). Their relative advantages and disadvantages are briefly discussed.

The third chapter presents the influence of the power amplifier and mixer nonlinearities on modern digital communications systems, and provides an overview of the single carrier QPSK transmitter architecture. This includes the digital baseband generation, filtering and the operation of the digital QPSK modulator. In addition, the nonlinear distortion caused by the nonlinear devices such as the mixer and power amplifier as a result from the envelope fluctuations of the filtered signal is discussed. The LINC is then considered as a prospective Linearisation technique to reduce the nonlinear distortion caused by both the mixer and the power amplifier in the single carrier QPSK transmitter with two-stage upconversion. Finally, prototype systems have been constructed to demonstrate the scheme and its capability.
Chapter 4 provides an overview of OFDM concept and demonstrates the effect of the high PAPR of the signals when nonlinear amplifiers are used in OFDM transmitters, followed by reviewing the background of some of the existing solutions for PAPR reduction, and a brief comparison between them is presented. The LINC technique is then used to linearise the power amplifier in an OFDM transmitter. Finally, a prototype system is constructed and the data is presented in the frequency domain.

Chapter 5 is devoted to digital baseband compensation algorithms for IQ imbalance in direct conversion OFDM LINC transmitter. The compensation technique helps to correct for the analogue imperfections, which in turn improves the overall LINC performance.

In a practical implementation, the data is collected using a measurement set-up station, and the technique is performed on prototype systems to evaluate its performance.

Finally, chapter 6 concludes the dissertation and provides future research directions.

1.4 Related Publications

Two papers are included in this thesis.


Chapter 2

Linearisation Techniques

2.1 Introduction

Linearity properties and power efficiency are competing objectives in RF and microwave power amplifier design, especially in digital mobile communication systems, where the total efficiency of the system is significantly determined by the efficiency of the high power amplifier at the time of transmitting. This means that the operating time for example in a handset is greatly dependent on the efficiency of the HPA, while high power efficiency is also preferred in base stations in order to achieve low power consumption and avoid problems of overheating. Given that both linearity and efficiency aspects are generally desirable, there is much interest among the international research community in techniques, which can compensate for nonlinear distortion generated by the HPA, thereby allowing it to be used at higher drive levels for optimum power efficiency. As a result of the nonlinearity distortion caused by the HPA, the transmitted signal is distorted and ultimately increases the bit error rate.

An obvious solution is to filter the amplified transmitted signal. However, filtering cannot easily suppress emissions from the transmitted signal that are located on the same frequency or near to the desired frequency band.

A very common although trivial solution to nonlinear distortion, is backing-off the operating point of the power amplifier far from the saturation region. This is normally measured by the input back-off and output back-off parameters that correspond to the distance in dB (Decibels) between the average and saturated input and output power. This solution restricts the driving input to the HPA to operate within a reduced region of the whole dynamic range, and as result more stages are required in the amplifier to maintain a given level of the power.
transmitted, and hence greater DC (Direct Current) input power is consumed. Thus reducing the power efficiency of the transmitter, which is certainly a critical consideration in communication systems design.

Complementary to the above solutions, the use of Linearisation techniques has been shown to be essential in reducing the nonlinear distortion, and enabling the HPAs to operate near saturation with good power efficiency. The more widely known techniques are feedforward, predistortion/adaptive baseband predistortion, cartesian loop, EER and LINC/CALLUM.

2.2 Feedforward Concept

Feedforward [26] involves comparing the power amplifier input and output signals to derive an error or distortion term in a signal-cancelling loop. This residual error is then amplified in a separate, low power amplifier before being subtracted from the main amplifier output in an error-cancelling loop as shown in Figure 2-1

![Figure 2-1: Block diagram of feedforward](image-url)
Chapter 2. Linearisation Techniques

If the low power, error or equivalently known as auxiliary amplifier is perfectly linear and the error-cancelling loop is perfectly balanced, then the overall result is distortionless, and the output signal is a relatively clean amplified version of the input. However, in practice the cancellation loops are only partially effective, and the technique is compromised.

In a practical feedforward implementation, there will be imbalance in the error-cancelling loop, which will limit the distortion reduction. For example, a 1 dB gain error and a 10° phase error limits the distortion suppression to just 14 dB. It was demonstrated in [26] that an improvement of say, 30 dB would require the balancing to be within 0.3 dB and 1°. Even if such stringent requirements can be met, the overall linearity can never be better than that of the error amplifier, which must therefore operate in Class-A and will consequently be power inefficient. These problems are further compounded by errors in the signal-cancelling loop, which will increase the power handling requirements of the error amplifier.

Whilst feedforward is capable of improving linearity over a wide bandwidth, where some experimental results are reported [27] for a C band feedforward linearised amplifier demonstrating a minimum of 20 dB IMD suppression across the 5.9 - 6.4 GHz satellite band, it requires more analogue hardware to implement, and it may also be susceptible to drift in gain and group delay characteristics due to the likelihood of deterioration of performance with ageing, temperature variations and other variable conditions. However, a recent resurgence of interest in feedforward correction techniques has led to their application in broadband applications.

2.3 Predistortion Concept

The basic idea of predistortion is to cancel the distortion in the power amplifier by predistorting the transmitted signal with the inverse function of the amplifier. The signal is thus predistorted before being applied to the amplifier. If the predistorter has a non-linearity which is the exact inverse of the amplifier nonlinearity, then the distortion introduced by the amplifier will exactly cancel the predistortion, leaving a distortionless output. Figure 2-2 shows the basic predistortion concept.
Such predistortion may be implemented at RF, IF or at baseband. Baseband linearisers based on the use of LUTs (Look-up Tables) held in memory are more common with the ready availability of VLSI (Very Large Scale Integrated Circuit) and microprocessors, and can offer a compact solution. Until recently, however, it has been easier to generate the appropriate predistortion function with RF or IF circuitry.

In its simplest analogue implementation, a practical predistorter can be a network of resistors and nonlinear elements such as diodes or transistors. Several examples of this technique have appeared in the literature, example [28], where the reduction in third order IMD that has been reported, is typically in the range 7-15 dB.

Another predistortion based method, which shows reasonable improvement have been analysed in [29] for satellite communications systems, here a pair of FET (field effect transistor) amplifiers are used as the predistorter. In this arrangement, the input signal is unequally split between the two amplifiers, such that one of them is driven into compression. The compressed output is then scaled and subtracted from the linear output to produce the inverse of the compression characteristic, as required. Reduction in IMD of around 4-5 dB has been measured using this technique, but only when the main amplifier is operated with at least 0.3 dB of output back-off.
Predistortion linearisation is a mature technology, and have been favoured for microwave frequencies and above applications because of their wideband performance, and ability to function as standalone units. However, the poor performance is due to the fact that the amplifier characteristics are not constant, but vary with time, frequency, power level, supply voltage and environmental conditions. It's necessary to adjust the predistort waveform so that it can track changes in amplifiers characteristics. This technique is known as adaptive baseband predistortion.

Adaptive baseband predistortion as the name suggests uses DSP techniques, in this system predistortion is applied at baseband before upconversion to RF [30]. A feedback path is generally provided to support real-time adjustment of the predistortion coefficient in order to maintain a high level of linearity as shown in Figure 2-3.

![Figure 2-3: Block diagram of adaptive predistortion](image)

It was demonstrated in [30] that for a two-tone test, the third harmonic distortion product is reduced by 30 dB. While the reported digital adaptive predistortion techniques provide more
precision and flexibility, the associated instantaneous bandwidth limitations are severe, and they tend to be very computationally or memory intensive. However, development of faster algorithm for adaptive baseband predistortion techniques is always in progress [31].

Another factor that hinders the deployment of digital adaptive predistortion technique is a lack of interface standards between RF outputs and inputs. Although some progress has been made over the past few years. As such interface are not upgraded of currently existing transmission schemes such as OFDM.

2.4 Cartesian Loop Concept

The cartesian loop is a well known feedback linearisation technique [32]. It demodulates a portion of the PA output signal and negatively feeds this back at baseband through differential amplifiers to the I/Q modulator with the complementary error signal, thus compensating for the PA distortion as shown in Figure 2-4.

![Figure 2-4: Block diagram of cartesian loop](image)

Figure 2-4: Block diagram of cartesian loop
Chapter 2. Linearisation Techniques

The cartesian loop technique provides linearisation of a complete transmitter as opposed to just the power amplifier. This technique combines the upconversion and power amplification processes by taking the baseband IQ signals and translating them into an RF carrier frequency at a high power level. The result is that any nonlinearity in the upconverter, driver-amplifier chain and RF power amplifier are negated. In addition, if another stage of frequency translation is needed in the process, then nonlinearities in this device are also removed.

The cartesian loop technique is widely used in low frequency amplifiers, where stability of the feedback loop is easy to maintain, and improvements of more than 30 dB in spectral regrowth are possible using this technique, but it is inherently narrowband due to stability [33,34], which depends upon loop gain, phase shift introduced by the PA and its compensation, time delays in the circuit and loop filter bandwidth. Loop delays themselves could be a source of output distortion. To overcome the imperfections of analogue components, a digital cartesian loop is investigated in [34] where the processing is shifted to baseband in order to provide a better control and compensation of loop delay. However, the proposed digital cartesian technique tends to be computationally intensive, and it is still in its early stages and there is no prototype system to demonstrate the performance of the system.

2.5 Envelope Elimination and Restoration Concept

EER also known as Khan technique is based upon the equivalence of any narrowband signal to simultaneous amplitude and phase modulations [35]. In a modern implementation, both the envelope and the phase modulated carrier are generated at baseband using DSP techniques.

Kahn technique transmitter operates with high efficiency over a wide dynamic range and therefore produces a high average efficiency for a wide range of signals and power back-off levels. Figure 2-5 demonstrates the basic concept of EER.
In EER the envelope of the RF signal to be amplified is extracted, amplified linearly and separately, and applied as modulation to the DC power supply of the RF power amplifier. The technique has the theoretical potential to achieve 100% DC to RF power conversion efficiency at all envelope levels of the modulation signal. In practice, the efficiency falls short of this, however, the actual figure may still be in the region of 65% [36].

Despite these impressive figures, there are a number of practical limitations to the linearity available from the system. Most significantly, where low envelope levels are used, the RF power transistor may cut off, introducing significant distortion into the system, and the phase coherency between the envelope and the main phase signals in the restoration process. An additional delay process is then needed, which in turn limits the operation bandwidth of the technique.

EER was initially developed for SSB (Single Sideband) and TV (Television) transmitters, but there are number of articles [37], which consider its implementation with multicarrier modulation schemes for future generations of wireless systems.
2.6 Linear Amplification with Nonlinear Components

Concept

The LINC technique is based on the so called the out-phasing technique [13], which was introduced in the mid 1930's to overcome the increasing problems with high cost and low power efficiency of high power AM (Amplitude Modulation) broadcast transmitters. When the technique was rediscovered in the early 1970's by Cox [14], it became more known as LINC. Cox suggested a solution that was suitable for modulation schemes exhibiting both amplitude and phase variations. The LINC scheme avoids the nonlinear characteristic of the power amplifier by feeding it with a constant envelope signal. Two phasors with equal amplitudes are generated from the input signal in the signal component separator. These phasors are amplified separately in high power amplifiers and finally recombined to form an amplified replica of the input signal as shown in Figure 2-6.
Chapter 2. Linearisation Techniques

Figure 2-6: Block diagram of a LINC architecture

Earlier papers [14,38] suggested a completely analogue solution where the signal components separator operated at some intermediate frequency or directly at the carrier frequency. In a modern implementation [39], DSP techniques are used to produce the constant envelope signals. Theoretically, the LINC technique offers the potential of high power efficiency since the input to the PAs are constant envelope signals. Moreover, the LINC scheme is inherently stable since it's an open loop system, and thus, it's suitable for broadband applications. However, the complexity of the system and the problem of the imbalance between the signal paths, which is hard to achieve in an open loop approach prevented the LINC technique from becoming widely accepted.

To overcome the major stumbling block with the LINC approach, which is the generation and control of the signal sources with sufficient accuracy and tracking capability to reproduce the linear power output. A novel implementation of two channel RF synthesis, termed CALLUM was developed [40]. The CALLUM method can cope with these issues by
Chapter 2. Linearisation Techniques

encompassing the generation and combining of the synthesis signals in a closed loop control system.

The CALLUM technique can be regarded as a particular implementation of cartesian feedback in the LINC system. There are a few variations on the implementation of the system in the literature. However, the basic structure of CALLUM [40] is illustrated in Figure 2-7.

![Figure 2-7: Block diagram of CALLUM](image)

The CALLUM system produces a linear, high level RF output signal in the same manner as the LINC method. An arbitrary input signal varying in phase and amplitude is decomposed into two constant envelope signals, which can each be amplified by grossly nonlinear but highly efficient amplifiers and then recombined at high level to synthesize the required output. The constant envelope signals are derived from two VCOs (Voltage Controlled Oscillators). The original LINC system operated in an open loop configuration and was
sensitive to small gain and phase imbalances in the two paths. The CALLUM system has a closed loop configuration and can continuously correct for such imbalances. It is a complete transmitter scheme, which both upconverts and linearly amplifies a baseband signal represented in cartesian form, within a closed loop system.

A major obstacle in the design of such a system is to maintain the stability of the feedback loop [34], which requires careful design in order to prevent low frequency poles from reducing stability margins. This restricts the CALLUM method to narrowband modulating signals.
2.7 Conclusion

Each of these linearisation techniques has its own advantages and disadvantages, nonetheless all these methods, depending on the circumstances, can contribute linearisation improvement benefits at or nearer to saturation PA operation. Furthermore, many linearisation techniques that are intended for use in reducing the nonlinear distortion caused by the power amplifiers, are also applicable to other components such as frequency translating devices or even the whole transmitter as was demonstrated in the case of cartesian loop technique.

Some of the drawbacks, however, include the need for down-conversion and demodulation of the output of the power amplifier, especially for linearisation techniques that have closed loop topology such as the adaptive baseband predistortion, CALLUM and cartesian loop. The need for high speed computation and/or convergence, susceptibility to modeling errors, use of non real time applications as in the case of adaptive digital predistortion. Furthermore bandwidth limitation due to feedback loop means that these linearisation schemes become less and less competitive as the required input signal bandwidth increases, which is a key consideration for multicarrier OFDM based broadband applications.

Feedforward linearisation on the other hand simultaneously offers wide bandwidth and good IMD suppression. The price, however, for this performance is the higher complexity due to the presence of extra RF components, and the requirement of amplitude and phase match in two different loops.

Predistortion linearisation have been favoured for microwave frequencies and above applications because of their wideband performance and ability to function as standalone units. However, the poor performance is due to the fact that the amplifier characteristics are not constant and varies with different elements including the operating frequency.

EER linearisation technique can deliver high power efficiency and suitable for wideband applications, but the linearity achieved depends upon the accuracy of reproduction of the input signal’s amplitude and phase information.
The LINC technique converts the input signal into two constant envelope signals that are amplified by HPAs then combined before transmission. Theoretically, the technique can provide excellent efficiency over a broad range of output powers since the branch amplifiers are constantly operating at optimum, maximum swing, each amplifier is always operating at peak efficiency and can offer wide bandwidth, making suitable for broadband applications. Consequently, it's very sensitive to gain and phase imbalances in the two signal paths, which are mainly caused by the RF components in the transmitter including the analogue IQ modulators.
Chapter 3

Single Carrier Transmitter Linearisation

3.1 Introduction

In narrowband wireless communication systems, adjacent channel interference specifications require the use of tightly band limited signaling formats. To meet these requirements, pulse shapes that span many symbols are used. Typically, some sort of Nyquist pulse shape such as an RRC (Root Raised Cosine) filter is used. These pulse shapes cause a substantial increase in the peak power of traditional linear modulation formats such as m-PSK.

Nonlinear amplification of such a signal will cause distortion in the output signal and will regenerate side lobes, thus destroying the spectral containment of the original signal, degrading system performance and causing disturbance to neighbouring channels (spectral leakage, regrowth).

The effects of nonlinear distortions from power amplifiers on adjacent channel interference as well as bit error rate performance are investigated in detail in [41] for both QPSK and π/4 shift QPSK signaling formats for portable radio communications. It was shown in [41] that when a pulse shaping filter with a roll-off factor α of 0.5 is used, the peak to average power ratio is 3.2 dB for π/4 shift QPSK and 4 dB for QPSK, and these values increase with a smaller values of α since the amplitude fluctuations is getting larger.

Ideally, constant envelope signaling such as FM and CPM (Continuous Phase Modulation) would allow for maximum amplifier efficiency. However, such formats are not very bandwidth efficient. Thus, there is a need to produce bandwidth efficient signals, which also have a low PAPR.
Chapter 3. Single Carrier Transmitter Linearisation

This chapter proposes the LINC [14] method for improving the spectral regrowth behaviour of modulated QPSK signal. In this technique, the transmitter is made of two branches having a high power amplifier each, operated with constant envelope signals, with a combining structure at the output. The constant envelope feature of these signals enables high efficiency amplification since the power amplifiers can be operated deep into saturation without corrupting the information in signal transmission.

Mixers are key components in communication systems for frequency translating signals. In transmitters, mixers upconvert the baseband signal to IF or to RF signal for transmission via an antenna. However, the inherent nonlinearity of mixers in communication systems creates similar undesired effects to the ones caused by the PA like IMD, spreading the spectrum to a wider bandwidth. Conventional linearisation techniques such as cartesian loop method have the advantage to include the upconversion stage in the linearisation process but they are still issues regarding the stability of the system, which were analysed in [42]. Other reported linearisation techniques for mixers within radio systems applications are mainly dealt with the downconversion stage in the receiver side. This is because the mixer is the key components in the receiver, which determines IMD levels. Some of these techniques are based on feedforward [43,44] and predistortion [45] methods. High levels of IMD reduction can be obtained, in the region of 30 dB [44]. However, for the reported feedforward method [44] the performance is critically dependent on the amplitude matching of the IMD products and the mismatching characteristics of the two mixers. In addition, it would be impractical to use neither feedforward method reported in [44] nor predistortion method reported in [45] at the transmitter side for both frequency translation and power amplifier linearisation.

In comparison to feedforward and predistortion approaches, the proposed LINC technique has the potential of greatly simplifying the design of circuitry due to the elimination of a large number of RF components by moving the processing from analogue to digital. Hence with this option, the upconversion processing and power amplification operations are incorporated within the linearisation system. This allows both devices to operate in closer to saturation unlike many conventional transmitter linearisation techniques where only the PA is linearised.
This chapter will begin with a brief overview of a two stage upconversion single carrier transmitter architecture. It will then be followed by the description of the proposed LINC for overcoming nonlinearity of the transmitter at baseband level. The detailed analysis, some simulation results, and finally prototype systems with practical measurement results considerations will be presented illustrating the performance of the technique in improving spectral regrowth for both frequency translation and power amplifier devices. Even though couple of papers [14,46] have pointed out to the linearisation of power amplifiers and frequency translating components using the LINC technique, with respect to the theoretical behaviour there is not yet, to the author’s knowledge, practical measurement results showing the performance of the LINC approach used for both frequency translation devices and power amplifier linearisation.

3.2 Single Carrier Transmitter Architecture

QPSK also referred to as 4-ary PSK modulator is used routinely to modulate multicarrier systems such as OFDM, but is highly developed stand-alone transmission format for point-to-point digital transmission in both military and commercial communications systems. Terrestrial microwave radio links and satellite communication systems frequently employ QPSK as their modulation format, is proven in multiple field trials, provides a good balance between bandwidth efficiency and robustness to channel impairments, and offers simple implementation with open architecture.

In the configuration of a two-stage upconversion narrowband [47] QPSK transmitter, the baseband signal is first converted to an IF signal and then to an RF signal, because of the stringent image rejection requirements of the transmitter, a BPF filter is often used in the RF stage followed by power amplifier and an antenna as shown in Figure 3-1.
3.2.1 Digital QPSK Modulator

There is an increasing tendency to implement the modulator digitally i.e. to generate the intermediate frequency digitally using digital IQ vector modulator, which leads to better time and environment stability, greater flexibility and higher performance than traditional analogue techniques. This factor becomes more important as the LINC transmitter [25,48] is more sensitive to IQ imbalances of the analogue vector modulators. The configuration of a conventional digital QPSK modulator is shown in Figure 3-2.
The symbol builder in the digital modulator takes the incoming unipolar binary bit stream at a bit rate of \( f_b \) produced in the data-generation process, collects them into symbols at a rate of \( f_s \). In the case of QPSK the symbol builder collects two bits per symbol and maps them to the four quadrants of the I and Q plane as shown in Figure 3-3.
The relation between symbol rate and bit rate is as follow:

\[ f_s = \frac{f_b}{2} \]  

(3.1)

Where

\[ f_s = \frac{1}{T_s} \]  

(3.2)

\( T_s \) being the symbol duration. The modulated signal has four distinct phases \( (\pi/4, 3\pi/4, 5\pi/4 \) and \( 7\pi/4 \) ), each representing one symbol, and each symbol contains 2 bits of information.

Although Gray coding is not presented in Figure 3-3, the mapping of the bits into symbols in the MATLAB programs carried out in this work was done in accordance with the Gray code. This code ensures that neighbouring points in the constellation only differ by a single bit and helps to minimise the overall bit error rate.
The symbol builder then creates two bipolar binary waveforms that are passed through two spectrally shaped baseband filters and then modulated by the inphase and quadrature carriers and combined to produce the intermediate frequency output signal.

Some degree of over-sampling is usually applied to the signals before filtering. Over-sampling is the process of increasing the number of samples per symbol. The QPSK modulator shown above produces one sample for each symbol (two bits). An over-sampling ratio of four results in four samples per symbol [49] and a longer waveform, easing the transition band requirements of the reconstruction filters.

Finally, the digital QPSK signal is converted to analogue and filtered at the output of the modulator to limit further its power spectrum. This prevents spill-over into adjacent channels and also removes out-of-band spurious signals caused by the modulation operations.

### 3.2.1.1 Pulse Shaping Filters

The Nyquist or pulse shaping filters are applied to baseband signals to reduce the transmitted bandwidth and increase spectral efficiency. These filters are FIR (Finite Impulse Response) filters with taps that represent the sampled impulse response of the desired filter and they can be accomplished using the mathematical concept of convolution [47].

The properties of the filter are controlled by the roll-off factor $\alpha$ defining the excess bandwidth relative to the symbol rate as well as the amplitude range of the signal. Another property is that for each symbol transmitted there exists a time point in the signal waveform where the influence from preceding and succeeding symbols is zero, i.e., there is no ISI (Intersymbol Interference). ISI is best thought of as the energy from one symbol being spread into adjacent symbols. This has two effects; firstly, there is less energy left in the symbol for the demodulator to make a decision as to the value of that symbol. Secondly, energy from adjacent symbols will further increase the chances of an erroneous decision if those adjacent symbols are of a different value.
Chapter 3. Single Carrier Transmitter Linearisation

Usually the filter is split, half being in the transmit path and half in the receiver path. In this case root Nyquist filters commonly called root-raised cosine are used in each part, so that their combined response is that of a Nyquist filter. The impulse responses of the raised cosine filter with symbol rate $T_s$ and a roll-off factor $\alpha$ is given by:

$$h(t) = \frac{(\sin(\frac{\pi t}{T_s}) (\cos(\frac{\pi \alpha t}{T_s})))}{(\frac{\pi t}{T_s}) (1 - (\frac{2\pi \alpha t}{T_s})^2)}$$  \hspace{1cm} (3.3)

The single-carrier QPSK modulator includes a pair of root raised cosine filters with a roll-off ($\alpha = 0.35$), which controls the sharpness of the filters and gives a direct measure of the occupied BW (Bandwidth) of the system. The relation between the BW and $\alpha$ is:

$$BW = \left(\frac{1}{T_s}\right)^* (1 + \alpha)$$  \hspace{1cm} (3.4)

(*) denotes multiplications. Alpha is sometimes called the “excess bandwidth factor” as it indicates the amount of occupied bandwidth that will be required in excess of the ideal occupied bandwidth, which would be the same as the symbol rate. Smaller values of $\alpha$ require less bandwidth.

If the filter had a perfect brick wall characteristic with sharp transitions i.e. an $\alpha$ of zero, the occupied bandwidth would be equal to symbol rate. In a perfect world, the occupied bandwidth would be the same as the symbol rate, but this is not practical since it’s proven impossible to implement a filter with an alpha of zero. Figure 3-4 shows the effect of $\alpha$ on filter bandwidth and envelope fluctuation.
Chapter 3. Single Carrier Transmitter Linearisation

As it can be seen from the constellation diagram in Figure 3-4(a), all the states have the same amplitude, and so QPSK does not use amplitude to modulate the carrier. However, as the carrier transitions from one state to another moves diagonally across the centre point of the axes, the amplitude of the carrier will temporarily change. Thus QPSK is not a constant envelope modulation scheme. In this case any signal path through which a QPSK signal travels must have a degree of linearity in order to avoid creating distortion. Meanwhile, the constellation diagram of the QPSK signal in Figures 3-4(b) and 3-4(c) show that filters with alphas of 0.75 and 0.35, respectively, smooth the transitions and narrow the frequency spectrum required.

Furthermore, Figure 3-4 shows that different filter $\alpha$ also affects the transmitted power. In the case of the unfiltered signal in Figure 3-4(a), with an $\alpha$ of infinity, the maximum or peak power of the carrier is the same as the nominal power at the symbol states. However, As can be seen from the constellation diagram in Figures 3-4(b) and 3-4(c), the smaller alpha takes more peak power because of the overshoot in the filter's step response. This produces trajectories, which loop beyond the outer limits of the constellation. At an alpha of 0.35, about the usual value for most radio systems today, there is a need for significant excess
power beyond that needed to transmit the symbol values themselves. Hence, a key factor in the use of the raised cosine filter is that changing the roll-off factor $\alpha$ of the filter changes PAPR of the filtered signal. The PAPR is a popular statistic for describing signals with amplitude variation.

Generally, a signal with higher PAPR requires amplifiers with higher linearity to handle the average power requirements and the peak amplitude excursions without generating excessive out of band distortion.

### 3.2.1.2 QPSK Output Signal

The filtered signals are fed to digital IQ modulator (see Figure 3-2) for directly translating the baseband signal into an intermediate frequency signal. The output bandpass signal can be written as:

$$S(t) = I(t)\cos(w_c t) + Q(t)\sin(w_c t)$$

where $I(t)$ and $Q(t)$ are the in-phase and quadrature component and $w_c$ is the carrier frequency in rad/sec. Figure 3-5 shows amplitude fluctuation of the QPSK output bandpass signal $S(t)$ with roll-off factor $\alpha=0.35$. 
Chapter 3. Single Carrier Transmitter Linearisation

Figure 3-5: Simulated time domain output signal of the QPSK modulator

The characterisation of continuous-time signals given above are usually carried over to discrete-time signals. Such signals are obtained by sampling a continuous-time signal uniformly at a sufficiently high rate and expressed as:

\[
S[n] = I[n]\cos\left(\frac{2\pi F_c}{F_{\text{samp}}} n\right) + Q[n]\sin\left(\frac{2\pi F_c}{F_{\text{samp}}} n\right)
\]  

(3.6)

The sampling rate \( F_{\text{samp}} \) in Hz is high enough to satisfy \( F_{\text{samp}} > 2F_c + BW \), where \( F_c \) and \( BW \) are the intermediate carrier frequency and the bandwidth in Hz, respectively.

In practice, a DDS (Direct Digital Synthesiser) produces sinusoids at a given frequency, and digital mixers in QPSK modulator are simply multipliers that generate quadrature modulation. The outputs from the multipliers are summed and fed to a DAC (Digital to Analogue Converter).
3.2.1.3 Spectral Efficiency of QPSK Signals

The spectral efficiency of a modulation scheme is defined as the ratio of the bit rate to the bandwidth, $f_b/BW$, expressed in bit/s/Hz. In the case of QPSK, the bandwidth of the signal is $f_s^*(1 + \alpha)$. Hence, its spectral efficiency is:

$$\eta_{QPSK} = \frac{f_b}{BW} = \frac{f_b}{(1 + \alpha)f_s} = \frac{2}{1 + \alpha} \tag{3.7}$$

since $f_s = f_b/2$.

For the minimum bandwidth case ($\alpha = 0$) the theoretical spectral efficiency of QPSK systems is 2 bit/s/Hz. However, for practical filters with a roll-off factor of $\alpha = 0.35$, the spectral efficiency is 1.5 bit/s/Hz. Figure 3-6 shows the simulated spectrum of the filtered inphase signal and the spectrum of the DSB (Double Sideband) QPSK modulator output signal with a roll-off factor $\alpha = 0.35$. The system bandwidth is 1.35 times 3 dB bandwidth, where 3 dB bandwidth is equal to symbol rate.
Figure 3-6: Simulated spectra
(a) Inphase signal after being passed through the baseband filter
(b) QPSK modulated signal
3.2.1.4 Measured Spectra of QPSK Modulator Output Signal

Once the QPSK modulated carrier is combined it’s then scaled, and the data file generated by MATLAB program is then loaded to a TTI-TGA1244 four channel arbitrary waveform generator with selectable output filter, operating at a maximum sampling frequency of 40 MHz and maximum operating voltage of 20 V (Volts) peak to peak. It should be noted that the operating sampling frequency and the output voltage of the arbitrary waveform generator dictate the actual output signal characteristics including the carrier frequency and bandwidth of the signal, which in turns changes the bit rate, and the magnitude of the output signal. The arbitrary waveform generator is used to feed the analogue IF signal on one of the four channels to an Agilent E4407B spectrum analyser operating in a frequency range of 9 kHz to 26.5 GHz, a 20 dB attenuator was connected to the input of the spectrum analyser as shown in the measurement set-up of Figure 3-7.

![Measurement set-up for QPSK output modulator](image)

The TGA1244 has a true variable-clock architecture, it can also operate in direct digital synthesis mode and it’s capable of producing up to four waveforms, which can be either fully independent or linked using simple or complex relationships including full interchannel triggering, summing and phase control. The channels can be phase-locked with user defined phase angles, allowing the generation of multiphase waveforms or locked waveforms of different frequencies. It can generate signals within the frequency range 0.001 Hz to 16 MHz. Waveforms may be defined with 12 bit vertical resolution and from 4 to 65536
horizontal points. GPIB (General Purpose Interface Bus) interface is used as standard interface for the downloading of data files generated by MATLAB programs from the personal computer. In this set-up, the measurement is been made at the final stage of the QPSK modulator using one channel. Figure 3-8 is a spectrum analyzer plot of the modulator DSB output centered at a carrier frequency of 2.171 MHz and a baud rate of 1030 kbaud. The 1.03 MHz 3 dB bandwidth, which is half bit rate and 1.39 MHz stopband edges (excess bandwidth/system bandwidth) can be read from the diagram.

![Figure 3-8: Measured spectra of DSB QPSK output signal](image)

From the measured spectra of the QPSK modulated output signal, the side lobes are almost 60 dB below the carrier, and there is no widening of the signal, which is often referred to as spectral regrowth. If the inphase and quadrature signals of the QPSK modulator are unfiltered, the digital system in which the carrier is modulated with rectangular pulses may give rise to infinite sidebands following a \( \frac{\sin(x)}{x} \) law. Interestingly, the sidebands may still be regenerated owing to nonlinear effects in the high power amplifier and mixer. These effects are discussed in the following section.
3.2.2 RF Stage

In the single carrier transmitter (see Figure 3-1 section 3.2), the QPSK modulated signal is generated digitally using a digital IQ modulator. The digital signal with varying envelope is then converted to analogue format and upconverted by a mixer to an RF signal before feeding into the power amplifier. QPSK modulation was used because of its ability to increase the data transmission rate. However, the benefit of using linear modulation schemes such as QPSK can only be maintained if the fidelity of the complex modulating signal can be preserved through the transmitter, particularly through the upconverter and the PA stage.

3.2.2.1 Nonlinearities

Generally the nonlinearities are mainly caused by two nonlinear components in a transmitter, mixers and power amplifiers. The mixer is a high frequency device, whose modelling is similar to the power amplifier. It is usually to operate the mixer at its maximum output in order to reduce the amount of gain required in succeeding RF amplification stage. However, nonlinearity in the mixer limits the maximum size of the output signal by producing IMD in the desired signal band causing spectral regrowth. The interference that it produces will in turn drive the PA and result in unwanted products in the final transmitted signal. Unwanted emissions from an interfering transmitter may fall within a desired channel resulting in degradation of a wanted signal operating in that channel.

The nonlinear distortion is characterized, measured, and specified by various techniques, depending upon the specific signal and application.

The C/I (Carrier to Intermodulation) ratio compares the amplitude of the desired output carriers to the IMD products. ACPR (Adjacent Channel Power ratio) compares the power in an adjacent channel to that of the signal. It is currently the most widely used measure of linearity. EVM is the distance between the desired and actual signal vectors. ACI it is typically defined as the ratio of the spectral density of the modulated signal to the maximum level of spectral density outside the channel in question when is used as a measure of the
spectral purity for a given transmitter. Often in this thesis it’s the improvement in ACI that is presented rather than the actual ACI level for a given linearisation circuitry.

Estimation of spectral regrowth and nonlinear distortion has been approached in a variety of ways. The excitation of a nonlinear circuit by a large number of input tones based on the spectral balance method was analysed in [50], and the resulting algorithm was used to predict ACPR. Another method that was used to characterise spectral leakage to adjacent channels have been reported in [51], where a least squares fitting of a power series to AM-AM (amplitude modulation to amplitude modulation) and AM-PM (Amplitude Modulation to Phase Modulation) transfer data was used to predict amplitude and phase transformation through nonlinear microwave power transistors. It was found that ACPR can loosely correlate to the third order IP3 (Intermodulation Product), although the presence of strong fifth-order nonlinearity either due to loading or intrinsic device characteristics can impact ACPR in the TDMA (Time Division Multiple Access) digital system. In [52] a method was developed to predict ACPR based on a time domain analysis technique and bandpass nonlinearity theory. AM-AM and AM-PM transfer characteristics are used to directly predict samples of the output complex envelope based on samples of an input complex envelope and the algebraic expression for a bandpass nonlinearity given by the describing function and corresponding nonlinear phase and amplitude.

Volterra series provides the most general form of analytical analysis of nonlinear circuits. However, it is typically a laborious process to derive a formulation even for the simplest of circuits. Nevertheless, it is a useful analysis technique for assessing spectral regrowth and has received more attention in literature. Analytical expressions for gain compression and phase distortion from a third-order Volterra nonlinear transfer function model was reported in [53], and these generated expressions were to predict spectral regrowth of a MESFET (Metal Semiconductor Field Effect Transistor) power amplifier. The method presents a connection between IMD and spectral regrowth, but is limited by the increasing complexity of the Volterra analysis for transfer functions above third order. In [54] a method where the modulated input signal as a sum of sinusoids was used and third order Volterra analysis was applied to find all third order terms that end up about the carrier frequency. A commercial Volterra software package was used to simulate the resulting spectral regrowth generated
Chapter 3. Single Carrier Transmitter Linearisation

about the carrier. However, the results were not compared to measurement or theoretical data to determine limitations of the analysis or the number of sinusoids needed to accurately represent the modulated input signal. In [55], spectral regrowth analysis from Volterra kernel models was extracted from CW (Continuous Wave) measurements of microwave nonlinear circuits, and in [55] time-varying Volterra series was used to analyse spectral regrowth generated by a microwave mixer circuit showing that a large number of tones could be efficiently used with Volterra analysis to solve for the spectral regrowth.

Although such analysis were helpful in underlying existing methods for estimating the nonlinear distortion, the objective of this thesis is the suppression of spectral regrowth to reduce adjacent channel interference and minimisation of in-band distortion to improve BER.

### 3.3 Single Carrier QPSK LINC Transmitter

The LINC technique [14] amplifies two constant amplitude signals, which represent an input signal to be amplified. The LINC technique uses a signal components separator [56] to split the input signal \( S(t) \) into two signals \( S_1(t) \) and \( S_2(t) \), which are constant envelope, phase varying signals. The LINC may be supplied a complex baseband digitally sampled signal [57]. The baseband signals can be a representation of any linear modulation scheme.

The LINC architecture of radio transmitter often takes the form shown in the simplified block diagram of Figure 3-9.
A problem occurs when the conventional LINC transmitter in Figure 3-9 is considered for communication systems. A very high degree of accuracy in the phases and amplitudes of the two constant envelope signals $S_1(t)$ and $S_2(t)$ [58] is required in order to achieve proper operation. This accuracy is partly dependent on the gain and phase imbalances as well as dc offset in the analogue IQ modulators [25]. Imperfections in the analogue IQ modulator result in unwanted envelope fluctuations resulting in changes in the characteristics of the signals $S_1(t)$ and $S_2(t)$ that are supposed to be constant envelope signals. As a result, when the signals $S_1(t)$ and $S_2(t)$ enter the power amplifiers, nonlinear distortion will appear at the transmitter output. In [48], the simulation of a typical LINC arrangement where the analogue IQ modulators have identical misalignments of 0.5 dB gain, 5 degree phase and 5% dc offset for a TWT (Traveling Wave Tube) amplifiers operating at 2 dB output back-off, it only generate a signal which is 40 dB above its intermodulation distortion, down from a 50 dB when minimal errors and TWT amplifiers operating at 2 dB output back-off were used. This is not sufficient for most radio transmitter applications where 60 dB is often required. Compensation techniques that have appeared in the literature to compensate for the IQ
imbalance will be reviewed in chapter five when the direct conversion OFDM system using analogue IQ modulator is considered for broadband applications.

The analogue IQ modulators are usually used for their wide bandwidth. However, for applications which require narrower bandwidth and relatively low intermediate frequency, digital IF transmitter, which was described in the beginning of this chapter and depicted in Figure 3-10, provides accurate IQ generation by allowing carrier to be implemented purely in the digital domain, which provides high degree of signal processing accuracy.

![Figure 3-10: Block diagram of proposed digital IF transmitter](image)

A common problem with this arrangement is that RF mixing process will produce many spurious products, as well as the main sum and difference frequency components. These arise through mixing of the LO (Local Oscillator) harmonics with harmonics of the IF input signal, often referred to as 'm' x 'n' products. These harmonics can be reduced by ensuring the IF port is operated well below compression. However, it's recommended to operate the mixer at its maximum output in order to reduce the amount of gain required in succeeding RF amplification stage. Spurious products, which fall at offsets far from the wanted

40
frequency-band, can be suppressed through filtering, but those close to the carrier will not be attenuated. On the other hand, since the LINC technique will be used to linearise the power amplifier. In this case the IF signal will itself be a constant envelope signal, which has the advantage that the RF mixer can be operated near compression without causing spectral regrowth.

To explain the principle in more detail consider a general bandpass [56] input signal:

\[ S(t) = a(t) \cos(2\pi f_c t + \phi(t)) \]  

(3.8)

Where \( a(t) \) is the amplitude, \( \phi(t) \) is the instantaneous phase and \( f_c \) is the frequency of the source signal \( S(t) \).

This signal can be split in two constant envelope signals [56,57], \( S_1(t) \) and \( S_2(t) \), such that

\[ S_1(t) = \frac{a_{\max}}{2} \sin(2\pi f_c t + \phi(t) + \alpha(t)) \]  

(3.9)

and

\[ S_2(t) = \frac{a_{\max}}{2} \sin(2\pi f_c t + \phi(t) - \alpha(t)) \]  

(3.10)

where

\[ \alpha(t) = \sin^{-1}\left(\frac{a(t)}{a_{\max}}\right) \]  

(3.11)

with \( a_{\max} \) represents the maximum magnitude value of \( a(t) \). Earlier papers, e.g. [14] suggested analogue solutions where the signal components separator operated at some intermediate frequency or directly at the carrier frequency. The complexity, however, of these systems prevented the LINC technique from becoming widely accepted.

Today, the evolution of digital signal processing techniques has made it possible to implement the signal components separator using software or digital hardware.
To implement the signal components separation using digital signal processing, a perpendicular projection, $E(t)$ is added to and subtracted from the input signal $S(t)$ to form the two constant envelope signals $S_1(t)$ and $S_2(t)$ as seen in Figure 3-11 [59].

\begin{figure}
\centering
\includegraphics[width=0.5\textwidth]{vector_diagram}
\caption{Vector diagram showing the relation between the input signal and the constant envelope signals}
\end{figure}

From Figure 3-11 the constant envelope signals $S_1(t)$ and $S_2(t)$ are given as:

\begin{align}
S_1(t) &= S(t) + E(t) \\
S_2(t) &= S(t) - E(t)
\end{align}

(3.12) and (3.13)

where

\begin{align}
|S_1(t)| &= |S_2(t)| = a_{\text{max}}
\end{align}

(3.14)
Chapter 3. Single Carrier Transmitter Linearisation

\[ 2S(t) = S_1(t) + S_2(t) \quad (3.15) \]

Using either the graph in Figure 3-11 or from equations (3.12) and (3.13),

\[ E(t) = S_1(t) - S(t) \quad (3.16) \]

and equivalently

\[ E(t) = S(t) - S_2(t) \quad (3.17) \]

Multiply equation (3.16) by equation (3.17), and the outcome is:

\[ E(t)^2 = (S_1(t) - S(t))(S(t) - S_2(t)) \tag{3.18} \]

\[ = S_1(t)S(t) - S_1(t)S_2(t) - S(t)S(t) + S(t)S_2(t) \]

\[ = S(t)(S_1(t) - S_2(t)) \tag{3.18} \]

\[ = S_1(t) - S_2(t) \tag{3.18} \]

Since \( 2S(t) = S_1(t) + S_2(t) \) in equation (3.15)

\[ E(t)^2 = S(t)(2S(t) - \frac{S_1(t)S_2(t)}{S(t)} - S(t)) \tag{3.19} \]

\[ = S(t)^2(2 - \frac{S_1(t)S_2(t)}{S(t)^2} - 1) \]

\[ = S(t)^2(1 - \frac{S_1(t)S_2(t)}{S(t)^2}) \]

\[ = -S(t)^2 \left( \frac{S_1(t)S_2(t)}{S(t)^2} - 1 \right) \]

Since \( |S_1(t)| = |S_2(t)| = a_{\text{msg}} \) thus,
Chapter 3. Single Carrier Transmitter Linearisation

\[ E(t) = jS(t) \frac{a_{\text{max}}^2}{\sqrt{|S(t)|^2 - 1}} \]  

(3.20)

where \(|S(t)|\) is smaller or equal to \(a_{\text{max}}\). Representing \(E(t)\) in cartesian form,

\[ E(t) = j(I(t) + jQ(t)) \frac{a_{\text{max}}^2}{\sqrt{|S(t)|^2 - 1}} \]

(3.21)

\[ = jI(t) \frac{a_{\text{max}}^2}{\sqrt{|S(t)|^2 - 1}} - jQ(t) \frac{a_{\text{max}}^2}{\sqrt{|S(t)|^2 - 1}} \]

\[ = -Q(t) \sqrt{\frac{a_{\text{max}}^2}{|S(t)|^2 - 1}} + jI(t) \sqrt{\frac{a_{\text{max}}^2}{|S(t)|^2 - 1}} \]

Let

\[ C = \frac{a_{\text{max}}}{\sqrt{|S(t)|^2 - 1}} \]

(3.22)

Where \(C\) is a constant, and \(I(t)\) and \(Q(t)\) are the inphase and quadrature components, respectively, of the input signal \(S(t)\).

By using equations (3.12) and (3.13) in cartesian complex form,

\[ I_1(t) + jQ_1(t) = I(t) + jQ(t) + (-Q(t) \cdot C + jI(t) \cdot C) \]

(3.23)

\[ = I(t) + jQ(t) - Q(t) \cdot C + jI(t) \cdot C \]

\[ = I(t) - Q(t) \cdot C + jQ(t) + jI(t) \cdot C \]

\[ = (I(t) - Q(t) \cdot C) + j(Q(t) + I(t) \cdot C) \]

and

44
Chapter 3. Single Carrier Transmitter Linearisation

\[
I_2(t) + jQ_2(t) = I(t) + jQ(t)(-Q(t) \cdot C + jI(t) \cdot C)
\]
\[
= I(t) + jQ(t) + Q(t) \cdot C - jI(t) \cdot C
\]
\[
= I(t) + Q(t) \cdot C + jQ(t) - jI(t) \cdot C
\]
\[
= \left( I(t) + Q(t) \cdot C \right) + j(Q(t) - I(t) \cdot C)
\]

(*) denotes multiplications. Following from equations (3.23) and (3.24), and to satisfy equation (3.15), we get:

\[
I_1(t) = \frac{(I(t) - C \cdot Q(t))}{2}
\]
\[
Q_1(t) = \frac{(Q(t) + C \cdot I(t))}{2}
\]
\[
I_2(t) = \frac{(I(t) + C \cdot Q(t))}{2}
\]
\[
Q_2(t) = \frac{(Q(t) - C \cdot I(t))}{2}
\]

where \((I_1(t), Q_1(t))\) and \((I_2(t), Q_2(t))\) are the inphase and quadrature components of the two constant envelope signals \(S_1(t)\) and \(S_2(t)\), respectively. The above algebraic was devised by the author as to my knowledge; no detailed analysis has been recorded to refer the readers to. Figure 3-12 shows the proposed digital IF LINC transmitter with detailed algorithm for digital signal components separator.
The two constant envelope signals $S_1(t)$ and $S_2(t)$ that are representing the intermediate frequency signals $IF_1$ and $IF_2$, respectively, are upconverted individually using two RF mixers then amplified and sent to the power combiner, the resultant signal is the desired amplified replica of the original signal $S(t)$. Hence, the nonlinear characteristics of the nonlinear mixers and power amplifiers are completely avoided by feeding each of them with constant envelope signal.

It should be noted that the inphase and quadrature components $(I_1(t), Q_1(t))$ and $(I_2(t), Q_2(t))$ of the constant envelope $S_1(t)$ and $S_2(t)$, respectively, are scaled before feeding each of them to the digital IQ modulators in order to satisfy equation (3.25).
3.3.1 Experimental Results

Transmitter measurements are typically made at the power amplifier port. In this case, however, it was necessary to examine the transmitter at various test points. These are: the constant envelope signal $S_1(t)$ in order to determine if the LINC algorithm is functioning properly. Next, the measurements made at the output port of the mixer for both linearised and unlinearised signals. These signals are then compared to evaluate the effectiveness of the LINC in reducing spectral regrowth. Finally, measurements are made at the power amplifier output port, and comparison of the power spectrum results are made to determine the effectiveness of the LINC.

In the following experiments, the DUT (Devices Under Test) are a mini-circuit power amplifier model ZHL-42W and mini-circuit mixer model ZEM-4300. The characteristics of the power amplifier [60] and of the mixer [61] are shown in Table 3.1 and Table 3.2, respectively.

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Gain (dB)</th>
<th>DC Power (Volt)</th>
<th>DC Power (mA)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$F_L$ 2000</td>
<td>30</td>
<td>+15</td>
<td>880</td>
</tr>
</tbody>
</table>

Table 3-1: Characteristics of the power amplifier, model ZHL-42W

Where $F_L$ and $F_U$ are the low and upper frequency in the operating frequency band.
The input to the RF mixer is a 1.03 MHz 3 dB bandwidth signal centered at a carrier frequency of 2.171 MHz. The input to the power amplifier is the same as the input to the mixer except the signal is shifted in frequency by 1.8 GHz to an RF signal using local oscillator, which is generated by the Agilent LO generator, model E4422B.

In all cases, the baseband input data is a QPSK modulation signal with a roll-off factor alpha of 0.35 as it was presented previously in section (3.2.1.4).

### 3.3.1.1 Measurement Results of the Constant Envelope Signals

In the QPSK digital IF LINC transmitter, the input source signal $S(t)$ represents the filtered inphase and quadrature components of the QPSK baseband modulation signals. Recall in section 3.2.1.1, Figure 3-4(c) where alpha was 0.35, from the constellation of the inphase versus the quadrature components of the QPSK modulation, the phase varies from 0 degree to 360 degree with varying envelope. Here, we are trying to demonstrate that the input signal $S(t)$ is decomposed to two signals $S_1(t)$ and $S_2(t)$ with a varying phase range from 0 degree to 360 degree but with constant envelope. Figure 3-13 shows a block diagram for measuring the constellation of one of the constant envelope signals, in this case is the upper branch $S_1(t)$.

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Conversion Loss @ LO=1.8GHz (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>LO/RF</td>
<td>IF</td>
</tr>
<tr>
<td>300 - 4300</td>
<td>DC - 1000</td>
</tr>
<tr>
<td>7.4</td>
<td>7.03</td>
</tr>
<tr>
<td>6.85</td>
<td></td>
</tr>
</tbody>
</table>

**Table 3-2: Characteristics of the mixer, model ZEM-4300**
The baseband modulation, the signal components separator algorithm and digital mixing are programmed using MATLAB software programming language, then the program is downloaded to the arbitrary waveform generator, which connected to the HP (Hewlett-Packard) model 54602B four channel 150 Mhz oscilloscope. Figure 3-14 shows the measurement set-up.
Figure 3-15 shows the measured constellation of the constant envelope signal $S_1(t)$.

As it can be seen from the diagram shown on the scope and presented in Figure 3-15, the circular constellation is slightly noisy. This may be due to the effect of the DACs and the analogue reconstruction filters that are embedded within the arbitrary waveform generator and/or may be due to the effect of the scope itself. However, as it was expected the constellation is circular, hence, the signal components separator algorithm in the QPSK digital IF LINC transmitter is functioning properly to produce a constant amplitude signals.

3.3.1.2 Measurement Results for Mixer Linearisation

The two constant envelope signals IF1 and IF2 are frequency translated with nonlinear RF mixer1 and RF mixer2, respectively, generating signals RF1 and RF2, which are recombined
to form a frequency translated replica of the input signal. Both RF mixers have the same characteristics.

Figure 3-16 shows the proposed system for mixer linearisation using the LINC technique.

![Diagram of mixer linearisation using LINC technique](image)

**Figure 3-16: Block diagram for frequency translation linearisation**

In accordance with the LINC, the inphase and quadrature baseband components I(t) and Q(t) are fed to the signal components separator, which in turn generates the inphase and quadrature components (I₁(t), Q₁(t)) and (I₂(t), Q₂(t)) of the two constant envelope signals S₁(t) and S₂(t), respectively. The four components are modulated to IF signals using a pair of digital IQ modulators each generating a constant envelope signal namely S₁(t) and S₂(t). The constant envelope signals are converted to analogue signals, upconverted to RF signals using a pair of analogue mixers, filtered and then recombined to generate a frequency-shifted replica of the input signal. Since the input signals to the analogue RF mixers are constant, they can operate close to saturation without causing nonlinear distortion at the output signal.

In order to determine the effectiveness of the LINC technique in linearising the mixer, a two-stage upconversion prototype system was carefully assembled as shown in Figure 3-17.
Chapter 3. Single Carrier Transmitter Linearisation

Data file from MATLAB program that includes all the QPSK modulator baseband processing and the signal components separator algorithm is loaded to the arbitrary waveform generator, which in turn generates two analogue intermediate frequency constant envelope signals IF1 and IF2 with power level to drive the mixers into compression. The arbitrary waveform generator is connected to the personal computer through a GPIB.

The two constant envelope signals IF1 and IF2 are mixed with LO1 and LO2 to form RF1 and RF2 signals, respectively. Both mixers are mini-circuit, model ZEM-4300, and they were operating near compression point. The power available at the IF ports of the mixers was delivered by the arbitrary waveform generator. It should also be noted that LO1 and LO2 signals have same characteristics, i.e. they are in phase, operating with similar frequency and both have equal amplitude. The LO signals were generated from a power splitter, which is connected to an LO signal generator from Agilent, model E4422B.
Two bandpass filters were required to block the image band caused by mixing. However, due to the fact that bandpass filters were not available, for image rejection, a zoom function on the spectrum analyser was used to zoom into the wanted band instead. The author regrets that BPFs were not used to determine their effect on the overall system. Nevertheless, the general idea and the experiment as a whole should perform as predicted since the unwanted image band is far from the desired band.

The two mixed signals RF1 and RF2 are then recombined with a power combiner to form an upconverted replica of the input signal, which then fed to the input port of the spectrum analyser. During the measurements the input port of the spectrum analyser was connected to a 20 dB attenuator.

Figure 3-18 shows spectrum analyser plots of the frequency-translated signal before and after linearisation.
Figure 3-18: Effectiveness of LINC technique in suppressing spectral regrowth, when the nonlinear mixer is used.
(a) Spectra without the LINC technique
(b) Spectra with the LINC technique

Results in Figure 3-18 showing reduction of spectral spreading when mixers were driven deep into compression point in QPSK transmitter. The LINC technique improves the spectral of the frequency translated QPSK signal by almost 20 dB, but this relies on both mixers to maintain high degree balancing of the two paths, in other words, accurate amplitude and phase match is required from both mixers in order to obtain a replica of the original signal. Nevertheless, the mixers are operated at their maximum output in order to reduce the amount of gain required in succeeding RF amplification stage without producing major IMD in the desired signal band.
3.3.1.3 Measurement Results for QPSK Digital IF LINC Transmitter

Once the mixer linearisation measurement results were recorded, the frequency translated signals RF1 and RF2 are then amplified with the two high efficient but nonlinear power amplifiers (see Fig 3-12) generating an amplified replica of the input signal, which is obtained by summing the two upconverted and amplified output constant envelope signals.

Figure 3-19 shows prototype system for QPSK digital IF LINC Transmitter

![Prototype system for QPSK digital IF LINC transmitter](image)

Figure 3-19: Prototype system for QPSK digital IF LINC transmitter

Duplicating the procedure in the previous prototype system given in Figure 3-17, then the upconverted signals RF1 and RF2 are amplified using two Mini-Circuits ZHL-42W high power amplifiers as shown in Figure 3-19. The combined amplified RF output signal is attenuated by 20 dB in order to compensate for the gain of the amplifiers, and measured using a spectrum analyser.
Since the input signal to each nonlinear high power amplifier has a constant envelope, the power amplifiers are operated near saturation in order to obtain high efficiency.

Figure 3-20 shows spectrum analyser plots of the QPSK digital IF transmitter before and after linearisation.

The measurement results show that the LINC technique improved the spectral regrowth by almost 20 dB, when nonlinear mixer and power amplifier used in the QPSK digital IF transmitter were operating near compression.

From Figure 3-20(b), it can be seen that although the upconverted and amplified signal is not exactly a copy of the signal presented in Figure 3-8, section 3.2.1.4, significant reduction of spectral regrowth can still be obtained by using the proposed LINC method.
The reason that the proposed LINC method was not very effective in completely suppressing spectral regrowth is its sensitivity to amplitude and phase imbalances between the two signal paths. Recall from equations (3.12) and (3.13) that if the two signal branches are perfectly matched, \( E(t) \) will ideally cancel its anti-phase counterpart when the two signals \( S_1(t) \) and \( S_2(t) \) are upconverted, amplified and recombined, leaving only the upconverted and amplified source signal at the transmitter output. Using a conventional hybrid combiner is a convenient solution as it can provide high isolation and well-defined impedances, but it requires the input signals to be identical to avoid power losses. If the two input signals are uncorrelated the loss will be 3 dB and for the LINC transmitter the loss is sometimes even worse. More efficient combining techniques [62] exist, but they require the amplifiers to act as ideal voltage sources as the load impedance for each amplifier varies with such signal combiners.

Although in the LINC structure considered in this chapter the IQ modulation is performed digitally, which means better IQ matching can be achieved, an automatic control system based on feedback loop is advised for this scheme to achieve high degree of balance between the two signal paths. However, as the signal bandwidth gets wider, feedback systems begin to exhibit stability effects due to the delay in the loop.
3.4 Conclusion

It was demonstrated in this chapter that the filtered QPSK signal proved to be sensitive to nonlinear distortion. As a result, the transmitted signal suffers a degree of spectral regrowth originally removed during filtering. In many cases, transmitted signals must conform to stringent specifications or government regulations regarding spurious spectral emissions. The occurrence of spectral regrowth can lead to RF emissions exceeding specified spectral masks.

This chapter considered the LINC to linearise two-stage upconversion transmitter for narrowband applications. By adding a digital signal components separator in the baseband, both mixer and power amplifier were allowed to operate into their nonlinear region, thereby significantly increasing their power efficiency. The power efficiency provides longer battery life for mobile terminal users and reduces heat in base stations. The test carried out for QPSK modulated carrier has shown almost 20 dB ACI improvement at the upconverted and amplified output signal.

In addition, by using digital IF signal, the effect of analogue IQ modulators imperfections on LINC technique was reduced, ac coupling is also possible because the analogue IF signal is no longer centered at dc. As a result, the dc offset of the analogue circuits before the mixer is eliminated and consequently, will not cause LO leakage. The digital linearisation technique of the transmitter potentially offers many other benefits, including programmability and reconfigurability, absence of tuning or aging problems, as well as easy integrability and testing. Future advances in ASIC (Application Specific Integrated Circuit), FPGA (Field Programmable Gate Array) and DACs technology will no doubt eventually allow for an increased bandwidth and totally soft configuration.

In summary, the use of LINC technique and a digital IF stage results in a prototype transmitter with excellent improvement in linearity, which can be useful in modern and future wireless communications systems, for both mobile terminals and base station applications.
Chapter 4

OFDM LINC Transmitter

4.1 Introduction

In recent years the growth of multimedia services and broadband applications in digital data transmission has led to ever increasing demands on throughput capacity of wired as well as wireless communications systems. The coexistence of many services and simultaneous users on the support of single frequency networks has become feasible since the implementation of techniques like OFDM [1] and COFDM (Coded Orthogonal Frequency Division Multiplexing) were possible at low cost thanks to the evolution of digital signal processing techniques and the expansion of its capabilities. For instance, current technical standards are already considering bit rates as high as 54 Mbps for broadband digital communications while recent investigations on wireless indoor multimedia communications have attempted to reach rates of up to 155 Mbps [63]. These trends, in addition to the development of more powerful hardware for specific new digital signal processing applications, has widely extended and made possible the use of bandwidth efficient modulation schemes, such as m-PSK and m-QAM, for coded digital broadband systems. In this context, nonconstant amplitude modulations, in particular the OFDM schemes have been adopted as the standard transmission technique in the framework of a variety of next generation communications services. Indeed, OFDM systems are already operational for DAB (Digital Audio Broadcasting) [2,9,64] and DVB-T (Terrestrial Digital Video Broadcasting) [7,65] communications, while it has also been adopted for the new high bit rate WLAN/MAN (Wireless Local/Metropolitan Area Networks) standards, as IEEE802.11a-1999 and HIPERLAN II ETSI specifications [6,66], as well as the Japanese MMAC (Multimedia Mobile Access Communications).
In OFDM systems, the entire channel is divided into N orthogonal narrow subchannels or subcarriers, and the high-rate data are transmitted in parallel through the subchannels at the same time. Therefore, the symbol duration is N times longer than that of single carrier systems.

The advantages offered by OFDM techniques are manifold. The most outstanding of them being its robustness in transmitting information over frequency selective channels where coding techniques [67] can be implemented to reduce the system degradation due to multipath fading.

Multipath fading of wireless channels leads to ISI, which limits the transmission rate of single carrier systems. In single carrier communication systems, ISI is usually dealt with by a time domain channel equaliser [10]. However, when the data rate increases, the symbol duration reduces and the equaliser becomes complex. OFDM is an elegant solution to the severe ISI problem and requires no extensive equalisation.

Additionally, it is worthy to mention the simple implementation using DSP techniques associated to OFDM transmission and reception processes, which can be implemented as low-pass signal treatments based on the FFT (Fast Fourier Transform) algorithm.

In contrast to its appealing properties, an important number of studies [2-4] in this area have shown that a simple OFDM scheme performs poorly under the influence of nonlinear distortions present in along the transmission chain. The OFDM signal is characterised by a highly nonconstant envelope that results from the combination of several amplitude subcarriers at each symbol. This is expressed by a large PAPR that makes these systems extremely sensitive and vulnerable to the nonlinear distortion typically introduced by the high power amplifier [68]. The nonlinear distortion caused by the power amplifier determines, on one hand, the power spectral density outside the used channel leading to an impact on adjacent channel suppression and transmit signal spectrum. On other hand, the introduced inband distortion influences the modulation accuracy.

As derived analytically in [69], the maximal theoretical PAPR of OFDM modulated signal $S(t)$ sampled at the symbol rate with N subcarriers is:
Chapter 4. OFDM LINC Transmitter

\[ PAPR(S(t)) = \frac{\max|S(t)|^2}{E|S(t)|^2} = 10\log_{10} N \]  

(4.1)

Where \( E\{.\} \) denotes statistical averaging, and it’s usually expressed in dB.

This result, however, is quite pessimistic, for example, when 1705 subcarriers are used, the PAPR is approx 32.32 dB, most of the time the practical value of PAPR is much smaller. In fact, as described in [70] the PAPR of OFDM signals is in the range of 8-9 dB for a small number of subcarriers and 13-14 dB for a large number of subcarriers. Therefore, since it’s not practical to use equation (4.1) to determine the PAPR of OFDM signal, the CDF (Cumulative Distribution Function) is often used to evaluate the envelope of OFDM signals. CDF is the distribution function of the PAPR. The probability that the PAPR is below some threshold \( r \) can be written as:

\[ \Pr(\text{PAPR} \leq r) = (1-\exp(-r))^N \]  

(4.2)

Figure 4-1, taken from [70], shows the CDF of the PAPR in OFDM signal for different numbers of subcarriers.
As it can be seen from Figure 4-1 [70], that even when the subcarriers number $N$ was increased to 100000, the probability of the PAPR exceeding 15 dB is only about $10^{-8}$, which may be low for many applications.

Although the PAPR of OFDM signals in general is much higher than the PAPR of single carrier QPSK modulated signal [41], it's practically not as sensitive to the increase in the number of subcarriers as it was estimated in equation (4.1).

However, even though the practical values of the PAPR of OFDM signals are lower than was estimated, higher transmitter output back-off is required in comparison to a single carrier system. Hence, the PAPR reduction in OFDM system is desirable.

The LINC technique is considered in this chapter for OFDM transmitter linearisation. Detailed analysis of the LINC technique was presented in the previous chapter. The most outstanding advantage of the LINC scheme is the generation of the constant envelope signals from a signal with varying amplitude. Since amplitude variations do not have to be dealt with, it is possible to use high power amplifiers, which will amplify signals linearly by using the two phase modulated signals. The nonlinearity of the amplifiers is no longer a problem.
Chapter 4. OFDM LINC Transmitter

in the amplification of multiple signals or those containing large PAPR because the amplitude of the two generated signals become constant amplified amplitudes as they are amplified by amplifiers. Moreover, since the LINC scheme is based on an open loop configuration, it can remain stable and handle wideband signal, making it suitable for broadband applications such as OFDM systems.

Unlike the conventional transmitters, to the author's knowledge, during this work neither prototype systems nor practical measurement results demonstrating the effectiveness of the LINC technique for OFDM system have been previously reported.

This chapter organised as follows: The OFDM system used throughout this study is defined and the different blocks of the transmission chain are described. After a brief review of some well known reported techniques that used to reduce the PAPR in OFDM systems, prototype OFDM LINC transmitter and practical measurement results are then presented showing the effectiveness of the LINC in reducing spectral regrowth in OFDM signals.

4.2 OFDM Transmitter

OFDM is characterised by transmitting the signal on a large number of subcarriers (frequency division multiplexing) and thereby allowing each subcarrier to transport only a moderate bit rate. The transmission system considered in this work is mainly based on the OFDM scheme used for digital video broadcasting applications and reported in [7,65,67,71]. The basic model of an OFDM transmitter [67] is shown in Figure 4-2.
Chapter 4. OFDM LINC Transmitter

![Simplified block diagram of OFDM transmitter](image)

Figure 4-2: Simplified block diagram of OFDM transmitter

Basically, the OFDM signal is made up of a sum of N complex orthogonal subcarriers indexed with \( k = \{0, 1, 2, 3, 4, \ldots \ldots N-1\} \). In applications such as digital video broadcasting systems, each subcarrier is modulated using either QPSK, 16-QAM or 64-QAM. If we let QPSK data be \( c_k \) and \( f_c \) be the RF carrier frequency, then the transmitted signal \( S(t) \) [65] for one OFDM symbol, \( t=0 \) to \( t=T_s \), is specified as:

\[
S(t) = \Re\left\{ \sum_{k=K_{\min}}^{K_{\max}} c_k e^{j2\pi f_c(t - \Delta)/T_s} \right\}
\]  

(4.3)

Where

- \( \Re \) is the real signal
- \( k' \) is the carrier index relative to the centre frequency

\[
k' = K - \frac{K_{\max} + K_{\min}}{2}
\]  

(4.4)
Chapter 4. OFDM LINC Transmitter

\( f_c \) (Hz) is the RF centre carrier of the transmitted signal

\( k \) is the carrier number

\( K_{\text{min}} \) is the minimum carrier number

\( K_{\text{max}} \) is the maximum carrier number

\( \xi_k \) is the modulation of carrier \( k \)

\( T_u \) (\( \mu s \)) is the inverse of the carrier spacing or the useful symbol time

\( \Delta \) (\( \mu s \)) is the guard interval.

\( T_s \) (\( \mu s \)) is the total symbol duration:

\[
T_s = T_u + \Delta \quad (4.5)
\]

The OFDM signal is basically defined by the elementary time element \( T \):

\[
T = \frac{T_u}{N} \quad (4.6)
\]

where \( N \) is the IFFT size, and \( 1/T \) is often referred to as the system clock frequency for the OFDM system. \( \Delta \) and thus \( T_s \) are integer multiples of \( T \).

The data in the OFDM modulator is split into \( N \) streams, which are independently modulated on parallel closely spaced subcarrier frequencies. Since each subcarrier is QPSK modulated, hence the data rate \( R \) (bit/s) is:

\[
R = \frac{N_{\text{symbol}} \times 2 \text{bits/symbol}}{T_s} \quad (4.7)
\]

\( (*) \) denotes multiplications.
4.2.1 Guard Interval and Coding

The subcarriers are orthogonal and it is therefore possible to decode the signal even though there is some frequency overlapping of the individual subcarriers. Although the symbol rate of each subcarrier is moderate, ISI would still occur if no special measures were taken. To avoid the ISI, a guard interval is inserted before each symbol. The guard interval consists of a cyclic continuation of the useful symbol. It will ensure that the orthogonality of the subcarriers can be retrieved for the received signal, as long as these are inside the guard interval. The guard interval can have different values: 1/4, 1/8, 1/16 and 1/32 of the active symbol length. Clearly, from equation (4.5) and (4.7), the penalty of a long guard interval (= 1/4 of active symbol duration) is a lower data capacity.

However, even though the guard interval will maintain orthogonality of the received subcarriers, in applications such as digital video broadcasting, echoes will cause fading. It is therefore often necessary to use the C (Coding) in COFDM [67]. Coding is essential for the performance of the system. By choosing an efficient forward error correction coding scheme and interleaving it is possible to achieve sufficiently low error rates in fading channels. The details of the mechanisms for interleaving and error correction are explained more detailed in [65].

4.2.2 IFFT and Serial to Parallel Conversion

OFDM achieves orthogonality in the frequency domain by allocating each of the separate information signals onto different subcarriers. OFDM signals are made up from a sum of sinusoids, with each corresponding to a subcarrier. The baseband frequency of each subcarrier is chosen to be an integer multiple of the inverse of the symbol time, resulting in all subcarriers having an integer number of cycles per symbol. As a consequence the subcarriers are orthogonal to each other.

There is a clear resemblance between equation (4.3) and the IDFT (Inverse Discrete Fourier Transform):
Chapter 4. OFDM LINC Transmitter

\[ x_n = \frac{1}{N} \sum_{q=0}^{N-1} X_q e^{j2\pi nq/N} \]  \hspace{1cm} (4.8)

Since various efficient FFT algorithms exist to perform the DFT and its inverse, it is a convenient form of implementation to generate \( N \) samples \( x_n \) corresponding to the useful part, \( T_s \) long, of each symbol. The IFFT size is \( 2^m \), for example, when \( m = 11 \), which is the values considered in this work and often used for most communication systems such as DVB-T 2K mode. The IFFT size will then be 2048, and this value determines the maximum number of subcarriers to be used. In practice a number of subcarriers at the bottom and top end of the OFDM spectrum will be left out in order to allow for separation between channels (guard band).

The data to be transmitted is typically in the form of a serial data stream. In OFDM, each symbol typically transmits 40-4000 bits, and so a serial to parallel conversion stage is needed to convert the input serial bit stream to the data to be transmitted in each OFDM symbol. The data allocated to each symbol depends on the modulation scheme used and the number of subcarriers. For example, in this work, the subcarrier is QPSK modulated and each subcarrier carries 2 bits of data. Using a transmission of 1705 subcarriers then the number of bits per symbol would be 3510. Since only 1705 subcarriers are used for data transmission, then the outer subcarriers are unmodulated and set to zero amplitude. These zero subcarriers provide a frequency guard band before the Nyquist frequency and effectively act as an interpolation of the signal and allows for a realistic roll off in the analogue anti-aliasing reconstruction filters.

4.2.3 RF Stage and Simulated OFDM Output Signal \( S(t) \)

The OFDM signal is generated at baseband using complex samples, and then modulated up to the required frequency using an analogue IQ modulator. The analogue IQ modulator frequency shifts the OFDM signal from DC to the required RF frequency, and converts the
complex signal into a real signal $S(t)$. The time and frequency response of the real signal $S(t)$
are shown in Figure 4-3(a) and Figure 4-3(b) respectively.

![Graphs showing time and frequency domain responses](image)

Figure 4-3: OFDM output signal $S(t)$
(a) Time domain
(b) Frequency domain

As it can be observed from Figure 4-3(a), the value of the aforementioned PAPR of the RF signal in the time domain response is high. This is one of the main disadvantages of OFDM scheme because of the large amplitude swings of the signal will introduce very high intermodulation levels when the signal is nonlinearly amplified. In the frequency domain, the transmission spectrum as shown in Figure 4-3(b) will broadened and the intermodulation distortion will give rise to adjacent channel interference and inband interference, the most evident effect of which is an impairment in terms of bit error rate deterioration, which have been investigated separately in [72, 73]. As a result, linear behaviour of the system over a large dynamic range is needed and the efficiency of the output power amplifier is reduced.
4.3 PAPR Reduction Techniques

A number of approaches have been proposed for reducing the PAPR of OFDM signals. Some are based on the classical linearisation techniques that were reviewed in chapter two. Others approaches are based on the reduction of the PAPR of the transmitted signal. Those methods are coding, clipping and tone injection.

4.3.1 Coding

Coding techniques make use of the fact that only certain input data sequences will generate OFDM modulated signals with high PAPR. So it is possible to try to encode input data in a way that once modulated, the resulting signal exhibits a low PAPR.

4.3.1.1 Complementary Golay Sequences

Complementary Golay sequences [74] are a set of sequences that once modulated has the property of limiting the PAPR.

Detailed theoretical study of the characteristics of complementary Golay sequences is presented in [74] and their performance is evaluated by simulation and practical implementation in [75]. The simulation results in [75] show that the code can reduce the PAPR of the transmitted signal to 3 dB, while practical results shows that the PAPR of the transmitted signal is actually reduced from 9 dB to only 6 dB.

The use of this coding scheme has several advantages. First, theoretically, the PAPR is limited to 3 dB and independent of the characteristics of the input data and the number of subcarriers. Second, the same DSP that implements OFDM can be used to implement the coding scheme too. Third, there is a coding gain so their error correction capabilities may be used to improve the system margin.

The main disadvantages, however, are the fact that all the subcarriers must be modulated in the same way, which must be an m-PSK modulation scheme. Secondly, in addition to the
need of a complex decoder, it is not clear whether pilots can be used with this scheme. Pilots are usually used in OFDM systems for synchronization.

### 4.3.1.2 Hadamard and Trellis Codes

In [76], the use of a Hadamard transform before the IFFT in the transmitter, and later in the receiver prior to FFT is proposed. Simulated results in [76] show that by using Hadamard transform, the PAPR can be reduced by about 1.2 dB for a QPSK modulated OFDM system with 32 subcarriers. The reduction in PAPR, however, was slightly larger for 16QAM modulated OFDM system with 32 subcarriers.

The advantages of this method are its simplicity and the fact that the same DSP that implements the FFT algorithms can also implement the encoding-decoding procedure.

Disadvantages are the low reduction, around 1.2 dB, although same reduction is achieved by other schemes that are much more complex. The number of subcarriers also has a minor effect on PAPR reduction performance.

### 4.3.1.3 Trellis sequences

In [77] trellis code shaping is discussed as a possible candidate to be used for PAR reduction. Simulation results in [77] show that the PAPR reduction achieved with this method for a QAM modulated OFDM signal is very low but it has the advantage of low complexity since it does not require additional FFT.

### 4.3.2 Orthogonal Pilot Sequences

Pilot Sequences are primarily provided for channel estimation and can also be employed for frequency synchronization. However, in [78] a new use of the pilots for PAPR reduction is proposed. At the transmitter, the procedure will be to check all possible pilot sequences included in the set per OFDM symbol and then to choose for transmission the pilot sequence that generates the lower PAPR per symbol.
This technique is still requires the transmission of some additional information to the receiver to determine which pilot sequence was transmitted. It is a desirable condition that the elements of this set are orthogonal, because in that case, there will be no need of transmitting the additional side information required to inform about which element of the set has been chosen.

Results in [78] show that PAPR reduction of 2 to 3 dB can be achieved for QPSK modulated OFDM schemes using a small set of pilots. In this way, the complexity in the transmitter is low. Also, when this proposal is only applied to OFDM symbols exceeding a certain PAPR threshold, which are the ones that should be more desirably reduced, the PAPR reduction increases to even larger values. The main advantage of this scheme is the fact that the same set of pilots that usually used for channel estimation or frequency synchronization can also be reused to reduce the PAPR in OFDM system. Regarding disadvantages, this scheme depends on input data, and in some sense on the number of subcarriers. In other words, as the number of subcarrier increases the effectiveness of the scheme decreases.

### 4.3.3 Selective Mapping

In [79] an approach based on selective mapping scheme was proposed. In traditional selective mapping, statistically independent alternative transmit sequences are generated which represent the same information. Finally, the sequence that lowers the PAPR is then selected for transmission. In order to perform the appropriate inverse operation in the receiver, it is necessary that the transmitter send information about which sequence was chosen. This means that side information is needed. Also, the transmitter must perform different IFFTs and then calculate resulting PAPR in order to select the sequence that exhibits the lowest PAPR.

In [79], all possible sequences are completed with a prefix with some information called label. These labels are scrambled and additionally convolutionally encoded before insertion into the symbol. Then, the transmitter performs the IFFT and selects the sequence with lowest PAPR. In this way, the side information is lower and besides it is protected against errors. Results in [79] show PAPR reduction of 1 to 2 dB for a small number of subcarriers.
Chapter 4. OFDM LINC Transmitter

The main advantage of this scheme is the error protection of the side information. Also, it may be used with a different modulation type in each subcarrier. However, this scheme has several disadvantages. The major one, however, is complex in the transmitter to perform large number of IFFTs and calculate symbol PAPRs. All these calculations must been performed in real time.

4.3.4 Partial Transmit Sequences

This technique can be considered as a particular case of selective mapping [79]. The procedure is as follows. First, the OFDM symbol is divided into M blocks, example, for N = 256 and M = 16, they are 16 blocks of each 16 subcarriers. Then, each block is multiplied by a different phase, and the combination that causes the lowest PAPR is then chosen for transmission. The optimal partial transmit sequences searches the best phase combination in order to obtain the lowest PAPR, nevertheless, this is a complex procedure even for a small number of blocks.

In [80,81], a sub-optimal algorithm to find a proper phase combination is proposed, which avoids performing a search for all the phase combinations. However, the receiver must have the knowledge of which phase combination was sent, so some side information must be transmitted and consequently, the efficiency decreases.

This algorithm is very simple as it begins with all the phases equal to 1, then, the PAPR is calculated. The first block phase is switched and PAPR is again calculated. If new PAPR is lower than the former, this phase is stored; otherwise the phase is switched back. The algorithm proceeds by switching phases for all the blocks.

In [80] a new detection scheme is proposed in order to avoid the use of side information: only two possible phases are used, that is, two different constellations are allowed for each block, one rotated and one without rotation. The algorithm puts some markers into these blocks that must be rotated. The detection scheme first eliminates the modulation information by raising to the fourth the frequency symbols in the case of QPSK. Then, with the modulation removed, data symbols, in frequency domain, are differentially detected by
computing for each block. It was shown in [80] that a reduction in the range of 2 to 3 dB can be achieved.

The advantages of this scheme are its conceptual simplicity and reasonable PAPR reduction. The main disadvantage, however, is that it is needed to perform 2M IFFTs for each symbol in real-time. Also, the reduction is proportional to M. In addition, the scheme is only valid for PSK modulations and it is not possible to use a different modulation for each subcarrier.

### 4.3.5 Tone Injection

In this case, the constellation of the modulation in each subcarrier is extended, so that different constellation points can represent the same value. Then, each time the constellation point that generates the lower PAPR is used. It can be considered as the introduction of a new tone and hence the name.

In [82], the original $2^M$QAM constellation is extended to $2^{(M+m)}$QAM. In the decoder, the original constellation is known, so if an extended symbol is received, the decoder knows that an extension has happened and decodes correctly.

The main advantage of this technique is that it can be used with QAM constellations and 2 to 3 dB improvements in PAPR reduction was reported in [82]. However, the PAPR reduction with this new scheme depends on the number of tones injected [82], that is, as $m$ is increased a higher PAPR reduction is achieved. Hence, the disadvantages are the fact that the complexity increases for high reductions.

### 4.3.6 Clipping

Coding and tone injection are a desirable method to reduce the PAPR for small number of subcarriers since they do not introduce any distortion to the signal. However, as the number of subcarriers increases, techniques such as coding become intractable since the memory needed to store the codebook grows exponentially with the number of subcarriers. A simple
method for reducing PAPR, whose complexity does not depend on the number of subcarriers, is digital clipping [83].

Digital clipping of OFDM signal, however, causes significant spectral leak into adjacent channels. These out-of-band components must be filtered to prevent adjacent channel interference. In [84], a 128 subcarriers OFDM signal is clipped by a 1.4 clipping ratio and filtered by a 103-tap FIR filter to reduce spectral leakage. The results show that the clipped and filtered signal reduces the spectral leakage by almost 25 dB.

The disadvantages are the act of filtering in digital clipping causes peak regrowth. In addition there is in-band interference due to clipping itself. This introduces additional problems because the information contained in the clipped portion is completely lost.

4.3.7 Predistortion

In contrast to the above techniques, the classical linearisation schemes such as predistortion techniques are gaining attention in the recent literature [85] for OFDM transmitter linearisation. Besides, some classical linearisation techniques such as digital predistortion method can be used in conjunction with the above-mentioned techniques, for example, in conjunction with coding method.

Analysis of an adaptive predistorter were presented in [86], it reported improvements in the range of 10-12 dB were obtained. However, the improvements seem to decrease sharply by increasing the number of symbols. In addition, stability analyses of the system estimator and the predistorter in the proposed adaptive predistorter need to be performed in the future to determine the effectiveness of the scheme.

4.4 OFDM LINC Transmitter

As mentioned earlier, there is no prototype system demonstrating the ability of the LINC for OFDM systems. This section proposes the OFDM LINC transmitter, which will be constructed in prototype system to evaluate its effectiveness. The experiments will be
conducted to demonstrate that the LINC technique is superior to the above-mentioned methods when it used to linearise the power amplifier in OFDM transmitters.

The QPSK modulated OFDM LINC transmitter presented in the rest of this work is constructed on the basis of the architecture in section 3.3 and section 4.2. In this experiment 1705 QPSK modulated OFDM subcarriers were used to transmit information, each transmitting 2 bits in $T_s$ microseconds. The output signals, $I(t)$ and $Q(t)$ from OFDM modulator are then fed to the digital signal components separator.

Figure 4-4 shows a block diagram of the proposed QPSK modulated OFDM LINC transmitter.
In accordance with the LINC technique, the digital inphase and quadrature components $I_1(t)$, $Q_1(t)$, $I_2(t)$ and $Q_2(t)$ of the two constant envelope signals $S1(t)$ and $S2(t)$ are converted to analogue signals, filtered and upconverted to RF signals using a pair of analogue IQ modulators, then fed to two nonlinear power amplifiers. The output signals from the nonlinear power amplifiers are then summed using a conventional power combiner to form a replica of the input signal.

4.4.1 Experimental Results

In this system, a personal computer, which is connected to a four-channel arbitrary waveform generator, is used to generate from the high PAPR QPSK modulated OFDM signal the baseband components $I_1(t)$ and $Q_1(t)$, $I_2(t)$ and $Q_2(t)$ of the two constant envelope signals $S1(t)$ and $S2(t)$, respectively. The arbitrary waveform generator has four DACs followed by reconstruction filters. MATLAB computer programs are used for baseband generation and data manipulations.

The RF chain consists of a two high power amplifiers model ZHL-42W, two power supplies used to supply the DC power to the power amplifiers, two analogue IQ modulators for frequency upconversion, an LO signal generator, which used to supply 1.8 GHz LO signals to the analogue IQ modulators and finally an attenuator which is connected to the spectrum analyser. Characteristics of some of the hardware used in this prototype system are given in the previous chapter.

Each of the two analogue IQ modulators has been assembled from two mixers, a 90° power splitter and an inphase power combiner as shown in Figure 4-5.
Figure 4-5: Analogue IQ modulator

LO port is the local oscillator input to the analogue IQ modulator, I and Q ports are the inphase and quadrature input baseband signals, respectively. The RF port is the upconverted output signal from the analogue IQ modulator. For simplicity, the top layout of the analogue IQ modulator will be used in the rest of this work.

Figure 4-6 shows block diagram of the prototype system used to predict the effectiveness of the LINC technique for OFDM transmitter.
To begin, in order to determine the nonlinear distortion caused by the power amplifier in OFDM signal, MATLAB computer program, which includes the QPSK modulated OFDM baseband processing, is converted to a data file and loaded to four-channel arbitrary waveform generator.

The arbitrary waveform generator in turn generates two analogue baseband signals $I_1(t)$ and $Q_1(t)$ (upper branch of Figure 4-6). The two analogue baseband signals are then fed to the inphase and the quadrature ports of the analogue IQ modulator A. The analogue IQ modulator frequency shifts the baseband signals from DC to the required RF frequency and generates an RF signal with varying phase and amplitude.
Chapter 4. OFDM LINC Transmitter

The varying phase and amplitude RF signal is then amplified using high power amplifier as seen in the upper branch of Figure 4-6, and fed directly to the spectrum analyser.

The high power amplifier was operating deep in saturation at its maximum power output, so a 30 dB attenuator is connected to the input port of the spectrum analyser in order to compensate for its gain.

Once the effect of nonlinear distortion introduced by the high power amplifier in QPSK modulated OFDM signal is recorded. The main prototype system given in Figure 4-6 was then constructed, and a second MATLAB computer program, which includes both the QPSK modulated OFDM baseband processing and the signal components separator processing for LINC is converted to a data file and again loaded to four-channel arbitrary waveform generator.

The arbitrary waveform generator generates four analogue baseband signals $I_1(t)$, $Q_1(t)$, $I_2(t)$ and $Q_2(t)$ for both upper and lower branches as seen in Figure 4-6. The four analogue baseband signals are then fed to the inphase and the quadrature ports of the analogue IQ modulator A and the analogue IQ modulator B, respectively, generating two RF constant envelope signals, namely, $S_1(t)$ and $S_2(t)$.

The RF signals $S_1(t)$ and $S_2(t)$ having varying phase but constant envelopes are amplified using two high power amplifiers. The combined amplified RF output signal $S(t)$ with varying phase and envelope is then attenuated by 30 dB in order to compensate for the gain of the amplifiers, before it was measured using a spectrum analyser.

Since the input signal to each nonlinear high power amplifier has a constant envelope, the power amplifiers are operated deep into saturation in order to obtain maximum power efficiency.

4.4.1.1 Parameters for OFDM LINC Prototype System

The parameters for OFDM LINC prototype transmitter were defined to some extent with reference to the specifications of DVB-T 2K mode standard [65], but at the same time, in
accordance with the characteristics of the arbitrary waveform generator with a maximum sampling frequency of 40 MHz. Obviously, this is far too slow for many broadband applications. Table 4-1 shows the main parameters used for OFDM LINC prototype system.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>OFDM LINC Prototype Transmitter</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of carriers $K$</td>
<td>1705</td>
</tr>
<tr>
<td>Value of carrier number $K_{\text{min}}$</td>
<td>0</td>
</tr>
<tr>
<td>Value of carrier number $K_{\text{max}}$</td>
<td>1704</td>
</tr>
<tr>
<td>Symbol Duration $T_s$</td>
<td>930 µsec</td>
</tr>
<tr>
<td>Guard interval duration $\Delta$</td>
<td>0 µs</td>
</tr>
<tr>
<td>Carrier spacing $1/T_u$</td>
<td>1075 Hz</td>
</tr>
<tr>
<td>Spacing between carriers $K_{\text{min}}$ and $K_{\text{max}}$</td>
<td>1.83 MHz</td>
</tr>
<tr>
<td>Modulation</td>
<td>QPSK</td>
</tr>
<tr>
<td>RF carrier $f_c$</td>
<td>1.8 GHz</td>
</tr>
<tr>
<td>Data bit rate</td>
<td>3.67 Mb/s</td>
</tr>
</tbody>
</table>

Table 4-2: Main parameters for OFDM LINC prototype transmitter

As mentioned earlier in section 4.2.1, the guard interval can be chosen to 1/4, 1/8, 1/16 and 1/32 of the active symbol length. However, since the channel characteristics are not dealt
with in this work, neither coding nor guard interval were considered in the MATLAB programs carried out for this work.

You will also notice that the RF centre frequency in the simulated results in Figure 4-3(b), section 4.2.3, is much lower than the 1.8 GHz RF centre frequency used in the following experiments. This, however, has no effect on the experiments carried out during this work as only baseband processing is needed for the prototype systems and not RF processing. Low RF frequency is used in the simulations mainly due to the restricted simulation time and memory capability.

4.4.1.2 Measurement Results

Figure 4-7 shows spectrum analyser plots of OFDM signals centered on RF carrier frequency $f_c$, before and after linearisation.
The effect of nonlinear distortion introduced by the high power amplifier in QPSK modulated OFDM transmitter is shown in Figure 4-7(a), the spectral regrowth improvements as shown in Figure 4-7(b), which was achieved by applying the LINC technique can be thus measured with respect to the nonlinear distorted signal.

The improvements in spectral regrowth observed by applying the LINC technique to QPSK modulated OFDM transmitter when the high power amplifiers were driven deep into saturation is approximately 15 dB. However, according to the theoretical analysis of LINC technique in section 3.3. The LINC OFDM system should produce an amplified version of the simulated transmitted signal $S(t)$ as seen in Figure 4-3 (b), section 4.2.3. The reason for this is arise from the requirements in accurate balance between the upper and lower branch as seen in the prototype system. This is a major drawback of the LINC.
However, it's worthy to mention that although the practical results show that the improvement in spectral regrowth obtained in this experiment falls far behind the expected theoretical results. In comparison to the PAPR reduction reported techniques, which were briefly reviewed in section 4.3, the LINC scheme performs far better in minimising the effect of nonlinear distortion caused by the power amplifier in OFDM system. Moreover, the LINC technique is not restricted to any type of modulation scheme or any number of subcarriers and does not require any decoding at the receiver.

The imbalance between the two branches is partly due to the misalignments in the analogue IQ modulators [21,25]. In the linearised OFDM transmitter (Figure 4-6), two analogue I/Q modulators are used. Being analogue, these IQ modulators usually have imperfections that result in an imperfect match between the two RF channels. For example, gain mismatch might cause the I signal to be slightly smaller than the Q signal.

Digital IQ modulators are an alternative that have the advantage of minimising this imbalance since they tend to be more accurate due to improved matching between the digital processing of the I and Q channels, and the phase accuracy of the digital IQ modulators. Nevertheless, they are not often considered for broadband applications such as OFDM systems as they tend to suffer from limited bandwidth. In addition, they require extra RF frequency upconversion stages as well as image rejection filters.

Since this misalignment in the analogue IQ modulator constitutes a nuisance for LINC OFDM systems, it must be compensated. Therefore, in next chapter we aim to propose a scheme to reduce the misalignments to a minimum between the signal paths.
4.5 Conclusion

The review of formal aspects of OFDM signal generation and the development of expression for modeling the transmission process were presented and allowed an understanding of the characteristics of the transmitted signal. In addition, brief review of well-known PAPR techniques was included. Although there are an extensive bibliographical list of contributions to the study of OFDM characteristics and algorithms to perform PAPR reduction, the information available for OFDM LINC system is not so wide.

In this chapter, it was demonstrated that the proposed OFDM LINC transmitter is well suited for broadband applications, optimally employing the potential merits. The LINC technique adopts digital algorithm for generating two output signals with constant envelope from a QPSK modulated OFDM signal with high PAPR. This system is highly suited for OFDM signal amplifications since the inputs to the amplifiers are constant so the amplifiers can operate to have maximum efficiency for a given RF power. For experimental verification, a practical prototype OFDM LINC transmitter was constructed and tested to show the effectiveness of the LINC. Indeed, improvements in the spectral regrowth were almost 15 dB when the amplifiers were operating deep into saturation. This improvement could be even higher if a smaller output back-off is being adopted. Moreover, the OFDM LINC system is expected to be more effective if an accurate balance between the two branches could be achieved. However, in the conventional case, the highly accurate balance cannot be achieved due to the poor characteristics of the analogue RF components.

In comparison to coding, orthogonal pilot sequences, partial transmit sequence, selective mapping and tone injection methods, the LINC technique performs far better. In addition, it shows to be independent of the characteristics of the input data and it allowed the possibility to use a different modulation scheme for each subcarrier.

Although digital clipping and filtering method reported better reduction in spectral regrowth than the LINC technique, this improvement was proportional to the clipping factor, and the higher clipping factor causes more lost of information, which in turn increases bit error rate.
Adaptive predistorter has also been shown to be effective. However, the improvements achieved using this method seem to decrease sharply by increasing the number of subcarriers. In addition, the stability of the system is of main concern, especially for broadband applications.

In summary, the experimental results confirm that OFDM LINC transmitter provides good performance in the linearity in comparison to the abovementioned techniques and can be easily embodied in digital domain. Thus, this prototype transmitter is a hopeful solution for recent and future broadband applications. Regardless of this, there exist many other open issues concerning both OFDM transmission and LINC architecture, some of which we intend to address in the following chapter.
Chapter 5

IQ Imbalance Compensation

5.1 Introduction

The OFDM transmitter presented in the previous chapter (see Figure 4-2), the inphase and quadrature baseband signals are directly frequency shifted to RF signal using the analogue IQ modulator. This structure is known as direct upconversion architecture. In this architecture only one LO is required and there is no image frequency, and thus, no need for image frequency rejection filters. Therefore, it is well suited to monolithic integration.

One practical complication in using direct conversion architecture is the IQ imbalance introduced by the analog IQ modulators. Unfortunately, both the direct conversion OFDM system [20-24][87,88] and LINC structure [25,48] are very sensitive to IQ imbalance of the analog IQ modulators. This sensitivity leads to either heavy analogue processing specifications, and thus, an expensive system or large degradation in the system performance. In fact, even with careful design of analog circuits, practical quadrature signal mixing always introduces certain amount of amplitude and phase errors.

Unlike nonlinear distortions, reducing the input signal level cannot eliminate IQ imbalance. The relative dc offsets, i.e., carrier leakage, will get worse as the signal levels are reduced, and additional dc nulling circuits may be necessary [89,90]. In a general term, the gain and phase misalignments produce a counter-rotating signal of amplitude \( v \), is given by [90]

\[
v = \frac{\sqrt{\delta_k^2 + \Phi^2} \exp(j \tan^{-1} \frac{\Phi}{\delta_k})}{2}
\]  

(5.1)
Where $\Phi$ is the phase mismatch, and $\delta k$ is the gain mismatch factor between the inphase and quadrature signals. The spectrum of the counter-rotating signal represents the unwanted sideband in the transmitted spectrum signal.

In [48], the simulation results of a conventional LINC transmitter where two analogue IQ modulators were used with misalignment parameters of 0.5 dB gain and 5 degree phase error, show degradation of 10 dB in spectral regrowth. Hence, the ideal performance of LINC method relies on a robust IQ imbalance compensation scheme that can reduce the effect of the non-idealities of the analogue IQ modulators. In reality, however, the performance of the LINC can also be affected by the mismatch in the signal paths caused by other components in the transmitter.

The effect of IQ imbalance in direct conversion OFDM systems is well documented in the literature and different approaches exist in analyzing the gain and phase imbalance, for example, in [91], the theoretical analysis estimated that the gain and phase misalignments in OFDM system cause the symbol at the subcarrier $k_{\text{max}}$ to be multiplied by the complex factor $\gamma$, is given by [91]

$$\gamma = \frac{1 + (1 + \delta_k)(\cos(\Phi) - j\sin(\Phi))}{2} \quad (5.2)$$

In addition, a spurious component will be present, which is equal to the conjugate of the symbol at $k_{\text{min}}$ subcarrier multiplied by another complex term $\lambda$, is given by [91]

$$\lambda = \frac{1 - (1 - \delta_k)(\cos(\Phi) - j\sin(\Phi))}{2} \quad (5.3)$$

The symbol at the $k_{\text{max}}^{\text{th}}$ subcarrier, therefore, will include an interference related to the symbol at the $k_{\text{min}}^{\text{th}}$ subcarrier, and vice versa. Simulation results in [91] show 14 dB degradation in the performance when a gain mismatch of 0.5 dB and a phase mismatch of 5 degrees were introduced in a 16-QAM modulated OFDM signal.
Consequences of the analytical model of IQ imbalance in [87] for both single carrier (non-OFDM) and multicarrier OFDM based systems, show that the OFDM based system is more sensitive to IQ imbalance effect than the single carrier (non-OFDM) based system.

Although several compensation algorithms have been proposed to compensate for IQ imbalance in OFDM systems, two promising IQ compensation methods are the BSS (Blind Source Separation) [92], and a HD (Hard Decision) method [93], operating in the time and frequency domain, respectively.

The BSS method is based on adaptive algorithm, and thus can suffer from the delay caused by the adaptation time but doesn't require any information concerning the underlying system, and hence the name, the term "blind" stress the fact that no prior information about the underlying mathematical model is necessary for the estimation. Meanwhile, the HD is based on the training sequences being sent at the beginning of each packet and requires information on the underlying system i.e., the number of subcarriers and the modulation scheme. However, it waives any knowledge of the data being transmitted, regardless of its character. Although, the performance of HD method is much lower than that of BSS method [87], its advantage is the computational simplicity of the correction algorithm, and since it's operating in the frequency domain, it may make use of the already implemented FFT. Future analysis is needed to determine whether those techniques can perform well with the LINC structure as well as the OFDM structure.

Reported IQ imbalance techniques for the LINC architecture tend to reduce the effect of misalignments between the two branches. In [94] a phase correction method was proposed for a LINC transmitter employing OQPSK (Offset-QPSK). In this method, a multiplier detects the phase imbalance, and the phase of one branch is controlled, and by adding or subtracting a certain phase increment to the controlled branch, it compensates for the phase error. The technique shows a significant improvement in spectral regrowth. In addition, since the phase imbalance is detected by multiplying the two power amplifiers outputs. Hence it also reduces the phase error introduced by other analog components in the forward path of the branches.
Chapter 5. IQ Imbalance Compensation

The drawback of the above algorithm is that the accuracy is compromised if the imbalance is a result of the combination of both gain imbalance and phase imbalance. Therefore, a method was proposed in [95] to correct the gain imbalance as well as the consequent phase imbalance. In this method the CDMA (Code Division Multiple Access) output signal is attenuated and down converted and then fed to a LPF before a baseband controller is used to compensate for both the gain and phase errors. With this method, an improvement of 16 dB in spectral regrowth was achieved.

However, the drawback of those two methods [94,95] is the delay caused by the feedback configuration, and thus, stability analyses of the system needs to be performed in order to determine the effectiveness of both schemes, especially for wide band applications such as OFDM systems, and careful design is also required to prevent the additional imbalance introduced by the required extra analogue hardware in the down-conversion path used in the closed loop. Although, the two schemes proved to be effective in reducing the effect of the IQ imbalance between the two branches in the LINC structure, they do not address the effect of the IQ imbalance within each branch, and thus won’t be suitable for direct conversion OFDM based systems.

Alternatively, a compensation algorithm can be designed which operates at the baseband and able to compensate for both the IQ imbalance in one branch, that is, the OFDM system, and the imbalance between the two branches, that is, the LINC system. Therefore, this chapter introduces a low-complexity digital compensation scheme to combat the IQ imbalance caused by the analogue IQ modulator in the direct conversion OFDM transmitter. The compensation processing is done at the baseband so it does not require any additional analogue hardware. The proposed digital IQ imbalance compensation method is then used in conjunction with the LINC technique to maximise the performance of the transmitted OFDM signal. Finally, a prototype system, which cascade the LINC and IQ imbalance compensators is constructed to provide proof of concept of the method. The results are then compared to those achieved exclusively by the LINC technique, which were presented in the previous chapter.
Chapter 5. IQ Imbalance Compensation

5.2 OFDM Transmitter with IQ Imbalance Compensation

The proposed compensation technique of the IQ imbalance is based on the relation between the input and the output of the analogue IQ modulator information. Once this information is available, the digital IQ imbalance compensator estimates the difference between the ideal output baseband signals from the OFDM modulator and the measured data of the analogue IQ modulator and perform error correction. The IQ imbalance compensator can be implemented using either software or digital hardware. In this work, however, MATLAB program is used to develop the IQ imbalance compensator algorithm.

5.2.1 Analog IQ Modulator Measurements

To begin, measurements are required to fully characterize the analogue IQ modulator, i.e., the input and output relation of the device. The analog IQ modulator was assembled in a similar procedure described in section 4.4.1. Then a suitable measurement station has been set-up to carry out the measurements on the analogue IQ modulator, and is schematically depicted in Figure 5-1.

![Figure 5-1: Analogue IQ modulator measurement set-up](image_url)
Using HP-V EE (Hewlett Packard-Visual Engineering Environment), a graphical programming language from Agilent Technologies, optimised for building test and measurement applications, controlling a HP-8510C network analyser and two programmable power supplies, model HP-E3631A, triple output DC power supply. This measurements procedure can then be automated.

Figure 5-2 shows the HP-V EE program that was used for controlling the instruments, measuring the constellation of the analogue IQ modulator and storing the data into file to be used with MATLAB programs for the proposed IQ imbalance compensation technique.
Figure 5-2: HP-VEE program for measuring the analogue IQ modulator characteristics
As it can be seen from Figure 5-2, the program consists of a double for-next loop that steps through the full range of bias levels i.e. the in-phase and quadrature components of the analogue IQ modulator on power supply E3631A_2 for every voltage step taken by power supply E3631A_1. Once the desired voltage levels have been passed to the power supplies, the program requests the value of the S21 (Scattering Parameter) marker. This gives the value of the analogue IQ modulator’s complex transmission coefficient for the assigned inphase and quadrature control voltage levels. This data, along with the two voltage levels, are then appended to a text file to be used with MATLAB programs. Once the apparatus is set up, the program can run overnight. The operating frequency was set to a single point of 1.8 GHz.

Figure 5-3 shows the measured characteristics of the analogue IQ modulator on a polar plane.

![Figure 5-3: Measured constellation of the analogue IQ modulator A](image-url)
The polar plot is useful because the entire constellation of the analogue IQ modulator is covered. Instead of storing data directly into a data file, the constellation is also displayed in vector form. As it can be seen from Figure 5-3, the magnitude of the vector is the distance from the centre of the display, and the phase is displayed as the angle of vector referenced to the horizontal line from the centre to the right-most edge. The constellation set appears to shift to the right. This is due to the IQ imbalance of the analogue IQ modulator.

5.2.2 IQ Imbalance Compensation Algorithm

After the measured data have been collected, MATLAB program is then used to extract specific constellations for generating the OFDM baseband signals. This involves specifying a set of desired transmission states and then searching through the data to find which measured values are closest to them as shown in Figure 5-4.
As it can be seen from Figure 5-4, $I_r(t)$ and $Q_r(t)$ represent the baseband components of the measured complex transmission coefficient, $M_r$ represent the measured magnitude, and $I_i(t)$ and $Q_i(t)$ represent the ideal transmission coefficient of the analogue IQ modulator. The measured data is loaded to column arrays known as LUTs. For each of OFDM magnitude value $(I_i^2(t)+Q_i^2(t))^{1/2}$, the program will check every measured stored data to determine if it has a lower error function.

The closest measured point $M_o$ is defined to be the one for which the magnitude of the distance separating it from the ideal point is minimised. This gives the following error function:

$$EV(t) = ((I_i(t) - I_r(t)), (Q_i(t) - Q_r(t)))$$

$$= (E_I(t), E_Q(t))$$

Where $EV(t)$ represents the error vector, $E_I(t)$ and $E_Q(t)$ represent the inphase and quadrature components of the error vector $EV(t)$, respectively.

The above formula gives the expression for $EVM(t)$.

$$EVM(t) = \sqrt{E_I^2(t) + E_Q^2(t)}$$

Where $EVM(t)$ is the error vector magnitude. Similarly, the expression for $EVP(t)$ is found to be:

$$EVP(t) = \tan^{-1} \left( \frac{E_Q(t)}{E_I(t)} \right)$$

Where $EVP(t)$ is the error vector phase. Once the measured point that minimises this error becomes known, its associated voltage levels also become known. The program then
substitutes the new values with the ideal values as the new input baseband signals to the analogue IQ modulator.

5.2.3 Prototype System and Results

A suitable prototype system is constructed to determine the effectiveness of the IQ imbalance compensation scheme for the direct conversion OFDM transmitter as shown in the diagram of Figure 5-5.

![Prototype system of OFDM transmitter with digital IQ imbalance compensation](image)

The personal computer is used to develop the OFDM baseband signals and the IQ imbalance compensation algorithm using MATLAB software, and generates the data file. The data file is then loaded to an arbitrary waveform generator, which is connected to the computer via a GPIB. The arbitrary waveform generator generates the two new analogue baseband signals $I_r(t)$ and $Q_r(t)$.

The new baseband signals $I_r(t)$ and $Q_r(t)$ are fed directly to the in-phase and quadrature ports of the analogue IQ modulator, respectively. The 1.8 GHz frequency carrier signal is generated using an Agilent LO signal generator model E4422B. The resulting output RF signal $S(t)$ is injected into a 30 dB attenuator and displayed on the spectrum analyser.
As it was mentioned in the previous chapter, the parameters used for the baseband processing of OFDM transmitter in this work were with reference to the specifications of DVB-T 2K mode standard, which usually have a bandwidth of 7.61 MHz. However, due to the arbitrary waveform generator low sampling frequency, the parameters of the OFDM signal including the bandwidth used in the experiments carried out in this work had to be scaled. These arbitrary waveform generators were at the time among the fastest and cost effective available for general-purpose laboratory use.

5.2.3.1 Results

In order to determine the effect of the IQ imbalance on OFDM transmission signal, first, the simulation of a direct conversion OFDM transmitter using an ideal IQ modulator was carried out. The simulated result will then be compared to the measured results. Figure 5-6 shows the simulated results illustrating the output spectra of an ideal direct conversion QPSK modulated OFDM signal with symbol duration $T_s$ of 930 µs and 1705 subcarriers.

![Simulated spectra of an ideal direct conversion QPSK modulated OFDM signal](image)

Figure 5-6: Simulated spectra of an ideal direct conversion QPSK modulated OFDM signal
As it can be seen from Figure 5-6, the first side lobes of the spectrum are almost 45 dB below carrier. The 3 dB bandwidth is readily from the graph. The 18 MHz centre frequency carrier used in the simulation process, as mentioned earlier, is mainly due to the restricted simulation time and memory capability of the personal computer.

Figure 5-7 shows the measured spectra of the QPSK modulated OFDM signal without the IQ imbalance compensation technique and with IQ imbalance compensation technique.

![Figure 5-7: Measurement results illustrating the output spectra for direct conversion OFDM signals](image)

(a) OFDM spectra without IQ imbalance compensation  
(b) OFDM spectra with IQ imbalance compensation

As it can be seen from Figure 5-7(a), the first side lobes in the measured spectra of the direct conversion QPSK modulated OFDM signal without IQ imbalance compensation technique is almost 35 dB below carrier, in other words, there is a spectral regrowth of almost 10 dB in
comparison to the simulated results of Figure 5-6 when an ideal IQ modulator was used. This degradation is partly due to the IQ imbalance caused by the analogue IQ modulator.

However, as it can be seen from Figure 5-7(b), there is an improvement of almost 5 dB in spectral regrowth, which corresponds to a 40 dB below carrier when the QPSK modulated OFDM transmitter adopts the IQ imbalance compensation technique. This is 5 dB less than the predicted result of Figure 5-6. Part of this loss in the performance can be attributed to other components such as DACs and reconstruction filters embodied in the arbitrary waveform generator. Regardless of this loss, the digital IQ imbalance compensation technique proved to be effective in reducing the non-idealities effects of the analogue IQ modulator in direct conversion OFDM transmitter.

5.3 OFDM LINC Transmitter with Digital IQ Imbalance Compensation

Taking advantage of the effectiveness of the proposed digital IQ imbalance compensation technique and its simplicity. It was then decided to impose it on the direct conversion OFDM LINC transmitter. However, since the LINC architecture requires two analogue IQ modulators, in this case, a measurement of the characteristics of the second modulator named analogue IQ modulator B is then necessary.

By duplicating the measurement procedure presented in section 5.2.1 for the analogue IQ modulator A, the measured constellation of the analogue IQ modulator B is then presented in Figure 5-8.
Chapter 5. IQ Imbalance Compensation

Although the constellation set of the analogue IQ modulator B depicted in Figure 5-8 looks similar to that of the analogue IQ modulator A presented in Figure 5-3, a closer look will reveal that there is a slight difference in their characteristics. This implies that when those two analog IQ modulators are used in the LINC system, the LINC system will not only suffer from the IQ imbalance within the analogue IQ modulators but also from the imbalance between the two signal paths (upper and lower branch) since the characteristics of the two analogue IQ modulators do slightly differ from each other. This will cause incomplete cancellation of unwanted elements when the two constant envelope signals $S_1(t)$ and $S_2(t)$ are combined. As a result, a large number of unwanted spurious products will appear in the output spectrum, increasing spectral regrowth. In other words, recall from equations (3.12) and (3.13) presented in chapter 3, that is, if the two signal paths (upper and lower branch) in the LINC transmitter are perfectly matched, $E(t)$ will ideally cancel its anti-phase counterpart when the two constant envelope signals $S_1(t)$ and $S_2(t)$ are combined, leaving only a replica of the amplified signal $S(t)$. However, since the characteristics of the two analogue IQ
modulators are slightly different. This will cause gain and phase mismatch between the two branches, and thus, the signal $E(t)$ will not completely cancel its anti-phase counterpart.

Hence, by adopting the digital IQ imbalance compensation technique, we are not only aiming at reducing the effect of IQ imbalance within the two branches but also reducing the misalignments between the two branches.

A schematic diagram of the proposed direct conversion OFDM LINC transmitter with digital IQ imbalance compensation technique is depicted in Figure 5-9.
In the proposed direct conversion OFDM LINC transmitter with digital IQ imbalance compensation technique, the ideal inphase and quadrature QPSK modulated OFDM components $I_1(t)$ and $Q_1(t)$ are directly fed to the signal components separator, which in turn generates the ideal inphase and quadrature components $I_{1i}(t)$ and $Q_{1i}(t)$, $I_{2i}(t)$ and $Q_{2i}(t)$ of the two constant envelope signals $S_1(t)$ and $S_2(t)$, respectively. The baseband components $I_{1i}(t)$ and $Q_{1i}(t)$, $I_{2i}(t)$ and $Q_{2i}(t)$ are then applied to the two IQ imbalance compensation blocks named IQ imbalance compensation A and IQ imbalance compensation B, one for each branch of the LINC transmitter. Both IQ imbalance compensators follow the same procedure described in section 5.2.2 simultaneously.

Each digital IQ imbalance compensator generates two new baseband components $I_{1r}(t)$ and $Q_{1r}(t)$, $I_{2r}(t)$ and $Q_{2r}(t)$, respectively. The new components are applied to the two analogue IQ modulators, which in turn frequency shift the baseband signals from DC to the required RF frequency and each generates a new constant envelope RF signal named $S_1(t)$ and $S_2(t)$. The two signals $S_1(t)$ and $S_2(t)$ with constant envelopes but varying phase are amplified separately and recombined to form an amplified replica of the input signal $S(t)$.

Since the IQ imbalance will be reduced to a minimum in each branch. This in turn will reduce the imbalance between the two branches, and thus, improve the efficiency of the LINC architecture.

5.3.1 Prototype System and Results

Figure 5-10 shows block diagram of the prototype system used to predict the effectiveness of the digital IQ imbalance compensation technique for OFDM LINC transmitter.
Figure 5-10: OFDM LINC transmitter with digital IQ imbalance prototype system

For clarity, you will notice that there is a similarity between the prototype system of Figure 5-10 and the prototype system of Figure 4-6. Indeed, the instruments and the RF components are the same but the baseband processing is different since the baseband processing in the former adopts both the LINC processing and the IQ imbalance compensation algorithm while the baseband processing in the latter adopts only the LINC processing.

In the prototype system of Figure 5-10, MATLAB computer program that includes the QPSK modulated OFDM baseband processing, the signal components separator processing for LINC and the IQ imbalance compensation algorithm is converted to a data file and again loaded to four-channel arbitrary waveform generator.
The arbitrary waveform generator generates four analogue baseband signals $I_1(t)$, $Q_1(t)$, $I_2(t)$ and $Q_2(t)$ for both the upper and lower branches. The four analogue baseband signals are then fed to the inphase and quadrature ports of the analogue IQ modulator A and analogue IQ modulator B, respectively, generating two RF constant envelope signals, $S_1(t)$ and $S_2(t)$, that are amplified separately and recombined.

Finally, the combined RF output signal $S(t)$ with varying phase and envelope is then attenuated by 30 dB in order to compensate for the gain of the amplifiers, before it was displayed on a spectrum analyser. Two amplifiers are used to drive the HPAs into compression.

5.3.1.1 Results

Figure 5-11 shows the measurement results illustrating the output spectra for direct conversion OFDM LINC transmitter with digital IQ imbalance compensation technique.

![Figure 5-11: Measurement results illustrating the output spectra for OFDM signals](image)

(a) with LINC technique but without IQ imbalance compensation
(b) with LINC technique and IQ imbalance compensation
Chapter 5. IQ Imbalance Compensation

For clarity, Figure 5-11(a) is a duplicate of Figure 4-7(b), which represents the measured spectra of the QPSK modulated OFDM LINC transmitter but without the IQ imbalance compensation technique. That Figure was duplicated here in order to compare it with the measured spectra of the QPSK modulated OFDM LINC transmitter with digital IQ imbalance compensation scheme as depicted in Figure 5-11(b) and to determine the effectiveness of the IQ imbalance compensation scheme when it’s adopted by OFDM LINC transmitter. Hence, as it can be seen from Figure 5-11 that the digital IQ imbalance technique improves the performance of the OFDM LINC signal by almost another 8 dB, and thus, the digital IQ imbalance compensation technique is not only proved to be effective in reducing spectral regrowth in one branch, that is, direct conversion OFDM transmitter without a LINC structure but also between the two branches, that is, with the LINC structure.

Hence, the LINC architecture with digital IQ imbalance compensation technique proved to be an effective solution in reducing spectral regrowth in a direct conversion OFDM transmitter.

However, the requirements with respect to the spectrum level outside the channel bandwidth as determined in the standard DVB-T [65] through spectrum emission templates, should be at least 36 dB lower than the level of the spectrum at frequencies inside the nominal bandwidth. According to Figure 5-11(b), the requirement of an attenuation of 36 dB is not obtained. This can be attributed to the effect of the characteristics of the power combiner, cables, adaptors and the mismatch between the characteristics of the amplifiers themselves. This problem can be overcome by adopting an appropriate output back-off, which will help to reduce the level of the spectrum at frequencies outside the nominal bandwidth and achieve the required ACP (Adjacent Channel Protection).


Chapter 5. IQ Imbalance Compensation

5.4 Conclusion

The chapter addressed the problem of IQ imbalance in direct conversion OFDM transmitter. By using HP-VEE software and hardware instruments, an input-output relation of the analogue IQ modulator was derived. This input-output relation was then used to develop a compensation algorithm for the IQ imbalance in the digital domain. An important property of the proposed IQ imbalance compensation technique is that it reduces the effect of IQ imbalance in both the direct conversion OFDM transmitter as stand alone system and OFDM LINC system jointly at the baseband. In other words, the OFDM transmitter is not necessarily required to use the LINC scheme in order to achieve better IQ matching. This is an advantage for systems where the OFDM system is designed to use an alternative PAPR reduction method.

Although the method was proved to be effective in improving the spectral regrowth and it’s simple to implement, prior knowledge of the characteristics of the analogue IQ modulator is required. In addition, those phase and gain errors are random in nature, and thus, an adaptation configuration may need to be considered in the future.

Another disadvantage of the proposed IQ imbalance compensation technique is that since the analogue IQ modulator’s distribution of transmission states can be rather sparse in some areas as evidence from Figure 5-3 and 5-8, it will then be necessary to sweep the inphase and quadrature control voltages with a fine increment, especially when a high level modulation scheme is required for the OFDM transmission signal. In a practical scenario, this will lead to a very large measurement arrays, increasing the LUT memory resources.

However, regardless of those drawbacks, this chapter presented a working prototype system for a direct conversion OFDM LINC transmitter with digital IQ imbalance compensation scheme suitable for a wide range of broadband applications, with good improvement in spectral regrowth.
Chapter 6

Conclusion

6.1 Contributions of this Thesis

The number of applications where linear and efficient transmitter units are required increases every year. As a matter of fact, nowadays, transmitter linearisation is a field of great interest for the new high-speed wireless communications systems.

In this work two types of transmitters were investigated. Two-stage upconversion architecture and direct upconversion architecture. For a two-stage upconversion transmitter, the LINC technique was used to generate signals with constant envelope for both mixer and power amplifier that are employed in the RF stage of the single carrier QPSK (non-OFDM) transmitter, and thus eliminating the nonlinear distortion generated in the frequency-translating process in addition to that generated in the power amplifier itself. The technique was published under the title “Linear Frequency Translation and Amplification with Nonlinear Components”.

For a direct upconversion OFDM transmitter, the effect of the gain and phase imbalance caused by the analogue IQ modulator was investigated, and from the measured results, it was demonstrated that these effects not only reduce the performance of the direct conversion OFDM system but they severely reduce the efficiency of the LINC technique itself causing spectral regrowth. A digital IQ imbalance compensation method was then developed, which helped to correct for these mismatches, and thus extending the LINC ability to compensate for the nonlinear distortion caused by the power amplifier and the analogue imperfections caused by the analogue IQ modulators in the direct conversion multi-carrier QPSK modulated OFDM transmitter simultaneously. With this scheme, all the processing is
accomplished at baseband. Again, the technique was published under the title “OFDM LINC Transmitter with Digital I/Q Imbalance Compensation”.

The performance of the techniques was demonstrated through experimental measurements using prototype systems that have been constructed for both single carrier QPSK (non-OFDM) and multi-carrier QPSK modulated OFDM transmitters, and they consist of (i) a processing and control unit, namely a personal computer, (ii) an Agilent E4422B RF signal generator, (iii) a TTI-TGA1244 four channel arbitrary waveform generator with 12-bit vertical resolution and 256k point waveform memory, (iv) an Agilent E4407B spectrum analyzer, (v) a HP 54602B four channel 150Mhz oscilloscope and (vi) RF components.

MATLAB programs have been used for all the required baseband processing. This includes the generation of the LINC signals, the computation and compensation of the IQ imbalance and the generation of the final baseband signals for single carrier QPSK (non-OFDM) and multi-carrier QPSK modulated OFDM transmitters.

A suitable measurement station has been set-up to carry out the measurements on the analogue IQ modulators, and it consists of an Agilent 8510C microwave RF vector network analyser, a HP E3631A triple output power supply and a computer that used for programming and reading the instruments via a GPIB using HP-VEE programs to control them and perform the task of data collection.

This PhD represents a distinct contribution to the area of transmitter linearisation. It also adds to the current knowledge in high-speed wireless communications systems, transmitters architecture, PAPR reduction techniques, frequency translation techniques and finally measurement techniques.

Table 6-1 presents some of the major contributions made in this thesis.
In conclusion, the aims of the research as outlined in the first chapter have been achieved.

### 6.2 Future Work

A couple of open possibilities regarding the improvement and further evaluation of the proposed techniques can be listed at this point.

Due to the scarcity of spectrum and the new technical possibilities, attention has been drawn to the millimetre-wave band. It has become a hot topic as a research area for broadband communications. Some research programs have been looking at the possibility of using OFDM for broadband systems at carrier frequencies in the range of 40 GHz and up, which is nearly unused and allows for large bandwidth applications. Thus there is an open possibility to evaluate the proposed LINC with IQ imbalance technique for such frequency bands.

The proposed techniques for transmitters linearisation are based on open loop configuration, and thus any changes in the characteristics of RF components will in turn affect the performance of the systems. Hence adaptation may need to be investigated and considered for future work.
References


