Digital Front End for Base-station RF

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ABSTRACT OF THE THESIS

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Professor Moncef Tayahi

The digital front-end (DFE) is the most critical stage in a wireless base-station. The DFE along with the analog to digital converter (ADC) is responsible for bridging the analog RF and IF processing on one side and the digital baseband processing on the other side. The most important reason for replacing analog with digital signal processing is the ability to softly reconfigure the channels in the base station RF in real time, thus allowing for the implementation of various signal conditioning, compensation and mitigation channel non-linear responses.

This thesis presents a versatile Digital Front-End architecture, which has been simulated using MATLAB/Simulink. The architecture includes the design of robust Digital Up-Conversion (DUC) blocks in the transmit downlink and Digital Down-Conversion (DDC) blocks present in the receiver uplink paths in a wireless base station RF. Crest factor reduction (CFR) schemes help reduce the Peak to Average Power Ratio (PAPR) of the signal entering the base-station and have been implemented widely for code division multiple access (CDMA) and Long Term Evolution (LTE) systems, this is important because if the signal with the high PAPR is allowed to pass through the power amplifier (PA) it will result in the amplifier operating in its nonlinear region creating non-linear distortions in amplitude and phase, and the only other way to avoid this is to back off the signal to the linear region of the amplifier thus reducing its efficiency.

Finally, we implement Digital Predistortion (DPD) which is a method by which one first stimulates a non-linear power amplifier (PA) with baseband samples and then observes the result
of that stimulus at its output. Without this process we will need to use a power amplifier with a higher input power rating which needs to be backed off to operate in its linear region thus reducing the efficiency of the PA used and increasing its cost. As the PA is the heart of the base-station RF, we show that the main function of the DFE is to ensure a PA linearized output with a high efficiency.
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# TABLE OF CONTENTS

ABSTRACT .................................................................................................................................................. ii

ACKNOWLEDGEMENTS .............................................................................................................................. iv

1. INTRODUCTION ....................................................................................................................................... 1
   1.1. MOTIVATION ...................................................................................................................................... 1
   1.2. OUTLINE ........................................................................................................................................... 2

2. DIGITAL UP CONVERSION & DIGITAL DOWN CONVERSION (WCDMA & LTE) ...................... 3
   2.1. INTRODUCTION ................................................................................................................................. 3
   2.2. DIGITAL UP CONVERSION .............................................................................................................. 3
   2.3. DIGITAL DOWN CONVERSION ......................................................................................................... 16

3. CREST FACTOR REDUCTION ................................................................................................................. 20
   3.1. INTRODUCTION ................................................................................................................................. 20
   3.2. CFR USING WINDOWS AND FILTERS ........................................................................................... 22
   3.3. RESULTS .......................................................................................................................................... 27

4. DIGITAL PREDISTORTION ..................................................................................................................... 46
   4.1. INTRODUCTION ................................................................................................................................. 46
   4.2. PREDISTORTION TECHNIQUES WITH RESULTS ........................................................................ 48

5. CONCLUSIONS ........................................................................................................................................ 70

REFERENCES ............................................................................................................................................. 72
LIST OF FIGURES

Figure 1: Block diagram showing DUC for WCDMA signal ...................................................... 04
Figure 2: RRC filter using Chebyshev Window ........................................................................ 06
Figure 3: RRC filter using Rectangular Window .................................................................... 07
Figure 4: RRC filter using Hann Window ................................................................................ 07
Figure 5: RRC filter using Blackman Window ........................................................................ 07
Figure 6: RRC filter using Hamming Window ......................................................................... 08
Figure 7: WCDMA signal used ............................................................................................... 09
Figure 8: DUC for WCDMA signal using a single up-sampled-by-8 filter .................................. 10
Figure 9: DUC for WCDMA signal using an up-sample-by-4 filter followed by a half-band ... 11
Figure 10: DUC for WCDMA signal using a cascade of three half-band interpolators .......... 11
Figure 11: DUC filter chain for WCDMA signal .................................................................... 12
Figure 12: LTE signal used ..................................................................................................... 12
Figure 13: Block diagram showing DUC for LTE signal .......................................................... 13
Figure 14: DUC filter chain for LTE signal ............................................................................. 13
Figure 15: DUC for LTE signal using RRC filter followed by 2 half band interpolation filters... 14
Figure 16: NCO signal ............................................................................................................ 15
Figure 17: Block diagram showing DDC for WCDMA signal .................................................. 15
Figure 18: DDC for WCDMA signal using two half band interpolators & a channel filter ...... 17
Figure 19: DDC filter chain for WCDMA signal .................................................................... 17
Figure 20: Block diagram showing DDC for LTE signal .......................................................... 18
Figure 21: DDC filter chain for LTE signal ............................................................................. 18
Figure 22: DDC for LTE signal using one half band interpolator & a channel filter ............. 19
Figure 23: Difference between clipped and windowed signal ................................................ 22
Figure 24: Block diagram of the windowing method ................................................................. 24
Figure 25: WCDMA signal used ............................................................................................ 28
Figure 26: Depiction of Algorithm I ........................................................................................ 28
Figure 27: Cancellation pulse used for WCDMA signal ........................................................... 28
Figure 28: WCDMA signal before and after CFR for PAPR reduction to 8dB using cancellation pulse with 60 taps ....................................................................................... 29
Figure 29: WCDMA signal before and after CFR for PAPR reduction to 8dB using cancellation pulse with 200 taps .......................................................................................... 30
Figure 30: CCDF curves for WCDMA signal showing PAPR reduction to 8dB .................. 30
Figure 31: Time domain response of WCDMA signal before and after CFR ....................... 31
Figure 32: LTE signal used .................................................................................................... 31
Figure 33: LTE signal before and after CFR for PAPR reduction to 7dB using cancellation pulse with 200 taps ............................................................................................. 32
Figure 34: CCDF curves for LTE signal showing PAPR reduction to 7dB ........................... 32
Figure 35: Time domain response of LTE signal before and after CFR .............................. 33
Figure 36: LTE signal before and after CFR for PAPR reduction to 7dB using cancellation pulse obtained from Blackman window with length 200 ...................................................... 33
Figure 37: CCDF curves for LTE signal showing PAPR reduction to 7dB using Blackman Window ................................................................. 34
Figure 77: DPD of WCDMA signal using NARMA model for a PA using Ghorbani model

Figure 76: Characteristics of PA using Ghorbani model

Figure 75: DPD of LTE signal using NARMA model for a PA using Saleh model

Figure 74: DPD of WCDMA signal using NARMA model for a PA using Rapp’s model

Figure 73: DPD of LTE signal using NARMA model for a PA using Rapp’s model

Figure 72: DPD of WCDMA signal using NARMA model for a PA using Ghorbani model

Figure 71: Block diagram showing DPD using NARMA model

Figure 70: LTE signal output from the PA on using DPD with K=13 & Q=6

Figure 69: LTE signal output from the PA on using DPD with K=9 & Q=4

Figure 68: LTE signal output from the PA on using DPD with K=7 & Q=4

Figure 67: LTE signal output from the PA on using DPD with K=5 & Q=2

Figure 66: LTE signal output from the PA on using DPD with K=13 & Q=6

Figure 65: WCDMA signal output from the PA on using DPD with K=13 & Q=6

Figure 64: WCDMA signal output from the PA on using DPD with K=9 & Q=5

Figure 63: WCDMA signal output from the PA on using DPD with K=5 & Q=2

Figure 62: WCDMA signal at the input & output of the PA

Figure 61: Block diagram showing DPD of PA using the method of least squares

Figure 60: LTE signal before and after CFR for PAPR reduction to 6.6dB using cancellation pulse with number of taps= 650

Figure 59: LTE signal before and after CFR for PAPR reduction to 7.1dB using cancellation pulse with number of taps= 650

Figure 58: DPD for LTE signal using LUT technique when SSPA PA is used

Figure 57: DPD for LTE signal using LUT technique when TWT PA is used

Figure 56: Characteristics of Rapp model [31]

Figure 55: Characteristics of Saleh model [30]

Figure 54: Block diagram showing principle of predistortion using LUT

Figure 53: CCDF curves for 2 carrier LTE signal showing PAPR reduction to 7.1dB

Figure 52: PAPR reduction to 7.1dB for a 2 carrier LTE signal

Figure 51: Extended spectrum view

Figure 50: CCDF curves for LTE signal showing PAPR reduction to 7.1dB

Figure 49: LTE signal before and after CFR for PAPR reduction to 6.6dB using cancellation pulse with number of taps= 650

Figure 48: CCDF curves for LTE signal showing PAPR reduction to 6.6dB

Figure 47: LTE signal before and after CFR for PAPR reduction to 8dB using cancellation pulse

Figure 46: Time domain response of the cancellation pulse

Figure 45: LTE signal before and after CFR for PAPR reduction to 8dB

Figure 44: Time domain response of the cancellation pulse

Figure 43: LTE signal before and after CFR for PAPR reduction to 7.5dB using cancellation pulse

Figure 42: Time domain response of LTE signal before and after CFR

Figure 41: Cancellation pulse for LTE signal with 650 taps

Figure 40: Cancellation pulse for LTE signal with 550 taps

Figure 39: CCDF curves for LTE signal showing PAPR reduction to 7.5dB

Figure 38: LTE signal before and after CFR for PAPR reduction to 7.5dB using cancellation pulse with number of taps= 550

Figure 37: Cancellation pulse for LTE signal with 550 taps

Figure 36: Cancellation pulse for LTE signal with 650 taps

Figure 35: LTE signal output from the PA on using DPD with K=13 & Q=5
Figure 78: DPD of LTE signal using NARMA model for a PA using Ghorbani model ............... 69
## LIST OF TABLES

Table 1: WCDMA Design Parameters ........................................................................................................ 8
Table 2: LTE Design Parameters ............................................................................................................. 13
CHAPTER 1: INTRODUCTION

1.1 MOTIVATION

In the recent days the wireless industry has been trying to rapidly reduce Capital Expenditure (CAPEX) and Operating Expenditure (OPEX) [2]. It has been determined that up to 60 percent of the overall CAPEX cost is due to the radio elements within a typical base station RF. In addition to this, the base station RF also contains the power amplifiers which are responsible for a large percentage of the OPEX cost that is incurred.

The CAPEX can be reduced through the use of low cost non-linear power amplifiers, and highly integrated digital RF transceivers. This thesis demonstrates efficient design techniques for DUC (Digital Up Conversion), DDC (Digital Down Conversion), CFR (Crest Factor Reduction) and DPD (Digital Predistortion) processing. OPEX can be further reduced through the use of advanced, efficient algorithms, as OPEX is directly related to the power amplifier efficiency in the base station. Presently, only a small proportion of the DC power being consumed by the base station is converted into radiated energy. The transmitted signal generally determines the efficiency at which the power amplifier operates. WCDMA/LTE signals normally tend to have a high Peak-to-Average Power Ratio (PAPR) or Crest Factor. This tends to impose severe restrictions on the operations of the power amplifier. As a result of these peaks the power amplifier is generally backed off from its most efficient operating point. Thus to increase the power amplifier efficiency, CFR algorithms are used to decrease the PAPR of the transmitted signal before it enters the power amplifier. As a result of this, the power amplifier can now be operated with less back off and as a result an increased efficiency.

In addition to the method mentioned above Digital Pre-Distortion (DPD) can be used to increase the efficiency of the power amplifier. Over here instead of using digital signal processing to
reduce the dynamic range of the transmitted signal as is done in CFR, DPD algorithms tend to linearize the power amplifier itself.

1.2 THESIS OUTLINE
The thesis is structured as follows:
Chapter 2 discusses the background and need for Digital Up Conversion (DUC) and Digital Down Conversion (DDC) is shown and the design of digital-up-converter and digital-down-converter filters for WCDMA (Wideband Code Division Multiple Access) and LTE (Long Term Evolution) signals is given.

Chapter 3 gives some details on Crest Factor Reduction (CFR) along with some details on the methods that are used for CFR; following this discussion the simulated results of the proposed algorithms are presented.

Chapter 4 is dedicated to the discussion on Digital Pre-Distortion (DPD); the detailed background is given along with the techniques being used in the literature for achieving DPD compensation. This is followed up with the methods that have been simulated, some details on these methods are given and the simulation results are shown.

Chapter 5 discusses the summary of the results obtained in each of the chapters and gives some suggestions on future work that can be done.
CHAPTER 2: DIGITAL UP CONVERSION & DIGITAL DOWN CONVERSION  
(WCDMA & LTE)

2.1 INTRODUCTION:

The main purpose of Digital Up Converters (DUC) and Digital Down Converters (DDC) that are widely used in communication systems is for converting the sample rate of signals [3]. The process of digital up conversion is required for when a signal is translated from baseband to intermediate frequency (IF) band. On the other hand digital down conversion is used when a signal is converted from intermediate frequency band to baseband. The design structure of DUCs and DDCs typically includes frequency shifting using mixers and sampling rate converters. The actual structure of a DUC or DDC depends mainly on the conversion ratio that needs to be achieved.

2.2 DIGITAL UP CONVERSION

Introduction to Digital Up Conversion:

An important part of digital front end (DFE) used for RF communication systems is the Digital Up Converter (DUC) [4]. The main function of the DUC is to convert one or more channels of data from baseband format to a pass-band signal consisting of modulated carriers belonging to a set of one or more specified radio or intermediate (RF or IF) frequencies. This is achieved in two steps: (i) increasing of the sampling rate by interpolation, providing spectral shaping and rejection of interpolation images by means of filtering, and (ii) shifting the signal spectrum to the desired carrier frequencies using RF mixers and Voltage Controlled Oscillators (VCOs).
The block diagram shown below depicts an outline of the process being performed [5]:

![Block diagram showing DUC for WCDMA signal](image)

As seen in the process shown above there are 2 stages:

A. Pulse shaping and interpolation stage

B. Frequency translation of the single or multiple-carrier baseband WCDMA/LTE signal from 0 Hz to a set of specified center frequencies.

**A. Pulse Shaping Stage:** The pulse shaping stage comprises many pulse shaping filters.

The main purpose of these pulse-shaping filters is to keep the signal in its allotted bandwidth, maximizing its data transmission rate and minimizing transmission errors [6]. The pulse-shaping filter ideally has the following 2 properties:

i. A high stop band attenuation so as to reduce the inter channel interference to the maximum extent that is possible.

ii. Minimized inter symbol interferences (ISI) to obtain a bit error rate as low as possible.
One of the pulse shaping filters is the RRC (Root Raised Cosine) filter that is used to avoid intersymbol interference and to constrain the amount of bandwidth that is required for transmission. The RRC filter is the most suited to do the pulse shaping required as it has a transition band that is shaped like that of a cosine curve and its response meets the Nyquist Criteria. The first of the Nyquist criterion states that for achieving ISI-free transmission the impulse response of the shaping filter should have zero crossings at multiples of the symbol period. This criterion is met by the RRC filter.

A sinc pulse in the time domain does meet these requirements as it has a frequency response which is like a brick wall but this filter cannot be realized practically; however it is possible to approximate it by sampling the impulse response of the ideal continuous filter. For this to be realized the sampling rate has to be at least twice the symbol rate of the digital data to be transmitted. Thus it is necessary for the filter to interpolate the data by at least a factor of two and sometimes even more to simplify the complexity of the analog circuitry.

By definition, the rectangular pulse meets the first criterion one; it is zero at all points that exist outside of the present pulse interval. Hence it clearly cannot bring about interference during the sampling time of other pulses. But the problem with the rectangular pulse is that it has significant energy that is spread over a fairly large bandwidth; as a result of, the rectangular pulse is unsuitable for modern transmission systems. This is why pulse shaping filters are important and the rectangular pulse is clearly not the best choice for band-limited data transmission. We conclude that the pulse shape, which limits bandwidth, decays quickly, and provides zero crossings at the pulse sampling times, is the raised cosine pulse.
In this work the transmitted pulse shaper is thus an RRC filter with roll-off factor $\alpha = 0.22$ obtained from section 6.6.2.1 of 3GPP TS 25.104 [7], the spectral mask requirement for the $P \geq 43$ dBm case. When convolved with the matched RRC filter, the overall response of the raised cosine filter has no inter-chip interference (ICI) as the zero crossings occurs at chip intervals and it’s time domain response is infinite.

A simple technique involves using the window approach to design a filter within a reasonable length. It has been determined that a 45-tap symmetric RRC filter with Chebyshev windowing (sidelobe parameter set to 55 dB) meets all of the objectives required of the filter (all other windows have been tried and results are shown below). It is seen that the Chebyshev window provides better sidelobe suppression when compared to a rectangular window at the expense of some widening of the main-lobe. In addition the Chebyshev window is a kind of an adjustable window that requires a lower filter order to achieve the same performance as that which can be achieved by various windows, such as Hann, Blackman, and Hamming. The frequency responses of the RRC filter with an interpolation factor of two for different windows are shown below:

Chebyshev Window: sidelobe parameter set to 55, number of taps=45, the black line shows the spectral mask

![Figure 2: RRC filter using Chebyshev Window](image)
Rectangular Window: Number of taps = 105

Hann Window: Number of taps = 65

Blackman Window: Number of taps = 85
Hamming Window: Number of Taps - 61

![Figure 6: RRC filter using Hamming Window](image)

**DUC for WCDMA signal**

*Table 1: WCDMA Design Parameters: Below is a table summarizing the WCDMA parameters used in the simulation throughout the thesis*

<table>
<thead>
<tr>
<th></th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Carrier Bandwidth</td>
<td>5.0 MHz</td>
</tr>
<tr>
<td>2</td>
<td>Number of Carriers</td>
<td>1</td>
</tr>
<tr>
<td>3</td>
<td>Baseband Chip Rate</td>
<td>3.84 MCPS</td>
</tr>
<tr>
<td>4</td>
<td>IF Sample rate</td>
<td>61.44 MSPS (3.84x16)</td>
</tr>
<tr>
<td>5</td>
<td>Input Signal Quantization</td>
<td>16 bit I &amp; Q (Complex)</td>
</tr>
<tr>
<td>6</td>
<td>Output Signal Quantization</td>
<td>16 bit I &amp; Q (Complex)</td>
</tr>
</tbody>
</table>
WCDMA Signal Used:

![WCDMA Signal Used](image)

Figure 7: WCDMA signal used

The interpolation filter chain that is designed for the DUC needs up sample the baseband data by a factor of 61.44/3.84=16, as up sampling creates spectral images at multiples of the sampling rate, the main function of these interpolation filters is to remove these images and to keep the signal within its spectral mask. There are several techniques that can be used to perform the specific rate change. The first technique is to directly design an interpolator with an up sample factor of 16 having pulse shaping capability in one step. But this design is impractical as it is very difficult to design such a filter to meet the required spectral mask within a reasonable filter length. In addition such a design will result in an extremely high computational complexity. The second option is to split the rate conversion process into several interpolation stages.

Normally it is much more practical to design an RRC channel filter by using an up sampling factor of 2 and lower order to meet the system performance requirement. Even after the signal is passed through the channel filter, the signal will still have to be up-sampled by a factor of 8 and will need to have the aliasing effects removed in this process.
Once the signal is passed through the channel filter it is next passed through interpolation filters to remove the aliasing effects produced by up-sampling. Half band filters are generally preferred as interpolation filters. Half band filters are a type of FIR filter which have their transition region centered at one quarter of the sampling rate, Fs/4. The end of its pass band and the beginning of the stop band are equally spaced on either side of Fs/4. For the purpose of implementing an interpolation filter with an up-sample rate of two, the half band filter is preferred as its hardware requirement is less which in turn reduces the power consumed as well. The reduction in the hardware complexity is due to the fact that every odd indexed coefficient in the time domain is zero other than for the center tap and the even indexed coefficients are symmetric.

For DUC Filter Design the 4 following configurations are possible, each of these is preceded by an up-sample by 2 RRC filter:

1) A single up-sampled-by-8 filter:

Here the filter used is a constrained equi-ripple low pass filter, filter order is 112, fs=61.44 MHz, fpass=2 MHz & fstop=2.3 MHz, the black line represents the spectral mask.

![Figure 8: DUC for WCDMA signal using a single up-sampled-by-8 filter](image)
2) An up-sample-by-4 filter followed by a half-band filter:

First filter is a constrained equiripple low pass filter, filter order is 56, \(fs=30.72\) MHz, 
\(f_{\text{pass}}=2\) MHz & \(f_{\text{stop}}=2.3\) MHz, the half band filter is of order 17, \(fs=61.44\) MHz, \(fc=2\) MHz

![Figure 9: DUC for WCDMA signal using an up-sample-by-4 filter followed by a half-band filter](image)

3) A cascade of three half-band interpolators:

Here 3 half band interpolators are used, first half band filter is of order 31, \(fs=15.36\) MHz, 
\(fc=2\) MHz, second half band filter is of order 19, \(fs=30.72\) MHz, \(fc=2\) MHz, third half band filter is of order 15, \(fs=61.44\) MHz, \(fc=2\) MHz

![Figure 10: DUC for WCDMA signal using a cascade of three half-band interpolators](image)
Summary

From all the methods that have been shown above, the final method required the least computation and it is the method of choice in other papers as well and as can be seen in the final design the signal within the bandwidth is not suppressed at all as in the previous 2 cases. This filter architecture choice is shown below:

![DUC filter chain for WCDMA signal](image1)

DUC for LTE Signal

The LTE signal that is used is shown below:

![LTE signal used](image2)
Table 2: LTE design parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 Carrier Bandwidth</td>
<td>20.0 MHz</td>
</tr>
<tr>
<td>2 Number of Carriers</td>
<td>1</td>
</tr>
<tr>
<td>3 Baseband Chip Rate</td>
<td>30.72 MCPS</td>
</tr>
<tr>
<td>4 IF Sample rate</td>
<td>245.76 MSPS (30.72x8)</td>
</tr>
</tbody>
</table>

The block diagram shown below depicts an outline of the process being performed for an LTE signal:

![Block diagram showing DUC for LTE signal](figure13)

The optimum DUC Filter Design chain for an LTE signal is given below:

![DUC filter chain for LTE signal](figure14)
The filter design chain consists of an up-sample by 2 RRC filter followed by 2 half band interpolation filters:

The RRC filter is designed as a Chebyshev Window: sidelobe parameter set to 33, number of taps=45, fs=30.72 MHz, fc=7 MHz; first half band filter is of order 22, fs=61.44 MHz, fc=9 MHz; second half band filter is of order 10, fs=122.88 MHz, fc=9 MHz, the line in black shows the spectral mask.

**Figure 15:** DUC for LTE signal using RRC filter followed by 2 half band interpolation filters

**Summary**

Thus the figure above shows the LTE signal with the ideal filter chain design for a LTE signal.

**Frequency Translation**

After pulse shaping and up-converting the baseband signal, the signal spectrum is shifted from centered at 0 Hz to an intermediate frequency in the range of [-Fs/2, Fs/2], here the frequency was selected as 9.5 MHz.
This process requires a Direct Digital Synthesizer (DDS) and a mixer. The function of the DDC was emulated by designing a numerically controlled oscillator (NCO) in Matlab.

![Figure 16: NCO signal](image)

### 2.3 DIGITAL DOWN CONVERSION

The main purpose of DDC is to translate one or more intermediate IF channels from a set of specified center frequencies to 0 Hz. It also involves decimation and matched filtering to remove any adjacent channels and thus maximize the received signal-to-noise ratio (SNR).

Block Diagram showing the process is given below:

![Figure 17: Block diagram showing DDC for WCDMA signal](image)
The DDC mainly consists of 2 parts like in DUC, they are:

- Frequency translation of the channel back to 0MHz from 9.5MHz
- Decimation

From the 3GPP specification for WCMDA signals, the input to the DDC is a real signal sampled at 61.44 MSPS and the complex baseband outputs are produced at a sampling rate of $2 \times F_{\text{chip}} = 7.68$ MSPS where $F_{\text{chip}}$ is 3.84 MSPS [8]. Therefore the factor of down sampling is $61.44/7.68 = 8$.

Practically as in the case of interpolation, it is better to perform the process of decimation in a cascade of small rate stages rather than in one single step with a large rate change. While designing systems, a smaller alteration in the rate results in a wider transition band, which in turn leads to much fewer taps when designing the filter. Now when looking at the system from the perspective of hardware implementation, it is understood that the usage of multiple stages results in latter decimation stages operate at a lower sample rate, in a similar fashion as the DUC. Thus, the composite decimator requires hardware of lower complexity than a single stage, and the computational efficiency is significantly increased as well.

The basic DDC Filter Design consists of:

Two half band interpolators (Decimation Filters) & a channel filter (RRC filter like in the DUC case):

First half band filter is of order 17, $f_s=61.44\text{MHz}$, $f_c=2\text{MHz}$, second half band filter is of order 23, $f_s=30.72\text{MHz}$, $f_c=2\text{MHz}$, RRC filter with roll of factor-0.22, Chebyshev window with side lobe suppression, $f_c=1.9$, $f_s=7.68\text{Mhz}$, filter order- 114, side lobe attenuation 50
Summary:

Different filter designs similar to the DUC part were also tried but the design shown above yielded the best results using filters with the least number of taps. This preferred filter architecture choice is shown below:
Digital Down Conversion for LTE signals

The block diagram given below shows the outline of the process being performed:

![Figure 20: Block diagram showing DDC for LTE signal](image)

The ideal DDC Filter Design chain for an LTE signal is given below:

![Figure 21: DDC filter chain for LTE signal](image)

The basic DDC Filter Design consists of:

One half band interpolator (Decimation Filter) & a channel filter (RRC filter like in the DUC case):
The half band filter is of order 14, \( f_s=245.76\text{MHz} \), \( f_c=9\text{MHz} \), the RRC filter has roll off factor-0.22, Chebyshev window with side lobe suppression, \( f_c=9 \), \( f_s=122.88\text{ MHz} \), filter order- 52, side lobe attenuation 37.

![Power Spectral Density](image)

**Figure 22**: DDC for LTE signal using one half band interpolator & a channel filter

**Summary:**

Different filter designs similar to the DUC part were also tried but the design shown above yielded the best results using filters with the least number of taps.
CHAPTER 3: CREST FACTOR REDUCTION

3.1 INTRODUCTION

A popular choice in the modern communication systems is Multi-carrier transmission, such as discrete multi-tone (DMT) in digital subscriber line (DSL) and orthogonal frequency division multiplexing (OFDM) in wireless LAN, wireless MAN and 3GPP long term evolution (LTE) [9]. In these multi-carrier systems, information data are generated in the frequency domain, transformed to the time domain, and sometimes spread to the space domain if multiple antennas are used.

These multi-carrier signals usually tend to display large variations in amplitude in the time-domain, which can approximately be described by the Gaussian distribution [10]. But this is an undesirable phenomenon as it involves nonlinear distortion, lowering of the power efficiency, and an increase in the implementation cost, when the signal with the high PAR is applied to the transmitter front-end. The input power (signal with high PAR) drives the transmitter front-end (say a power amplifier) into the nonlinear region, as a result the peaks of the signal will be clipped, and this generates nonlinear distortions both in the signal bandwidth and in the adjacent channels. Hence to avoid these distortions, the general procedure is to back off the signal down to the linear region.

When the PAR of the signal is not reduced, the power efficiency obtained from the PA is quite low. One more advantage of CFR is to ensure the efficient utilization of the dynamic range of digital-to-analog converter (DAC) or analog-to-digital converter (ADC), as the high peaks rarely occurs. Thus by using a powerful PAR reduction technique, we can expect high power efficiency,
low nonlinear distortion, and small dynamic-range requirement of the analog components. Hence it is useful to investigate the CFR as a useful mechanism to boost the performance of multi-carrier systems.

Distortion based CFR techniques have quite a few advantages when used for the system implementation, like simple structure, low computational complexity, and no transmission of side information. But it does have the disadvantage that these nonlinear CFR operations may greatly degrade the link quality [11]. Due to this additional operations are required to control the in-band and out-of-band distortions. As far as the in-band part is concerned, EVM analysis can be used to relate the performance degradation to various distortion mechanisms and the EVM is to be controlled to satisfy the 3GPP standard requirements. For the out-of-band part the spectral mask is a deterministic metric to constrain the out-of-band spectrum regrowth.

Some of the techniques that are used to achieve CFR Reduction are: Amplitude clipping and filtering [12, 13], Tone reservation [14], Active Constellation extension [15], Use of windows and filters for controlled clipping. Out of the techniques mentioned the amplitude clipping and filtering technique is the simplest and computationally the easiest method to implement but the distortion caused by amplitude clipping can be viewed as a source of clipping noise and it causes in-band and out-of-band radiation. The methods of tone reservation and active constellation extension are much more efficient and results in very little clipping noise but they do suffer from the disadvantage of being computationally complex. The method of using windows and filters for controlled clipping is the most simple and efficient as compared to the others as it doesn’t involve too much generation of spectral noise and as it is computationally simple. This is the method that is described in this thesis.
3.2 CFR using Windows and Filters

Conventional clipping, which is shown mathematically in the equation below results in sharp corners in a clipped signal, which in turn causes unwanted out-of-band radiation [16] - [21].

\[
B(x) = \begin{cases} 
  x, & |x| \leq A \\
  Ae^{j\Phi(x)}, & |x| > A 
\end{cases}
\]

Here A is the maximum value allowed for the clipped signal. Thus, to smooth these corners and to compensate for the out-of-band power problem, the clipping that is implemented is realized by multiplying the original signal with a suitable function. This procedure is called windowing. The difference between the conventional clipping and windowing is presented in figure below.

Figure 23: Difference between clipped and windowed signal
The main advantage of this method of windowing compared to the other methods combining conventional clipping and filtering is the absence of the peak regrowth, which is always present when filtering is involved. In addition, this windowing technique can be applied to a single carrier as well as to a multicarrier signal.

It is to be noted that the multiplication in the time domain corresponds to convolution in the frequency domain; hence it is observed that the spectral widening of the signal to be clipped can in turn be controlled by adjusting the spectral properties of the multiplying signal. Theoretically, the multiplying signal can be any arbitrary signal that gives the desired result, but, in practice, the signal must be generated in a systematic way. In practice, the windowing pulse or a pulse from a filter is selected such that its frequency response is within the bandwidth of the signal that needs to be clipped.

**Windowing algorithm**

The windowing algorithm can be expressed in terms of conventional clipping as shown below whereas conventional clipping can be expressed as a multiplication:

\[ x'(n) = c(n)x(n); \]  

(3.1)

where

\[ c(n) = \begin{cases} 1, & \text{if } |x| \leq A \\ A/|x(n)|, & \text{if } |x| > A \end{cases} \]  

(3.2)
Where $A$ is the maximum amplitude allowed for the clipped signal. The concept involved in the windowing method is to replace the function $c(n)$ with the function:

$$b(n) = 1 - \sum_{k=-\infty}^{\infty} a_k w(n - k) ,$$

(3.3)

Where $w(n)$ is the window or filter function and $a_k$ is a weighting coefficient. The block diagram of the windowing method is presented in the figure below:

![Block diagram of the windowing method](image)

Figure 24: Block diagram of the windowing method

In order to achieve the desired clipping level, the function $b(n)$ needs to satisfy the inequality given by the equation shown below:

$$1 - \sum_{k=-\infty}^{\infty} a_k w(n - k) \leq c(n),$$

(3.4)

for all values of $n$. 
In order to reduce the error in the time domain, i.e. to minimize the difference between \( c(n) \) and \( b(n) \), inequality 3.4 must be as near equality as possible. The difference that exists between \( c(n) \) and \( b(n) \) is dependent on the shape of the window or filter, the window length or filter \( Nw \) defined as a number of samples \( w(n) \) that are not equal to zero, the weighting coefficients \( a_k \) and finally the clipping threshold.

The spectral properties of the resulting clipped signal depend not only on the shape but also the length of the window or filter used. As it is necessary to keep the algorithm simple enough to be implemented, the shape and the length of the window or filter used and the clipping threshold should be kept constant. Once the shape and the length of the window or filter is chosen, the weighting coefficients \( a_k \) are to be determined. Instead of calculating \( a_k \) and then determining \( b(n) \), \( b(n) \) can be determined from \( c(n) \) directly shown below, also it is assumed that clipping probability and window length are so low that windows do not overlap in the time domain, hence to find \( b(n) \) first the part shown below is determined:

\[
\sum_{k=-\infty}^{\infty} a_k w(n - k)
\]

(3.5)

Next by convolving the function \( 1-c(n) \) with the window \( w(n) \), \( b(n) \) now becomes:

\[
b(n) = 1 - \sum_{k=-\infty}^{\infty} [1 - c(k)]w(n - k)
\]

(3.6)
Window/Filter selection

As stated earlier, multiplication in the time domain leads to convolution in the frequency domain. In order to limit the widening of the signal spectrum, the window/filter function that is selected should have a narrow spectrum. However, due to the strict localization in the frequency domain there is a wide response which results in the time domain, this means that the effect of the clipping spreads to the samples adjacent to the sample to be clipped, thus in turn increasing the EVM. Hence it can be concluded that the selection of the window is a tradeoff between the spectral properties of the signal and the EVM. The EVM is calculated as shown below:

$$EVM_{RMS} = \frac{\frac{1}{N} \sum_{k=1}^{N} (e_k)}{\sqrt{\frac{1}{N} \sum_{k=1}^{N} (I_k^2 + Q_k^2)}} \times 100$$

Where $e_k = (I_k - \hat{I}_k)^2 + (Q_k - \hat{Q}_k)^2$

$I_k$= In-phase ideal reference value of the $k^{th}$ symbol in the burst

$Q_k$= Quadrature phase ideal reference value of the $k^{th}$ symbol in the burst

$N$ = Input vector length

$I_k$ and $Q_k$ represent ideal (reference) values $\hat{I}_k$ and $\hat{Q}_k$ represent measured (received) symbols

The optimal choice mainly depends on the system to which the windowing is applied. As an example, we consider the use of some well-known windows [33] in reducing the CFR of LTE/WCDMA signal. The windows and filters that are used are: Hamming, Hanning, Blackman and triangular windows and the filter used is the constrained equiripple low pass filter (filter parameters are such that each pulse is the impulse response of a filter that is designed to match
the spectral content of the signal with $D_{\text{pass}} = -20\,\text{dB}$, $D_{\text{stop}} = -50\,\text{dB}$). It was found that the best results are obtained by using the constrained equiripple filter, Blackman and triangular windows.

### 3.3 RESULTS

An efficient algorithm to implement the peak windowing method is presented. Here, the key objective is to keep the computational complexity of the algorithm minimal for high-speed implementations and avoid iterative processes which would cause a delay. This most probably does not lead to an optimal performance in terms of PAPR and EVM.

The controlled clipping process is performed by using 3 algorithms given below:

**Algorithm I:**

- Set the clipping threshold required as needed.

- Then determine the position of the peaks above the threshold level and note their values above the threshold.

- Next, obtain the cancellation pulse parameters using a window or constrained equiripple low pass filter.

- The cancellation pulse is then scaled with respect to each peak corresponding to its extent above the threshold level and it is then subtracted from the signal peak, thus reducing the signal peak to the threshold value.

- The signal is scanned for the highest peak in the signal which is first cancelled, then the search is conducted for the second peak and so on till the desired PAPR values is reached.

The WCDMA signal used is shown below:
Some of its specifications are:
Bandwidth=5MHz; Sampling frequency=30.72MHz, Number of samples used=307200; Chip rate=3.84MSPS.

The figures shown below depict the algorithm shown above:

The black line shows the position of the peak, the curve in blue shows the signal before PAPR reduction, the curve in red shows the PAPR reduced signal.

The figure below gives the time domain response of the cancellation peak:
This cancellation pulse is the time domain response of constrained equiripple low pass filter (Dpass=-20dB, Dstop =-50 dB).

**Results of PAPR reduction to 8dB for a WCDMA signal:**

The figure below shows the signal before and after CFR when the cancellation pulse is a time domain response of a constrained equiripple low pass filter with parameters: Dpass=-20dB, Dstop =-50 dB, fc=1.92MHz, number of taps=60:

![Figure 28: WCDMA signal before and after CFR for PAPR reduction to 8dB using cancellation pulse with 60 taps](image)

The line in blue shows the original signal before CFR while the line in green shows the signal after CFR. The %EVM value obtained here is 2.53%. The %EVM value according to 3GPP specifications is 12.5% so as seen the EVM value is much below the threshold set by 3GPP specifications. The line in black shows the spectral mask. The figure below shows the signal curves before and after CFR when the constrained equiripple low pass filter has number of taps = 200:
The line in blue shows the original signal before CFR while the line in green shows the signal after CFR. The %EVM value obtained here is 3.2%.

The figure below gives the CCDF (Complimentary Cumulative Density Function) curves before and after CFR:

---

Figure 29: WCDMA signal before and after CFR for PAPR reduction to 8dB using cancellation pulse with 200 taps

Figure 30: CCDF curves for WCDMA signal showing PAPR reduction to 8dB
The figure below shows the time domain response of the signal before and after CFR:

![Time domain response of WCDMA signal before and after CFR](image)

**Figure 31: Time domain response of WCDMA signal before and after CFR**

The signal in blue is before CFR whereas the signal in red is after CFR, as it can be seen the signal after CFR is below the threshold value which is set to achieve PAPR value of 8dB. 800 peaks above the set threshold were cancelled to get the above results.

**Results of PAPR reduction to 7dB for a LTE signal:**

The LTE signal used is given below:

![Power Spectral Density](image)

**Figure 32: LTE signal used**
Some of its specifications are:

Bandwidth=20MHz; Sampling frequency=245.76MHz, Number of samples used=2.4576 MSPS;
Chip rate=3.84MSPS

The figure below shows the signal before and after CFR when the cancellation pulse is a time
domain response of a constrained equiripple low pass filter with parameters: Dpass=-20dB, Dstop
=-50 dB, fc=9MHz, number of taps=200:

![Power Spectral Density](image)

**Figure 33:** LTE signal before and after CFR for PAPR reduction to 7dB using cancellation pulse with 200 taps

The signal in black gives the original while the signal in red gives the LTE signal after CFR. 3500
peaks above the threshold value were cancelled. The %EVM=3.13%. The %EVM value
according to 3GPP specifications is 12.5% for a 16QAM signal so as seen the EVM value is
much below the threshold set by 3GPP specifications. The CCDF curves are shown below:

![CCDF curves](image)

**Figure 34:** CCDF curves for LTE signal showing PAPR reduction to 7dB
As seen from the above figure the PAPR is reduced by 3.8dB.

The figure below gives the time domain response of the signal before and after CFR:

![Time domain response of LTE signal before and after CFR](image)

**Figure 35: Time domain response of LTE signal before and after CFR**

The figure below shows the signal before and after CFR when the cancellation pulse is a Blackman window of length 200, the other windows did not yield as good results as the Blackman window:

![Power spectral density](image)

**Figure 36: LTE signal before and after CFR for PAPR reduction to 7dB using cancellation pulse obtained from Blackman window with length 200**
The signal in black gives the original while the signal in red gives the LTE signal after CFR. 3600 peaks above the threshold value were cancelled. The %EVM obtained here is 3.21%. The CCDF curves are shown below:

![CCDF Curve](image)

*Figure 37: CCDF curves for LTE signal showing PAPR reduction to 7dB using Blackman Window*

As seen from the above figure the PAPR is reduced by 3.7dB.

**Algorithm II**

- The clipping threshold level is set based on the PAPR target to be achieved.
- A search is conducted through the signal to determine the number of peaks.
- As a peak is encountered it is immediately reduced to the threshold level and the search continues through the signal to determine the next peak.
- The above 3 steps are repeated till all the peaks above the threshold are cancelled.
Results of PAPR reduction of around 7.5dB for a LTE signal:

The figure below shows the signal before and after CFR when the cancellation pulse is a time domain response of a constrained equiripple low pass filter with parameters: Dpass=0dB, Dstop = -80 dB, fc=9MHz, number of taps=550:

![Power Spectral Density graph](image)

The clipping threshold is set at 2.5 times the mean value. Here the numbers of peaks are cancelled in 2 stages. The number of peaks cancelled in first iteration= 10714 peaks, the number of peaks cancelled in the second iteration = 5191 peaks, the %evm value = 5.52%.

The CCDF curves are shown below:

![CCDF graph](image)
The curve in red is obtained after CFR while the blue curve represents the CCDF of the original signal.

The figure below gives the time domain response of the signal before and after CFR:

![Image of Time Domain Response]

**Figure 40: Time domain response of LTE signal before and after CFR**

The frequency and time domain response of the cancellation pulse which is a constrained equiripple low pass filter with parameters: $D_{pass}=0$ dB, $D_{stop}=-80$ dB, $f_c=9$ MHz, number of taps=550 is shown below:

![Image of Frequency Response]

**Figure 41: Cancellation pulse for LTE signal with 550 taps**

The frequency response of the cancellation pulse is shown above.
The time domain response of the cancellation pulse is shown above.

Algorithm III

- The clipping threshold level is set based on the PAPR target

- A search is conducted through the signal to determine the number of peaks, their locations and the magnitudes above the threshold level

- The cancellation pulses are placed at each of the peaks, then all the pulses are subtracted from the peaks
Results of PAPR reduction of 8dB for a LTE signal:

The figure below shows the signal before and after CFR when the cancellation pulse is a time domain response of a constrained equiripple low pass filter with parameters: $D_{pass}=0\,\text{dB}$, $D_{stop}=-80\,\text{dB}$, $f_c=9\,\text{MHz}$, number of taps=650.

The clipping threshold is set at 2.5 times the mean value. The $\%\text{evm}$ value = 2.75\%, the number of stages = 4, the details for each stage are:

Stage 1: Clipping Threshold = 2.7, number of peaks cancelled=5135

Stage 2: Clipping Threshold = 2.7, number of peaks cancelled=1384

Stage 3: Clipping Threshold = 2.7, number of peaks cancelled=190

Stage 4: Clipping Threshold = 2.7, number of peaks cancelled=14
The CCDF curve is shown below:

![CCDF Curve](image)

**Figure 44: CCDF curves for LTE signal showing PAPR reduction to 8dB**

The curve in red is obtained after CFR while the blue curve represents the CCDF curve of the original signal.

The frequency and time domain response of the cancellation pulse which is a constrained equiripple low pass filter with parameters: $D_{\text{pass}}=0\,\text{dB}$, $D_{\text{stop}}=-80\,\text{dB}$, $f_c=9\,\text{MHz}$, number of taps=650 is shown below:

![Cancellation Pulse](image)

**Figure 45: Cancellation pulse for LTE signal with 650 taps**
The frequency response of the cancellation pulse is shown above.

![Figure 46: Time domain response of the cancellation pulse](image)

The time domain response of the cancellation pulse is shown above.

**Results of PAPR reduction of 6.6dB for a LTE signal:**

The figure below shows the signal before and after CFR when the cancellation pulse is a time domain response of a constrained equiripple low pass filter with parameters: Dpass=0dB, Dstop = -80 dB, fc=9MHz, number of taps=650:

![Figure 47: LTE signal before and after CFR for PAPR reduction to 6.6dB using cancellation pulse with number of taps=650](image)
The clipping threshold is set at 2.3 times the mean value. The evm value = 8.97%, the number of stages = 5, the details for each stage are:

Stage 1: Clipping Threshold = 2.3, number of peaks cancelled=38925

Stage 2: Clipping Threshold = 2.3, number of peaks cancelled=11973

Stage 3: Clipping Threshold = 2.3, number of peaks cancelled=5983

Stage 4: Clipping Threshold = 2.3, number of peaks cancelled=2201

Stage 5: Clipping Threshold = 2.3, number of peaks cancelled=423

The CCDF curves are shown below:

The curve in red is obtained after CFR while the blue curve represents the CCDF curve of the original signal.
Results of PAPR reduction of 7.1dB for a LTE signal:

The figure below shows the signal before and after CFR when the cancellation pulse is a time domain response of a constrained equiripple low pass filter with parameters: $D_{pass}=0\text{dB}$, $D_{stop}=-80\text{ dB}$, $f_c=9\text{MHz}$, number of taps=650:

![Power Spectral Density](image)

The clipping threshold is set at 2.5 times the mean value. The \textbf{evm value} = \textbf{5.43\%}, the number of stages = 4, number of taps = 650, the details for each stage are:

Stage 1: Clipping Threshold = 2.5, number of peaks cancelled=18255

Stage 2: Clipping Threshold = 2.5, number of peaks cancelled=6660

Stage 3: Clipping Threshold = 2.5, number of peaks cancelled=2487

Stage 4: Clipping Threshold = 2.5, number of peaks cancelled=597
The CCDF curves are shown below:

![CCDF curves](image1)

**Figure 50: CCDF curves for LTE signal showing PAPR reduction to 7.1dB**

The curve in red is obtained after CFR while the blue curve represents the CCDF curve of the original signal.

An extended view of the spectrum is shown below:

![Extended spectrum view](image2)

**Figure 51: Extended spectrum view**
Results of PAPR reduction of 7.1dB for a LTE two carrier signal:

The figure below shows the signal before and after CFR when the cancellation pulse is a time domain response of a constrained equiripple low pass filter with parameters: Dpass=0dB, Dstop =-80 dB, fc=9MHz, number of taps=650:

![Power Spectral Density](image)

Figure 52: PAPR reduction to 7.1dB for a 2 carrier LTE signal

The clipping threshold is set at 2.5 times the mean value. The evm value = **8.34%**, the number of stages = 8, the details for each stage are:

- **Stage 1:** Clipping Threshold = 2.5, number of peaks cancelled=18242
- **Stage 2:** Clipping Threshold = 2.5, number of peaks cancelled=9259
- **Stage 3:** Clipping Threshold = 2.5, number of peaks cancelled=5826
- **Stage 4:** Clipping Threshold = 2.5, number of peaks cancelled=3922
- **Stage 5:** Clipping Threshold = 2.5, number of peaks cancelled=2617
- **Stage 6:** Clipping Threshold = 2.5, number of peaks cancelled=1591
- **Stage 7:** Clipping Threshold = 2.5, number of peaks cancelled=654
- **Stage 8:** Clipping Threshold = 2.5, number of peaks cancelled=293
The CCDF curves are given below:

![CCDF curves](image)

Figure 53: CCDF curves for 2 carrier LTE signal showing PAPR reduction to 7.1dB

The curve in red is obtained after CFR while the blue curve represents the CCDF curve of the original signal.

**Summary**

From the 3 algorithms that have been shown above, Algorithm I produces the best results with low %EVM values but it is not a computationally efficient algorithm as in each iteration the entire signal is searched for the highest peak above the threshold which is then cancelled, Algorithm II is more efficient this way as a lesser number of iterations are involved and all the peaks above the threshold are cancelled as the search through the signal is conducted during each iteration but it does require a few iterations as a new peak may arise when an existing peak is cancelled, Algorithm III is computationally the most efficient algorithm and it requires the least number of iterations and it does produce good results though it does have the disadvantage of yielding a high %EVM value, but as this EVM value lies within the 3GPP specifications this method can be selected for FPGA implementation.
CHAPTER 4: DIGITAL PREDISTORTION

4.1 INTRODUCTION

Power amplifiers are nonlinear components that are present in communication systems [23]. This nonlinearity causes spectral regrowth that is present beyond the spectral bandwidth which in-turn interferes with adjacent channels. In addition to this it also brings about distortions in the signal bandwidth, these causes a decrease in the bit error rate at the receiver end. The present transmission formats such as WCDMA & LTE are especially vulnerable to these distortions as they have a high PAPR value i.e., presence of large fluctuations in their signal envelopes. Thus as mentioned earlier backing-off the input signal to operate the power amplifier in the linear region is ill advised as the power amplifier efficiency will be very low for these high PAPR signals, it is typically less than 10% [24]; i.e., more than 90% of the dc power is lost and turns into heat.

Improving the power amplifier efficiency has several advantages like: it significantly reduces the electricity and cooling costs that fall upon the service providers, and considering that in the present day there are a huge number of base stations deployed worldwide this offers huge power savings.

One of the best and most cost effective techniques is the use of Digital Predistortion, in this method a digital predistorter is added in the baseband to create an expanding nonlinearity which in turn is complementary to the compressing characteristic of the power amplifier. In an ideal scenario, the cascade of the predistorter and the power amplifier results in a linear signal chain and the input is amplified by a constant gain. Thus with the presence of the predistorter, the power amplifier can be utilized up to its saturation point while still operating in the linear region, thus significantly increasing the PA’s efficiency. Actually speaking, the power amplifier characteristics may change over a period of time because of temperature drift, component aging,
etc. Therefore, the predistorter should be designed in such a way that it has the ability to adapt to these changes.

Most of the designs that exist for the PA in the present day treat it as a memory-less device; i.e., its current output depends only on the current input through a nonlinear mechanism. This instantaneous nonlinearity can be characterized by the AM/AM and AM/PM responses that are obtained of the power amplifier, in which the output signal amplitude and phase deviation of the power amplifier output are to be considered as functions of the amplitude of its current input. Over the past decade there has been an intensive research that has been conducted on the DPD techniques [25].

It has been observed that as the signal bandwidth gets wider, like as in WCDMA/LTE signals, memory effects are exhibited by power amplifiers. This effect can be widely seen in the high power amplifiers that are used in wireless base stations. The memory effects are created due to several factors such as: the thermal constants of active devices or due to the components present in the biasing network which exhibit frequency dependent behavior [26]. Due to this effect the current output of the power amplifier depends on not only the current input, but also past inputs as well. Thus it can be said that the power amplifier is a nonlinear system with memory effects. This memory effect of the PA should be taken into account as only a limited linearization performance can be achieved when memory-less pre-distorters are used [27], [28]. Thus it is necessary to design pre-distorters with memory structures as well. This dissertation investigates robust predistorter models that are capable of linearizing power amplifiers with memory effects. It also investigates system implementation issues related to these wideband digital predistortion systems. In this dissertation, we propose novel pre-distorters and their implementation through simulations for Power Amplifiers with and without memory effects.
4.2 PREDISTORTION TECHNIQUES:

The modeling techniques and predistorter design for memory-less power amplifiers, as well as power amplifiers with memory effects are presented below.

Method 1: Adaptive Linearization of Power Amplifiers using Look up Table Technique (LUT)

This predistortion technique uses memory look-up tables for the purpose of pre-distorting the baseband drive signal [29], also it is a computationally simple technique used for linearizing memory-less high power PA’s (HPA’s).

Let us start by giving a description on the memory-less model used for describing the nonlinear HPA being used for the simulations. The complex envelope of the input signal to the HPA can be described as:

\[ x(t) = \rho(t) \exp[j\phi(t)] \]  \hspace{1cm} (4.1)

Then the complex envelope of the output signal is given by:

\[ z(t) = A[\rho(t)] \exp[j(\phi(t) + \Phi[\rho(t)])] \]  \hspace{1cm} (4.2)

Where \( A(\rho) \) and \( Q(\rho) \) represent respectively the AM/AM and AM/PM conversion of the nonlinear amplifier. For the purpose of describing this technique we consider two different kinds of nonlinear HPA models:

- A travelling wave tube (TWT) which is described by the memory-less Saleh model where [30]

\[ A[\rho(t)] = A_{sat}^2 \times \frac{\rho(t)}{\rho^2(t) + A_{sat}^2} \]

\[ A[\rho(t)] = \frac{\prod}{3} A_{sat}^2 \times \frac{\rho^2(t)}{\rho^2(t) + A_{sat}^2} \]  \hspace{1cm} (4.3)

And \( A_{sat} \) represents the amplifier input saturation voltage.
- A solid-state amplifier (SSPA) modeled as [31, 32]:

\[
A[\rho(t)] = \frac{\rho(t)}{[1 + (\rho(t)/A_0)^{2p} \rho^{1/2p}]}; \Phi[\rho(t)] \geq 0
\]  

\[ 4.4 \]

Here the parameter \( p \) is the smoothness factor (for the simulations \( p \) is taken as 2) which controls the smoothness of the transition from the linear region to the limiting region, and \( A_0 \) is the maximum output amplitude [show the characteristic graph if you can find it]

PRINCIPLE OF PREDISTORTION

The principle of predistortion is shown in the figure below:

\[
z(t) = R(t).exp[j\Psi(t)] \xrightarrow{T[.]} r(t).exp[j\theta(t)] \xrightarrow{\text{PA}} z(t) = R(t).exp[j\Psi(t)]
\]

**Figure 54: Block diagram showing principle of predistortion using LUT**

Normally a transformation \( T[.\) is applied to the signal, this is done so that when the signal is amplified the nonlinear effects of the PA are compensated for.

The transformation for a memory-less amplifier is determined as follows. Considering \( x(t) = z(t) \), and also considering that the following conditions have to be satisfied:

\[
\rho(t) = A[r(t)] \quad \text{(4.5)}
\]

\[
\varphi(t) = \theta(t) + \phi[r(t)] \quad \text{(4.6)}
\]

Hence by inverting, the input-output relationship for the predistorter is obtained as:

\[
r(t) = A^{-1}[\rho(t)] \quad \text{(4.7)}
\]

\[
\theta(t) = \varphi(t) - \phi[r(t)] \quad \text{(4.8)}
\]
The inversion for both TWT and SSP amplifiers is shown below:

For TWT Amplifier:

\[
 r(t) = \begin{cases} 
 \frac{A_{sat}^2}{2\rho(t)} \left[ 1 - \sqrt{1 - \left( \frac{2\rho(t)}{A_{sat}^2} \right)^2} \right], & \rho(t) \leq \frac{A_{sat}}{2} \\
 A_{sat}, & \rho(t) > \frac{A_{sat}}{2} \end{cases} 
\]

(4.9)

\[
 \theta(t) = \begin{cases} 
 \varphi(t) - \frac{\pi}{6} \left[ 1 - \sqrt{1 - \left( \frac{2\rho(t)}{A_{sat}^2} \right)^2} \right], & \rho(t) \leq \frac{A_{sat}}{2} \\
 \varphi(t) - \frac{\pi}{6}, & \rho(t) > \frac{A_{sat}}{2} \end{cases} 
\]

(4.10)

For SSPA:

\[
 r(t) = \frac{\rho(t)}{\left[ 1 - \left( \frac{\rho(t)}{A_0} \right)^{2p} \right]^{1/2p}} 
\]

(4.11)

As the interest here is in the digital implementation of the predistorter it is necessary to quantize both the input and output samples of the predistorter. Hence it is seen that the predistortion can be obtained in an efficient manner using Look-up table techniques. For the case of the TWT the interval \((0, A_{sat}/2)\) of the input amplitude \(\rho\) and \((0, A_{sat})\) for the output is quantized. In a similar fashion for case of SSPA the interval \((0, A_{sat})\) for input and \([0, r(A_0)]\) for the output amplitude is quantized.

**The predistort Look-up tables**

In practice, the general method to select the look-up predistorter table is to represent the signal in polar co-ordinates [34]. As a result of this selection the output value of the table depends only on the signal amplitude, that is:
The desired input value $\rho$ forms the address from which the modified value, $r$, is derived.

**Adaptation Algorithm used:**

Let $(\rho_n, \phi_n)$ be the input signal, $(r_n, \theta_n)$ be the pre-distorted signal data point and $(R_n, \Psi_n)$ be the corresponding feedback signal where:

$$R_n = G \rho_n,$$  \hspace{1cm} (4.14)

The following recursions are known to converge to the true solutions of the non-linear PA:

$$H_R(\rho_{n+1}) = H_R(\rho_n) - \alpha (R_n - G \rho_n)$$  \hspace{1cm} (4.15)

$$H_\phi(\rho_{n+1}) = H_\phi(\rho_n) - \beta (\Psi_n - \phi_n),$$  \hspace{1cm} (4.16)

The values of the step size coefficients $\alpha$, $\beta$ determine stability, convergence rate and the sensitivity of the system to external errors.

**Simulation Results:**

The characteristics of the TWT power amplifier (Saleh Model) are shown below:

![Figure 55: Characteristics of Saleh model [30]](image-url)
The characteristic of the SSPA power amplifier (Rapp Model) is shown below, here the model parameter $p$ is selected as 2 for the simulations:

![Characteristics of Rapp model](image)

**Figure 56: Characteristics of Rapp model [31]**

Results for WCDMA Signal when TWT PA is used:

![Power Spectral Density](image)

**Figure 57: DPD for WCDMA signal using LUT technique when TWT PA is used**

From the above figure it can be seen that by employing this technique the PA distortions can come down by around 30dB.
Results for WCDMA Signal when SSPA PA is used:

From the above figure it can be seen that by employing this technique the PA distortions can come down by around 20dB.

Results for LTE Signal when TWT PA is used:

From the above figure it can be seen that by employing this technique the PA distortions can come down by around 20dB.
Results for LTE Signal when SSPA PA is used:

From the above figure it can be seen that by employing this technique the PA distortions can be reduced by around 20dB.

Method 2: DPD of PA using the method of least squares for PA with memory effects using the method of least squares

It is seen that in wideband wireless communication system, amplitude gain and phase shift of power amplifier are dependent on the bandwidth of transmission signal, this sensitivity results in the memory effect [33]. Hence, it is very essential to take into account both the nonlinear characteristics and the memory effect in the power amplifier model. This memory effect had not been taken into account in the previous method and is very harmful to be neglected while designing the DPD.

The model that is used most commonly for the analysis of a nonlinear system is the Volterra series, it has the advantage that it overcomes the weakness of Taylor series and considers the memory effects as well [35], the discrete Volterra series can be expressed as:
\[ y(n) = \sum_{k=1}^{K} y_k(n) \]  

(4.17)

Where

\[ y_k(n) = \sum_{m_1}^{Q-1} \sum_{m_k}^{Q-1} h_k(m_1, \ldots, m_k) \prod_{l=1}^{k} z(n - m_l) \]  

(4.18)

Where \( y(n) \), \( z(n) \) are used to represent the output and input signals of power amplifier. \( K \) represents the nonlinear order, and \( Q \) represents the memory depth, \( h_k(m_1, \ldots, m_k) \) is known as the core function.

Wideband RF power amplifiers can be treated as a band limited system, hence by ignoring different time cross term and even order components, Volterra series can be simplified as:

\[ y(n) = \sum_{k=1, \text{odd} q=0}^{K} \sum_{q=0}^{Q} h_{kq} z(n - q) | z(n - q) |^{k-1} \]  

(4.19)

Where \( K \) is nonlinear order and \( Q \) is memory depth. Generally the 3\(^{rd}\) and 5\(^{th}\) order intermodulations are determined to be the main components of nonlinear distortion, and memory effect mitigates with long time delay, thus the parameters that are commonly set are \( K = 5 \), \( Q = 2 \).
**Design of the predistorter:**

The indirect learning architecture [36, 37] that is used is shown in Figure 1 below:

![Block diagram showing DPD of PA using the method of least squares](image)

The predistorter input depicted by $x(n)$, the baseband PA output is depicted by $y(n)$. In the feedback path that is shown as “Predistorter Training”, $y(n)/G$ is its input, where the gain of the PA is denoted by $G$, and its output is given by $z^\wedge(n)$. The predistorter itself is actually an exact copy of the predistorter training block. The output is given by $y(n) = G \cdot x(n)$, and the error signal $e(n) = z(n) - z^\wedge(n)$ is 0.

Now the idea is to reduce the error between $y(n)$ and $G \cdot x(n)$, in order to achieve this the predistorter parameters are selected so as to minimize the error $e(n)$. Ideally the predistorter parameters are the inverse characteristic of the PA, thus the objective is to prepare the predistorter which has the inverse characteristic of PA directly.

Now to denote the baseband input into a nonlinear system by $x(n)$. The baseband output corresponding to this is $y(n)$ which is modeled by [38]:
\[ y(n) = \sum_{k=1}^{K} \sum_{q=0}^{Q} b_{kq} x(n-q) \left| x(n-q) \right|^{k-1} \]  
\[ (4.20) \]

where \( x(n) \) and \( y(n) \) are, respectively, the input and output of the nonlinear system, and \( b_{kq} \) are the coefficients of the nonlinear system. Now to define:

\[ \Phi_{kq}(n) = \left| x(n-q) \right|^{k-1} x(n-q) \]  
\[ (4.21) \]

Equation (4.20) then becomes

\[ y(n) = \sum_{k=1}^{K} \sum_{q=0}^{Q} b_{kq} \Phi_{kq}(n) \]  
\[ (4.22) \]

Now defining the \( N \times 1 \) input data vector \( \mathbf{x} = [x(t_1), \ldots, x(t_N)]^T \), the \( N \times 1 \) output data vector \( \mathbf{y} = [y(t_1), \ldots, y(t_N)]^T \), and the \( K \times (Q+1) \) parameter vector \( \mathbf{b} = [b_{10}, \ldots, b_{K0}, b_{11}, \ldots, b_{KQ}]^T \).

Next define \( \mathbf{\varphi}_{kq} = [\varphi_{kq}(0), \ldots, \varphi_{kq}(N-1)]^T \), \( \mathbf{\varphi}_q = [\varphi_{1q}, \ldots, \varphi_{Kq}] \), and the \( N \times K \) input matrix \( \mathbf{\Phi} = [\mathbf{\varphi}_0, \mathbf{\varphi}_1, \ldots, \mathbf{\varphi}_Q] \). Then equation (4.22) can be represented by:

\[ \mathbf{y} = \mathbf{\Phi} \mathbf{b} \]  
\[ (4.23) \]

Thus from the set of input/output measurements, a simple least-squares (LS) estimator is obtained for the parameter vector \( \mathbf{b} \):

\[ \hat{\mathbf{b}}_{\text{LS}} = (\mathbf{\Phi}^H \mathbf{\Phi})^{-1} \mathbf{\Phi}^H \mathbf{y} \]  
\[ (4.24) \]

**Algorithm Description:**

The predistorter and the power amplifier can both be modeled by Equation (4.23), and they are the two primary nonlinear components that are present in the predistortion system. The parameter vector of the predistorter present in the feedback path is determined from the set of input/output data. From Figure 61, it can be seen that the input signal to predistorter in the feedback path is \( y(n)/G \), and hence by using the PA input signal \( z(n) \) as the output of predistorter in the feedback path, the ideal parameter vector of the predistorter can be obtained using Equation (4.24). After this now the ideal predistorter parameter vector is transferred to the predistorter in the forward
path so as to distort the input signal \( x(n) \). For obtaining better performance, this procedure mentioned above can be repeated, an advantage of this iterative procedure is that more diverse values can be obtained for \( z(n) \) and \( y(n) \) which is in-turn useful for obtaining a more accurate model parameter estimate.

**The algorithm uses the following steps:**

- **Step 1:** Acquire \( N \times 1 \) input data vector \( x \) and \( N \times 1 \) output data vector \( y \)
- **Step 2:** Using input data vector \( x \), calculate input matrix \( \Phi_x \). Then calculate predistorter output data vector \( z \) (where \( z = \Phi_x b \), use initial value of \( b = [1 0 0 \ldots 0]^T \))
- **Step 3:** Using the PA output data vector \( y \), calculate the predistorter input matrix \( \Phi_y \)
- **Step 4:** Using \( \Phi_y \) and \( z \), calculate new \( b \) (where \( b \) gives the coefficients of the non-linear system), then calculate output data vector of “Predistorter Training” \( z \)
- **Step 5:** With \( z \) as the PA input data vector, acquire new PA output data vector \( y \)

**The results from simulations are shown below for a WCDMA Signal:**

The input and output in all the figures below have been normalized to 0dB

![Power Spectral Density](image)

Figure 62: WCDMA signal at the input and output of the PA

In figure 62 the input signal is shown along with the distorted signal from the Power Amplifier
In figure 63 it is seen that by applying predistortion algorithm, the distortions of the PA came down by 20dB, for K=5 & Q=2 (K- nonlinear order, Q-Memory Depth)

In figure 64 it is seen that by applying predistortion algorithm, the distortions of the PA came down by 35-40dB, for K=9 & Q=4
In figure 65 it is seen that by applying predistortion algorithm, the distortions of the PA came down by 50-60dB, for K=13 & Q=6

The results from simulations are shown below for a LTE Signal:

In figure 66 above the input signal is shown along with the distorted signal from the Power Amplifier
In figure 67 it is seen that by applying predistortion algorithm, the distortions of the PA came down by 15-20dB, for K=5 & Q=2 (K- nonlinear order, Q-Memory Depth)

In figure 68 it is seen that by applying predistortion algorithm, the distortions of the PA came down by 25dB, for K=7 & Q=4 (K- nonlinear order, Q-Memory Depth)
In figure 69 it is seen that by applying predistortion algorithm, the distortions of the PA came
down by 30dB, for K=9 & Q=4 (K- nonlinear order, Q-Memory Depth)

In figure 70 shown above it is seen that by applying predistortion algorithm, the distortions of the
PA came down by 40dB, for K=13 & Q=6 (K- nonlinear order, Q-Memory Depth)
Summary

Thus it can be seen from the above graphs that by increasing the nonlinear order $K$ and the memory depth $Q$, the PA distortions come down significantly but the disadvantage is that the computational complexity increases greatly as the size of the matrices involved in the calculations increase.

Method 3: Using NARMA Model

This method uses a digital adaptive predistorter with a nonlinear auto-regressive moving average (NARMA) structure in which the parameters can be easily obtained by using a NARMA model of the PA [39, 40, 41]. This model takes into account both the linear and nonlinear parts, thus preventing a cascaded linear–nonlinear decomposition as in the case of other models like Hammerstein or Wiener. The adaptation process involved for this DPD depends on the PA NARMA model, and its stability, despite its nonlinear feedback structure, may be calculated and preserved. The digital predictive predistortion system used here is based on the block diagram shown in Fig. 71,

![Block diagram showing DPD using NARMA model](image)
Here the DPD linearization is performed at the baseband by adaptively forcing the PA to function linearly. The functioning of the DPD is simple and direct. First it is necessary to identify the low-pass complex envelope PA behavioral model, once identified a set of coefficients determining the PA behavior is calculated at baseband using the PA input \((x_A)\) and output \((y_A)\) discrete complex envelope data.

In this method a kind of NARMA structure to identify the PA low-pass complex envelope behavioral model is used. The output for this NARMA PA behavioral model is described in the equation below:

\[
y_A(k) = \hat{f}_0(x_A(k)) + \sum_{i=1}^{N} \hat{f}_i(x_A(k - \tau_i)) - \sum_{j=1}^{D} \bar{g}_j(y_A(k - \tau_j))
\]  

(4.25)

Where \(\hat{f}_i\) and \(\bar{g}_j\) are nonlinear functions which are evaluated as a Cartesian product between input/output samples (present or delayed) and then implemented with power series or by using lookup tables here the power series method is used to represent them as:

\[
f_i(x_A(k - \tau_i)) = (|x_A(k - \tau_i)|)(x_A(k - \tau_i))
\]  

(4.26)

\[
g_j(y_A(k - \tau_j)) = (|y_A(k - \tau_j)|)(y_A(k - \tau_j))
\]  

(4.27)

\(\tau_i\) and \(\tau_j\) are the most significant delays of the input and output, respectively, contributing at describing the PA memory effects. After determining the nonlinear functions \(\hat{f}_i\) and \(\bar{g}_j\) that describe the PA behavioral model, we now consider \(y_D\) as the desired linearized PA output. It is defined as the signal to be transmitted multiplied by a linear gain:

\[
y_D(k) = x_T(k). G_{\text{linear}}
\]  

(4.28)

Now \(x_A(k) = x_T(k)\) if no baseband predistorter is considered. Next using (4.25) it is possible to derive:

\[
\hat{f}_0(x_A(k)) = y_A(k) - \sum_{i=1}^{N} \hat{f}_i(x_A(k - \tau_i)) + \sum_{j=1}^{D} \bar{g}_j(y_A(k - \tau_j))
\]  

(4.29)
On solving (4.29) it is possible to obtain the necessary amplifier input in order to achieve a certain output. Now we consider $y_D(k)$ (desired output) as a prediction of the future value of $y_A(k)$ (current output) and with respect to this the input value of the PA $x_A(k)$ that permits achieving the desired output $y_A(k)= y_D(k)$ is calculated. Finally, the digital PD output $x_A(k)$ (and PA input) is given as:

$$(x_A(k)) = f_{o}^{-1} \left( y_D (k) - \sum_{i=1}^{N} \hat{f}_i (x_A(k - \tau_i)) + \sum_{j=1}^{P} \hat{g}_j (y_A(k - \tau_j)) \right)$$ \hspace{1cm} (4.30)

The adaptive process followed by this digital adaptive predistorter in order to perform the PA linearization consists in the following:

- Identifying new $f_i$ and $g_j$ nonlinear functions
- Testing the stability of both PA and digital PD NARMA models
- Inverting the nonlinear function $f_o$ since it is necessary to find the digital PD output as shown in the above equation

**Simulation Results:**

**Results when power amplifier based on Rapp model is used (parameter $p = 2$):**

For a WCDMA signal:

![Power Spectral Density](image)

*Figure 72: DPD of WCDMA signal using NARMA model for a PA using Rapp's model*

The above figure shows that by using the NARMA model, the PA distortions penalty went down by 35–40 dB.
For an LTE Signal:

![Power Spectral Density](image)

The above figure shows that by using the NARMA model, the PA distortions penalty went down by 35-40 dB for an LTE signal as well.

**Results when power amplifier based on Saleh model is used:**

For a WCDMA signal:

![Power Spectral Density](image)

The above figure shows that by using the NARMA model, the PA distortions penalty went down by around 35-40 dB like for the previous power amplifier model.
For a LTE Signal:

The above figure shows that by using the NARMA model, the PA distortions penalty went down by around 35-40 dB like for the previous model.

Results when power amplifier based on Ghorbani model is used:

Characteristics of Power Amplifier based on Ghorbani Model in comparison with Rapp’s model:

Figure 75: DPD of LTE signal using NARMA model for a PA using Saleh model

Figure 76: Characteristics of PA using Ghorbani model
The AM/AM and AM/PM characteristics for this model are given as:

\[ F_{AM/AM}(t) = \frac{x_1 \times t^{x_2}}{1 + x_3 \times t^{x_2}} + x_4 t \]
\[ F_{AM/PM}(t) = \frac{y_1 \times t^{y_2}}{1 + y_3 \times t^{y_2}} + y_4 t \]

The Ghorbani’s model is also used for describing SSPA’s, for the GaAs FET SSPA characterized by Ghorbani the parameters are given by:

\( x_1 = 8.1081; \ y_1 = 4.6645; \ x_2 = 1.5413; \ y_2 = 2.0965; \ x_3 = 6.5202; \ y_3 = 10.88; \ x_4 = -0.0718; \ y_4 = -0.003 \)

For a WCDMA Signal:

![Power Spectral Density](image)

Figure 77: DPD of WCDMA signal using NARMA model for a PA using Ghorbani model

The above figure shows that by using the NARMA model, the PA distortions penalty went down by around 35-40 dB like the case of the previous model.
For a LTE signal:

The above figure shows that by using the NARMA model, the PA distortions penalty went down by around 35-40 dB like for the previous model.

Summary

From the above figures it is observed that by using this technique of predistortion the PA distortion penalty goes down by around 35-40dB for different types of power amplifiers both when using WCMDA signals and LTE signals as well.
CHAPTER 5: CONCLUSIONS

In this thesis we have discussed the design and simulation of components of a Digital Front End (DFE) for a RF Base station. In the second chapter the process of Digital-up-conversion (DUC) and Digital-down-conversion (DDC) for WCDMA and LTE signals has been discussed and the design of the necessary filter chains for interpolation and decimation processes has been presented. For DUC it has been determined that the ideal interpolation filter chain consists of 3 half-band filers preceded by a RRC filter for a WCDMA signal and 2 half-band filers preceded by a RRC filter for LTE signal respectively. Similarly for DDC the ideal decimation filter chain consists of 2 half band filters followed by a RRC filter for a WCDMA signal and a half-band filter followed by a RRC filter for LTE signal respectively.

The third chapter discusses the need for Crest Factor Reduction in WCDMA and LTE signals and gives details on the “windowing method” that has been used. Three algorithms have been discussed in detail and the simulation results for each of the algorithms have been presented and the complexities involved for each of the algorithms have been discussed as well for both WCDMA and LTE signals. The fourth chapter discusses the need for Digital Predistortion (DPD) and its importance in correcting the nonlinearities associated with power amplifiers, several power amplifier models have been discussed with and without memory effects and three methods for achieving DPD to correct these nonlinearities have also been presented.

As far as future work is considered, for the DUC and DDC portion of the DFE a more optimal RRC filter can be designed with a lesser number of taps thus in-turn reducing the hardware complexity. For the Crest Factor Reduction part of the DFR an improvement that can be made is to design a cancellation pulse that is more optimum than the one that has been used, this might in-turn reduce the %EVM value. Now looking into the DPD portion of the DFE the improvement
that can be achieved in this area is to work on the NARMA method by using a PA with memory effects.
REFERENCES


