DIGITAL COMPENSATION FOR THE NON LINEARITY OF RF AMPLIFIERS IN WIDEBAND WIRELESS APPLICATIONS

A THESIS

SUBMITTED IN FULFILLMENT OF THE REQUIREMENT

FOR THE AWARD OF DEGREE OF

DOCTOR OF PHILOSOPHY

IN

ELECTRONICS AND COMMUNICATION ENGINEERING

BY

RUCHI PASRICHA

DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

THAPAR UNIVERSITY

2013

CERTIFICATE

I, **Ruchi Pasricha** hereby declare that the thesis entitled, **"DIGITAL COMPENSATION FOR THE NON LINEARITY OF RF AMPLIFIERS IN WIDEBAND WIRELESS APPLICATIONS"** submitted to Thapar University, Patiala, in partial fulfillment of the requirements for the award of the Degree of **Doctor of Philosophy in Electronics and Communication Engineering** is a record of original and independent research work done by me during 2008-2013 under the supervision and guidance of **Dr. Sanjay Kumar**, Professor, Department of Electronics and Communication Engineering, Thapar University, and it has not formed the basis for the award of any Degree/Diploma/Associateship/Fellowship or other similarly title to any candidate of any university.

Ruchi Pasricha

RUCHI PASRICHA

(Signature of Candidate) Date : 24.1.2013

This is to certify that above statement made by the candidate is correct to the best of my knowledge.

Dr. Sanjay Kumar,
Guide and Professor of ECED,
Thapar University,
Patiala.

ABSTRACT

The power amplifier’s nonlinearity broadens the input signal’s bandwidth. This is known as spectral re-growth which is undesirable since it causes interference with adjacent channels. The nonlinearity generates spectral re-growth, which leads to adjacent channel interference and violations of the out-of-band emission requirements mandated by regulatory bodies. It also causes in-band distortion, which degrades the bit error rate (BER) performance. To reduce the nonlinearity, the power amplifier can be operated at a lower power (“backed off”) so that it operates within the linear portion of its operating curve. Linearization techniques for nonlinear microwave power amplifiers have gained significant interest with the advent of spectrally efficient wireless communication systems. Most recent transmission schemes, such as Wideband Code Division Multiple Access (WCDMA) or Orthogonal Frequency Division Multiplexing (OFDM), are especially vulnerable to the nonlinear distortions due to high fluctuations in their power levels. To ensure that linearity is maintained to a high degree, it is necessary to examine the nature of amplifier distortion. Audio amplifier distortion is of concern for many years. The conventional feedback techniques used at audio frequencies are not applicable to many RF amplifiers due to following problems:

1. Stability at high bandwidth
2. Cost for high gains in RF stages.

There are two classes (memoryless model and model with memory) for Digital pre-distortion implementations. Memoryless models focus on the power amplifier that has a memoryless nonlinearity, i.e., the current output depends only on the current input

through a nonlinear mechanism. This instantaneous nonlinearity is usually characterized by the AM/AM and AM/PM responses of the power amplifier, where the output signal amplitude and phase deviation of the power amplifier output are given as functions of the amplitude of its current input. Both memoryless polynomial algorithm and Look-Up Table (LUT) based algorithm are two key algorithms for memoryless models.

A low complexity, direct-learning multilevel lookup table based adaptive digital predistortion technique has been proposed to linearize a power amplifier (PA).Compared with the conventional predistorters, the proposed technique shows fast adaptation speed which enables the predistorter to track time-varying PA nonlinearities.

An adaptive digital pre-distortion (ADPD) technique is modeled and simulated. The aim of this thesis is to develop a linearization technique which is less complex and requires less memory from FPGA implementation point of view. For the communications system architect and the RF power amplifier (PA) designer, the new wireless formats introduce a number of challenges. Designers must determine the performance gap between their existing 3G designs and tomorrow’s 4G operating environments, and whether these 3G designs will need to be redesigned, or a new vendor qualified. The hardware must also meet or exceed absolute performance metrics such as ACPR, EVM or throughput (e.g., BLER and BER), while also meeting internal product design goals. Because smart phones and other advanced wireless devices rely so heavily on battery power, getting the most efficiency out of a design is critical. The RF PA plays a particularly key role in choosing and designing the right PA to meet design goals which is a significant challenge.

A pre-distorter applies distortion to the input signal in order to drive the PA harder. The DPD-PA cascade attempts to combine two nonlinear systems into one linear result which allows the PA to operate closer to saturation. Beyond this point, no increase in power will suffice to linearize the PA. The PAPR of the signal greatly restricts optimal performance of the DPD system. A CDMA signal, for example, may have a PAPR as high as 13dB. A PA transmitting such a signal must operate with significant back-off to prevent peaks from occurring beyond saturation. There are two common types of DPD implementation, the first is an analog implementation using a physical nonlinear device. The second and perhaps more popular choice, is a digital signal processor (DSP) hardware implementation where the DPD function is defined algorithmically through software.

As the signal bandwidth gets wider, such as in WCDMA, mobile WiMAX and 3GPP LTE and LTE-Advanced (up to 100 MHz bandwidth, 5 component carriers of carrier aggregation), power amplifiers begin to exhibit memory effects. This is especially true for those high power amplifiers used in wireless base stations. The causes of the memory effects can be attributed to thermal constants of the active devices or components in the biasing network that have frequency dependent behaviors. As a result, the current output of the power amplifier depends not only on the current input, but also on past input values. In other words, the power amplifier becomes a nonlinear system with memory. For such a power amplifier, memoryless predistortion can achieve only very limited linearization performance. Therefore, digital predistorters must need to have memory structures.

The most important algorithm for models with memory for Digital predistortion implementation is Volterra series and its derivatives. The most general way to introduce memory is to use the Volterra series. However, the large number of coefficients of the Volterra series makes it unattractive for practical applications. Therefore, there are several Volterra's derivatives including Wiener, Hammerstein, Wiener–Hammerstein, Parallel Wiener structures and memory polynomial model are popular in digital pre-distorters. The so-called "memory polynomial" is interpreted as a special case of a generalized Hammerstein model and is further elaborated by combining with the Wiener model. A memory polynomial pre-distorter uses the diagonal kernels of the Volterra series. The memory polynomial predistorter is used to linearize power amplifiers with memory effects. The pre-distorter is constructed using the indirect learning architecture, thereby eliminating the need for model assumption and parameter estimation of the power amplifier.

The introduction of the work and power amplifier modeling is discussed in Chapter 1. A comparison of various linearization techniques and dissertation outline is discussed in this chapter.

Literature review, gaps in present study, motivation, objectives and problem formulation is discussed in Chapter 2.

Power Amplifier modeling using simplified complex memory polynomial has been presented in Chapter 3.Gain, phase, input characteristics, constellations of actual and modeled PA are compared.

Digital predistortion technique is described in Chapter 4. Simulation results show linearity errors for power amplifier, magnitude and phase errors for DPD and frequency plots for power amplifier input and output.

In chapter 5, an adaptive digital predistortion technique is modeled and simulated. The Spectrum of output with and without predistortion is shown. Calculations show that the proposed LUT based adaptive digital predistorter reduces EVM to 18.14% and ACLR by 54.19 dB.

In Chapter 6, Digital predistortion of power amplifiers with memory effects for LTE systems is presented. Value of EVM is reduced after employing DPD for LTE systems. A table for ACLR measurements shows deterioration in the ACLR values when DPD stage is employed.

In Chapter 7, conclusions and future scope of work are discussed. Major findings in work are included in this chapter.

ACKNOWLEDGEMENT

First and foremost, I praise and thank our ALMIGTHY GOD whose blessings have bestowed in me the will power and confidence to carry out my Ph.D.

I extend my thanks to our Director, Dr. Abhijit Mukherjee and Registrar Dr. Dilip Kumar Patnaik, for giving me this opportunity to pursue Ph.D.

I would like to place my heart-felt and sincere thanks to my research guide, Dr. Sanjay Kumar, M.E., Ph.D., Associate Professor, Electronics and Communication Engineering, Thapar University, Patiala. I feel it a pleasure to be indebted to my guide for his valuable support, advice and encouragement.

I owe my gratitude and profound thanks to the doctoral committee members Dr. Rajesh Khanna, Professor and Head of Department, Electronics and Communication Engineering, Thapar University, Patiala, Dr. Amit Kohli, Assistant Professor, Electronics and Communication Engineering, Thapar University, Patiala and Dr. A.K. Lal, Associate Professor, School of Mathematics & Engineering for their valuable suggestions and guidance.

I would like to greatly acknowledge with deep sense of gratitude to all the faculty members and office assistants in the Department of Electronics and Communication for their continuous motivation, moral and technical support during the research work. My special thanks go to Dr. Rajesh Khanna, Professor and Head of Department, Electronics and Communication Engineering, Thapar University, Patiala for his valuable suggestions for the journal work.

I would like extend special thanks to my colleagues, Department of ECE, Chandigarh Engineering College for their valuable suggestions, support and encouragement.

I also thank the Management of Chandigarh Group of Colleges, Landran, Mohali, India for sponsoring my research papers in Springer and AMIS.

It is great privilege to express my profound gratitude, deep love and affection to my dear husband Mr. Atul Singla and my daughters Shreya Singla and Sanvi Singla who have been with me always in all the difficulties that came across my life. Their love, patience, persistent encouragement, good understanding and prayers enabled me to complete the research work successfully. Finally, I am highly indebted to my parents, sister, brother, brother in law, sister in law, mother-in-law, father-in-law and the family members for their love, affection, sacrifices, endurance and prayers all through my life.

Ruchi Pasricha

TABLE OF CONTENTS

LIST OF FIGURES

List of Tables

List of Acronyms

BER	Bit Error Rate
WCDMA	Wideband Code Division Multiple Access
OFDM	Orthogonal Frequency Division Multiplexing
LUT	Look-Up Table
ADPD	Adaptive Digital Predistortion
DSP	Digital Signal Processing
FM	Frequency Modulation
GMSK	Gaussian Minimum Shift Keying
EVM	Error Vector Magnitude
ACLR	Adjacent Channel Leakage Ratio
PAPR	Peak to Average Power Ratio
FSK	Frequency Shift Keying
CF	Crest Factor
PCDE	Peak Code Domain Error
LMS	Least Mean Square
RLS	Recursive Least Squares
EER	Envelope Elimination and Restoration
LO	Local Oscillator
IF	Intermediate Frequency
AQM	Analog Quadrature Modulator

IMD	Inter Modulation Distortion
FPGA	Field Programmable Gate Error
ADC	Analog to Digital Convertor
ASIC	Application Specific Integrated Circuits
DP	Digital Pre Distortion
OPEX	Ongoing Operating Expenses
WiMAX	World Wide Interoperability for Microwave Access
LTE	Long Term Evolution
NMSE	Normalized Mean Square Error
CCDF	Complimentary Cumulative Distribution Function

CHAPTER 1

INTRODUCTION

In modern wireless communication systems, it becomes necessary to provide high data rate multimedia services to the numerous subscribers in time.CDMA-2000, WCDMA, OFDM etc. transmit non constant envelope signals. These wireless communication systems use the limited frequency resource and the output modulated signals vary rapidly with high peak to average ratio(PAR).It is therefore necessary to develop power amplifier with high efficiency and good linearity so as to counteract with the problem. However, due to high PAR, the PAs operate at a large backoff power level to achieve linearity and therefore this lowers down the efficiency, typically less than 10% i.e. more than 90% of dc power is lost in form of heat. The possible solution is to operate the amplifier near to saturation where they are highly non linear but efficient, and linearize them by using external circuitry. One of the key issues in PA used in new generation mobile communication systems is the linearity which is visible in its gain and phase characteristics. A linear amplifier has constant gain and phase response for an input power region.

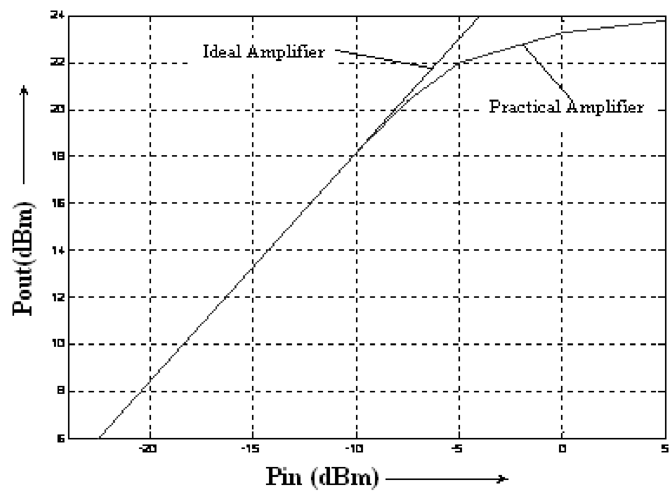

Fig. 1.1 Output power characteristics of a Power Amplifier [2]

The graph shows the linear curve for an ideal amplifier and practical amplifier. As we can see there is deviation in the behavior of practical amplifier as the two parameters-linearity and efficiency limits the performance of a PA. Fig. 1.1 shows output power versus input power characteristics of an ideal and a practical PA. A plot of power output of an amplifier with an ideal gain compression characteristics is shown in figure 1.2. The dashed line represents that ideal output changes abruptly from an area where no gain compression takes place to an area where massive gain compression(saturation) takes place. In terms of system requirements for this component, there is a sudden change from linear (usable) power to saturated (unusable) power. Fig. 1.2 (i) and 1.2 (ii) shows gain (AM/AM characteristics) and phase plots (AM/PM characteristics) of a PA. Solid lines are gain and phase characteristics of a practical PA and dashed lines indicate the ideally linear PA gain and phase characteristics.

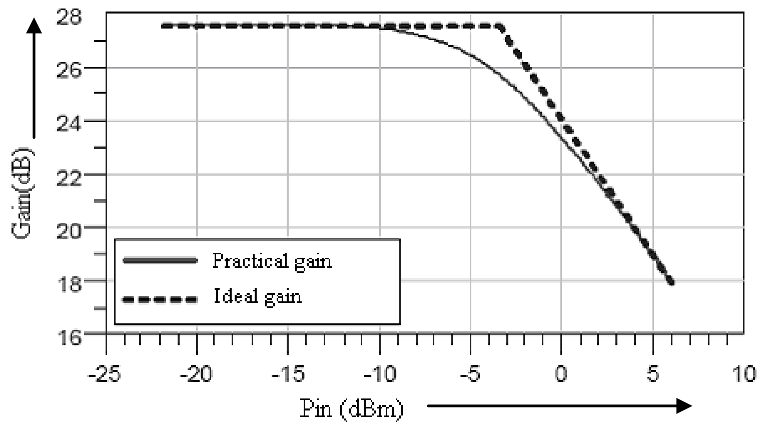

Fig.1.2 (i) Non-linear and ideally linear Power Amplifier Gain characteristics [2]

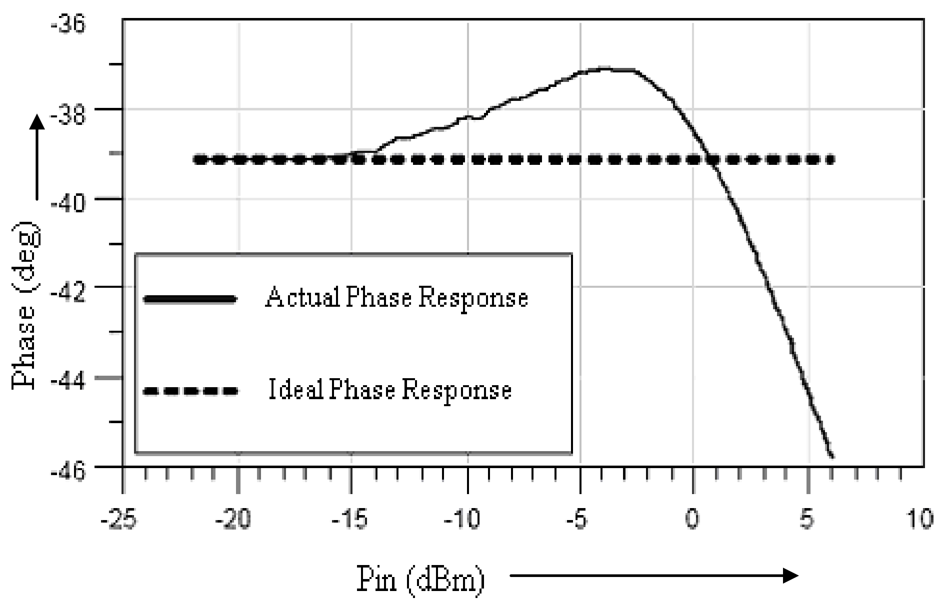

Fig. 1.2 (ii) Non-linear and ideally linear Power Amplifier Phase characteristics [2]

After reaching a relatively high output power value, the amplifier gain decreases gradually with increasing input power because the PA reaches its saturation point. Phase non-linearity increases with increasing input power and this can be observed in the graph of power amplifier phase characteristics. For the signals like FM (Frequency Modulation) or GMSK (Gaussian Minimum Shift Keying), which have constant envelope, PA linearity is not an important issue because the instantaneous input power stays constant and therefore there are no gain and phase variations for a specific operation point. However, the introduction of efficient transmission formats such as Wideband Code Division Multiple Access (W-CDMA) used in 3G cellular system has driven the need for spectrally efficient modulation schemes capable of handling the increased volume of mobile users. These schemes combine multiple carriers into a single wideband signal which is then amplified by a PA. Although W-CDMA signals are spectrally more efficient, they are subject to severe IMD distortion when the PAs are operated outside of their linear range. This distortion, often measured as Error Vector Magnitude (EVM) and Adjacent Channel Leakage Ratio (ACLR), is strictly limited by regulations as specified in [1].

Non-linearity not only causes in-band distortion but also degrades the bit error rate (BER) performance. Typically, the required linearity can be achieved either by reducing power efficiency or by using linearization techniques. For a Class-A PA, simply "backing off" the input power level can improve linearity. However, for high peak to average power ratio (PAPR) signals, this normally reduces the power efficiency down to 10% while increasing heat dissipation up to 90%. When considering the vast number of base stations that wireless operators need to account for, increasing power consumption, or in other words, power back-off is not a viable tradeoff. Therefore, an amplifier linearization has become an important technology and a desirable alternative to backing-off an amplifier in modern communications systems. Although nonlinear distortions can be equalized at the receiver side but it is complicated to implement due to the unknown effects of the channel. It is therefore easy to reduce the nonlinear distortions at the transmitter side.

1.1 Power Amplifier Modeling

Modeling of RF PA has been a subject of intense research for the last few years. But the modeling task of PA will not be accomplished until the parameters on which its performance can be evaluated are not known. A brief introduction to the basic insight of the PA parameters like IMD Products, ACLR, Efficiency, CF, EVM and PCDE, which can be used to evaluate its performance is given. The topic also presents the behavioral modeling approach to model PA. Non-linear Memory-less and Quasi Memory-less models of PA like Rapp model, Saleh model, Ghorbani model, Hyperbolic Tangent model has also been discussed. Also non-linear Models of PA with Memory like Wiener model, Hammerstein model, Parallel Hammerstein model and Memory Polynomial model most widely used in literature has been presented in this chapter.

1.1.1 Intermodulation Products

For two tone (f_1, f_2) input, the 3rd order term results in $2f_1 - f_2, 2f_2 - f_1, 2f_1 + f_2, 2f_2 + f_1$ harmonic components at the output of PA. So, in general, if we consider higher order terms also, we will get $mf_1 \pm nf_2$ harmonic components at the output of PA, where m and n are integers. These harmonic components are also known as IMD products.

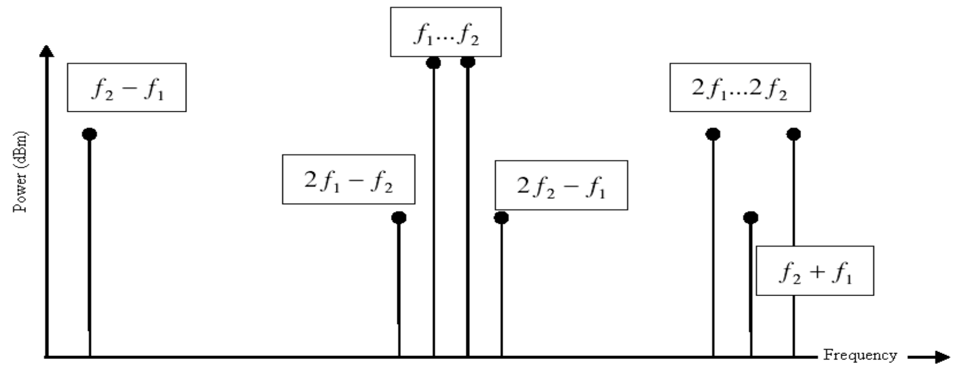

Fig. 1.3 Two Tone Test Spectrum Results of a non-linear Power Amplifier [2]

But out of these IMD's, the most serious is 3rd order IMD which is very close to the desired frequency as shown in Fig. 1.3 and is not easy to filter out. This product is usually characterized by the third order intercept point refereeing either to the input or the output(IIP_3 $orOIP_3$). This is best defined by looking at Fig. 1.3. It can be shown that the slope of the linear gain for input and output powers in dB is unity, likewise the slope of the third gain of the third order IMD component is 3[3], the point where the third order line intersects with the linear gain line is the third order intercept point.

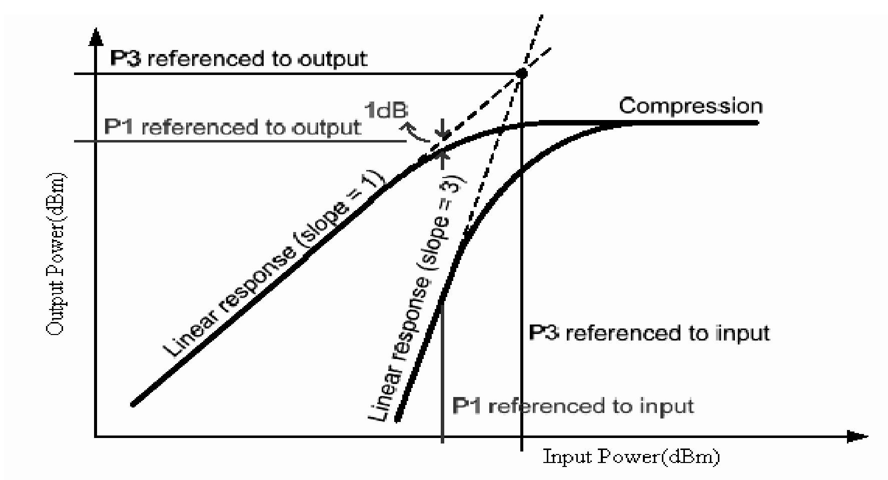

Fig. 1.4 Illustration of the first and third order intercept points [2]

Another figure of merit to characterize non-linearity is the 1dB compression point. For a non-linear device, the 1dB compression point is defined as the point where the difference between the non-linear device's output and the linear output is exactly 1 dB. 1dB compression point can be calculated as [4].

$$Y_{1dB} = G_1 V_1 + G_3 \left[\frac{3}{4} V_1^3 + \frac{3}{2} V_1 V_2^2\right] \tag{1.1}$$

The 1 dB compression point is typically 12 to 15 dB less than the 3rd order intercept point assuming that they are referenced at the same point.

1.1.2 Adjacent Channel Leakage Ratio

PA produces unwanted signal emission to adjacent channels, which are outlined in the related standards for W-CDMA and other modulation schemes. The term ACLR is a measure of adjacent channel emission and is very often faced as regulatory parameter [5]. For two tone input, ACLR is defined as [6].

$$ACLR_{dB} = IMR_2 - 6 + 10log_{10} \frac{n^3}{4\left[\frac{2n^3-3n^2-n}{24} + \frac{mod\left[\frac{n}{2}\right]}{8}\right] + \left[\frac{n^2-mod\left[\frac{n}{2}\right]}{4}\right]} \tag{1.2}$$

Where IMR_2 is two tone Inter Modulation Ratio (dB), defined as

$$IMR_2 = \frac{\frac{3}{4} G_3 V_1^2 V_2}{G_1 V_1} \tag{1.3}$$

and n is number of tones for the band of interest.

The idea behind this formula is to estimate ACLR for multi carrier modulation schemes, by expanding two tones inter modulation IMR to given number of in-band carrier. Permutation of each tones and interaction with all of the others will shape the spectral growth depending on $G_3, G_5 \dots . G_{2n-1}$odd order non-linearity coefficients. Also total power due to IMD components can be calculated as

$$P_{IMD} = \frac{\left[\frac{3}{4} G_3 V_1 V_2^2\right]^2}{2} \qquad (1.4)$$

1.1.3 Efficiency

One important parameter of PA is its Power Efficiency (η). Efficiency is a measure of how effectively an amplifier converts power drawn from the dc supply to useful signal power delivered to the output.

$$\eta = \frac{P_{ac}}{P_{dk}} \qquad (1.5)$$

Traditionally, PAs are categorized into different classes according to their historical precedence. Different PA classes can be divided into two major groups: linear and non-linear PAs. Class A, AB, B and C are some of the well-known linear PAs, which are distinguished primarily by their bias condition. Linear PAs have the advantage of high linearity that is important for variable envelope modulation schemes. However, linear amplifiers suffer from poor maximum power efficiency which limits their applications in low power devices. In practice, an efficiency of only below 20% can be achieved in those systems. In contrast, non-linear PAs (also known as switched mode PAs) can achieve better efficiency. As suggested by its name, non-linear PAs have poor linearity performance. Nevertheless, it is still acceptable for constant envelope modulation schemes (e.g. FSK). To overcome the problem of linearity for variable envelope systems, many linearization techniques have been proposed for non-linear amplifiers [7][8]. Therefore, due to their high efficiency and the development of linearization techniques, non-linear PAs have received more attention over linear topologies in mobile communication in the last decade. Class E and F are the most common classes of non-linear PAs. In comparison, Class E PA requires fast switching driver signal that is not required for Class F PA. For these reasons, Class-F PA has drawn more attention for its easier implementation and better integration with sub-micron CMOS technology. A Class F PA uses a output filter to control the harmonic content of its drain voltage or drain-current waveforms, thereby shaping them to reduce power dissipation by the transistor and thus to increase efficiency.

1.1.4 Crest Factor and its effect

CF is a measure of peak to average power ratio (PAPR) of the PA and is defined [9] as:

$$\mathrm{CF} = 10\log_{10}\left[\frac{\max(\mathrm{x}^2)}{\mathrm{E}[\mathrm{x}^2]}\right] \tag{1.6}$$

Where $\max\ x^2$ represents the maximum (peak) value of the signal x and $E\left[x^2\right]$ is expected (average) value of x. In order to increase the linearity the PA can be backed off from its saturation point. This back-off can be measured in terms of output back-off (OBO), which is defined as

$$\mathrm{OBO} = 10\log_{10}\left[\frac{\mathrm{P_{sat}}}{\mathrm{E[gnl(x)^2]}}\right] \tag{1.7}$$

Where g_{nl} is a non-linear function representing the non-linear gain response of the PA and P_{sat} is the maximum output power of the PA.

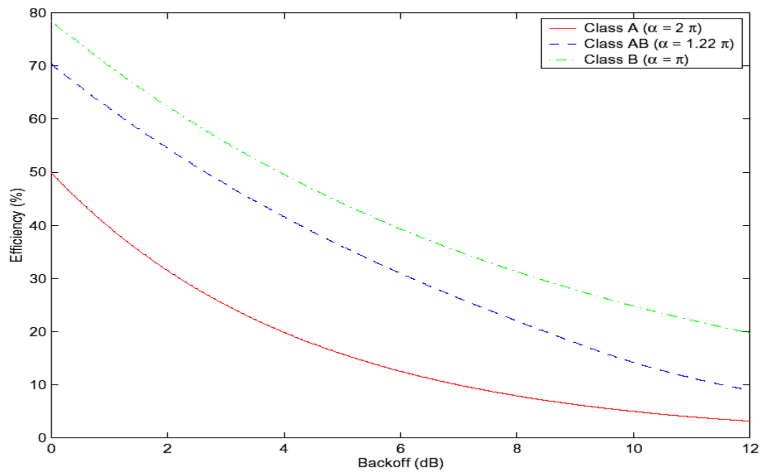

Fig. 1.5 Amplifier efficiency performances [2]

As shown in Fig. 1.5, the OBO defines how many decibels less than the maximum saturated output power the average output power is. If amplifier is assumed to be operating in linear region, then as clear from Fig. 1.4

$$OBO \approx CF \tag{1.8}$$

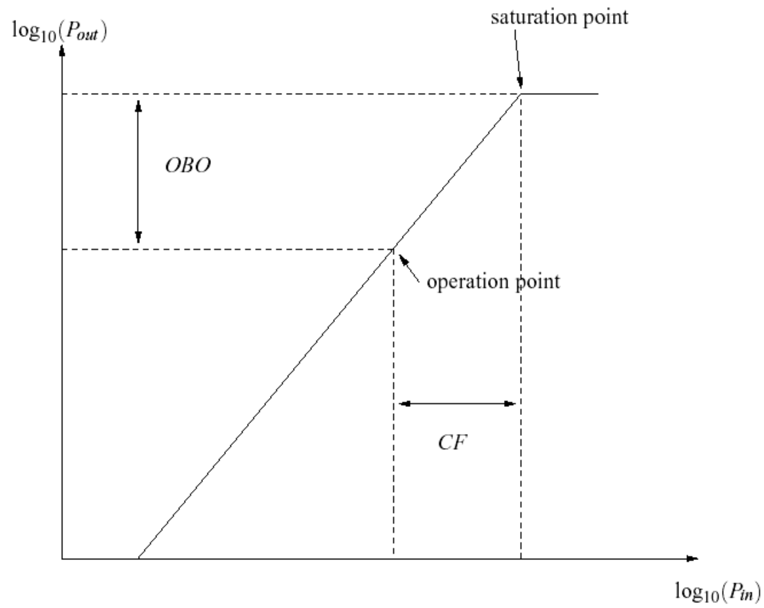

Fig. 1.6 Output Back-off Ratio

Thus CF has direct influence on OBO i.e. higher is the PAPR of the input signal more is the back-off required to achieve the same linearity.

1.1.5 Error Vector Magnitude

The modulation accuracy of the W-CDMA signals is measured by Error Vector Magnitude (EVM). EVM is a measure of the difference between the theoretical waveform and modified version of the measured waveform. EVM can be defined as the distance between the desired and actual signal vectors (error vector), normalized to a fraction of the signal amplitude.

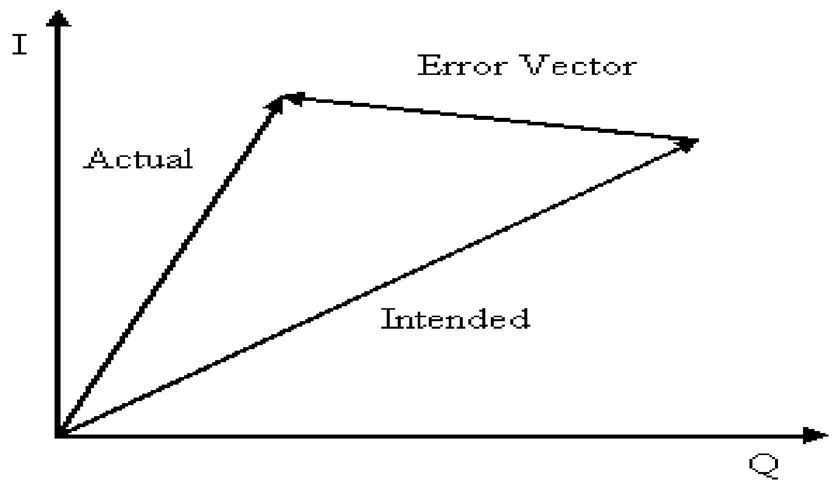

Fig. 1.7 Error Vector Magnitude

Mathematically, the error vector can be written as,

$$e = y - x, \tag{1.9}$$

where y is the modified measured signal and x is the ideal transmitted signal.

The EVM can be defined as the square root of the ratio of the mean error vector power to the mean reference signal power expressed as:

$$EVM_{RMS} = \sqrt{\frac{E[e^2]}{E[x^2]}} \tag{1.10}$$

Also for each symbol k, EVM can be defined as,

$$EVM(K) = \frac{|E(k)|}{\sqrt{\frac{1}{N}\sum_{k=1}^{N}|S(k)|^2}} \tag{1.11}$$

where E(k) is the error vector for symbol k, S(k) is the ideal signal vector of the symbol k and N is the number of symbols. Root-mean-square (RMS) value of EVM for a number of symbols is a widely used measure of system linearity and it can be defined as

$$EVM_{RMS} = \frac{\sqrt{\sum_{k=1}^{N} |E(k)|^2}}{\sqrt{\sum_{k=1}^{N} |S(k)|^2}} \tag{1.12}$$

EVM is an in-band distortion causing high bit error rates during reception of the transmitted data. Therefore EVM specifications must also be fulfilled in order to have proper communication.

1.1.6 Peak Code Domain Error

The quality of a W-CDMA signal can also be measured by the PCDE. PCDE is computed by projecting the power of the error vector onto the code domain at a specific spreading factor (SF). The code domain error (CDE) for every code in the domain is defined as the ratio of the mean power of the projection onto that code to the mean power of the composite reference waveform. This ratio is expressed in dB. The PCDE is defined as the maximum value for the CDE for all codes (Digital cellular telecommunications system, 1999). The projection of the error vector is calculated by despreading the error vector by all codes. After the despreading operation, there are SF error signals $e_{d,k}$. The PCDE is calculated as

$$\mathrm{PCDE} = 10\log_{10}\left[\frac{\max\left(\mathrm{E}\left[|\mathrm{e_{dk}}|^2\right]\right)}{\mathrm{E}[|\mathrm{s}|^2]}\right] \tag{1.13}$$

The relationship between the EVM and the PCDE can be written as [10]

$$PCDE = 10log_{10}\left[\frac{1}{SF}EVM^2\right] \quad (1.14)$$

1.2 Non-linear Memory-less and Quasi Memory-less Models of Power Amplifier

The AM/AM conversion for a non-linear system is the relation between the amplitude of the system's output and the amplitude of the system's input. The AM/PM conversion for a non-linear system is the relation between the phase change of the system's input and output, and the amplitude of the input signal. This is shown in Fig. 1.8.

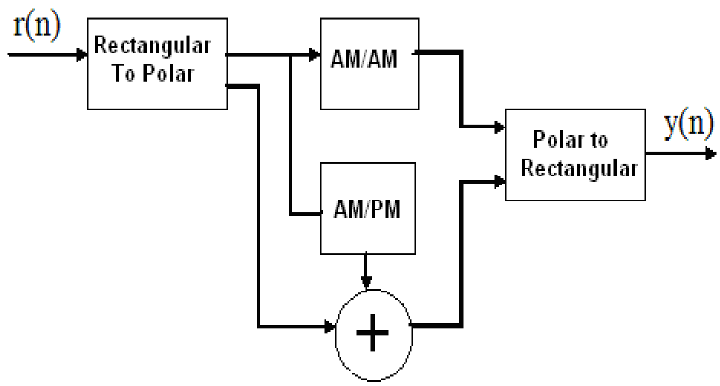

Fig. 1.8 Amplitude-Phase non-linear model structure for a complex base band input and output signals

Assuming the pass band input signal [11], the output of the AM/AM and AM/PM model $y(n)$ can be written as

$$y(n) = g\big(r(n)\big)cos(\omega_0 n + \phi(n) + f(r(n)) \quad (1.15)$$

where $g(n)$ represents AM/AM non-linearity and $f(n)$ represents AM/PM non-linearity.

1.2.1 The Rapp Model

The Rapp model uses three parameters, and models amplitude distortion but no phase distortion. The general expression of the AM/AM conversion is as follows [12]

$$g\big(r(n)\big) = \frac{r(n)}{\left[1+\left[\frac{r(n)}{O_{sat}}\right]^{2s}\right]^{1/2s}} \tag{1.16}$$

where$O_{sat}$is a parameter that sets the output saturation level and s is a parameter that sets the smoothness of the transition from linear to saturation states. It assumes linear performance until the saturation point is approached. When the saturation point is approached, a transition towards a constant saturated output is applied.

Advantages:

- The technique of this model is quite simple.
- The smaller the value of s, the smoother the transition.

Limitations:

- For higher values of s, the transition is not smooth.

1.2.2 The Saleh Model

The Saleh model is a quasi memory-less model. It uses four parameters to fit the model to measurement data. It’s AM/AM and AM/PM conversion functions are described by the following equations [13]

$$g\big(r(n)\big) = \frac{\alpha_a r(n)}{[1+\beta_a]r(n)^2} \tag{1.17}$$

$$f\left(r(n)\right) = \frac{\alpha_{\varphi} r(n)^2}{[1+\beta_{\varphi}] r(n)^2}` \quad (1.18)$$

where $\alpha_a, \beta_a, \alpha_{\varphi} and \beta_{\varphi}$ are the model's parameters.

Advantages:

- AM/AM and AM/PM characteristics are obtained.
- The model can be extended to incorporate frequency dependent behaviour.

Limitations:

- The equations which are extended to incorporate memory effects introduce significant loss in generality since it restricts the shape of in-phase and quadrature nonlinearities.

1.2.3 The Ghorbani Model

The Ghorbani model uses eight parameters to fit the model to the measurement data, this model is quasi-memory-less, and it's AM/AM and AM/PM conversions functions are described by the following equations [14]

$$g(r(n) = \frac{x_1 r(n)^{x_2}}{1+x_3 r(n)^2} + x_4 r(n) \quad (1.19)$$

$$f(r(n) = \frac{y_1 r(n)^{y_2}}{1+y_3 r(n)^2} + y_4 r(n) \quad (1.20)$$

$where x_1, x_2, x_3, x_4, y_1, y_2, y_3, y_4$ are the model's parameters.

Advantages:

- The Ghorbani model is best suited to FET Amplifier characteristics and matches them closely.

1.2.4 The Hyperbolic Tangent Model

The Hyperbolic Tangent model is quasi memory-less. It has five parameters these are IIP_3, linear gain G, upper power limit P_V, lower power limit P_L and the linear phase gain G_{PM} . One characteristic of this model is that its parameters are related to physical attributes like IIP_3. Its AM/AM conversion function is described by the following equation

$$g\big(r(n)\big) = \tanh\left[\frac{3}{IIP3} r(n)\right] G \qquad (1.21)$$

Where IIP_3 is the third order intercept point and $G$is linear gain.

The AM/PM conversion is linear and is specified by the slope of the AM/PM conversion in degree/dB. This linearity is bounded by two parameters which are upper power limit P_V and lower power limitP_L. If the power magnitude of the input is less than P_L then no phase distortion is added, and if the power magnitude of the input is greater than P_V then a constant phase shift of $G_{PM}(P_V - P_L)$ is applied. This can be represented by the following equation

$$f\big(r(n)\big) = \begin{cases} G_{PM}(P_V - P_L)|r(n)| > P_V \\ |r(n)|G_{PM}P_L \leq |r(n)| \leq P_V \\ 0 \qquad P_V < |r(n)| \end{cases} \qquad (1.22)$$

Advantages:

- Used for simple modeling of limiting behaviour of PA.

- It describes the large signal low quality response of bipolar transistor differential pair amplifier.

Limitations:

- Sharpness of transition from linear to limiting characteristics of the model is fixed in relation to the gain and cannot be adjusted without introducing additional parameters

1.3 Non-linear Models of Power Amplifier with Memory

In reality the output of the PA depends upon previous inputs as well as on the current input of the amplifier. This phenomenon is called memory effect. These memory effects are due to thermal effects and long time constants in DC bias circuits. It can be observed as asymmetries in lower and upper sidebands and bandwidth dependent variations in the magnitude of IMD products. For higher bandwidth applications, e.g. W-CDMA, the memory effects becomes severe, and cannot be ignored. Hence, memory-less and quasi memory-less models are not accurate enough. Therefore, a model which considers memory effects should be used for such applications. In the following sub section sections some of the most common models with memory are presented.

1.3.1 Volterra Series

To describe a polynomial non-linear system with memory, the Volterra series expansion has been the most popular model in use for the last three decades. The Volterra theory was first applied with non-linear resistor to a White Gaussian signal. In modern DSP fields, the truncated Volterra series model is widely used for non-linear system representations. The causal discrete-time Volterra filter is similarly based on the Volterra series and can be shown [15] as:

$$y(n) = h_0 + \sum_{k=0}^{\infty} h_1(k_1)x(n-k_1) + \sum_{k=0}^{\infty}\sum_{k=0}^{\infty} h_2\,(k_1,k_2)x(n-k_{1)}x(n-k_2) + \cdots +$$

$$\sum_{k=0}^{\infty} \ldots . \sum_{k=0}^{\infty} h_2\left(k_1, \ldots k_p\right)x(n-k_1) \ldots . x(n-k_p) \qquad (1.23)$$

Where $h_0$is a constant and $\left\{h_j\left(k_1, \ldots\ldots k_p\right), 1 \le j \le \infty\right\}$ is the set of jth-order Volterra kernel coefficients. Unlike the case of linear systems, it is difficult to characterize the non-linear Volterra system by the system's unit impulse response. Also as the order of the polynomial increases, the number of Volterra parameters increases rapidly, thus making the computational complexity extremely high. For simplicity, the truncated Volterra series is most often considered in literature.

The M-sample memory p^{th}-order truncated Volterra Series expansion is expressed as:

$$y(n) = h_0 + \sum_{k=0}^{M-1} h_1(k_1)x(n-k_1) + \sum_{k_1=0}^{M-1}\sum_{k_2=0}^{\infty} h_2\,(k_1,k_2)x(n-k_{1)}x(n-$$

$$k_2) + \cdots + \sum_{k_1=0}^{M-1} \ldots . \sum_{kp=0}^{M-1} h_2\left(k_1, \ldots k_p\right)x(n-k_1) \ldots . x\left(n-k_p\right) + \cdots \qquad (1.24)$$

There are several approaches to reduce the complexity of Volterra series. One approach is the basis product approximation[16]-[18], which represents the Volterra filter kernel as a linear combination of the product of some basis vectors to attempt to reduce the implementation and estimation complexity to that of the linear problem. Using the Volterra series, two major models have been developed to perform non-linear signal processing. The first model is the non-orthogonal model and is most commonly used. It is directly based on the Volterra series called the Volterra model. The advantage of the Volterra model is that there is little or no pre-processing needed before the adaptation. But because of the statistically non-orthogonal nature of the Volterra space spanned by the Volterra series components, it is necessary to perform the Gram-Schmidt/modified Gram-Schmidt procedure, which is crucial especially for the non-linear Least Mean Square (LMS) type algorithms and also for the non-linear Recursive Least Squares (RLS) type adaptive algorithms. The second model is the orthogonal model. In contrast to the Gram-Schmidt procedure, the idea here is to use some orthonormal bases or orthogonal polynomials to represent the Volterra series. The benefit of the orthogonal model is obvious when LMS-type adaptive algorithms are applied.

Advantages:

- Most general model
- Accurate

Limitations:

- Number of parameters needed increases dramatically.

1.3.2 Wiener Model

Wiener model consists of a LTI system followed by a non-linear model without memory as shown in Fig. 1.9

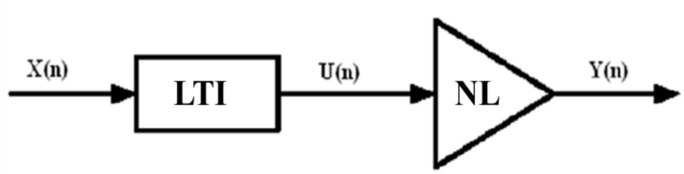

Fig. 1.9 Wiener model

The output of LTI system is given as

$$u(n) = \sum_{q=0}^{Q} h_q x\,(n-q) \tag{1.25}$$

This acts as an input to the non-linear model without memory. The output of non-linear system can be written as

$$u(n) = \sum_{q=0}^{Q} b_{2k-1}\, u(n)|u(n)|^{2(k+1)} \tag{1.26}$$

Putting equation 1.26 in equation 1.25, we get

$$u(n) = \sum_{q=0}^{Q} h_q \sum_{k=0}^{K} b_{2k-1}\, |x(n-q)|^{2(k+1)} x(n-q) \tag{1.27}$$

Advantages:

- Memory effects are modeled more accurately.
- Enhanced model can be employed to characterize power amplifier more accurately.

Limitations:

- Certain conditions are to be satisfied because this model is based on special preset structures.

1.3.3 Hammerstein Model

Hammerstein model consists of a LTI system preceded by a non-linear model without memory as shown in Fig. 1.10

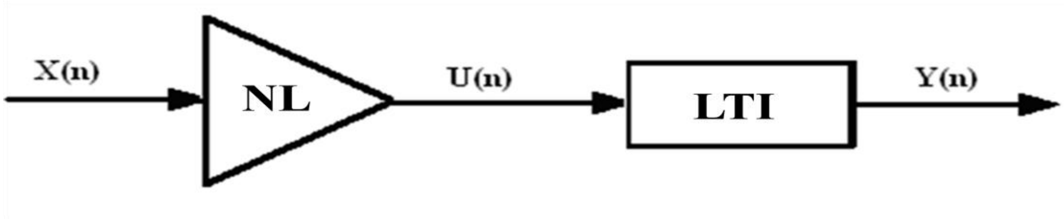

Fig. 1.10 Hammerstein model

The output of non-linear system is given as

$$u(n) = \sum_{k=1}^{k} b_k\, x(n)|x(n)|^{k-1} \tag{1.28}$$

This acts as input to the LTI system and the output of LTI system can be written as

$$y(n) = \sum_{q=0}^{Q} b_{2k-1}\, u(n)|u(n)|^{2(k+1)} \tag{1.29}$$

$$y(n) = \ u(n) * h(q) \tag{1.30}$$

Putting equation 1.28 and equation 1.29 in equation 1.30, we get

$$y(n) = \sum_{k=1}^{k} \sum_{q=0}^{Q} b_k h(q)\, x(n-q)|x(n-q)|^{k-1} \tag{1.31}$$

Advantages:

- Used for satellite communication channels
- Power amplifier at the satellite transponder is driven near saturation to exploit maximum power efficiency for the downlink.

1.3.4 The Parallel Hammerstein Model

The Parallel Hammerstein is an extension of the standard Hammerstein model. The model is illustrated in Fig. 1.11. The system in this case is modeled by [19][20].The output of Parallel Hammerstein model can be given as

$$y(n) = \sum_{k=1}^{k} H_{2k-1}(q)\,|x(n)|^{2(k-1)}x(n) \tag{1.32}$$

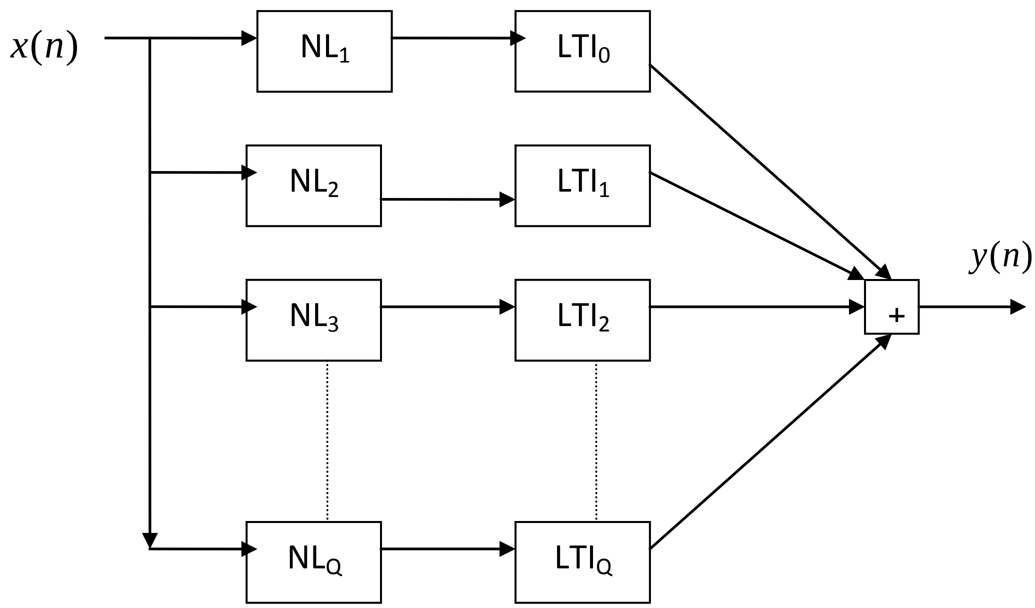

Fig 1.11 Parallel Hammerstein model

The main difference between the Parallel Hammerstein and the standard Hammerstein models is that in the Parallel Hammerstein model, different static non-linear orders are filtered with different LTI systems. For example, the term of the polynomial is filtered with $H_1(q)$, and the 2nd odd power term is filtered with $H_3(q)$ and so on.

Advantages:

- Compared to Wiener and Hammerstein models, this model is more general and therefore can provide a more accurate model for a PA.

- It can be easily converted to memory polynomial model.

1.4 Comparison of various linearization techniques

Newer transmission formats, such as W-CDMA have high PAPR i.e., large fluctuations in their signal envelopes. This means that the PA needs to be backed off far from its saturation point, which results in very low efficiencies, typically less than 10%; i.e., more than 90% of the dc power is lost in the form of heat. Considering the large number of wireless base stations deployed worldwide, improved PA efficiency can substantially reduce the power consumption and cooling costs incurred to the service providers. Thus, to obtain both linear amplification and high power efficiency, a linearizer is required.

The linearizer allows the amplifier to be operated at much higher operating point, since the distortion generated by the amplifier because of the peaks in input signals can be corrected up to the saturation level of the amplifier. Any input signal which drives the amplifier to hard saturation, the resulting distortions cannot be corrected since any increase in input power beyond this point will not result in an increase in output power.

In this section limitations and strengths of existing linearization techniques like Boot Up Bias, Dynamic Bias, Baseband Envelope Feedback, Polar Feedback, Cartesian Feedback, Envelope Elimination and Restoration (EER), Adaptive Feedforward, RF/IF Predistortion, DPD have been presented and compared to explore the idea to develop a better linearization technique for a multicarrier W-CDMA base station.

1.4.1 Boot Up Bias Technique

The simplest and most obvious way to improve the linearity is to drive the amplifier toward Class-A operation. As a result, the PA will operate in the small signal linear region and the corresponding out-of- band emission level will decrease. But this method comes with a price of lowering the overall efficiency of the PA, while reducing the total RF output power. Increasing the DC bias for a Class-A amplifier is an inefficient way to linearize a PA. However, if the bias level can adaptively change with the input envelope of the RF signal so that the PA dissipates as little power as possible while it maintains a reasonable out-of-band emission level, such a technique could be very practical.

1.4.2 Dynamic Bias Technique

In [21], the Dynamic Bias method has been used. It is shown that this method requires a fast speed wideband envelope detector and a DC-DC converter with high current capability, which is currently a challenge for the power supply industry. Also the performance of a Dynamic Bias system could be corrupted by undesired phase distortion occurring when relatively large changes in the bias level happen at a higher power level.

1.4.3 Baseband Envelope Feedback Technique

The RF feedback technique requires the components in the feedback path to operate at a higher frequency band or large bandwidth situation. The baseband signal is modulated onto the RF carrier and amplified by the PA, and then the PA output is taken, demodulated and added to the input such that the output of the main amplifier is linearized. In order to maintain system stability, the loop bandwidth must be within the MHz range.

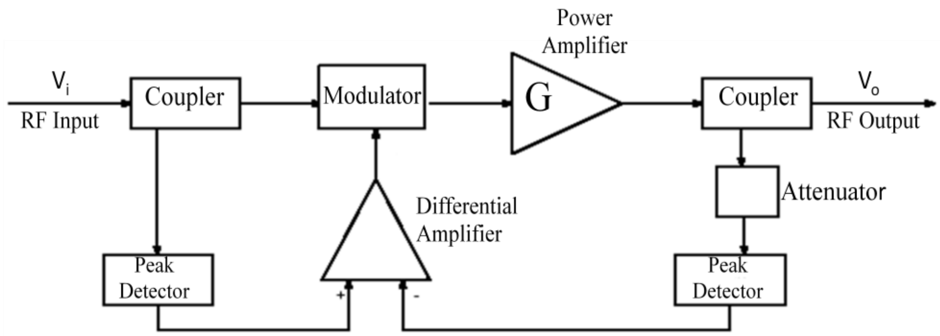

Fig. 1.12 Envelope Feedback to linearize Power Amplifier

Therefore, the main disadvantage of this system is the narrow bandwidth and, in some cases, complexity [22].

1.4.4 Polar Feedback Technique

This technique overcomes the inability of envelope feedback to correct for AM-PM distortion effects. Polar Feedback scheme provides relatively high efficiency since the PA can operate completely non-linearly and this method will be robust since it has both forms

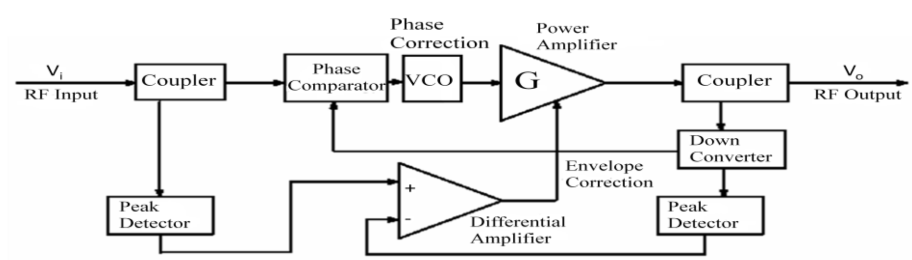

Fig. 1.13 Polar Feedback to linearize Power Amplifier

of feedback. Since both amplitude and phase are corrected in the polar feedback system, variations in temperature, load, and manufacturing should be mitigated. For a narrowband application, the improvement in two-tone IMD is typically around 30 dB[23].The disadvantage of polar feedback are different bandwidths required for the amplitude and phase feedback paths, which leads to a different level of improvement of the AM-AM and AM-PM characteristics and a poorer overall performance than that is achievable from an equivalent Cartesian-loop transmitter.

1.4.5 Cartesian Feedback Technique

Cartesian Feedback was first proposed by Petrovic [24]. In this technique the I and Q components modulate the carrier before passing it to a non-linear but efficient RF PA. The loop control characteristics are established by the gain and the compensation filters. Synchronization between the modulator and demodulator is essential and due to RF path differences in the forward and feedback paths, a phase adjuster is necessary to maintain the correct relationship between the input signals and feedback signals. As shown in Fig. 1.14 the input signal is separated into I and Q and fed to differential amplifier where input signals is subtracted from the feedback signal. The error signal is upconverted to RF using a local oscillator and then combined to produce the complex RF, which is amplified by the PA. The output of the PA is sampled and down converted and separated into I and Q using the same local oscillator used in up conversion process. The down converted output forms the feed back to the differential amplifiers.

A phase shift network is required to ensure that the up and down conversion processes are correctly synchronized. The main advantages of Cartesian over Polar Feedback is that a significant reduction in bandwidth requirement for the feedback loop allows more reduction of IMD and secondly simplicity of implementation. The experimental results in the literature have shown that 10-30 dB of improvement in IMD performance is achievable, however the stability criteria limits the maximum bandwidth to a few megahertz. Also the linearizing bandwidth is 5-10 times larger than the channel bandwidth [25]-[26]. Cartesian Feedback can automatically compensate for drifts in amplifier and non-linearities due to temperature and power supply variations. However, this technique is only conditionally stable and the setting of the adjuster with the aim of maintaining stability is one of the key problems.

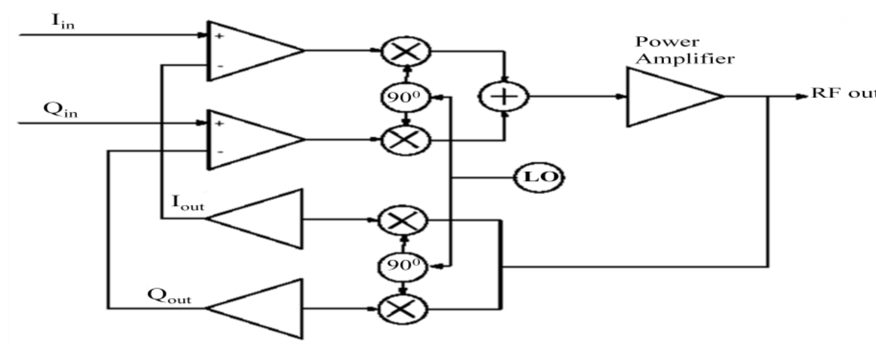

Fig. 1.14 Cartesian Feedback to linearize Power Amplifier

1.4.6 Envelope Elimination and Restoration Technique

The Envelope Elimination and Restoration (EER) linearization method was first proposed by [27]. The envelope of the RF input is first eliminated by a limiter to generate a constant amplitude phase signal. At the same time, the magnitude information is extracted by an envelope detector. The magnitude and phase information are amplified separately and then recombined to restore the desired RF output via a high efficiency switched-mode RF PA.

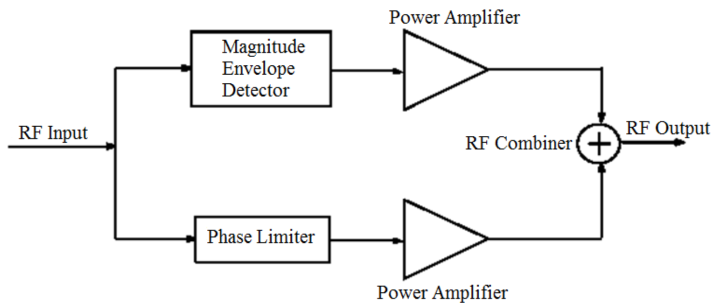

Fig. 1.15 Envelope Elimination and Restoration Technique to linearize the Power Amplifier

The key advantage of EER approach is that the RF PA always operates in an efficient switched mode. That is why the EER system can linearize the switched-mode RF PA without compromising its efficiency. Normally, the restoration is accomplished via biasing the PA's drain voltage. As the drain voltage is varied to correct the output amplitude of the PA, the phase also varies. Too much unintended phase modulation increases spectral re-growth above specifications. Another typical disadvantage of EER is the slowness of the envelope restoration feedback loop. Practically, EER only has on the order of 20-30 dB of dynamic range. Even when the bias level to the PA is zero, some AC power bleeds through.

1.4.7 Adaptive Feedforward Technique

It was invented as a means of distortion reduction in telephone repeaters [28]. Such an architecture has been used successfully to linearize many PAs. Feedforward linearization can deliver reasonable linearization performance (20 dB-40 dB improvement) over relatively wide bandwidths (3 MHz-50 MHz) and has the advantage of inherent stability

[29]-[30]. Amplitude and phase matching is a problem since amplifier characteristics tend to drift with temperature and time, and also vary with manufacturing tolerances.

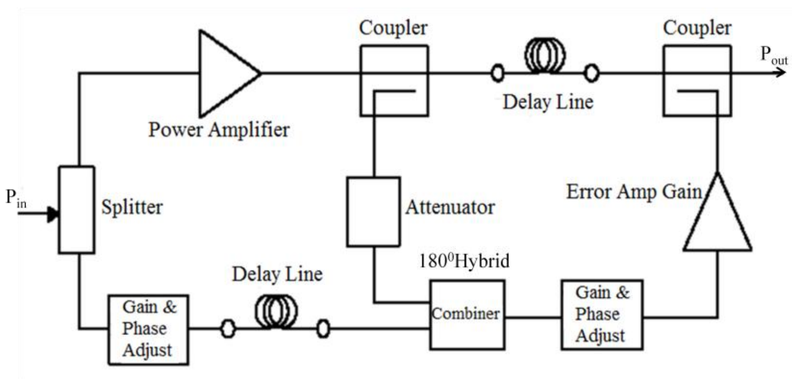

Fig. 1.16 Feedforward Technique to linearize Power Amplifier

1.4.8 RF/IF Predistortion Technique

Predistortion technique in its simplest form consists of a predistorter preceding the non-linear PA which has the inverse transfer characteristics of the PA. Fig. 1.17 shows Predistortion technique in its simplest form. It is an open loop system. However, most solutions presented in the literature have some kind of feedback to enable adaptation of the Predistorter. A large number of Predistorter networks have been reported in the literature.

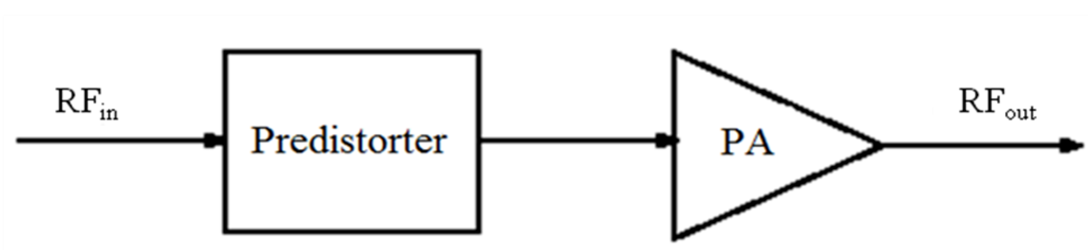

Fig. 1.17 Concept of Predistortion Technique to linearize Power Amplifier

Some networks use non-linear devices to input, while other networks curve-fit the distortion characteristics of the PA. An example of RF Predistorter is Cubic Predistorter, which eliminates the third order distortion by generating a correctly phased addition of a cubic component to the input signal to the PA. The advantage of the RF Predistorter is its ability to linearize the entire bandwidth of the PA, while the advantage of IF Predistorter is that same design can be used for a range of carrier frequencies by altering the Local Oscillator (LO) frequency.

1.4.9 Digital Predistortion Technique

The DPD method uses digital processing to synthesize the inverse transfer characteristic of a PA. The DPD is generally performed at the baseband. The distorted baseband signal is translated to a convenient intermediate frequency (IF) and then the RF signal is generated by mixing the IF with a LO. An alternative to generating IF frequency is a direct conversion to RF signal using an Analog Quadrature Modulator (AQM).

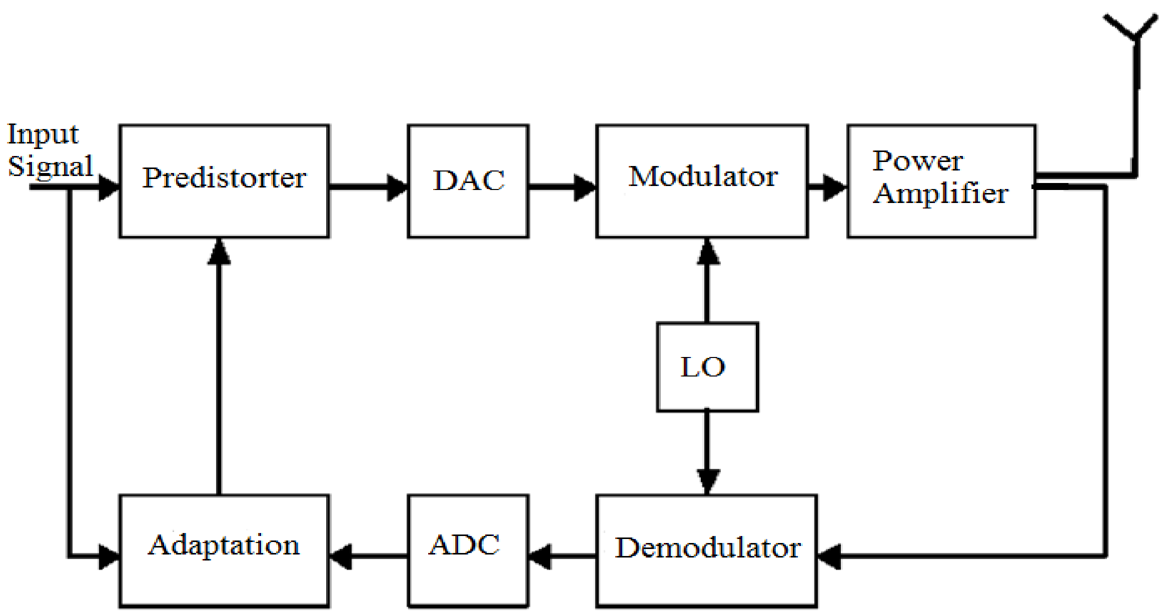

Fig. 1.18 Digital Predistortion Technique to linearize Power Amplifier

The DPD parameters are stored in a look up table or register table which can be updated with adaptive feedback. The Predistortion scheme works on the orthogonal I and Q components of the input and the feedback signals, thus providing both amplitude and phase correction. Furthermore, since the PA's non-linearity is a function of power, frequency, temperature and aging the look up tables must updated continuously, otherwise there will be degradation in IMD time because of temperature drift, component aging, etc. Therefore, the Predistorter should also have the ability to adapt to these changes.

1.5 Comparison of different Linearization Techniques

The literature search shows that from all the linearization techniques that have been developed, the Predistortion technique is the most commonly used in the new systems today.

Table 1.1: Qualitative comparison of different linearization techniques

Parameter	**Feedback Technique**	**Feedforward Technique**	**Predistortion Technique**
Bandwidth	Narrow	Wide	Very Wide
Efficiency	High	Low	High
Linearity	Good	Very Good	Good
Complexity	Medium	High	Medium

Table 1.2: Quantitative comparison of different linearization Techniques

Technique	Correction	Bandwidth	Efficiency	Flexibility	Cost
In-line Predistortion	2 to 3 dB	15 to 25 MHz	5 to 8%	Low	Very Low
Analog Predistortion	3 to 5 dB	15 to 25 MHz	5 to 8%	Low	Low
Feedforward	30 dB	25 to 60 MHz	6 to 10%	Medium	High
Digital Predistortion	15 to 20 dB	15 to 20 MHz	12 to 14%	High	Medium

Tables 1.1 and 1.2 show the comparison among different PA linearization techniques [2]. The DPD technique is moderately complex, as it offers good IMD reduction over a wide signal bandwidth and automatic adaptation maintains performance regardless variation in power supply, frequency, temperature and component aging.

1.6 Dissertation Outline

The thesis is organized as follows:

Chapter 2 deals with the literature survey and related works.

Chapter 3 presents the modeling of a Power amplifier for digital predistortion applications using simplified complex memory polynomial.

Digital Predistortion in WCDMA Power amplifier using embedded processor is discussed in Chapter 4.

Low Complexity look up table based adaptive digital Predistorter with low memory requirement is analyzed in Chapter 5.

Digital Predistortion of Power Amplifiers using Look-Up Table Mehod with Memory Effects for LTE systems are studied in Chapter 6.

Finally Chapter 7 concludes the research work.

Thesis organization is given in Figure 1.19.

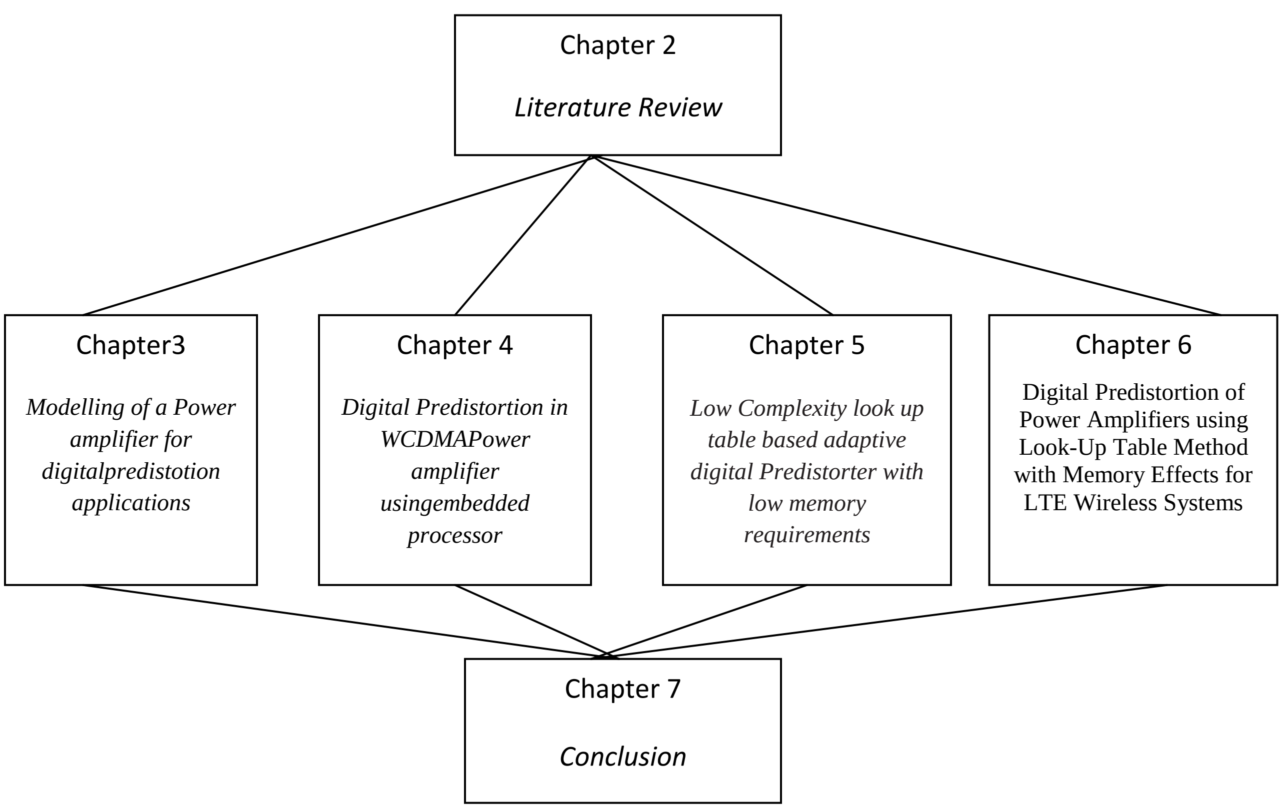

Fig. 1.19 Organization of the thesis

CHAPTER 2

LITERATURE REVIEW

2.1 PA Modeling

The modeling of RF PA can be performed by equivalent circuit modeling and behavioral modeling. In the equivalent circuit modeling, the analysis is carried out at component level and deep knowledge and insight in the circuit layout is required. The behavioral modeling approach utilizes the relationship between input and output signals in the PA i.e. instead of studying the actual cause of the signal appearance; we try to represent the dynamics mathematically. Non-linear amplifier behavioral models are categorized as under:

- Memory-less
- Quasi memory-less
- Models with memory

The causes of the memory effects can be attributed to thermal constants of the active devices or components in the biasing network that have frequency dependent behaviors. Due to the thermal memory effects, the variation in the envelope of a signal produces rapid changes in temperature in the active devices of amplifier. A change in temperature of amplifier affects its non linear gain. Due to changes in envelope, the IMD increases. As a result, the current output of the PA depends not only on the current input, but also on past input values [31]. In other words, the PA becomes a non-linear system with memory.

Numerous research paper have been presented on memory less models of Power Amplifier. A subsampling technique has been proposed [32] which can reduce the bandwidth of test signal without changing the waveform characteristics. The technique cancels the memory effects of amplifier without affecting its static behavior. Memory effects intensity metrics are introduced and are applied to evaluate the memory effects present in Doherty amplifier. A new modified Saleh models for memoryless nonlinear power amplifier was proposed by [33]. Saleh model uses AM to PM and AM to AM characteristics to model non linear amplifiers. The appropriate selections for amplitude and phase coefficients can provide a suitable model. Another advantage of Saleh model is that it could be altered to a frequency dependent model. As shown in the paper, there is improvement in modeling efficiency and performance of modeled Saleh over original Saleh for both polar and Quadrature models.

Non-linear signal processing algorithms have been growing interest in recent years [34]-[36]. To describe a polynomial non-linear system with memory, the Volterra series expansion has been the most popular model in use for the last three decades. It is an extension of simple Taylor series. The Volterra theory was first applied with non-linear resistor to a White Gaussian signal. In modern DSP fields, the truncated Volterra series model is widely used for non-linear system representations. There are several approaches to reduce the complexity of Volterra series. One approach is the basis product approximation [37], which represents the Volterra filter kernel as a linear combination of the product of some basis vectors to attempt to reduce the implementation and estimation complexity to that of the linear problem. A new method for the extraction of a behavioral model of an RF power amplifier using complex representations of input and output signal in time domain is presented [38] .A quasi memoryless model is used to extract the nonlinearity of the device. The proposed model provides more accurate results of actual power amplifier output signals in both time and frequency domains. In the system level analysis and design of wide band digital communication systems, the behavioral model of RF power amplifiers with memory effects are important [39].A Volterra based behavioral modeling is implemented through a bank of FIR filters and the coefficients can be extracted from time domain measurements or circuit envelope simulations. As compared to conventional Quasi memoryless model, the output of non linear distortion of power amplifier with memory effects has better accuracy. In another work done by [40], a simplified Volterra series is proposed to linearize broadband RF PA’s with memory effects. This indirect learning architecture is adopted to design the predistortion scheme and the recursive least square algorithm to identify the parameters of the predistorter. As shown by the simulation results, the proposed predistortion method can compensate the non linear distortion and memory effects of broadband RF PAs effectively.

An alternative form of Volterra system called Memory Polynomial model [41].This model consists of several delay taps and non-linear static functions. This model is a truncation of the general Volterra series, which consists of only the diagonal terms in the Volterra kernels. Thus, the number of parameters is significantly reduced compared to general Volterra series. The Parallel Hammerstein model includes the Hammerstein model as a special case. Hammerstein and Wiener models are the most specialized with the least number of coefficients, but are by no means the easiest to identify. The Memory Polynomial model, however, offers a good compromise between generality and ease of parameter estimation and implementation. A Generalized Memory Polynomial model for Digital Predistortion of RF Power Amplifiers was proposed [42] in which generalized Volterra representation is related to classical Wiener, Hammerstein, Wiener-Hammerstein and parallel Wiener structures. An envelope Memory Polynomial model is introduced to describe dynamic input output characteristics of RF power amplifiers. In this modeling approach, we use a truncation method and non linear time series method to determine the structure of model. A new generalized memory model is proposed that achieves better performance. A new predistortion architecture was described which predistortes AM/AM and AM/PM separately [43].When compared with traditional architecture, the new architecture can significantly reduce the ACPR and EVM of the output signals.

The term memory effects refer to the bandwidth-dependent non-linear effects often present in PAs. These encompass envelope memory effects and frequency memory effects. Envelope memory effects are primarily a result of thermal hysteresis and electrical properties inherent to PAs. Frequency memory effects are due to the variations in the frequency spacing of the transmitted signal [44].

To model a PA system for DPD design, it is important to model the behavior of PA non-linearity and memory effect accurately. There are many different types of models for system identification of non-linear systems. Some of the rich literature includes sources like [45]-[50]. A non-linear function of the parameter vector can be searched by using non-linear local optimization techniques. Due its ability to find unique global optimum point and computational easiness, the LSE technique is used to find the coefficients.

2.2 Linearization Techniques

There are a variety of RF PA linearization techniques. Till date, no single scheme dominates for general-purpose use. But this technique requires a fast speed wideband envelope detector and a DC-DC converter with high current capability. Also the performance of a dynamic bias system could be corrupted by undesired phase distortion occurring when relatively large changes in the bias level happen at a higher power level. Although this problem could be improved by simultaneously adapting a phase feedback loop [51], this adds another dimension of complexity, which is non-trivial in radio frequency (RF) application. A novel method of providing high efficiency power amplifier power control using variable envelope modulation scheme with enhanced data rates based on dynamic adaptive bias control was proposed [52].The architecture allows for substantially higher efficiency levels compared with conventional linear solutions.

Another simple way to perform linearization is to use feedback techniques. But for RF amplification, many stages are normally required to get enough gain, which reduces the overall efficiency since each stage uses power. More importantly, the delay per RF amplifier stage will cause instability if global feedback is used. Hence, not many practical applications employ RF feedback as a linearization approach. A new Wideband Adaptive digital predistortion technique employing feedback linearization was developed [53] for modulated signals with wide bandwidths by combining digital feedback predistortion linearization and memory effect compensation techniques. A linearization technique combined with iterative noise cancellation was proposed [54]. The simulation results based on mean square error and power spectral density criteria to evaluate PD performance are presented. The results show that iterative noise cancelation method significantly enhances the mean square error performance. ACLR improvement of about 8 dB for realistic wideband code division multiplex signal is shown in a feedback predistortion technique for power amplifiers based on amplitude and phase signal processing [55]. The results are achieved by employing input and output signals of power amplifier and observing the changes of amplitude and phase caused by nonlinear PA.

Baseband Envelope Feedback technique requires the components in the feedback path to operate at a higher frequency band or large bandwidth situation. In order to maintain system stability, the loop bandwidth must be within the MHz range. Therefore, the main disadvantage of this system is the requirement of large bandwidth in the feedback path. .A digital predistortion technique based on envelope feedback was proposed for the linearization of power amplifiers [56].The proposed technique employed two envelope detectors, estimating the envelope of PA output and that of the difference signal between the PA output and PA input.

Polar Feedback technique overcomes the inability of Envelope Feedback to correct for AM-PM distortion effects. But different bandwidths are required for the amplitude and phase feedback paths, which leads to a different level of improvement of the AM-AM and AM-PM characteristics and a poorer overall performance than that is achievable from an equivalent Cartesian-loop transmitter.

Cartesian Feedback was first proposed by Petrovic. In this technique the in-phase (I) and quadrature-phase (Q) components modulate the carrier before passing it to a non-linear but efficient RF PA. Cartesian Feedback can automatically compensate for drifts in amplifier non-linearities due to temperature and power supply variations. However, this technique is only conditionally stable and the setting of the adjuster with the aim of maintaining stability is one of the key problems. Amplifier non-linearities also affect stability as does excessive base band phase shift. Another limiting factor in this system is the non-linearities of the down converting mixers. But the main disadvantage of this scheme is the narrow bandwidth that is somewhat inherent in baseband feedback systems.

An open loop digital predistortion using Cartesian Feedback for Adaptive RF Power Amplifier linearization was proposed [57].An analog Cartesian Feedback is used to train a Cartesian look-up table, reducing DSP and power amplifier modeling requirements to minimum. The Bandwidth limitation can be overcome using this technique as Cartesian Feedback system does not operate continuously.

Envelope Elimination and Restoration (EER) technique was first proposed by Kahn. The EER system can linearize the switched-mode RF PA without compromising its efficiency. But EER suffers from the slowness of the envelope restoration feedback loop. Practically, EER only has on the order of 20-30 dB of dynamic range. Even when the bias level to the PA is zero, some AC power bleeds through. The linearization of highly efficient monolithic SiGe Class E power amplifiers using Envelope Elimination and Restoration technique has been studied at 900 MHz [58].The PAs achieve power added efficiency of 66%.The third order Intermodulation coefficient(IMD3), measured using two tone excitation signal is an important measure of linearity of a power amplifier [59].A two tone excitation signal of 10 MHz is used in this simulation. Compared to the PA without EER, the output power of the PA with EER declines less than 1 dBm on average.

Adaptive Feedforward was invented as means of distortion reduction in telephone repeaters by Black. Feedforward linearization [60-69] can deliver reasonable linearization performance (20 dB-40 dB) improvement over relatively wide bandwidths (3 MHz-50 MHz) and has the advantage of inherent stability. Amplitude and phase matching is a problem since amplifier characteristics tend to drift with temperature and time, and also vary with manufacturing tolerances. Adaptive techniques can enable the performance of the system to be maintained despite these effects; therefore, a digital signal processing (DSP) processor has to be used at this point to implement the adaptive algorithm.

For the linearization of power amplifiers, feedforward method is one of the most well known and widely applied methods. The gradient descent type methods have been developed in order to prevent the degradation in performance due to in accuracy in implementation. An analysis on the convergence of feedforward linearizer coefficients

[70] and resulting reduction of intermodulation distortion is observed when gradient descent type adaptations are used with a PA that exhibits memory. For the modeling of a PA, Hammerstein model is used.

Although lot of researchers have worked on analog and RF predistortion techniques, but due to less flexibility they are not much popular these days. The most promising and cost-effective linearization technique is adaptive baseband Digital Predistortion (DPD), which has recently demonstrated notable success in correcting the non-linearity of RF PAs. Due to Predistorters digital implementation, it currently benefits from the continuous improvements of DSP and field-programmable gate array (FPGA) circuitry.

Work on Adaptation on Digital Predistortion technique based on intersymbol interpolation was published by [71] in 1995 in Global Telecommunications conference. The work is centered on the investigating the performance of several adaptation strategies of a digital predistortion technique which effectively compensates for PA nonlinearities in bandwidth efficient digital communication systems. The research includes the study of Gradient algorithms in Cartesian coordinates and the predistorter optimization. An algorithm called genetic algorithm was developed [72] that adapts a polynomial pre-distortion function to minimize adjacent channel emissions. The paper describes the implementation of this algorithm and its performance is compared with the random search technique. This technique could be applied as a low cost solution to meet adjacent channel leakage ratio requirements.

A non uniform simple spacing technique having performance similar to the optimum amplifier was developed by [73]. This technique offers negligible degradation relative to the optimum approach. Work on High Precision RF Linearizer by Digital Predistortion technique was presented by [74]. A new adaptive digital predistortion linearization technique is developed for transmitters of IMT-2000 band mobile communication base stations. The simulation results have shown an improvement of -63 dB for third order intermodulation distortion suppression. Digital predistortion techniques can be employed to improve the performance of power amplifiers [75].

The look table based adaptive predistortion approaches are suitable and cost effective for the power amplifier linearization in wireless applications. But the existing digital adaptive predistortion approaches are sub optimum because a uniformly spaced look up table is adopted regardless of the characteristics of power amplifier and input signal statistics. The proposed technique [76] is capable of adapting the look up table spacing for PAs with the various non linear characteristics.

Work done by [77] deals with piecewise linear approximation of inverse characteristics of PA for digital predistortion techniques. In this a new approach for the implementation of memoryless digital predistorter is proposed. It uses a look up table with few entries which correspond to large step segment of amplitude range of input signal. A complex circular model is used to linearize the AM/AM and AM/PM characteristics of every segment. A new wideband digital feedback predistortion technique for modulated signals with a signal of wide bandwidth by combining digital feedback predistortion linearization and memory effect compensation technique was developed by [78].

A novel digital predistortion correction technique involving the addition of digital signal processing in time domain was put forward by [79].The results show increase in quality and performance of UWB signals thereby satisfying the requirements of UWB radar systems. A Look Up Table based digital predistortion technique for high voltage power amplifier in ultrasonic applications was proposed [80].The proposed system is able to reduce the second order harmonics and considerable improvement in power amplifier efficiency is observed. Another novel and efficient approach to optimize the linearity and efficiency of power amplifier used in mobile WiMAX applications is proposed by [81]. Thus, DPD provides significant accuracy and flexibility, better power efficiency and reduced implementation complexity. The table below shows the comparison of different techniques.

Table 2.1: Comparison of different linearization techniques [2]

Technique	Bandwidth (MHz)	%age efficiency	Correction	Flexibility	Cost
In-line Predistortion	15 to 20 MHz	5 to 8 %	2 to 3 dB	Low	Very Low

Analog Predistortion	15 to 20 MHz	5 to 8 %	3 to 5 dB	Low	Low
Cross Cancellation	10 to 20 MHz	10 to 12 %	15 to 20 dB	Medium	Medium
Feedforward	25 to 60 MHz	6 to10 %	30 dB	Medium	High
Digital Predistortion	15 to 20 MHz	12 to 14 %	15 to 20 dB	High	Medium

From comparison of different linearization techniques, it can be concluded that DPD is most promising and cost-effective linearization technique. A comparison of the literature has been given in Table 2.2, which shows the advantages and limitations of some important research work related to DPD.

S/N	Year	Author	Work	Weakness	Strength
1	1989	Bosch et al	Measurement and simulation of memory effects in pre-distortion linearizers.	Test equipment (twonetwork analyzers) are needed for the measurement.	Memory effectshas been Considered.
2	1990	Cavers	Amplifier linearization using a DPD with fast adaptation and low memory requirements.	Few details provided.	Introduces the gain based predistorter concept.
3	1991	Faulkner et al	Automatic adjustment of quadrature modulators.	Not specific for adaptive predistortion.	First paper to analyze the Problem of linearization.
4	1992	Wright et al	Experimental performance of an adaptive digital	-Very narrowband (8 KHz).	Experimental performance of linearizer.

			linearized power amplifier.	-Few details provided.	
5	1992	Faulkner et al	Spectral sensitivity of power amplifiers to quadrature modulator misalignment.	Does not provide a solution to the linearization problem.	Provides good analysis and results can be applied to a linearized amplifier
6	1992	Stapleton et Al	An adaptive predistorter for a PA based on adjacent channel emissions.	-Narrowband (8 KHz) -11 dB distortion Cancellation.	Introduces the concept of periodic update.
7	1995	Sundstrom et Al	Effects of reconstruction filters in digital pre-distortion linearizers for RF PA	Does not provide a solution to the linearization problem.	First paper to analyze digital pre-distortion problem.
8	1996	Cavers	The effect of quadrature modulator and demodulator errors on adaptive digital predistorters.	The solution does notaccount for equal delays in the real and imaginary signal.	Paper Provides good insight to analyze in depth digital pre-distortion issue.
9	1996	Goeckler	Use adaptive digital pre-distortion to simulate a linearization system. RF design.	Only simulated results shown.	Presented simulation investigation.
10	1997	Ren et al	Improvement of digital mapping predistorters for linearising transmitters.	-Only simulation results -The mapping predistorter requires large memory size.	Simulative studies on digital mapping predistorter.
11	1997	Andreoli et al	Linearizing digital RF amplifier.	-10 dB improvement across 10 MHz. -Few details provided.	Presented DPD for linearizing PA.
12	1998	Jeckeln et al	An L Band Adaptive DPD for PA using Direct I-Q modem.	-20 dB distortion Cancellation. -Narrowband (only 60 KHz)	Presented adaptive DPD
13	1999	Zavosh et al	Digital pre-distortion techniques for RF PA with CDMA Applications.	20 dB improvement across 3 MHz.	Presented DPD for CDMA.
14	2000	Jeckeln et al	Adaptive baseband/RF pre-distorter for PA through instantaneous AM-AM and AM-PM characterization using digital receivers.	Not 100% digital approach.	Presented DPD using analog IQ.
15	2000	Zavosh et al	Digital pre-distortion Linearizes RF PA's	-Narrowband. -Not the combined effects evaluated.	Analyses performance of DPD versus table size and quantization level.

16	2001	Kenington et Al	A GSM – EDGE High PA utilising digital linearization.	-Not 100% digital -800 KHz bandwidth	Worked on GSM-EDGE pre-distorter.
17	2001	Sills	Application of digital pre-distortion to BWA.	Only 5 dB distortion cancellation.	Presented wideband DPD
18	2001	Vuolevi et al	Measurement techniques for characterizing memory effects in RF PA	Methods is complicated and impractical for a digital linearizer.	Analysed thermal and electrical memory effects.
19	2002	Davis, et al	A software linearization method for enhancing noise loaded performance of amplifiers.	-Non-linear device static gain measurement. -10 dB improvement across 1 MHz.	Presented software linearization method.
20	2003	Hammi et al	Baseband digital pre-distortion using subband filtering technique.	No technique was proposed to compensate memory effects Compensation of PA.	Filters used to compensate for temperature changes.
21	2003	Jeckeln et al	Method for modelling amplitude and bandwidth dependent distortion in nonlinear RF devices.	Requires sampling of the input of the PA.	Memory effects have been Considered.
22	2003	Boumaiza et Al	Thermal memory effects modeling and compensation in RF PA and pre-distortion Linearizers.	Not all major memory effects are considered.	Provides method to compensate for thermal memory effects.
23	2004	Ding et al	Digital baseband predistorter constructed using memory polynomials.	-Polynomial approach is hard to implement. -No other memory effects correction.	Uses filters to correct for frequency memory effects.
24	2004	Franco	Minimizing PA memory effects.	Provides few details on the digital solution.	-Identifies sources of memory effects and provides analog and digital methods for correction. -Memory effects has been considered.
25	2005	Varahman et al	Adaptive digital pre-distortion for power amplifiers used in CDMA applications.	The proposed method is complex to implement.	Proposed wideband adaptive DPD technique using the concept of combination of memory polynomial pre-distorter and slope-dependent

					method.
26	2005	ChihebReb ai et al	Embedded software implementation of an adaptive baseband pre-distorter	Too Complex to implement.	Hybrid DSP/ FPGA based architecture has been used.
27	2005	ChihebReb ai et al	FPGA building blocks for an hybrid base band digital pre-distorter suitable for 3G power\ amplifiers.	The attention is only focused on the FPGA part.	a hybrid design based on FPGA and DSP is proposed for linearization pre-distortion.
28	2006	Shanying Wu et al	The effect of D/A accuracy on the performance of digital pre-distortion for RF power amplifiers	No DPD Design has been proposed.	Only the D/A conversion accuracy on the performance of DPD for RF PA is tested.
29	2007	Albert Cesari et al	FPGA Based DPD for RF PA with memory effects.	DPD parameter extraction procedures has implemented in the design.	Prototyping digital pre-distortion linearizers, and a scalable DPD architecture is proposed.
30	2007	G. Montoro et al	An LMS-Based adaptive pre-distorter for cancelling nonlinear memory effects in RF PA	Adaptation algorithm has been implemented using Multi LUT design, which makes the design complex and slow.	Nonlinear auto regressive moving average (NARMA) model has been proposed for PA and DPD.
31	2007	Sungchul Hong et al	Weighted polynomial digital pre-distortion for low memory effect doherty PA.	The design can only compensate low memory effects.	A simple and effective weighted polynomial digital pre-distortion algorithm, which consists of weighting, least square polynomial fit, and de-weighting has been proposed.
32	2008	Jangheon et al	A new wideband employing feedback linearization.	But again the proposed technique is complex to implement.	Wideband adaptive digital pre distortion technique using feedback.
33	2008	Raithwaite R.N. and Santa, A.	Wide bandwidth adaptive digital pre-distortion of power amplifiers using reduced order memory correction.	The proposed technique was tested for 20 MHz signal only.	Adaptive DPD using reduced order memory correction. Eigen value decompositions were used to reduce the order of the memory coefficient estimation.
34	2008	Tarasov et al	PA digital pre-distortion - fixed or adaptive.	No improvement in Signal to noise ratio and bit error rate.	Presented simulative investigations on DPD.

35	2009	Wang et al	Research of adaptive digital pre-distortion technology for wideband OFDM PA	Only 6dB improvement was obtained	Also presented simulative investigations on DPD.
36	2010	Jiang H. and Wilford P.A.	Adaptive DPD using separable functions	The technique is complex to implement and PA modelling was not proposed.	Proposed adaptive DPD using separable function.
37	2010	Rawat et al	Adaptive Digital pre-distortion of wireless PA/ transmitters using dynamic real-valued focused time-delay line neural networks.	Only 20dB improvement was obtained.	Proposed adaptive DPD using neural networks.
38	2010	Amandeep Singh Sappal et al	A novel black box based behavioral model of power amplifier for WCDMA applications.	S-parameters vary with frequency of operation.	A novel behavioral model based on a Black Box modeling is presented.
39	2011	ZhengGao et al	A digital-feedback pre-distortion technique for integrated high-voltage ultrasound transmitting power amplifiers.	Proposed technique only used for integrated high-voltage ultrasound transmitting PA	A digital-feedback pre-distortion technique has been proposed.
40	2011	Bo Ai et al	Novel pre-distortion of PA with proposed fractional order memory polynomial.	Only 11dB improvement has been achieved.	Both odd and even order memory effects have been included.
41	2011	Amandeep Singh Sappal et al	Fast complex memory polynomial based adaptive DPD.	The proposed model is implemented only at 1950 MHz.	Performance of proposed technique is better than its earlier implementations
42	2012	Calogero D. et al	Closed-loop digital pre-distortion system with fast real-time adaptation applied to a handset WCDMA PA module.	Performance has only been demonstrated on a handset WCDMA PA module.	PA output power is increased by upto 30.9 dBm, and increases upto 48.5%.
43	2012	Hong Jiang et al	Digital pre-distortion Using stochastic conjugate gradient method.	An iterative stochastic conjugate gradient (SCG) method introduced an additional complexity.	The DPD is applied to a Universal mobile telecommunications system (UMTS) system.
44	2012	Haiying Cao et al	Digital pre-distortion for high efficiency PA architectures using a dual-input modeling approach.	Only 7 dB improvement has been achieved	-Dual-input high efficiency PA architecture is proposed. -Dual-input approach helped to achieve maximized average power-added efficiency

					and minimized output distortion simultaneously.
45	2012	Lei Guan and Anding Zhu.	Optimized low-complexity implementation of least squares based model extraction for digital pre-distortion of RF PA	Only second order dynamic deviation reduction based model is used.	-1-bit ridge regression algorithm has been used to eliminate the ill-conditioning problem in the LS estimation. -The execution time of the algorithm is also much shorter than that using the standard LS employing a long sequence.
46	2012	Meenakshi Rawat and Fadhel M. Ghannouch i.	Distributed spatiotemporal neural network for nonlinear dynamic transmitter modeling and adaptive digital predistortion.	The delay caused by the DUT and propagation medium to ensure to an accurate behaviour prediction.	adaptive neural network approach for the behavioralmodeling of wireless transmitters has been proposed.
47	2012	Charles et al	Peak-power Controlling technique for enhancing digital pre-distortion of RF PA.	Only a class AB PA driven by wideband code division multiple access and WiMAX signals has been used.	The method can be considered as a joint CFR reduction and DPD.
48	2013	Jiwoo Kim et al	A Generalized Architecture for the Frequency-Selective Digital pre-distortion Linearization Technique.	But linearization has been done upto 3^{rd} IMD only.	Three LTE Signals of 5MHz bandwidth are used for input.
49	2013	Bradley Dean Lakiand Cornelis Jan Kikkert	Adaptive digital pre-distortion for wideband high crest factor applications based on the WACP optimization objective: an extended analysis.	This technique has resulted in high implementation complexity.	-The concept of weighting function taper was introduced and applied across the optimization schedules. -On air adaption phase is presented.
50	2013	Meenakshi Rawat et al	Three-layered biased memory polynomial for dynamic modeling and pre-distortion of transmitters with memory.	The proposed model could lead to instability if not conditioned accurately.	Behavioral model is used for three layered biased memory polynomial.

2.3 Gaps in Present Study

The first practical implementation of a gain based Digital Predistorter was proposed by James Cavers [82]. Prior to this method, the majority of the Digital Predistorters were based on the mapping predistorter principle, in which each possible signal level was directly mapped to another output level. As the signal bandwidth gets wider, such as in WCDMA signals, PA begin to exhibit memory effects.

Many papers describe methods to measure memory effects in RF PAs [83]-[85]. Those methods require the use of test equipment, and as a result are not practical in a Digital Predistorter implementation.

The first technique for the correction of quadrature modulators was given by Faulker [86] but this technique can't be used for adaptive predistortion. Faulkner [87] studied the effects of quadrature modulator errors on output spectrum of an amplifier. The paper described better analysis of the linearization problem but it does not provided a solution to the linearization problem.

Adaptive Predistorter using complex spectral convolution was studied by [88].They have introduced the concept of continuous update but the solution was applicable for narrowband signals only. The concept of complex gain Predistorter was presented by [89] but again solution was applicable for the narrowband signals only. The compensation for errors in digital receivers was discussed by [90] but his idea was applicable to direct conversion receivers only. The effects of amplitude, phase and frequency errors in DC receivers were studied by [91] but a solution to the linearization problem has not been recommended. The effects of reconstruction filters in Predistorters were studied by [92] but again a solution to the linearization problem has not been recommended.

The errors in quadrature modulator and demodulator in adaptive Predistorters were studied by [93]. This paper gives good insight into the analysis, but the solution does not account for equal delays in the real and imaginary signals. Simulation investigations on a Digital Predistorter were presented by [94]. But the paper presents only simulated results. The errors in quadrature modulator and demodulators in adaptive Predistorters were presented by [95] which provide better solution than his previous paper. Simulative studies on a digital mapping predistorter were presented by [96] but the presented mapping Predistorter requires large memory size. Adaptive Digital Predistorter for linearizing a PA were studied by [97] but only 20 dB improvement was noticed and solution is only applicable to narrowband signals. Effects of demodulator errors on Predistorter were studied by [98] but a solution to the linearization problem has not been recommended. Digital Predistorter for CDMA was presented by [99] but only 20 dB improvements were obtained for a signal of 3 MHz only. Modulator compensation for CDMA was presented by [100] but it was just a review of the paper early presented by [95].

Although analysis on a Digital Predistorter for linearizing a PA was performed to make a comparison between the performance of the Digital Predistorter versus table size and quantization level independently, but the combined effects of table size and quantization level on the performance of Digital Predistorter was not evaluated. Also the proposed method can be applied to narrow band signals only.

Work on GSM-EDGE Predistorter was done by [101] but the proposed solution was not 100% digital and was narrowband (800 KHz). A wideband Digital Predistorter was presented by [102] but only 5 dB improvements were noticed. The complex gain Predistorter was discussed by [103] but it was only an overview paper. A Digital Predistorter was presented by [104] but only 10 dB improvements was noticed across a signal of 1MHz. Simulative investigations on Digital Predistorter using sub-band technique were presented by [105]. Although filters were also used to compensate for temperature changes, no technique was proposed to compensate memory effects of PA. Other papers [106-107], propose memory effects models, although they provide some insight into the causes of memory effects but are again not practical in a Digital Predistorter implementation,. In recent years, some proposals have been made to measure and correct for memory effects in RF PAs. Not all major memory effects are considered, and only two of the papers show experimental results with actual memory effects correction. Although filters were used to compensate memory effects but proposed polynomial approach is hard to implement.

Wideband adaptive DPD technique was proposed by [108] which uses the concept of combination of Memory Polynomial Predistorter and slope-dependent method. But the proposed method is complex to implement. Adaptive DPD using reduced order memory correction was poposed by [109]. Eigen value decompositions were used to reduce the order of the memory coefficient estimation. But the proposed technique was tested for 20MHz signal only. Simulative investigations on DPD without improvement in Signal to noise ratio and bit error rate were presented by [110]. Simulative investigations on DPD were presented by [111]. But only 6 dB improvement was obtained using the proposed technique. Adaptive DPD using separable functions was proposed by [112]. But the proposed technique is complex to implement and PA modelling was not proposed. Adaptive DPD using neural networks was proposed by [113]. But only 20 dB improvement was obtained using the proposed technique.

2.4 Motivation and Problem Formulation

Power amplifiers are essential components in communication systems and are inherently nonlinear. The broadening beyond the signal bandwidth is due to the nonlinearity in power amplifier and it interferes with adjacent channels. Stringent limits on the ACI are imposed by the regulatory bodies, and thus the extent of PA nonlinearity must be controlled.PA linearization is often necessary to suppress spectral regrowth and reduce bit error rate(BER).

From the extensive literature review on the topics that relate to linearization of PA, it has been observed that DPD linearization technique can be of main concern. The main factors that limit the performance of the Digital Pre-distorter are the non-linearity and the memory effects of the PA. Thus exact characterization of the non-linear behavior of the PA and also its memory effects is a key issue in the successful implementation of a Digital Pre-distorter. The methods covered by the literature on DPD linearization are mostly test-equipment based. Also very few papers include analysis and results utilizing higher order non-linearity and memory. Also an important issue in the design of a Digital Predistorter is the required level of performance for a given application, which has been ignored in most of the literature.

2.5 Objectives of the Thesis

Keeping in view the above mentioned aspects, the objectives of the research were formulated which are listed as below:

- To study the various predistortion techniques for the digital compensation of RF amplifiers in wideband wireless applications.
- To develop an improved linearization technique for compensating the non-linear effects of RF amplifiers based on the exploration of the joint algorithm-architecture design space.
- To functionally simulate the proposed technique for performance evaluation.
- To implement the proposed system using various EDA tools on FPGA platform.

CHAPTER 3

MODELING OF A POWER AMPLIFIER FOR DIGITAL PRE-DISTORTION APPLICATIONS USING SIMPLIFIED COMPLEX MEMORY POLYNOMIAL

The linearization task for nonlinear microwave power amplifiers can be divided into two parts: exact modeling of the power amplifier (PA) to be linearized and development of linearization technique. In this chapter, PA modeling using simplified complex memory polynomial has been presented.

3.1 Introduction

Orthogonal frequency division multiple access (OFDM) systems allow the transmission of high data rates over broadband radio channels without the need of powerful channel equalizer. By using special modulation schemes, an OFDM system does not require a channel estimator. Thus, OFDM systems are less complex as compared to a single carrier transmission system. But major disadvantage of OFDM signals is that they have very large peak-to-average power ratio (PAPR). This high PAPR derives the power amplifier (PA) into non linear region and hence causes inter-modulation distortion (IMD), which causes spreading of power both within the band and in the adjacent frequency bands. Various PA linearization techniques have been discussed in literature [114-115] to reduce IMD, while maintaining the efficiency of the PA. Linearization task cannot be accomplished successfully until the PA to be linearized, is exactly modeled. This chapter focuses on the discussion of the characteristics and modeling of a PA with thermal and memory effects. The proposed PA model has already used in the development of Low Complexity Look Up Table based Adaptive Digital Pre-distorter with Low Memory Requirements [116].

A brief introduction to the background of the problem various performance indices on which the performance of a PA had already been discussed in Chapter 1.We will be modeling a PA with thermal and memory effects and the results of the proposed model have been shown.

3.2 Performance Indices of a Power Amplifier

The modeling task of PA will not be accomplished until the parameters on which its performance can be evaluated are not known. The basic parameters of the PA like IMD Products, Adjacent Channel Leakage Ratio (ACLR), Efficiency and Error Vector Magnitude (EVM), which can be used to evaluate its performance have already been discussed.

An amplifier is said to be linear if its output voltage is simply a constant times the input voltage.

$$V_{out}(t) = G.V_{in}(t) \tag{3.1}$$

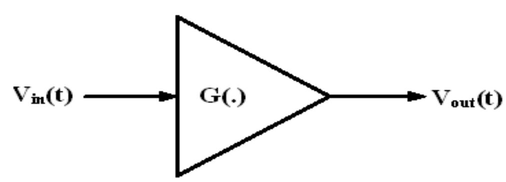

Fig. 3.1 Power Amplifier input-output diagram

But in reality they have non-linearities that make the output voltage a function of higher order terms of the input voltage. The output voltage $V_{out}(t)$ for such non-linear amplifier can be expressed mathematically as a Taylor series:

$$V_{out}(t) = G_1 V_{in}(t) + G_2 V_{in}^2(t) + G_3 V_{in}^3(t) \ldots\ldots + G_n V_{in}^n(t) \tag{3.2}$$

Thus, due to its non-linear characteristics, the PA is the main contributor for distortion products in a transmitter (T_x) chain. With Multi carrier modulation schemes like OFDM used in W-CDMA systems, even harmonic components could be filtered out, but 3rd and 5th order distortion component will exist with the fundamental channel. Thus equation 3.2 can be written as [117]:

$$V_{out}(t) = G_1 V_{in}(t) + G_3 V_{in}^3(t) + G_5 V_{in}^5(t) \ldots\ldots + G_{2n-1} V_{in}^{2n-1}(t) \quad (3.3)$$

If a two tone signal $V_{in}(t) = V_1 \cos \omega_1 t + V_2 \cos \omega_2 t$ is applied to the PA, then 3rd order term will produce $G_3 = (V_1 \cos \omega_1 t + V_2 \cos \omega_2 t)^3$, which is equal to

$$G_3 = (V_1^3 \cos^3 \omega_1 t + V_2^3 \cos^3 \omega_2 t + 3V_1^2 V_2 cos^2 \omega_1 t. \cos \omega_2 t + +3V_1 V_2^2 cos\omega_1 t. \cos^2 \omega_2 t) \quad (3.4)$$

Let us expand each term of equation 3.4 one by one:

No$V_1^3 \cos^3 \omega_1 t + V_1^3 \,(cos^2\omega_1 t. cos\omega_1 t) = V_1^3 \left[\left(\frac{1}{2} + \frac{1}{2} cos\omega_1 t\right)\right]$

$=\frac{1}{2} V_1^3 \cos \omega_1 t + \frac{1}{2} V_1^3 \,(cos2\omega_1 t. cos\omega_1 t)$

$=\frac{1}{2} V_1^3 \cos \omega_1 t + \frac{1}{4} V_1^3 \,[\cos\,(2\omega_1 t - \omega_1 t + \omega_1 t]$

$$=\frac{1}{2} V_1^3 \cos \omega_1 t + \frac{1}{4} V_1^3 \,[\cos(\omega_1 t) + \cos\,(3\omega_1 t)] = \frac{3}{4} V_1^3 \cos \omega_1 t + \frac{1}{2} V_1^3 \cos 3\omega_1 t \quad (3.5)$$

Similarly,$V_2^3 cos^3 \omega_2 t = \frac{3}{4} V_2^3 \cos \omega_2 t + \frac{1}{2} V_2^3 \cos 3\omega_2 t$ (3.6)

Also $3V_1^2 V_2 cos^2 \omega_1 t. cos\omega_2 t = (3V_1^2 V_2 cos\omega_2 t) \left(\frac{1+cos2\omega_1 t}{2}\right)$

$$= \frac{3}{2} V_1^2 V_2 . cos\omega_2 t + \frac{3}{2} V_1^2 V_2 [cos(2\omega_1 t - \omega_2 t) + cos(2\omega_1 t - \omega_2 t)] \quad (3.7)$$

Similarly,

$$3V_1 V_2^2 cos\omega_1 t \; cos^2\omega_2 t + \frac{3}{2} V_1 V_2^2 cos\omega_1 t [cos(2\omega_2 t - \omega_1 t) + cos(2\omega_2 t - \omega_1 t)] \quad (3.8)$$

From equations 3.5 to 3.8, we see that 3rd order term of equation 3.4 results in the frequency components $3f_1, 3f_2, 2f_1 - f_2, 2f_2 - f_1, 2f_1 + f_2, 2f_2 + f_1$ at the output of PA. The components $2f_2 - f_1$ and $2f_1 - f_2$ are in-band components which contributes for the in-band distortion at the output of PA.

3.3 Modeling of PA with Thermal and Memory effects

Any PA will show some dynamic deviations from its static characteristics. Such deviations have become to known as memory and thermal effects [118-119]. These effects are very troublesome for the process of pre-distortion. Thus, a PA model design simulations should be able to predict such memory and thermal effects and the designer must be bothered to include sufficient details of the bias circuitry, as well as the RF circuit, in the simulation file.

The Memory Polynomial model given by equation (3.9) can be used to incorporate both memory and thermal effects. The Memory Polynomial consists of several delay taps and non-linear static functions. This model is a truncation of the general Volterra series, which consists of only the diagonal terms in the Volterra kernels. Thus, the number of parameters is significantly reduced compared to general Volterra series. The Memory Polynomial model can be described as:

$$y(n) = \sum_{q=0}^{Q} \sum_{k=1}^{K} c_{2k-1,q} |x(n-q)|^{2(k-1)} x(n-q), \tag{3.9}$$

where

$x(n)$ is the input complex base band signal.

$y(n)$ is the output complex base band signal.

c_{kq} are complex valued parameters.

$Q$is the memory depth.

$K$is the order of the polynomial.

The even order terms are usually outside of the operational bandwidth of the signal and can be easily filtered out. This model considers polynomials with orders up to $2k \cdots - 1$, where K is a design parameter. For simplicity of implementation, the equation 3.9 can be rewritten as follows:

$$y(n) = \sum_{q=0}^{Q} F_q(n-q)$$

$$= F_0(n) + F_1(n-1) + F_2(n-2) + \cdots F_Q(n-Q) \tag{3.10}$$

where $F_q(n) = \sum_{k=1}^{k} c_{2k-1,q} |x(n)|^{2(k-1)} x(n)$

In its expanded form, equation 3.10 can be written as:

$$F_q(n) = c_{1,q} x(n) + c_{3,q} |x(n)|^2 x(n) + c_{5,q} |x(n)|^4 x(n) + \cdots c_{2K-1,q} |x(n)|^{2(k-1)} x(n) \tag{3.11}$$

Equation 3.11 can be implemented as shown below:

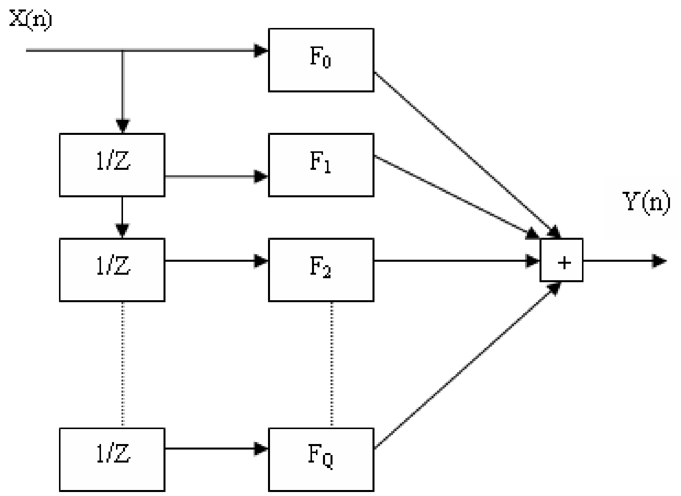

Fig 3.2 Implementation of Memory Polynomial given by equation 3.11

In modern communications applications, the actual modulation system is in use for testing a PA model. Obviously, such testing is essential, both during development and in production, for determining specification compliance on a product. But due to its simplicity in implementation, the two-carrier or two-tone test is still a convenient method for testing. So, in the present work the modeled PA has been tested using two carriers placed at 1.950 GHz and 1.955 GHz. The memory depth, $q = 3$ and polynomial of the 5th degree, $k = 3$ has been used. The input power has been varied from 0 to 20 dB. The set up shown in Figure 3.3 has been used for measurements. Due its ability to find unique global optimum point and computational easiness, the LSE technique [120] is used to find the coefficients.

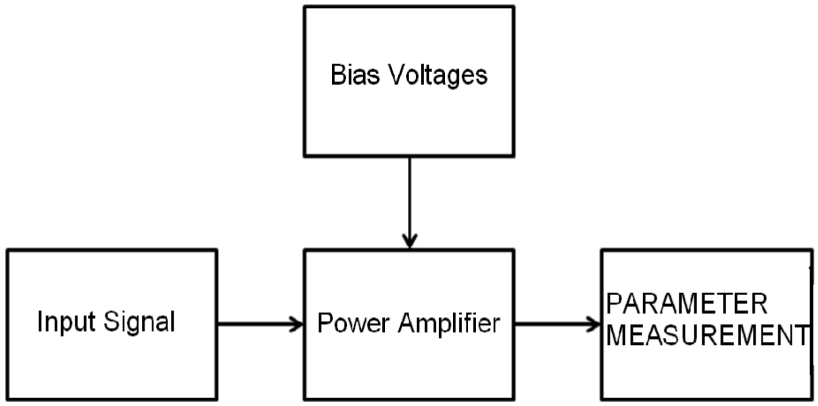

Fig. 3.3 Parameter measurement of PA

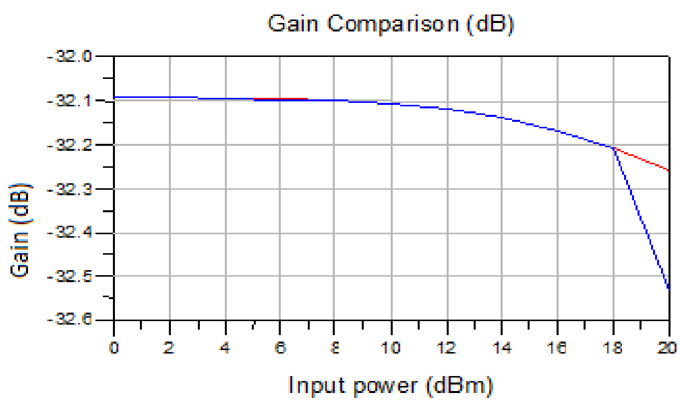

Fig 3.4 Comparison of Gain Compression of actual and modeled PA

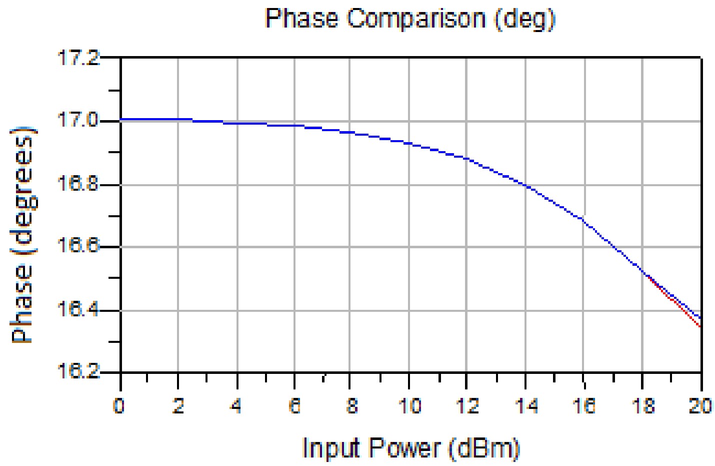

Fig. 3.5 Comparison of Phase Characteristics of actual and modeled PA

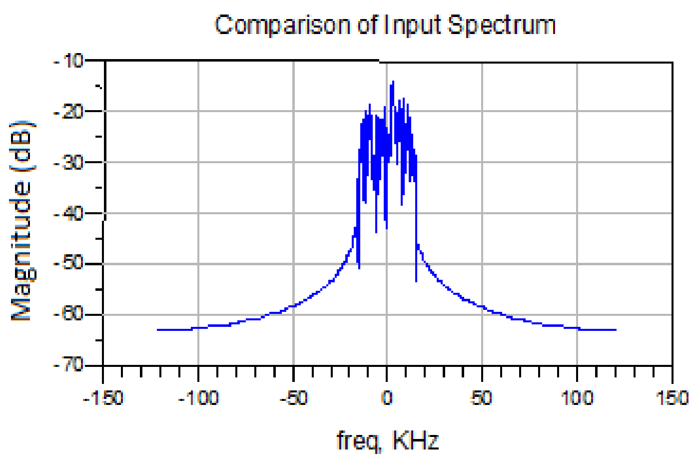

Fig. 3.6 Comparison of Input Characteristics of actual and modeled PA

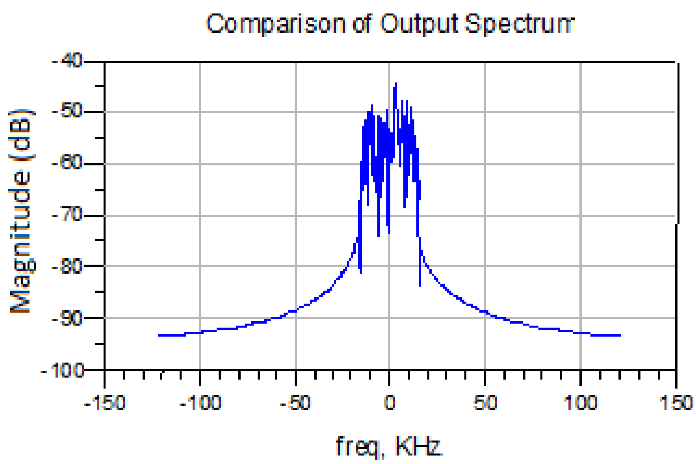

Fig. 3.7 Comparison of Output Characteristics of actual and modeled PA

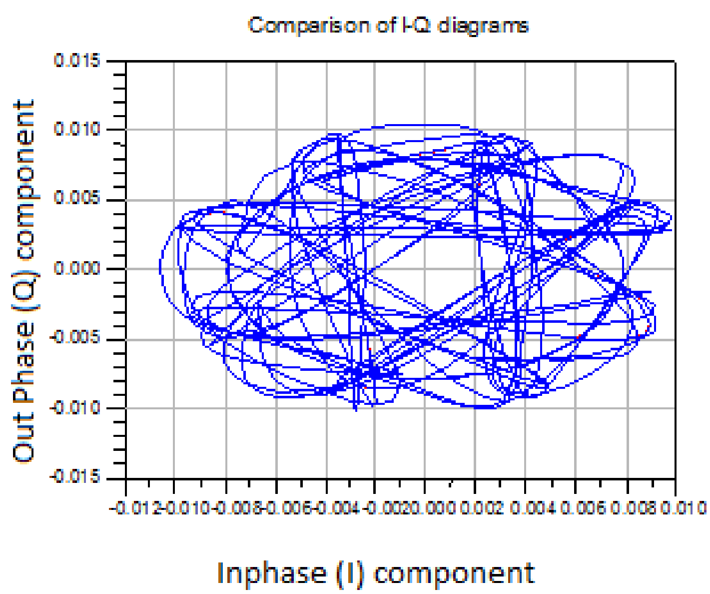

Fig. 3.8 Comparison of constellations of actual and modeled PA

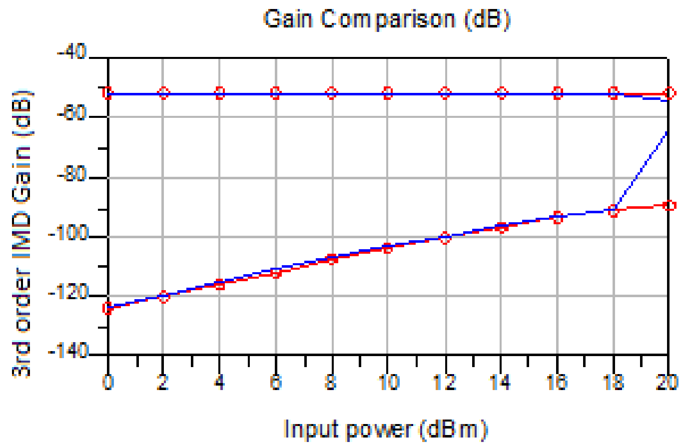

Fig. 3.9 3rd order IMD Gain reconstruction of actual and modeled PA

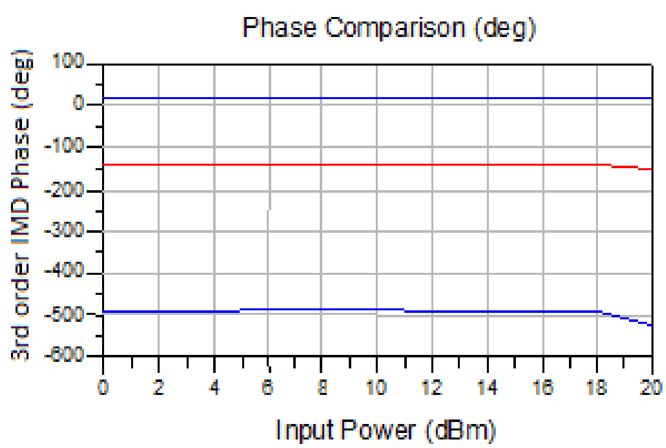

Fig. 3.10 3rd order IMD Phase reconstruction of actual and modeled PA

Figure 3.4 to 3.10 shows the comparison of Gain Compression, Phase Characteristics, Input Characteristics, Output Characteristics and constellations, 3rd order IMD Gain reconstruction and 3rd order IMD Phase reconstruction of actual and modeled PA respectively. From these figures, we could see that our modeled PA shows characteristics which are very much similar to the characteristics of the actual PA. Measurements also show that Lower channel ACLR for the actual amplifier is -16.627 whereas for modeled amplifier, its value as -16.626. Similarly upper channel ACLR for the actual amplifier has been calculated as -19.175 whereas for modeled amplifier, its value as -19.176. Also value of EVM has been calculated as 8.13% and 8.124% for the actual and modeled amplifier respectively. Thus actual amplifier and the modeled amplifier show almost similar characteristics, which show the validity of the modeled amplifier.

3.4 Conclusion

For accurate design and implementation of any PA linearization technique, exact modeling of PA is an important issue. In this chapter, modeling of PA using complex memory polynomial has been presented. The proposed memory polynomial is simple from implementation point of view. For the validity of the proposed PA model, its various characteristic like Gain Compression, Phase Characteristics, Input Characteristics, Output Characteristics and constellations, 3rd order IMD Gain reconstruction, 3rd order IMD Phase reconstruction, ACLR and EVM have been compared with the actual PA.

CHAPTER 4

DIGITAL PREDISTORTION IN WCDMA POWER AMPLIFIER USING EMBEDDED PROCESSOR

4.1 Introduction:

Linearity of the RF PA has become a critical design issue for non-constant envelope digital modulation schemes. This issue is particularly significant in spread spectrum applications such as CDMA and wideband CDMA (W-CDMA) base stations where the peak-to-average ratio of modulated RF signals can vary over a range of 3 to 12 dB [121]-[123]. Amplification of multicarrier (multichannel) signals requires adequate amplifier linearity in order to avoid significant cross modulation. Additionally, for bandwidth-efficient modulation the amplifier nonlinearity can produce substantial signal distortion and, hence, increased bit error rates (BER) [124]-[127].

Linearity is achieved, in part, through the use of more linear amplifiers such as class A amplifiers, and by operating the amplifier backed off from the saturation range so that the signal level is confined to the linear region of the amplifier characteristics. However, this approach results in low DC-to-RF conversion efficiency, which is particularly costly in base station applications. Furthermore, low DC to RF conversion necessitates high current operating points, resulting in undesired thermal effects.

To reduce the nonlinearity, the power amplifier can be backed off to operate within the linear portion of its operating curve. This means that the power amplifier needs to be backed off far from its saturation point, which results in very low efficiencies, typically less than 10% i.e., more than 90% of the dc power is lost and turns into heat. Considering the large number of wireless base stations deployed worldwide, improved power amplifier efficiency can substantially reduce the electricity and cooling costs incurred to the service providers. To improve the power amplifier efficiency without compromising its linearity, power amplifier linearization is essential.

In this chapter, an efficient method is applied for the linearization of power amplifiers (PAs) generally used in wireless base stations. PAs in the field today are predominately linearized by some form of feed forward technology, a concept originally proposed by [128].

A variety of other PA models exist and it is difficult to judge which PA model is the best, since it could depend on the type of the PA, the data format being transmitted, etc [129]. The organization of the chapter is as follows: The digital pre-distortion techniques are described in section 4.2. The DPD function implementation and various algorithms are discussed in section 4.3.The adaptive coefficient updating using NIOS embedded processor is discussed in detail in section 4.4. The various simulation and implementation results are summarized in sections 4.5 and 4.6. Finally, the conclusion of work is described in section 4.7.

4.2. Digital Predistortion

Digital predistortion is one of the most cost effective which can potentially compensate for PA nonlinearities. It adds a digital predistorter in the baseband to create an expanding nonlinearity that is complementary to the compressing characteristic of the power amplifier. Ideally, the cascade of the predistorter and the power amplifier becomes linear and the original input is amplified by a constant gain. With the predistorter, the power amplifier can be utilized up to its saturation point while still maintaining a good linearity, thereby significantly increasing its efficiency [130]. The simplified schematic of an adaptive digital Predistortion system is shown in figure 4.1.

A fully adaptive digital predistortion system requires the addition of a Predistortion circuit consisting of a digital predistorter and look-up table (LUT) to the transmission path in addition to a feedback path consisting of a demodulator, analog-to-digital converter (ADC) and adaptation circuit for updating the LUT. Most common implementations of digital predistortion utilize standard DSPs. Such processes typically operate with a word length of 16 or 32 bits, which provides sufficient accuracy for most applications. In specific applications, application-specific ICs (ASIC) are designed to implement the predistorter system, providing flexibility in controlling word length and power consumption.

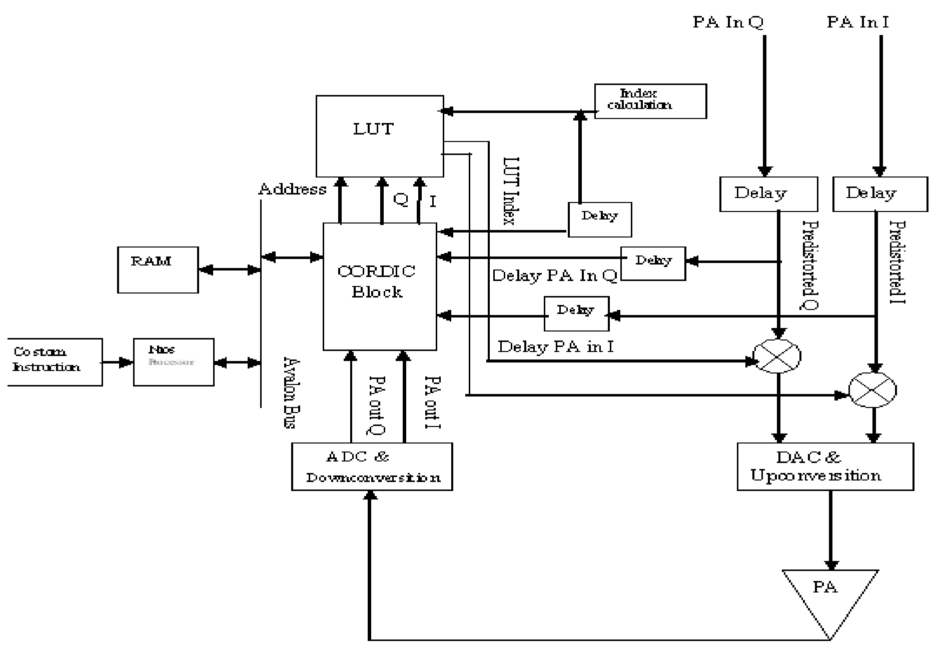

Fig. 4.1: Adaptive Digital Predistortion System

A system level block diagram of a digital Predistortion system is shown in figure 4.1. The digital predistortion system is part of the transmitter and requires the addition of several components to the standard transmitter components. The primary components augmenting the modulator, upconverter, and the power amplifier are Complex Gain Adjust block in the baseband signal path, Look-Up Table for storing a set of complex coefficients, Demodulator and downconverter whose input is drawn from the output of the power amplifier, adaptation algorithm processing block that calculates the complex coefficients that are stored in the lookup table. With the inclusion of digital predistortion, the digital complex baseband input signal samples are multiplied prior to the DAC by complex coefficients drawn from the look-up table. The look-up table coefficients implement the predistortion function, the adaptation algorithm determines the values of the coefficients by comparing the feedback signal and a delayed version of the input signal.

While checking the output, the main parameters to be taken into account are sampling rate used in the Digital Signal processing (DSP) portions of the transmitter, number of bits used for quantization by DAC and ADC, shape of the reconstruction filter at the output of the DAC, bandwidth of the low pass filter at the input of the ADC, stability of the feedback loop/adaptation algorithm, precision (fixed-point) of the look-up table and DSP processing, if any, memory available for the look-up table and adaptation algorithm, complexity of the adaptation algorithm and the amount of DSP horsepower required and accuracy of determination of the feedback delay. The adaptation of the memory polynomial model is done by feeding a set of primary signals into a group of parallel linear FIR filters, which implement the convolution for each input separately, and a least mean squares (LMS) algorithm is used in the adaptive linear FIR filters for its good convergence characteristic.

Let us consider a nonlinear system with memory effects with input $Z(n)$ and output $\mathrm{Y}(n)$

$$Y(n) = \sum_{q=0}^{q=Q} \sum_{k=0}^{k=k} C_{(2q+1)k} \times Z(n-q)|Z(n-q)|^{2k} \quad (4.1)$$

Where k = number of Volterra coefficients and q = size of memory

If q=0, the system described by equation (4.1) reduces to a memoryless nonlinear system. Because all practical PA’s shows memory effects, so it is best choice for us to use nonlinear system described by equation (4.1). The function of the adaptation algorithm is to derive the pre-distortion function i.e. the inverse characteristic of the amplifier response. The inverse of the nonlinear amplifier is adaptively tracked using a stochastic gradient method. Because the power amplifier characteristics vary slowly as a function of time, so despite slow convergence rate of LMS adaptive filters, the LMS approach is a reasonable choice for performing parameter tracking. At each iteration of the stochastic gradient algorithm, an update for the unknown vector is obtained from

$$W_{n+1} = W_n + \mu \times e_n \times Z_n \quad (4.2)$$

Where the error vector is defined as

$$e_n = z(n) - W_n \times Z_n \quad (4.3)$$

Here Z_n represents a vector containing all nonlinear products of input samples and is expressed as

$$Z_n \triangleq \begin{bmatrix} y(n) \\ y(n) \times |y(n)|^2 \\ y(n) \times |y(n)|^4 \\ y(n-1) \\ y(n-1) \times |y(n-1)|^2 \\ y(n-1) \times |y(n-1)|^4 \\ y(n-2) \\ y(n-2) \times |y(n-1)|^2 \\ y(n-2) \times |y(n-1)|^4 \end{bmatrix} \quad (4.4)$$

The DPD linearizer is implemented using equation (4.5). The equation is slightly modified version of equation (4.1).

$$Y(n) = \sum_{q=0}^{q=Q} \sum_{k=0}^{k=k} C_{(2q+1)k} \times Z(n-q)^{2k+1} \quad (4.5)$$

In our design we have selected K=2 i.e. only 1st, 3rd and 5th terms are considered and Q=2 i.e. only previous two inputs are considered. In its expanded form equation 4.6 can be written as

$$Y(n) = C_{10}Z(n) + C_{11}Z(n-1) + C_{12}Z(n-2) + C_{30}Z(n)^3 + C_{31}Z(n-1)^3 C_{32}Z(n-2)^3 C_{30}Z(n)^3 + C_{31}Z(n-1)^3 + C_{32}Z(n-2)^3 \quad (4.6)$$

Equation 4.6 is implemented using the minimum multiplier direct-form realization as shown

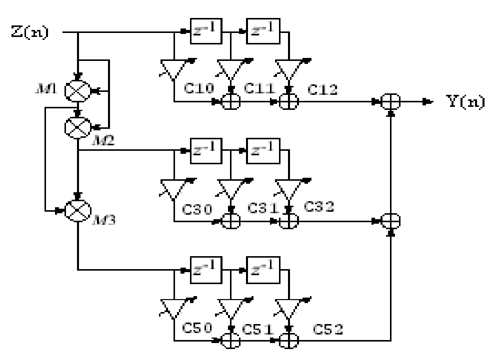

Fig. 4.2: Direct form realization of the modified memory polynomial

4.3 Implementation of Predistortion Function:

The Predistortion function, F, is assumed to be a function of the magnitude of the input signal [131]-[135]. The Predistortion function is implemented using a complex multiplier, a Look-Up Table and an Address Generation block that selects the appropriate coefficient from the look-up table, given the magnitude of the input signal as shown in figure 4.3. The coefficients stored in the look-up table are the value of the predistortion function at certain input signal magnitudes. Thus, the predistortion function is not implemented in an analytic manner; rather, it is only calculated at a specified number of points. The size of the lookup table employed determines the number of points at which the predistortion function is calculated. In addition, the distribution of the predistortion function points need not necessarily be evenly distributed across the range of the input signal magnitude. Instead, it may be desirable to distribute the predistortion function points across the range of the input signal magnitude using a squared (power) or logarithmic relationship. Figure 4.4 illustrates how the predistortion function may be indexed in the look-up table by the magnitude of the input signal and by the phase of the input signal.

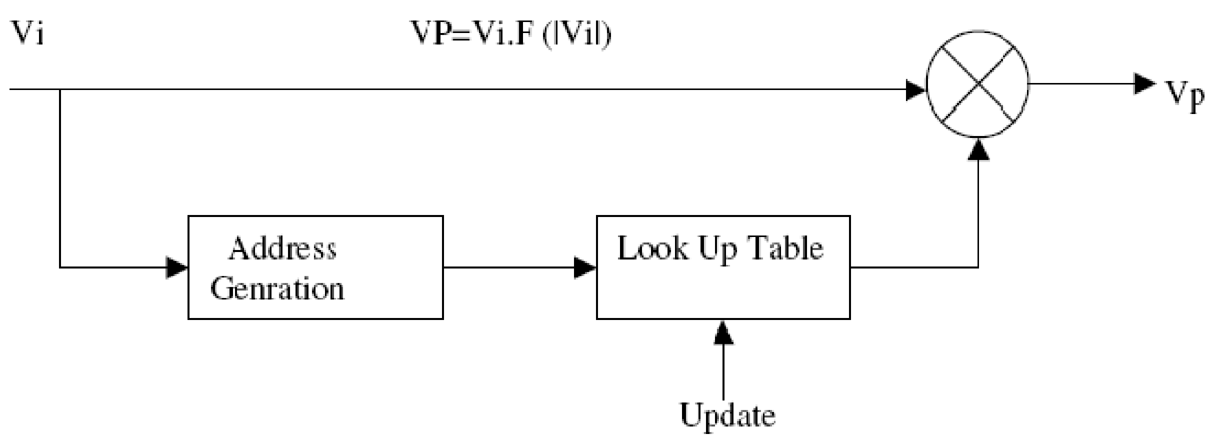

Fig. 4.3: Complex gain adjustment and Look up table

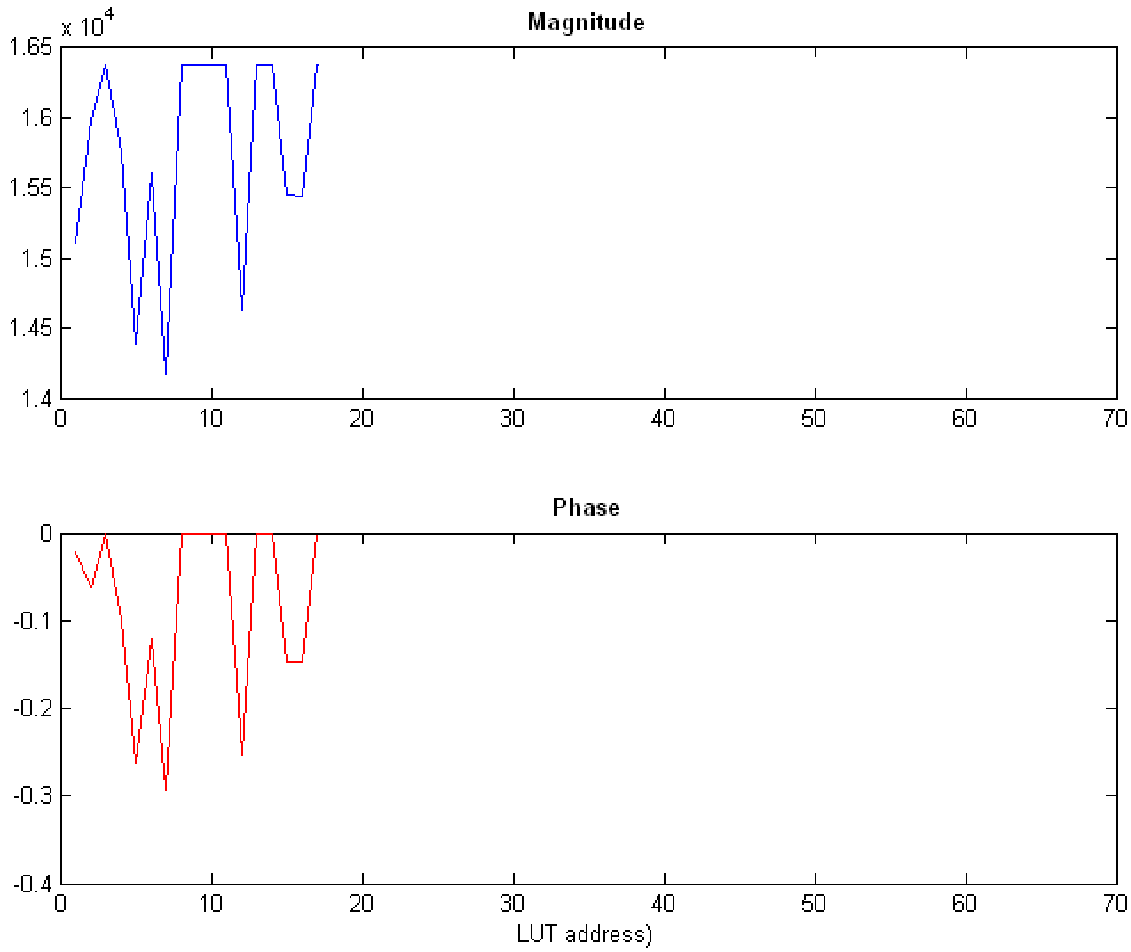

Fig. 4.4: Look-Up table entries for Magnitude and Phase indexing

4.3.1 Adaptation algorithm:

The function of the adaptation algorithm is to derive the predistortion function, F, i.e. the inverse characteristic of the amplifier response. The predistortion function may be derived using either a modulated signal input (random signal) or a known training signal input. The adaptation algorithm and its implementation are fundamentally different depending upon which type of input signal is utilized. The algorithms that are based upon the use of a modulated signal employ statistical signal processing and, typically, using curve fitting algorithm like least square approach ,to generate a smooth predistortion function. The complexity of the adaptation algorithm and its implementation can be significantly simplified, however, by using the alternative input signal.

4.3.2 Implementation

The algorithm is based upon the determination of the open loop gain, H, of the predistorter and amplifier combination at the power level associated with each look-up table entry.

$$1/H = V_i / / V_{fb} = G_{lin} / F(|V_i|).G(|V_p|) ===> 1+J*0 \quad (4.7)$$

The desired linear response of the predistorter and amplifier cascade requires that $F(|V_i|)$ $G(|V_p|) = k$ for all input. Hence, if G_{lin} is set to be equal to k, the desired open loop gain of the system is unity. If the calculated open loop gain is not equal to unity, the predistortion function must be adjusted in a manner to drive the open loop towards unity.

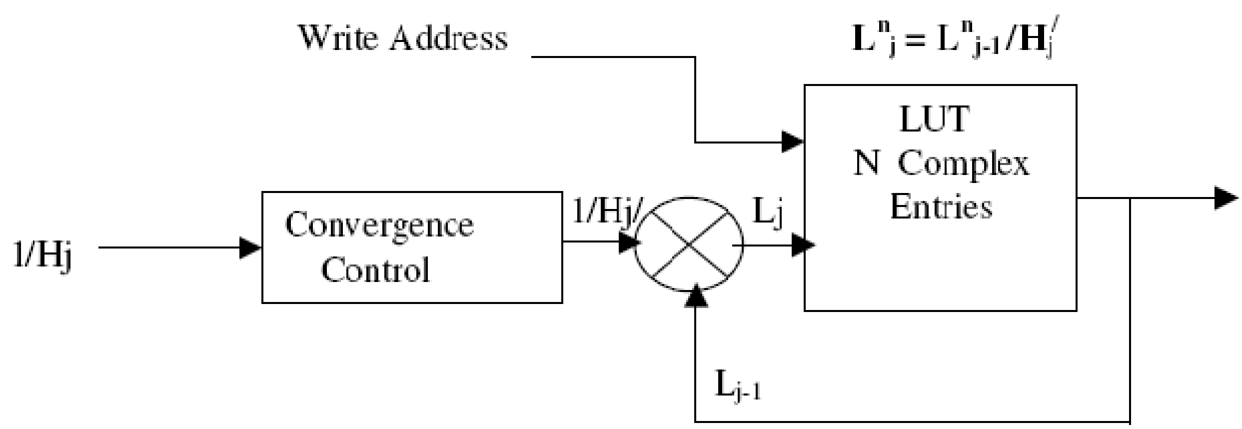

Fig. 4.5: Calculation of New Predistortion Function

The predistortion function is defined by a set of coefficients stored in the look-up table, L_n, where each n corresponds to an input signal magnitude which is mapped to a look-up table address. In order to drive the open loop gain to unity, the predistortion function coefficients are updated by dividing each coefficient by the calculated open loop gain (or by the calculated open loop gain adjusted to slow the rate at which the coefficients will change).

4.4. Adaptive Coefficient Update Using Embedded Processing:

In this application the PA characteristics do not change rapidly with time. The PA characteristics vary as a function of temperature drift and component aging, parameters that have long time-constants. To implement the DPD coefficient update, depending on system requirements, and in particular the required rate of coefficient adaptation, an FPGA embedded processor could be employed to realize the update. Figure 4.1 describes the algorithm for design. The incoming complex samples in I and Q, have correction factor applied from LUT and then send to RF module. The input power applied is able to derive the LUT addresses. For each real I and imaginary part Q, LUT must contain two values for each location. The samples are unconverted and then send to PA. The downconverted output of the PA is used to measure the error. The error is described by difference of input phase and magnitude and measure phase and magnitude. The delay block ensures that the input is compared to correct output value stored in LUT. The input data (in Cartesian form) is fed to index calculation block, which determines the index of LUT value. This LUT value modifies the delayed input data. The power of input data determines the LUT index. The input data is delayed by the delay block, prior to undergoing a complex multiplication with Cartesian LUT read from the LUT, to compensate for the delay through index calculation block and reading out of data from LUT.

4.4.1 Feedback processing:

CORDIC block: The CORDIC block is used to convert from polar to Cartesian and from Cartesian to polar. CORDIC is hardware efficient algorithm that allows trigonometric function to be perform using only shifts and adds as shown in figure 4.6.

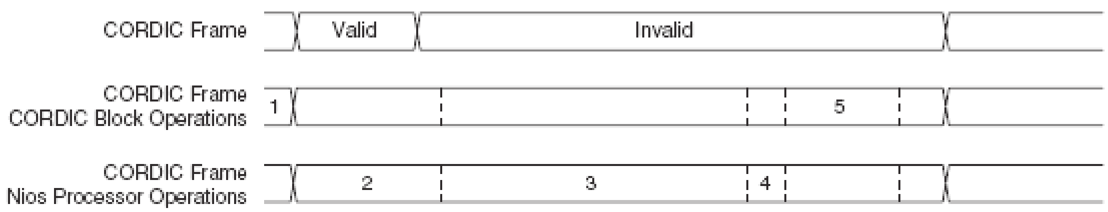

Figure 4.6: Synchronization between Nios and CORDIC

4.4.2 Nios Processor: The Nios Processor communicates with the CORDIC block via Avalon bus. The Nios Processor reads PA inputs and outputs in polar forms from CORDIC through Avalon. Using these values the Nios Processor implements the algorithm and calculates the new LUT value in polar form. Now these values are written in CORDIC by Nios Processor and converted to Cartesian form.

4.5. Simulation Results:

4.5.1 Linearity errors for Power Amplifiers (no predistortion)

Figure 4.8 gives an illustration of linearity errors in power amplifiers. Figure 4.1 is implemented in MATLAB. To make the work more interactive, a GUI has been developed in which various parameters of power amplifier, DPD algorithm, CORDIC algorithm etc. can be selected. By pressing the RUN button, figure 4.1 is simulated and outputs are generated which can be observed by opening the button provided in the action window of GUI.

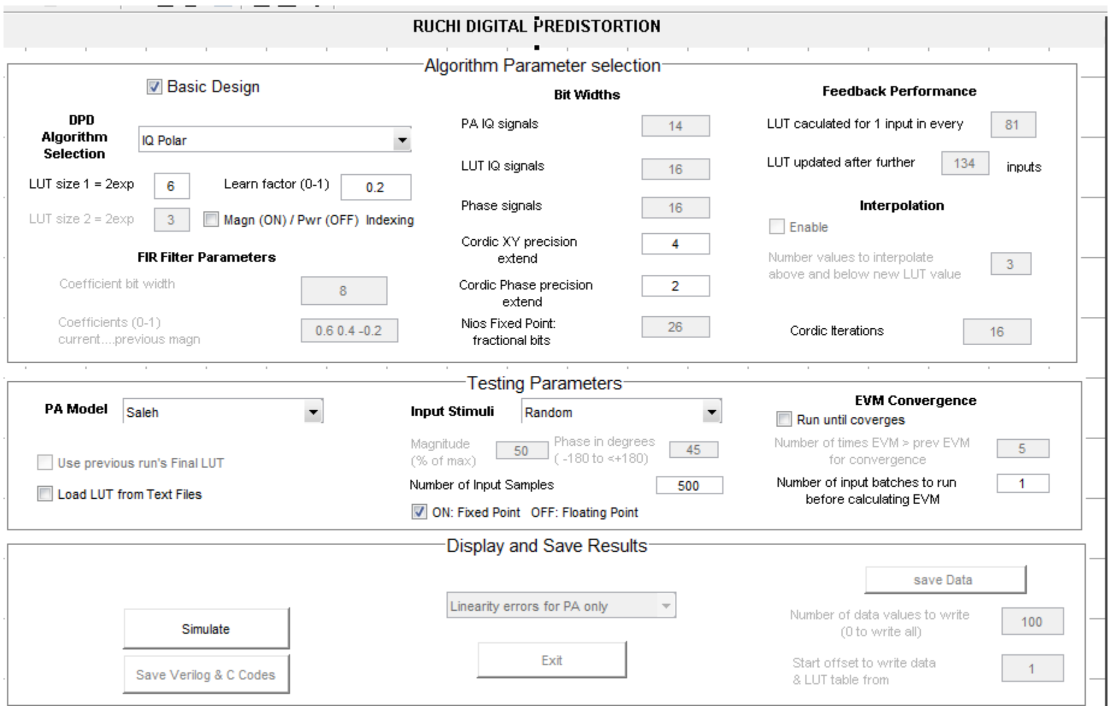

Fig. 4.7 Snapshot of GUI

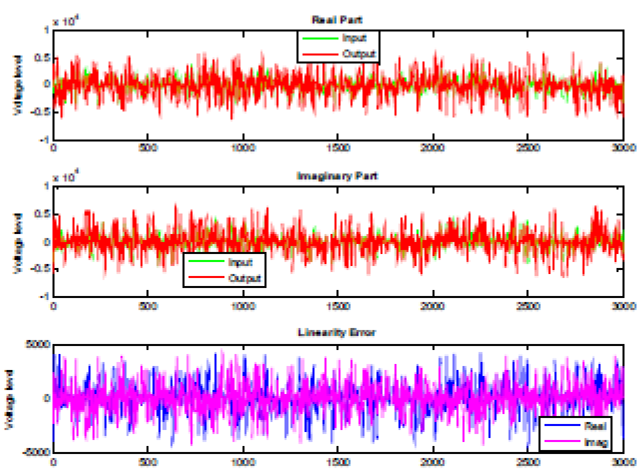

Fig. 4.8: Linearity errors for Power Amplifiers (no predistortion)

4.5.2 Normalized Linearity errors for Power Amplifiers (no predistortion)

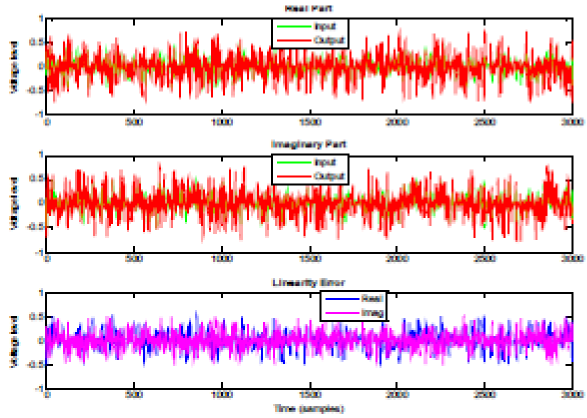

Fig. 4.9: Normalized Linearity errors for Power Amplifiers (no predistortion)

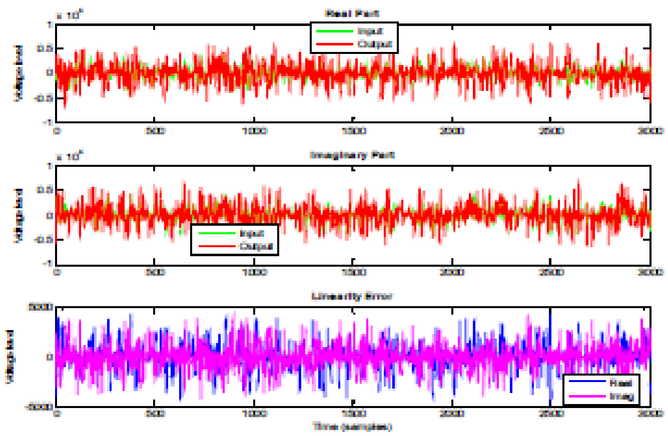

Fig. 4.10: Linearity Errors for DPD solution (Pre distorted)

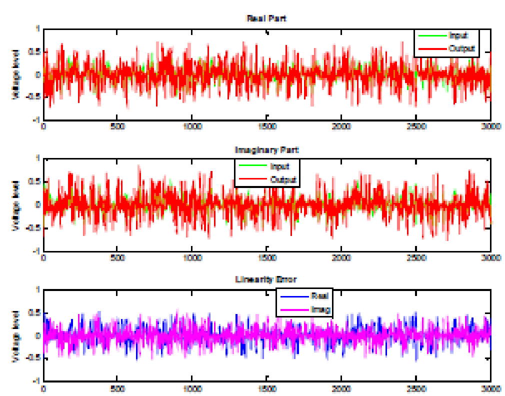

Fig. 4.11: Normalized Linearity Errors for DPD solution (Pre distorted)

Figures 4.9 to 4.11 gives results of the simulations performed in the MATLAB environment from the simulation set-up constructed for the DPD design.

4.5.3 Comparison of errors with Predistortion and without Predistortion

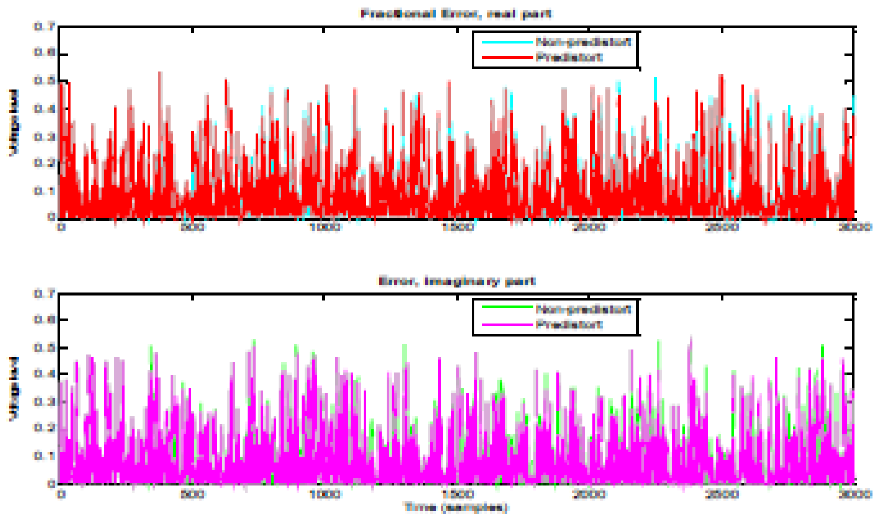

Fig. 4.12: Error comparison with and without Predistortion

4.5.4 Magnitude and phase errors for DPD solution

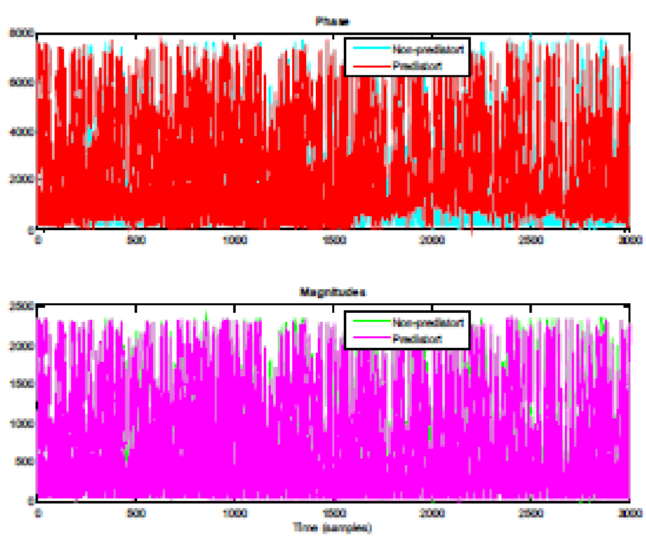

Fig. 4.13: Magnitude and phase error for the digital predistortion system

4.5.5 Signal Magnitude

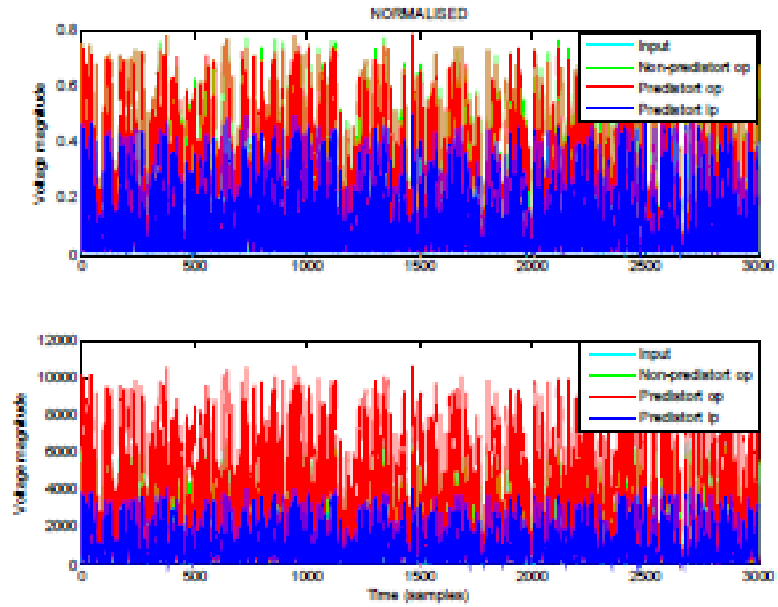

Fig. 4.14 Signal magnitude with and without predistortion

4.5.6 Frequency plots of Power amplifier input and output

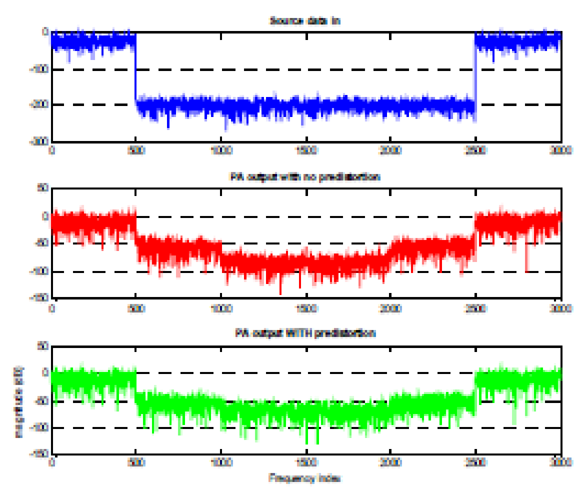

Fig. 4.15 Frequency plots of input signal and output signal with no predistortion and with predistortion to the Power Amplifier.

Figures 4.14 shows the input output simulation results with and without predistortion and figure 4.15 shows the power spectrum achieved in the overall DPD design. Figure 4.17

shows the effectiveness of DPD design in suppressing the spectral regrowth due to the non-linearity of the power amplifiers. The figure shows an overlay of the baseband signal spectrum, the PA output without linearization and with linearization.

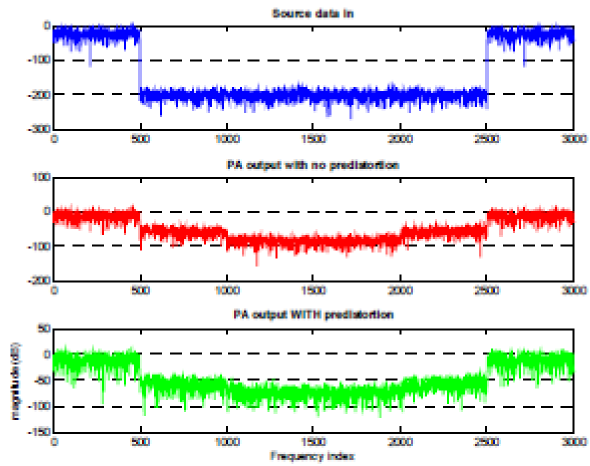

Fig. 4.16 Frequency plots of input signal and output signal with no predistortion and with predistortion to the power amplifier with increased filter

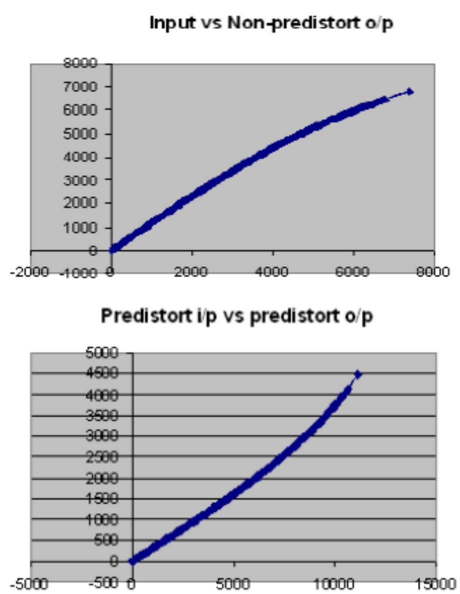

Fig. 4.17 Plot between input and non-predistorted output and plot between predistorted input and predistorted output

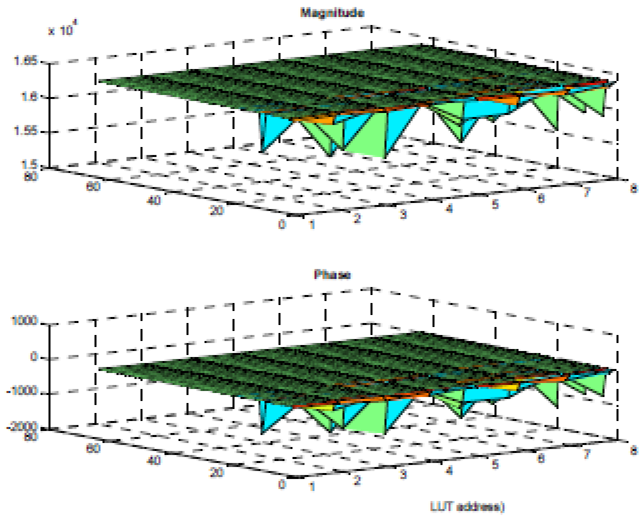

Fig. 4.18 LUT Content

The relationship between the input and the non predistorted ouput is shown in figure 4.17 and it clearly shows the non-linear region, whereas I the second plot in figure 4.17, it is clearly visible the amount of compensation applied. Figure 4.18 gives an idea of the magnitude and phase content of the look-up table.

4.5.7 Implementation Result

Figure 4.19 gives the ModelSim functional simulation results after the VHDL implementation of the DPD block. As we progress from the simple classical models to the more advanced variations of memory polynomials, the effectiveness of predistortion, in terms of reducing spectral regrowth, generally increases. However, there is an attending increase in computational complexity and this has to be weighed against other simpler expediencies, such as using a larger amplifier with increased backoff. Even for any particular scheme, performance will generally improve as the number of coefficients in the predistortion model is increased, at least up to some practical limit where a point of diminishing returns is reached. The design will lead to a more flexible and cost-effective solution because the hardware platform used is NIOS processor. An attempt has been made also to reduce the complexity of the design and at the same time maintaining the performance requirements intact. The overall power consumption of the hardware implementation is kept within limits by the proper use of low power design methodologies in the architectural implementation of the DPD model.

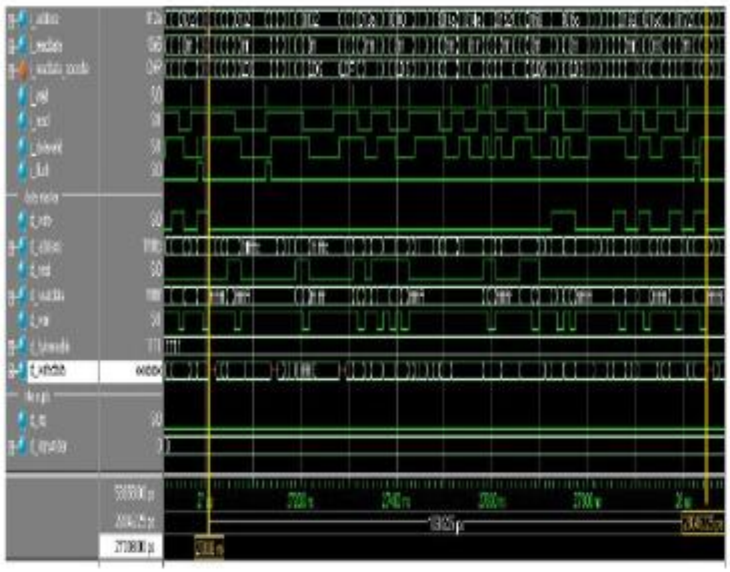

Fig. 4.19: Result of the Implemented system

4.6.7. Results:

EVM		Average EVM		EVM reduction over non_predistortion
Non predistorted	Predistorted	Non predistorted	Predistorted	
108.7263 %	102.6331 %	108.7263 %	102.6331 %	5.604%

Mean values over last 500 input samples:

ABS Phase error		ABS Magnitude error	
Non Predistorted	Predistorted	Non Predistorted	Predistorted
3421.21545	3077.30933	1049.64671	1007.00127

Reduction over non predistortion	
Mean ABS Phase error reduction	Mean ABS Magnitude error reduction
10.052%	4.063%

No Predistortion			
3rd Order		5th order	
Avg Sideband Magnitude(dB)	Max Sideband Magnitude(dB)	Avg Sideband Magnitude(dB)	Max Sideband Magnitude(dB)
57.124	-32.873	-84.975	-55.535

Predistortion			
3rd Order		5th order	
Avg Sideband Magnitude(dB)	Max Sideband Magnitude(dB)	Avg Sideband Magnitude(dB)	Max Sideband Magnitude(dB)
-55.360	-28.364	-70.594	-47.468

4.7. Conclusion

This chapter considered the design of digital predistortion systems to linearize power amplifiers with memory effects. By adding a digital predistorter in the baseband, the power amplifier is allowed to operate into its nonlinear region, thereby significantly increasing its efficiency. The efficiency gain translates into electricity and cooling cost savings for service providers and longer battery life for mobile terminal users. The challenge here is to address the memory effects exhibited by the higher power amplifiers or the power amplifiers for wideband signals. In addition, analog components in the transmitter have imperfections that need to be compensated as well.

CHAPTER 5

LOW COMPLEXITY LOOK UP TABLE BASED ADAPTIVE DIGITAL PREDISTORTER WITH LOW MEMORY REQUIREMENTS

5.1 Objective of the chapter

Linearization techniques for nonlinear microwave power amplifiers have gained significant interest with the advent of spectrally efficient wireless communication systems. In this paper a low complexity, direct-learning multilevel lookup table based adaptive digital predistortion technique has been proposed to linearize a power amplifier (PA). A loop delay compensation scheme has been used to achieve a significant reduction in convergence time and an improvement in linearization accuracy in the presence of an unknown loop delay. Compared with the conventional predistorters, the proposed technique show fast adaptation speed which enables the predistorter to track time-varying PA nonlinearities.

In the following sections, an adaptive digital pre-distortion (ADPD) technique is modeled and simulated. The aim of this chapter is to develop a linearization technique which is less complex and requires less memory from FPGA implementation point of view.

5.2 Basic approach used in Design of Adaptive Digital Predistorter

LUT based digital pre-distorters have low computational complexity, but they require significantly more memory space to store the model parameters than polynomial based digital pre-distorters. Thus LUT based digital pre-distorters have slow convergence speed as compared to polynomial based digital pre-distorters. Comparatively evaluation of a polynomial function is more computationally complex than a simple memory lookup table entry and compensation of higher orders of nonlinearity and memory effects requires a high order polynomial. For newer spectral efficiency transmission formats, a predistorter bandwidth of several tens of MHz might be required for implementation of these high order polynomials.

In this chapter a novel low complexity LUT based adaptive digital pre-distorter with reduced memory requirements has been proposed by using interpolation and efficient spacing of table entries, which leads to low LUT requirement. The proposed adaptive digital pre-distorter has much higher convergence rate as compared to other LUT based adaptive digital pre-distorters.

In the proposed design polar LUT based pre-distorter [136] has been used and is shown in Figure 5.1.

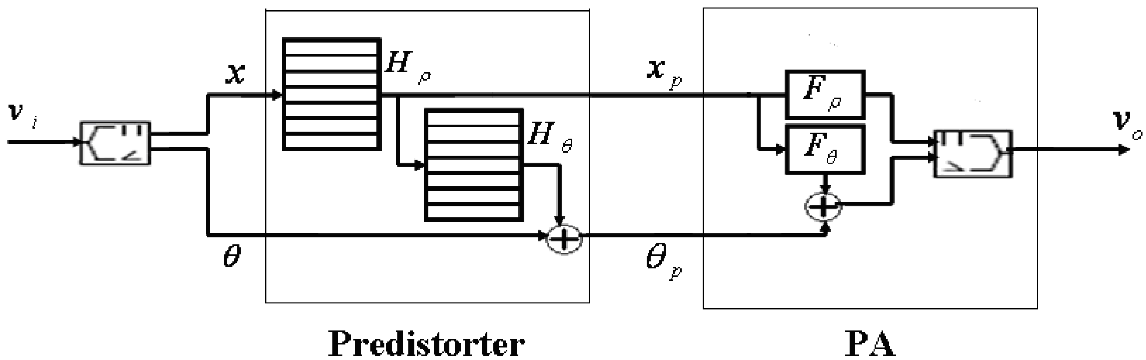

Fig 5.1 Polar LUT based Pre-distorter

Functions $f_p(x)$ and $f_\theta(x)$ denote the nonlinear AM-AM and AM-PM distortions of PA. The polar pre-distorter consists of two LUTs that approximate the inverse function of the PA's amplitude distortion $h_p(x) = f_p^{-1}(x)$ and the phase compensation function $h_\theta(x) = -f_\theta(x)$.

5.3 Proposed design

The block diagram shown in Figure 5.2 shows the simulation platform used in the proposed design.

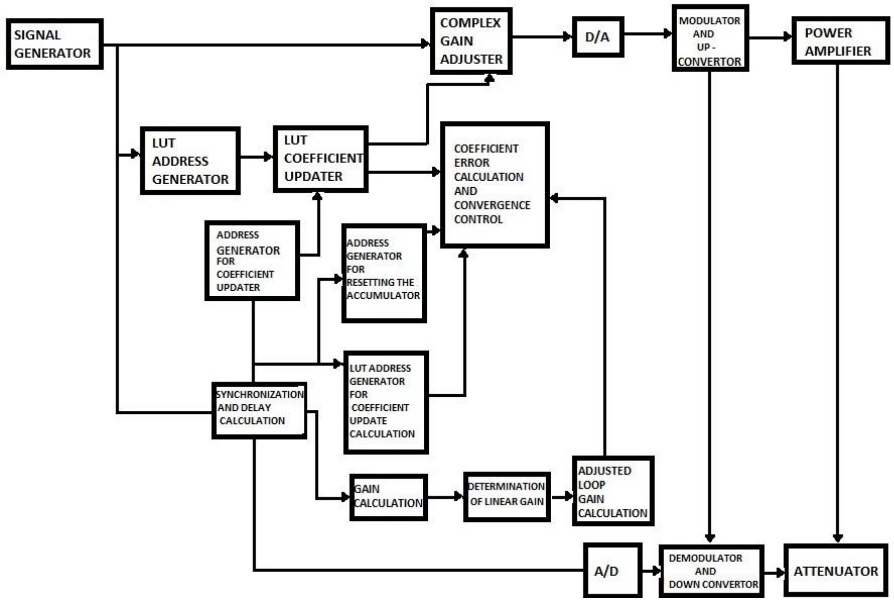

Fig. 5.2 Proposed LUT based Adaptive Digital Pre-distorter

This platform has been implemented using Agilent ADS software. Because, each block has been implemented at component level, so details of each component have not being given deliberately. The training signal is a tone having an increasing magnitude, i.e. a ramp. The tone is generated by uniformly increasing the amplitude of the in-phase and quadrature component of the complex baseband signal. The shape of training ramp is shown in Figure 5.3.

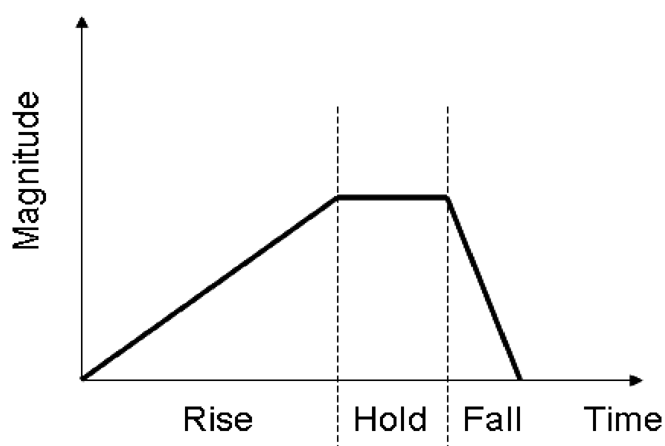

Fig. 5.3 Training Ramp

Only the rising portion of the ramp is used by the adaptation algorithm in the calculation of the predistortion function. The hold and trailing ramp portions are included to smooth the transition between power levels and minimize the impact of finite length filters on the latter part of the rising portion of the ramp.

The early implementations of LUT predistorters were mainly based on uniform spacing in power to reduce the complexity of the LUT address calculation. In the proposed implementation an optimal spacing scheme for LUT has been used. Optimal spacing has been achieved by applying a suitable compander to the amplitude signal prior to addressing the LUT. Here a set of companders are used that minimize the total residual nonlinear distortion power at the output of the PA. Thus combination of linear interpolation with optimal spacing results in higher improvements in PA non-linearity.

Here one thing that must be considered is that the compander is itself implemented as a LUT. It is therefore critical that the additional memory requirements do not offset the gains obtained from using optimal spacing. Thus the compander has to be implemented as a uniformly-spaced, linearly-interpolated LUT of size L, forming a piece-wise linear function. From simulation results it has been observed that as compared to uniform spacing the value of EVM improves by about 10 dB when the LUT size was kept as 256. Also it was observed that as the size of L has been increased beyond 256, no much improvement was found in EVM. Also when both uniform spaced and optimal spaced predistorters were implemented using VHDL synthesis, the gate count has shown almost two times reduction in case of optimally spaced compander.

The LUT address generation component translates the magnitude of the baseband input signal into a look-up table address using power addressing schemes. The look-up table is implemented using the LUT_RAM and the number of entries in the look-up table is taken as 256. The look-up table is initialized at the outset of the simulation using values read from a pair of text files. The text files have been stored in the data directory, so a path need not be provided along with the name of the text file. The accuracy of the open loop gain calculation dependents upon the accuracy of the estimation of the delay in the feedback path. The input signal must be delayed precisely by an amount equal to the delay in the feedback path. The delay in the feedback path is estimated by calculating the correlation between the magnitude of the input signal and the magnitude of the feedback signal. The use of the magnitude of the signals has the benefit of not requiring phase synchronization in the feedback path. Because the delay in the feedback path will not necessarily be equal to an integer number of DSP sample periods, interpolation is employed to more precisely align the input and feedback signals. Figure 5.4 shows the delay calculation between input signal and feedback signal.

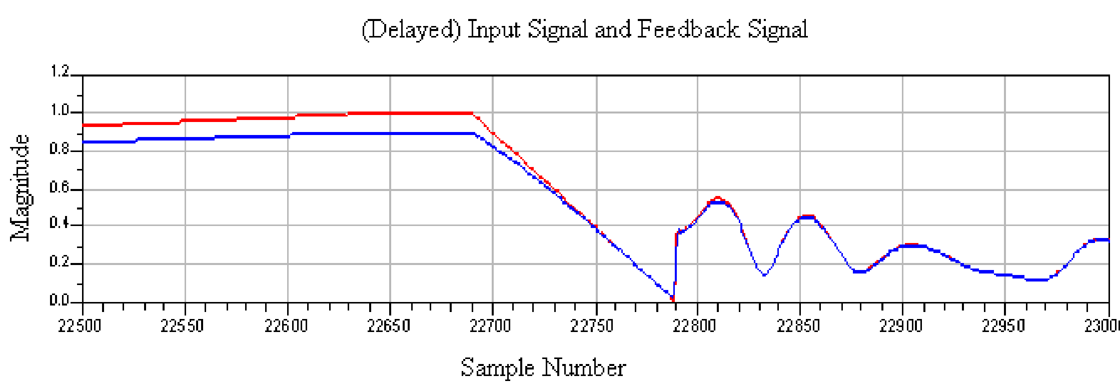

Fig. 5.4 Delay between input signal and feedback signal

The correlation between the input and feedback signal is performed on a modulated signal that precedes the training signal because the gain compression of the amplifier makes the accuracy of the correlation over the training signal suspect. In addition, because the envelope of the modulated signal will typically have a PDF such that it spends much of its time within the linear operating region of the amplifier, correlation using the modulated signal becomes more reliable. However, because the modulated signal is stochastic, the statistics of the modulated signal, as well as the size of the data block over which the correlation operation is performed will impact the accuracy of the delay estimation. In general, the accuracy of the estimation improves as the block size increases. Unfortunately, a larger block size requires more memory and takes longer to perform the estimate. The predistortion function cannot be exactly determined following the transmission of single training ramp and recalculation of the look-up table coefficients thereafter. A series of training ramps will have to be transmitted. Although significant improvement in the ACPR of the amplifier should be observed even after a single training ramp, yet simulations have shown that the pre-distortion function can converge to a solution after only six training ramps.

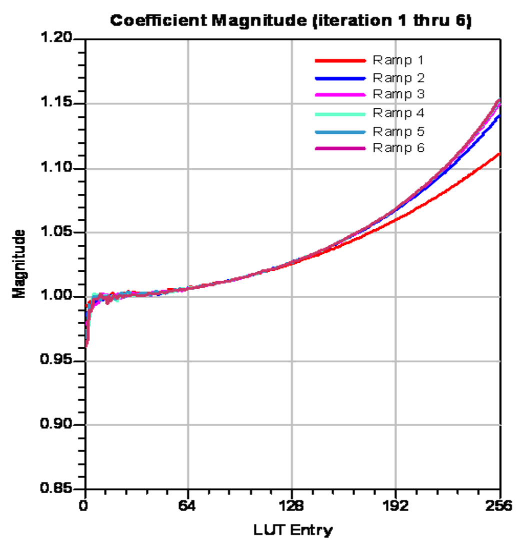

Fig. 5.5 Magnitude entries of LUT for different training ramps

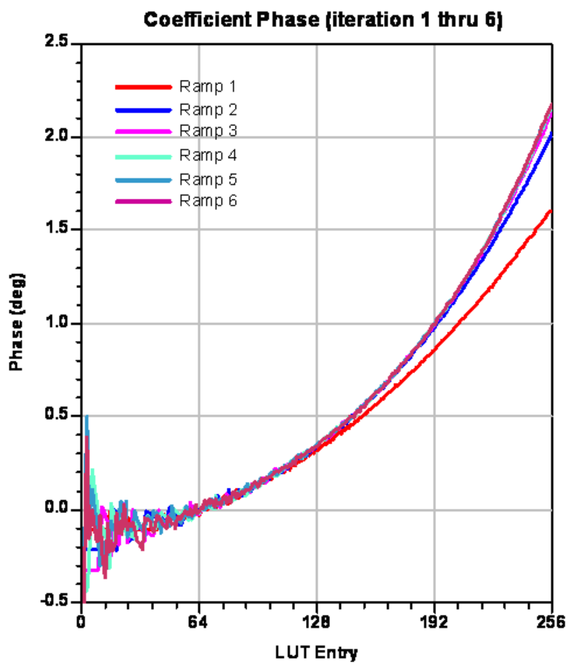

Fig. 5.6 Phase entries of LUT for different training ramps

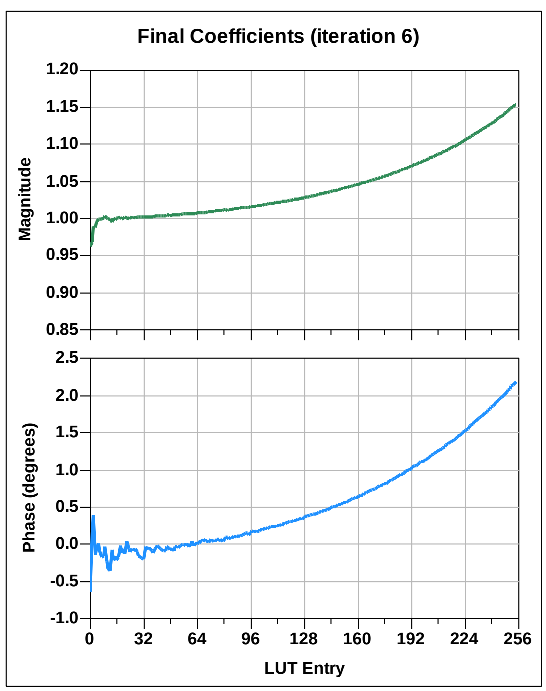

Fig. 5.7 Final Magnitude and Phase entries of LUT

Figure 5.5 and Figure 5.6 show the magnitude and phase LUT entries for different training inputs. The final magnitude and phase LUT entries have been shown in Fig 5.7.

To evaluate the performance of the proposed LUT based adaptive digital predistorter, the input and output signals of PA have been plotted as shown in Figure 5.7 and 5.8 respectively.

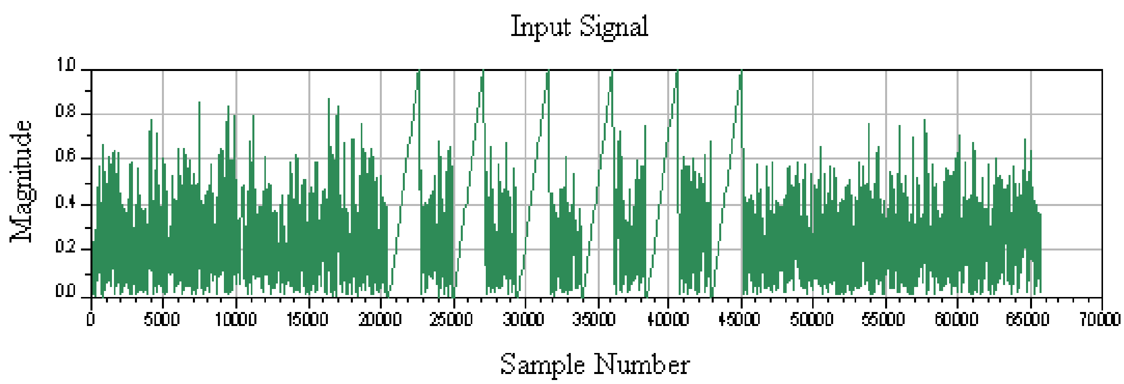

Fig. 5.8 Input Signal

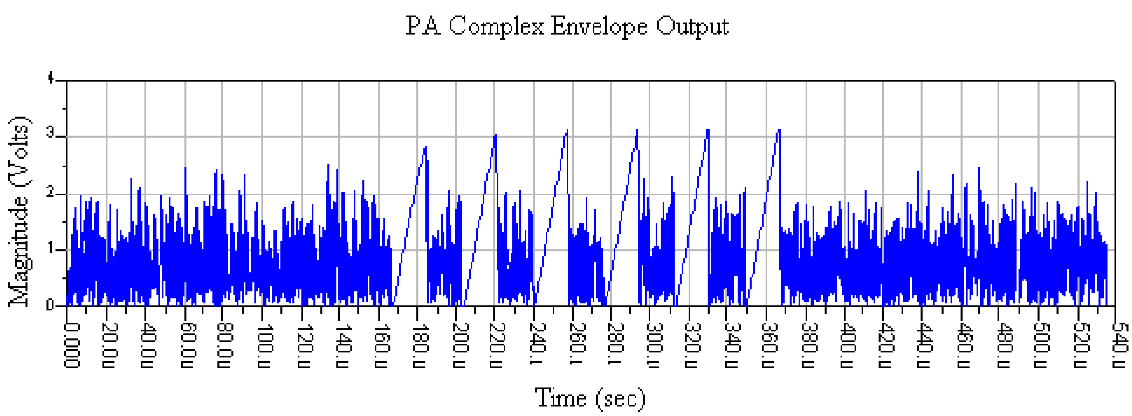

Fig. 5.9 Output Signal

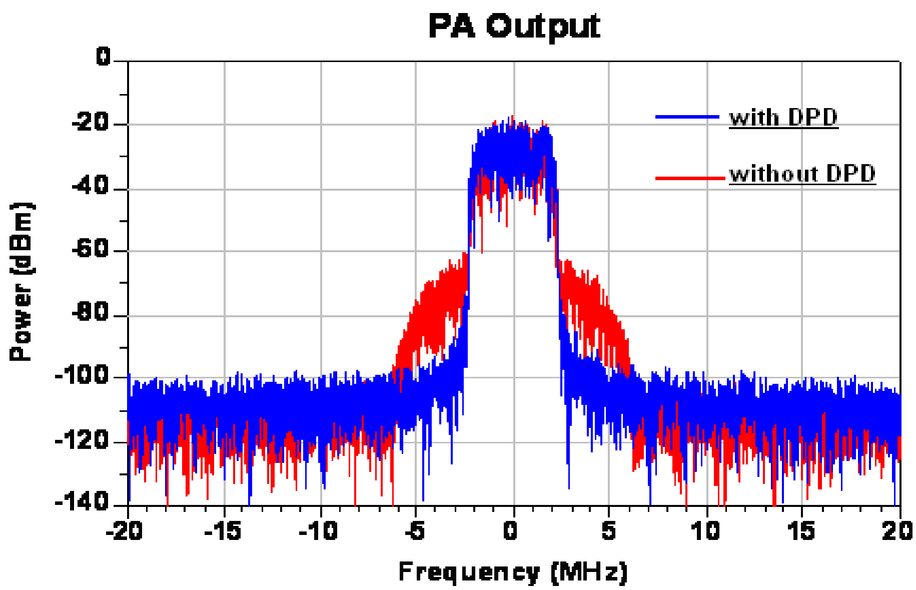

Fig. 5.10 Spectrum of output signal with and without Pre-distortion

The spectrum of the output with and without predistortion is shown in Figure 5.10. The spectrum shows the capability of the proposed design in suppressing signal re-growth. Calculations show that the proposed LUT based adaptive digital predistorter reduces error vector magnitude (EVM) to 18.14% and ACLR by 54.19 dB.

5.4 Conclusion

In the presented work a low complexity adaptive digital predistorter with much higher convergence rate as compared to other LUT based adaptive digital predistorters has been present. Six training ramps have been used to get high degree of convergence. Although the proposed design shows better performance in terms of reducing EVM and improving ACLR, yet future work will be focused on FPGA implementation of the proposed technique with less hardware requirements.

Chapter 6

Digital Predistortion of Power Amplifiers using Look-Up Table Method with Memory effects for LTE Wireless Systems

6.1 Introduction

Wideband signals such as those used in LTE systems are spectrally more efficient. These signals are applied to power amplifiers; which are essential components in communication systems but are inherently nonlinear. The nonlinearity generates spectral regrowth, which leads to adjacent channel interference and violations of the out of band emission requirements mandated by regulatory bodies. It also causes in-band distortion, which degrades the bit error rate (BER) performance [137]. To reduce the nonlinearity, the power amplifier can be operated at a lower power ("backed off") so that it operates within the linear portion of its operating curve. For the communications system architect and the RF power amplifier (PA) designer, the new wireless formats introduce a number of challenges. Designers must determine the performance gap between their existing 3G designs and tomorrow's 4G operating environments, and whether these 3G designs will need to be redesigned, or a new vendor qualified [138]. The hardware must also meet or exceed absolute performance metrics such as ACPR, EVM or throughput (e.g., BLER and BER), while also meeting internal product design goals. Because smart phones and other advanced wireless devices rely so heavily on battery power, getting the most efficiency out of a design is critical. The RF PA plays a particularly key role in choosing and designing the right PA to meet design goals which is a significant challenge. Also, newer transmission formats, such as wideband code division multiple access (WCDMA) and orthogonal frequency division multiplexing (802.11ac and LTE-Advanced), have high peak to average power ratios (PAPR); that is, large fluctuations in their signal envelopes. This means that the power amplifier needs to be backed off well below its maximum ("saturated") output power in order to handle infrequent peaks, which results in very low efficiencies, typically less than 10%. With more than 90% of the DC power being lost and turning into heat, the amplifier performance, reliability, and ongoing operating expenses (OPEX) are all degraded [139-140].

A predistorter applies distortion to the input signal in order to drive the PA harder. The DPD-PA cascade attempts to combine two nonlinear systems into one linear result which allows the PA to operate closer to saturation. Beyond this point, no increase in power will suffice to linearize the PA. The PAPR of the signal greatly restricts optimal performance of the DPD system. A CDMA signal, for example, may have a PAPR as high as 13dB. A PA transmitting such a signal must operate with significant back-off to prevent peaks from occurring beyond saturation. There are two common types of DPD implementation, the first is an analog implementation using a physical nonlinear device. The second and perhaps more popular choice, is a digital signal processor (DSP) hardware implementation where the DPD function is defined algorithmically through software [141].

There are two classes (memoryless model and model with memory) for Digital pre-distortion implementations. Memoryless models focus on the power amplifier that has a memoryless nonlinearity; i.e., the current output depends only on the current input through a nonlinear mechanism. This instantaneous nonlinearity is usually characterized by the AM/AM and AM/PM responses of the power amplifier, where the output signal amplitude and phase deviation of the power amplifier output are given as functions of the amplitude of its current input. Both memoryless polynomial algorithm and Look-Up Table (LUT) based algorithm are two key algorithms for memoryless models [142].

As the signal bandwidth gets wider, such as in WCDMA, mobile WiMAX and 3GPP LTE and LTE-Advanced (up to 100 MHz bandwidth, 5 component carriers of carrier aggregation), power amplifiers begin to exhibit memory effects. This is especially true for those high power amplifiers used in wireless base stations. The causes of the memory effects can be attributed to thermal constants of the active devices or components in the biasing network that have frequency dependent behaviors. As a result, the current output of the power amplifier depends not only on the current input, but also on past input values. In other words, the power amplifier becomes a nonlinear system with memory. For such a power amplifier, memoryless predistortion can achieve only very limited linearization performance. Therefore, digital predistorters must need to have memory structures [143].

The most important algorithm for models with memory for Digital predistortion implementation is Volterra series and its derivatives. The most general way to introduce memory is to use the Volterra series. However, the large number of coefficients of the Volterra series makes it unattractive for practical applications. Therefore, there are several Volterra's derivatives including Wiener, Hammerstein, Wiener–Hammerstein, parallel Wiener structures and memory polynomial model are popular in digital pre-distorters. The so-called "memory polynomial" is interpreted as a special case of a generalized Hammerstein model and is further elaborated by combining with the Wiener model. A memory polynomial pre-distorter uses the diagonal kernels of the Volterra series. The memory polynomial predistorter is used to linearize power amplifiers with memory effects. The pre-distorter is constructed using the indirect learning architecture, thereby eliminating the need for model assumption and parameter estimation of the power amplifier [144]-[145].

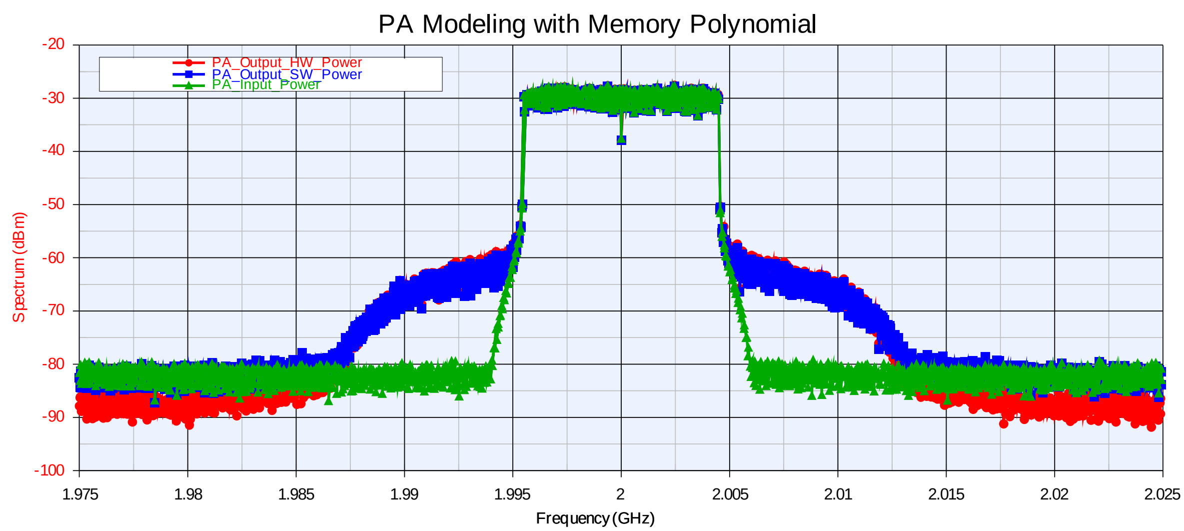

Fig. 6.1 PA modeling with memory polynomial: Power Spectrum vs Frequency

Figure 6.1 shows the spectrum of input and output of the actual and modeled PA. Red line shows the spectrum of actual PA and blue line shows the spectrum of PA modeled using memory polynomial. From this figure it has been concluded that the modeled PA has almost similar characteristics to that of the actual PA.

6.2 Proposed Model for Digital Predistortion in LTE Systems

In the DPD architecture, x(n) is the input signal to the pre-distortion unit, whose output z(n) feeds the power amplifier to produce output y(n). The most general form of nonlinearity with Q+1 taps of memory is described by the Volterra series, which consists of a sum of multidimensional convolutions. In the training branch, the Volterra series pre-distorter can be described by:

$$z(n) = \sum_{k=1}^{K} z_k(n) \qquad (6.1)$$

where

$$z_k(n) = \sum_{m_1=0}^{Q} \dots \sum_{m_k=0}^{Q} h_k(m_1 \quad , \dots , m_k) \prod_{l=1}^{k} y(n - m_l) \tag{6.2}$$

is the k-dimensional convolution of the input with Volterra kernel h_k. This is a generalization of a power series representation with a finite memory of length Q+1. The z(n) also can be written as follows:

$$z(n) = h_0 + \sum_{m_1=0}^{Q} h_1(m_1)y(n - m_1) + \sum_{m_1=0}^{Q} \sum_{m_2=0}^{Q} h_2(m_1, m_2)y(n - m_1)y(n - m_2) + \cdots \tag{6.3}$$

The DPD is described by a memory polynomial

$$z(n) = \sum_{k=1}^{K} \sum_{q=0}^{Q} a_{kq} x(n - q)|x(n - q)|^{k-1} \tag{6.4}$$

where the x(n) and z(n) are complex input signal and output signal of DPD model. The polynomial includes both the odd and even order terms. K and Q are the highest nonlinear and memory orders. a_{kq} is the complex coefficient of the polynomial that is to be extracted in this model .The memory polynomial can be described by the topology below.

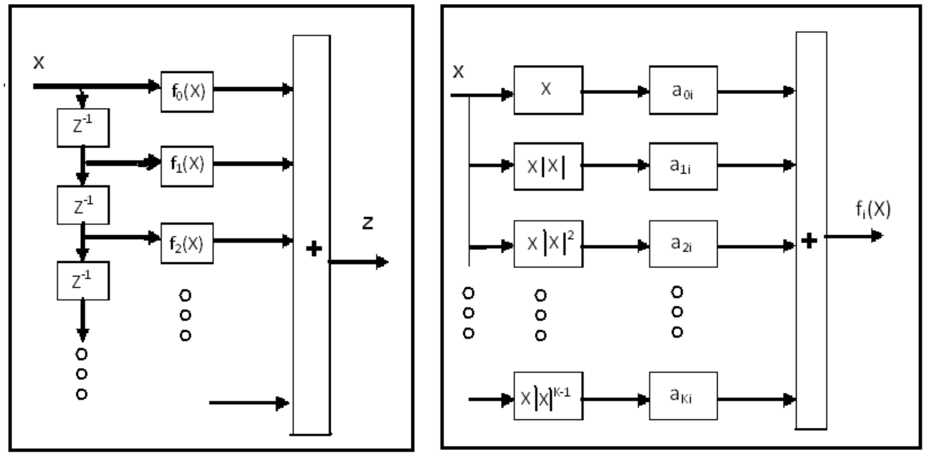

Fig. 6.2. Structure of the memory polynomial

The coefficients of the polynomial are extracted by indirect learning structure indicated in the following figure [146].

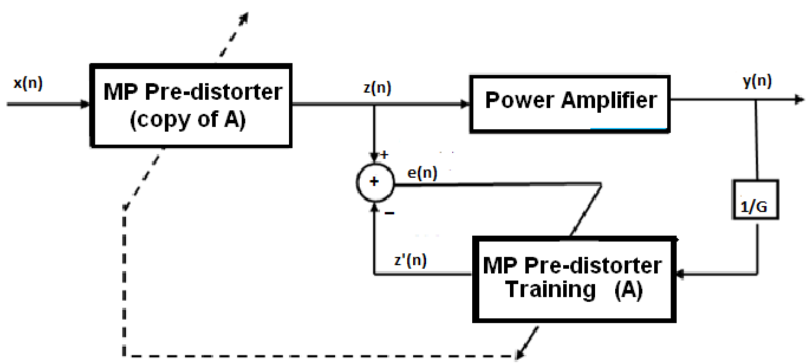

Fig. 6.3. Indirect learning structure to extract the coefficients of the polynomial

The algorithm uses two identical memory polynomial models for the predistorter and training. The real predistorter block is the exact copy of the training block. It has x (n) as input and z(n) as output. In the training branch, the memory polynomial pre-distorter can be described by:

$$z(n) = \sum_{k=1}^{K} \sum_{q=0}^{Q} a_{kq} y(n-q)|y(n-q)|^{k-1} \tag{6.5}$$

where y(n) and z(n) are the input and output of the pre-distorter in the training branch, respectively, and a_{kq} are the coefficients of the pre-distorter. If Q=0, the structure in the equation degenerates to a memoryless polynomial. Since the model in equation 6.5 is linear with respect to its coefficients, the predistorter coefficients a_{kq} can be directly obtained using a least-squares algorithm by defining a new sequence:

$$u_{kq}(n) = \frac{y(n-q)}{G}\left|\frac{y(n-q)}{G}\right|^{k-1} \tag{6.6}$$

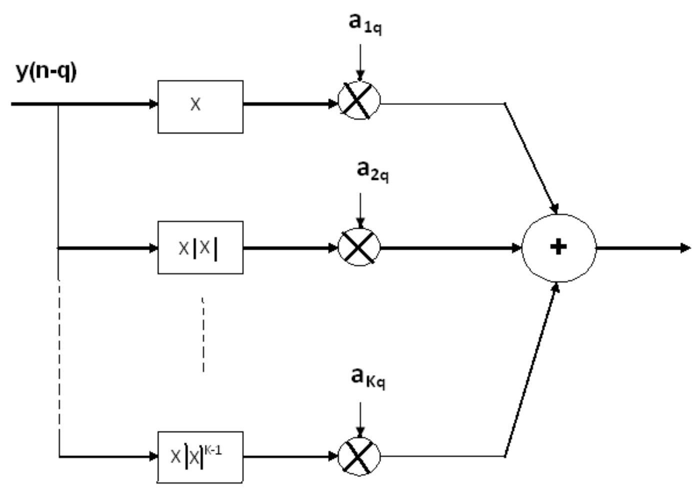

Fig. 6.4. Memory polynomial structure describing the behavior of nonlinear power amplifier (PA) for coefficient extraction.

At convergence, we have

$$z = Ua \tag{6.7}$$

where

$$z = [z(0), z(1), \ldots, z(N-1)]^T \tag{6.8}$$

$$U = [u_{10}, \ldots, u_{K0}, \ldots, u_{1Q}, \ldots u_{KQ}] \tag{6.9}$$

$$u_{kq} = [u_{kq}(0), u_{kq}(1), \ldots, u_{kq}(N-1)]^T \tag{6.10}$$

$$a = [a_{10}, \dots, a_{k0}, \dots, a_{1Q}, \dots, a_{KQ}]^T \quad (6.11)$$

The least squares solution is given by

$$\hat{a} = (U^H U)^{-1} U^H z \quad (6.12)$$

where $(U)^H$ denotes the complex conjugate transpose matrix.

After getting memory polynomial coefficients $a = [a_{10}, \dots, a_{k0}, \dots, a_{1Q}, \dots, a_{KQ}]^T$ and loading these coefficients into a nonlinear filter, the memory polynomial predistorter is able to function properly [147]-[148]. Figure 6.5 describes the simplified block diagram of the overall DPD system.

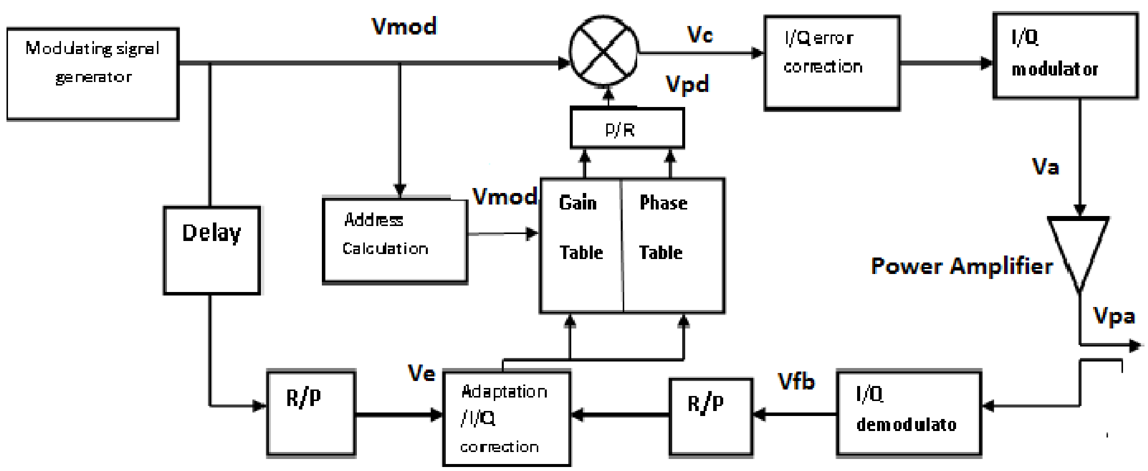

Fig. 6.5. Simplified block diagram of the overall DPD system

In the model identification, the PA output is normalized by the small-signal gain and then used as the input of the DPD model [149]. The PA input is taken as the output of the DPD model. Only part of the signal is used to do model identification. Accurate synchronization of the PA input and output signal is implemented in this algorithm [150]-[151]. The PA input and output after normalization and delay adjustment are given as output of this model.

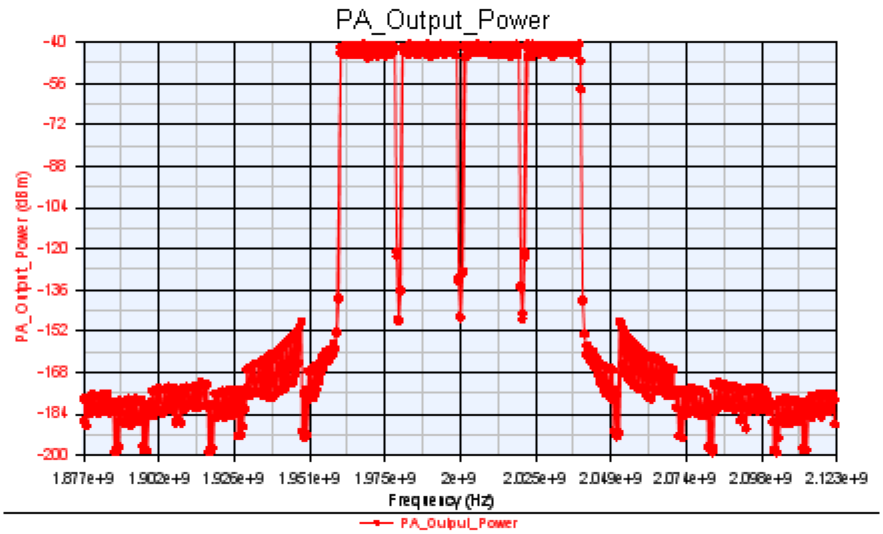

Fig. 6.6.Graph of PA Output Power with Frequency

Normalized mean squared error (NMSE) is used to evaluate how close are the PA input and the DPD model output. Highest nonlinear order K and memory order Q are key parameters that decide how well the DPD model works. To meet the requirements of the NMSE, K and Q values were increased step by step, Finally the values of K and Q chosen were 5 and 7 respectively. These values were required to justify the memory effects of RF power amplifier deployed in wideband communication systems especially LTE systems. Those samples with amplitude less than threshold would not be used in model extraction. No sample was discarded because the threshold level used was 0. This model is used to extract the memory polynomial model that describes the behavior of nonlinear power amplifier (PA). The PA output is normalized by the small signal gain and shifted to accurately synchronize with the PA input. The result is given as shifted PA output. After the normalization, the small-signal gain is 1 (0 dB). Behavioral models of power amplifier (PA) were traditionally developed for narrowband applications. The models are extracted based on the AM/AM and AM/PM characteristics without considering memory effects. A complex polynomial of instantaneous input power is used to approximate the PA gain. However, with the increase of signal bandwidth, memory effects cannot be ignored anymore. The AM/AM and AM/PM characteristics are not constant, but change as the function of both the present and the past input signals. Several models were proposed to describe the PA memory effects [152].

The average input power should be kept to the level that the peak output power of the PA does not exceed 3dB gain compression point. Otherwise, the extraction error will be large. The system was designed for an LTE bandwidth of 20MHz, base DFT size of 2048 and transmission bandwidth configuration, expressed in units of resource blocks is chosen as 100. The total power dynamic range is the difference between the maximum and the minimum transmit power of an OFDM symbol for a specified reference condition and is set to 20 dB, the cyclic prefix used is normal and the modulation scheme used is 64-QAM.

The Error Vector Magnitude is a measure of the difference between the ideal symbols and the measured symbols after the equalization. This difference is called the error vector. The value of EVM recommended for LTE systems for 64-QAM is 8%. Adjacent Channel Leakage power Ratio (ACLR) is the ratio of the filtered mean power centered on the assigned channel frequency to the filtered mean power centered on an adjacent channel frequency and the minimum requirement of ACLR in LTE systems is set to 45 dB.

The basic unit of EVM measurement is defined over one sub-frame (1ms) in the time domain and subcarriers (180 kHz) in the frequency domain:

$$\mathrm{EVM} = \sqrt{\frac{\sum_{t\in T}\sum_{f\in F(t)}\left|Z'(t,f)-I(t,f)\right|^2}{\sum_{t\in T}\sum_{f\in F(t)}\left|I(t,f)\right|^2}} \tag{6.13}$$

where

T is the set of symbols with the considered modulation scheme being active within the sub frame,

F(t) is the set of subcarriers within the subcarriers with the considered modulation scheme being active in symbol t,

I (t, f) is the ideal signal reconstructed by the measurement equipment in accordance with relevant Tx models,

Z'(t, f) is the modified signal,

Implicit in the definition of EVM is an assumption that the receiver is able to compensate a number of transmitter impairments. The signal under test is equalized and decoded according to:

$$Z'(t,f) = \frac{\mathrm{FFT}\left\{z(v-\Delta\tilde{t})e^{-j2\Pi\Delta\tilde{f}v}\right\}.e^{j2\Pi f\Delta\tilde{t}}}{\tilde{a}(t,f).e^{j\tilde{\phi}(t,f)}} \tag{6.14}$$

where

z(v) is the time domain samples of the signal under test, $\tilde{t}$I is the sample timing difference between the FFT processing window in relation to nominal timing of the ideal signal., $\tilde{f}$ is the RF frequency offset, $\tilde{\phi}(t,f)$ is the phase response of the T_X chain and $\tilde{a}(t,f)$ is the amplitude response of the T_X chain. The NMSE value is expressed in dB

$$\mathrm{NMSE[dB]} = 10 * \log_{10}\left[\frac{\sum|x_i - y_i|^2}{\sum|x_i|^2}\right] \tag{6.15}$$

The DPD module pre-distorts the signal in a reverse manner. The input of the DPD model is the normalized output of the PA, and the desired output of the extracted DPD model is the input of the PA. For DPD model NMSE evaluation, x_i is measured PA input signal, y_i is the extracted DPD model output signal. For PA model NMSE evaluation, x_i is measured PA output signal, y_i is the extracted PA model output signal. The smaller the NMSE, the more accurate the model. The recommended NMSE is less than -35dB. If the NMSE is too large, the extracted model cannot work effectively or even fails. Increasing the nonlinear order or memory order may improve NMSE. If the NMSE is still not low enough after nonlinear/memory order adjustment, lowering the operating point of PA is suggested.

The power amplifier model is used to adjust the value of the baseband signal and the input signal of the extracted DPD model to the same range. The PA output normalized by small-signal gain is taken as the input of DPD model. When the extracted DPD model is implemented, the actual input signal is the baseband signal. To make the DPD model work correctly, the baseband signal should be adjusted to the same range as the normalized output of the PA. The Gain is the adjustment factor. Typically, the gain adjustment is implemented between the signal source and digital predistorter model as shown below. The value of Gain is the adjustment factor calculated in this model.

6.3 Results and Discussion

Engineers migrating to 4G require a solution that makes implementing DPD fast and practical for 4G communications systems i.e. one that can be used by engineers at all levels of expertise and requires minimal equipment. In modern communication systems, spectrally efficient wideband RF signals have a peak-to-average power ratio (PAPR) as high as 13 dB. CFR preconditions the signal to reduce signal peaks without significant signal distortion. By reducing PAPR, CFR allows the PA to operate more efficiently at higher power levels, without impacting compliance with spectral mask and EVM specifications. CFR acts on the signal itself, whereas DPD corrects for the PA nonlinearity, allowing the signal to be run even higher.

CCDF (Complementary Cumulative Distribution Function) measurements provide important information for engineers involved in the design and manufacturing of system components used in fourth generation (4G) networks. 4G networks use OFDM that result in higher quality voice as well as greater data rates for cellular services. The peak to average ratio of 4G components is dependent on the sub channels used on a particular channel within a spread spectrum signal format. This can impact the distortion of a transmitted 4G signal. Performing CCDF measurements on 4G systems provide power characteristics of amplified, filtered, and mixed spread spectrum signals. The CCDF measurements performed in our case are shown in figure 6.7 and it provides information on the amount of time the signal spends at or above a specific power level – with the power level being in dB relative to the average power level. The Y axis is the percent of time the signal power is at or above the power specified by the X axis. The X axis represents dB above the average power level; this displays the peak to average ratios as opposed to absolute power levels. For example when t = 1% on the y axis, the corresponding peak to average ratio is 6.90 dB. This means the signal power exceeds the average by at least 6.90 dB for 1% (correct use) of the time. The peak and mean powers obtained are 4.724 dBm and -3.919 dBm, which gives rise to peak to average power ratio of 8.643 dBm and is calculated as PAPR = Peak Power - Mean Power.

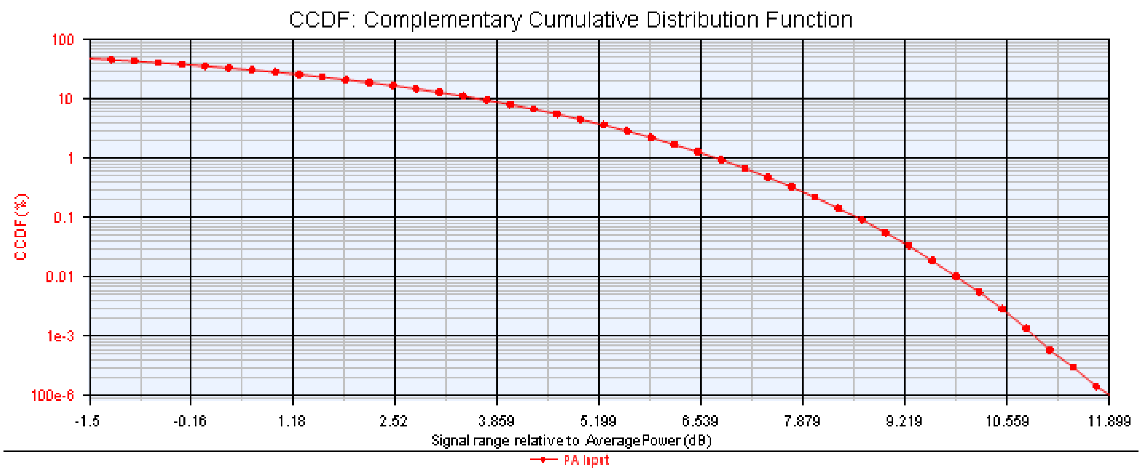

Fig. 6.7. Graph for the complimentary cumulative distribution function

Figure 6.7 shows an LTE-Advanced signal with 2x20 MHz carrier aggregation (40 MHz total signal bandwidth). The PA output spectrum is shown with and without DPD. Note the use of oversampling for an actual measurement bandwidth that is wider than the original signal. Without this additional bandwidth to quantify the out-of-band energy (distortion products), it is not possible to correct for this spectral regrowth later.

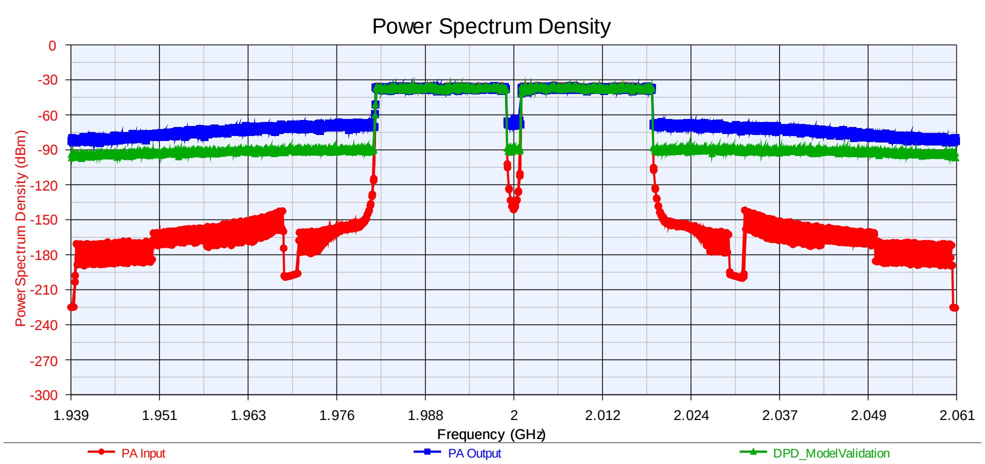

Fig. 6.8. DPD performance of LTE-A with 2x20 contiguous carrier aggregation.

The PA input is displayed in green, the distorted PA output without DPD is displayed in blue, and the linearized DPD+PA output is shown in red.

The DPD AM-AM shows the DPD AM-AM characteristics terms of two curves in Figure 6.8. The red curve is the AM-AM of the samples that are used to do model extraction. The blue curve is the AM-AM of the samples that are not used for model extraction. Figure 6.9 gives the DPD model validation as far as the AM-AM characteristics are concerned.

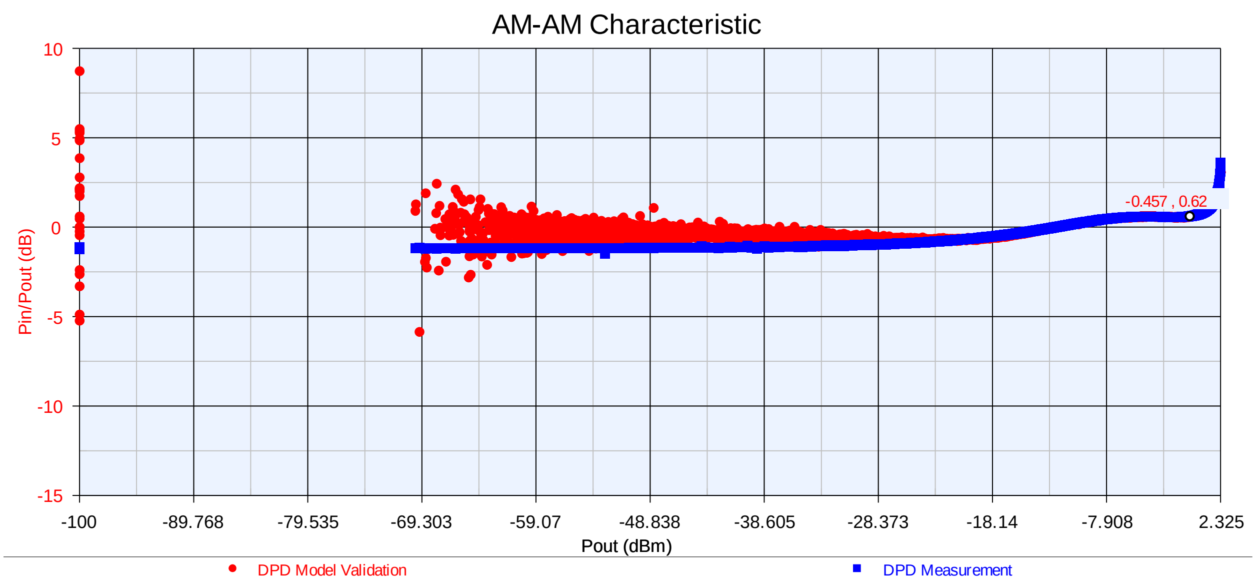

Figure 6.9 AM-AM Characteristics

The power alignment value is obtained as 0.652 and that value calculates back-off power. The NMSE value of –48.64 dB is computed as it is necessary to first look at the NMSE value and then the DPD spectrum. Figure 6.10 provides the power spectrum before DPD and after DPD. The EVM measurements show that when the original value before PA was -26.805, it was

-24.924 after the power amplifier stage when the DPD was not employed. And when the DPD is employed, the value of EVM was obtained as -26.801 which shows that by employing the DPD for LTE systems, the errors are minimized to a great extent.

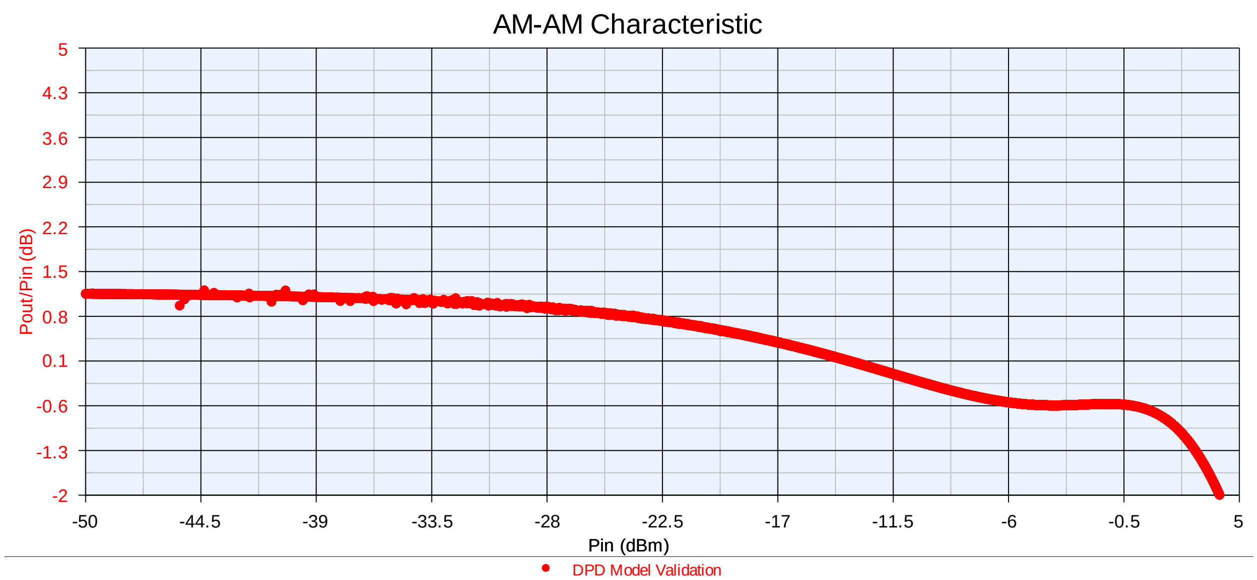

Fig. 6.10. DPD model validation for AM-AM characteristics

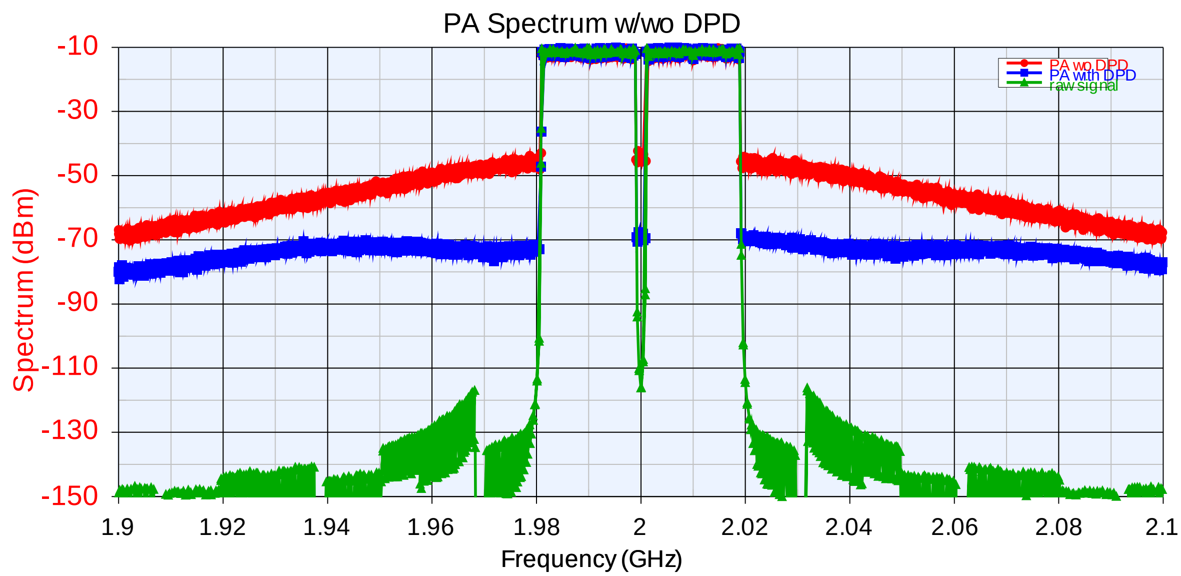

Fig. 6.11. Graph of PSD wrt Frequency

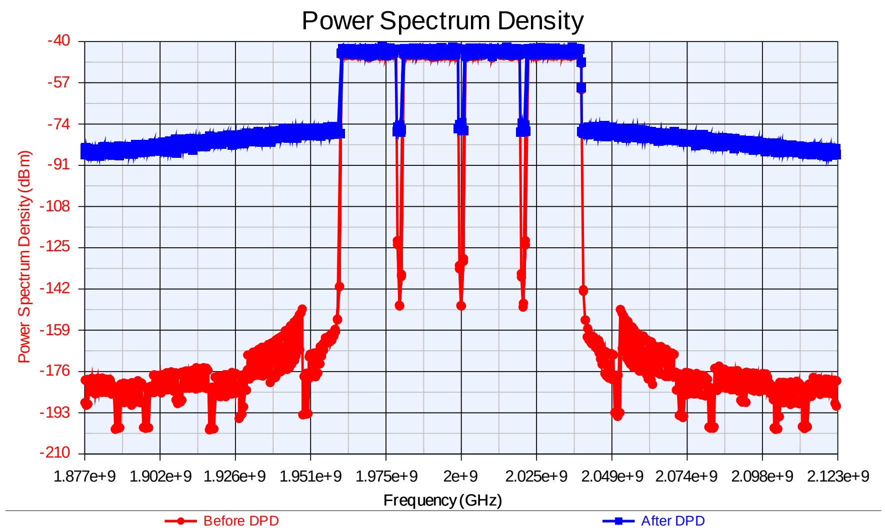

Fig. 6.12. DPD Model validation

Table 6.1 shows the adjacent channel leakage ratio (ACLR) measurements at three stages which involve the stage before PA, after the PA and then finally after the PA and after the DPD was employed. In LTE systems, the minimum requirement of ACLR has been set to 45dB. The table clearly shows the deterioration in the ACLR values after the PA stage and when DPD is employed, the ACLR values crosses beyond the threshold limits.

	ACLR L_2BW(dB)	ACLR L_BW(dB)	ACLR U_2BW(dB)	ACLR U_BW(dB)
Original ACLR	64.69	48.93	47.859	61.334
ACLR after PA	63.462	44.812	44.473	60.751
ACLR after PA+DPD	64.462	48.783	47.769	61.474

Table 6.1. ACLR Measurements

6.4 Conclusion

As engineers migrate to advanced wireless communication systems such as LTE-Advanced or 802.11, choosing and designing the right PA to meet design goals at the lowest possible cost becomes more difficult, both for base stations and mobile devices. Because DPD enables the PA to be operated with high efficiency, near saturation and without significant signal distortion, it allows engineers to address many base station/mobile device PA design challenges. This chapter considered the design of digital predistortion systems to Linearize power amplifiers with memory effects. By adding a digital predistorter in the baseband, the power amplifier is allowed to operate into its nonlinear region, thereby significantly increasing its efficiency. The efficiency gain translates into electricity and cooling cost savings for service providers and longer battery life for mobile terminal users. The challenge here is to address the memory effects exhibited by the higher power amplifiers or the power amplifiers for wideband signals. In addition, analog components in the transmitter have imperfections that need to be compensated as well.

Chapter 7

Conclusion and Future Scope

For accurate design and implementation of any PA linearization technique, exact modeling of PA is an important issue. Typically, the required linearity can be achieved either by reducing power efficiency or by using linearization techniques. For a Class-A PA, simply “backing off” the input power level can improve linearity. However, for high peak to average power ratio (PAPR) signals, this normally reduces the power efficiency down to 10% while increasing heat dissipation up to 90%. When considering the vast number of base stations that wireless operators need to account for, increasing power consumption, or in other words, power back-off is not a viable tradeoff. Therefore, an amplifier linearization has become an important technology and a desirable alternative to backing-off an amplifier in modern communications systems. Although nonlinear distortions can be equalized at the receiver side but it is complicated to implement due to the unknown effects of the channel. It is therefore easy to reduce the nonlinear distortions at the transmitter side.

Modeling of PA using complex memory polynomial has been discussed. The thesis also presents the behavioral modeling approach to model PA. Non-linear Memory-less and Quasi Memory-less Models of PA like Rapp model, Saleh model, Ghorbani model, Hyperbolic Tangent model has also been discussed. Also non-linear models of PA with Memory like Wiener Model, Hammerstein Model, Parallel Hammerstein Model and Memory Polynomial Model most widely used in literature has been presented. Its various characteristic like Gain

Compression, Phase Characteristics, Input Characteristics, Output Characteristics and constellations, 3rd order IMD Gain reconstruction, 3rd order IMD Phase reconstruction, ACLR and EVM have been compared with the actual PA.

The thesis considered the design of digital predistortion systems to Linearize power amplifiers with memory effects. By adding a digital predistorter in the baseband, the power amplifier is allowed to operate into its nonlinear region, thereby significantly increasing its efficiency. The efficiency gain translates into electricity and cooling cost savings for service providers and longer battery life for mobile terminal users. A low complexity, direct-learning multilevel lookup table based adaptive digital predistortion technique has been proposed to linearize a power amplifier (PA). A loop delay compensation scheme has been used to achieve a significant reduction in convergence time and an improvement in linearization accuracy in the presence of an unknown loop delay. Compared with the conventional predistorters, the proposed technique show fast adaptation speed which enables the predistorter to track time-varying PA nonlinearities. A novel low complexity LUT based adaptive digital pre-distorter with reduced memory requirements has been proposed by using interpolation and efficient spacing of table entries, which leads to low LUT requirement. The proposed adaptive digital pre-distorter has much higher convergence rate as compared to other LUT based adaptive digital pre-distorters. In the proposed design polar LUT based predistorter has been used. Optimal spacing has been achieved by applying a suitable compander to the amplitude signal prior to addressing the LUT.

The challenge here is to address the memory effects exhibited by the higher power amplifiers or the power amplifiers for wideband signals. In addition, analog components in the transmitter have imperfections that need to be compensated as well. The spectrum of the output with and without pre-distortion is shown in Chapter 5. The spectrum shows the capability of the proposed design in suppressing signal re-growth. Calculations show that the proposed LUT based adaptive digital pre-distorter reduces error vector magnitude (EVM) to 18.14% and ACLR by 54.19 dB.

The CCDF measurements performed in the last chapter provides information on the amount of time the signal spends at or above a specific power level – with the power level being in dB relative to the average power level. The Y axis is the percent of time the signal power is at or above the power specified by the X axis. The X axis represents dB above the average power level; this displays the peak to average ratios as opposed to absolute power levels. We can observe that when t = 1% on the y axis, the corresponding peak to average ratio is 6.90 dB. This means the signal power exceeds the average by at least 6.90 dB for 1% (correct use) of the time. The peak and mean powers obtained are 4.724 dBm and -3.919 dBm, which gives rise to peak to average power ration of 8.643dBm. The power alignment value is obtained as 0.652 and that value calculates back-off power. The NMSE value of –48.64 dB is computed.When the DPD is employed, the value of EVM was obtained as -26.801 which shows that by employing the DPD for LTE systems, the errors are minimized to a great extent.

The future work can be extended in a number of directions, including: Designing a fast adaptive memory polynomial predistorter based on the orthogonal polynomial theory, performing tests on different types of power amplifiers and establishing connections between the memory behaviors of the power amplifier, combining predistortion with peak-to-average ratio reduction techniques to further improve the efficiency of the power amplifier. Although the proposed design shows better performance in terms of reducing EVM and improving ACLR, yet future work will be focused on FPGA implementation of the proposed technique with less hardware requirements.

References

[1] Kenney, J. Sand Leke, A. (1991) Wireless Report: Power Amplifier spectral regrowth for digital cellular and PIIS applications. *Microwave Journal, Texas Instruments Inc., 12500 TI Blvd., Dallas, Texas*, 74-92.

[2] Amandeep Singh Sappal, Manjeet Singh Patterh And Sanjay Sharma, "To Develop A Linearization Technique for Mitigating the RF Power Amplifier's Non-Linearity Effects In A Multicarrier W-CDMA Base Station' Phd Thesis, Punjabi University, Patiala, 2011

[3] Baudoin, G. and Jardin, P. (2000) A new adaptive baseband pre-distortion algorithm for linearization of power amplifiers: application to EDGE-GSM transmitters. *Proc. of the European Conference on Wireless Technology*, 2000. pp. 163–166.

[4] Baudoin, G. and Jardin, P. (2001) Adaptive polynomial pre-distortion for linearization of power amplifiers in wireless communications and WLAN. *Proc. of the International Conference on Trends in Communications*, 2001. pp. 157–160.

[5] Bauernschmitt, R., Bitzer, T. and Woele, G. (2001) Digital predistortion for the linearization of power amplifiers in W-CDMA basestations. *Conference on Microwaves, Radio Communication and Electromagnetic Compatibility*, 2001. pp. 350–355.

[6] Bauman, R.M. (1983) Adaptive Feed-forward System. U.S.P. 4, 389, 618.

[7] Kenington, P.B. (2000) *High-Linearity RF Amplifier Design*, Artech House.

[8] Pothecary, N. (1999) *Feedforward Linear Power Amplifiers*, Boston: Artech House, ISBN: 1-58053-022-2.

[9] Cripps, S.C. (2002) *Advanced Techniques in Power Amplifier Design*, Artech House.

[10] Väänänen, O. (2006) Digital *Modulators with Crest Factor Reduction Techniques*. Ph.D. thesis, Helsinki University of Technology, Electronic Circuit Design Laboratory Report 42, Espoo.

[11] Eun, C.S. (1995) *Design and comparison of Non-linear compensators*. Ph.D. Thesis, University of Texas, Austin.

[12] Rapp, C. (1991) Effects of HPA Nonlinearity on a 4.DPSK/OFDM signal for a Digital Sound Broadcasting System. *Proceedings of the Second European Conference on Satellite Communications*, Liege, Belgium, 1991. pp. 179.184.

[13] Saleh, A.A.M., (1981) Frequency independent and frequency dependent nonlinear models of TWT amplifiers, *IEEE Trans. Communications*, COM.29: 1715.1720.

[14] Ghorbani, A. and Sheikhan, M. (1991) The Effect of Solid State Power Amplifiers (SSPA's) nonlinearities on MPSK and M-QAM Signal Transmission. *Sixth Int'l Conference on Digital Processing of Signals in Comm.*, pp. 193.

[15] Mathews V.J. and Sicuranza, G.L. (2000) *Polynomial Signal Processing*, New York: Wiley and Sons Inc.

[16] Metzios, B.G. (1994) Parallel Modeling and Structure of Nonlinear Volterra Discrete Systems. *IEEE Trans. Circuits and Systems, 41(5)*: 359–371.

[17] Newak, R.D. and VanVeen, B.D. (1996) Tensor product basis approximation for Volterra filters. *IEEE Trans. Signal Processing, 44*: 36–59.

[18] Paniker, T.M. and Mathews, V.J. (1996) Parallel-cascade realization and approximation of truncated Volterra systems. *Proc. IEEE ICASSP*, pp. 1589–1592.

[19] Ding L. (2004) *Digital Predistortion of Power Amplifiers for Wireless Applications*. Ph.D. thesis, Georgia Institute of Technology.

[20] Inc Isaksson, M., Wisell, D. and Rönnow, D. (2006) A Comparative Analysis of Behavioral Models for RF Power Amplifiers. *IEEE Trans on Microwave Theory and Techniques,54(1)*.

[21] Cripps, S.C. (1999) *RF Power Amplifiers for Wireless Communications*, Artech House.

[22] Bruss, S. (April 23, 2003), Linearization Methods.

[23] Raab, F.H. and Asbeck, P. (2003) RF and Microwave Power Amplifier and Transmitter Technologies. *High Frequency Electronics*

[24] Petrovic, V. (1983) Reduction of Spurious Emission from Radio Transmitters by Means of Modulation Feedback. *Proceedings of IEEE Conference on Radio Spectrum Conservation Techniques.* pp. 44-49.

[25] Cardinal, J.S. and Ghannouchi, F.M. (1995) A new Adaptive Double Envelope Feedback (ADEF) Linearizer for Solid State Power Amplifier. *IEEE Transactions on Microwave Theory and Techniques, 43(7)*: 1508 – 1515.

[26] Johansson, M. and Sundstrom, L. (1994) Linearization of RF Multicarrier Amplifier using Cartesian Feedback. *Electronics Letters,30(14)*: 1110 – 1112.

[27] Kahn, L. (1952) Single-sided Transmission by Envelope Elimination and Restoration. *Proceedings IRE.* pp. 803–806.

[28] Black, H.S. (1928) Translating System. U.S.P. 1, 686 792

[29] Kenington, P.B. (1992) Power Amplification Techniques for Linear TDMA 10 Base Stations. *Proceedings of GLOBECOM '92, Orlando, USA.* pp.74-78.

[30] Stewart, R.D. and Tusubira, F.F. (1988) Feedforward Linearization of 950MHz Amplifiers. *IEEE Proceedings*. *135(5)*: pp. 347- 350.

[31] Taijun, L., Boumaiza, S. and Ghannouchi, F.M. (2004) Dynamic behavioral modeling of 3G power amplifiers using real-valued time-delay neural networks. *IEEE Trans on Microwave Theory and Techniques,52 (3)*: 1025-1033.

[32] Oualid Hammi,Scott Carichner,Bill Vassilakis and Fedhel M.Ghannouchi,2003

[33] O'Droma M., Meza S., Lei Y, (2009) New modified saleh models for memoryless nonlinear power amplifier behavirol modeling. *IEEE Communications Letters, 13(6) : 399-401.*

[34] Bershad, N.J., Celka, P. and Vesin, J.M. (1999) Stochastic Analysis of Gradient Adaptive Identification of Nonlinear Systems with Memory for Gaussian Data and Noisy Input and Output Measurements. *IEEE Trans. Signal Processing, 47(3)*: 675–689.

[35] Billings, S. A., (1980) Identification of nonlinear systems: A survey. *IEEE Proceedings Pt. D.*, 127(6): 272–285.

[36] Mathews V.J. (1991) Adaptive Polynomial Filters. *IEEE Signal Processing Mag., 8(3)*: 10–26.

[37] Metzios, B.G. (1994) Parallel Modeling and Structure of Nonlinear Volterra Discrete Systems. *IEEE Trans. Circuits and Systems, 41(5)*: 359–371.

[38] Dooley J., O' Brien B, Brazil T.J. (2006) Prior Knowledge input Neural Network for Microwave Power Amplifier Modelling.*IEEE Conference publications, Page(s): 156 – 159*

[39] ZhuA., WrenM., Brazil T.J.,(2003) An efficient Volterra based Behavioral model for Wideband RF Power Amplifiers,Microwave Symposium Digest, 2003 IEEE MTT-S International Volume: 2 , Page(s): 787 – 790.

[40] ZheJin, ZhihuanSong, Jiaming He, (2008) Volterra series based predistortion for broadband RF power amplifiers with memory effects, Journal of Systems Engineering and Electronics, 19(4), Page(s): 666 – 671.

[41] Hyunchul, K. andKenney, J.S. (2003) Behavioral Modeling of Nonlinear RF Power Amplifiers Considering Memory Effects. *IEEE Trans. on Microwave Theory and Techniques,51(12).*

[42] MorganD.R., ZhengxiangMa,JaehyeongKim, Zierdt M.G., Pastalan J.,(2006) A Generalized Memory Polynomial Model for Digital Predistortion of RF Power Amplifiers, IEEE Transactions on Signal Processing, 54(10), Page(s): 3852 – 3860.

[43] KaiFu, ChoiLookLaw, Cunfeng Gu, Than Tun Thein, Myat Thu Kyaw,(2009) A new power amplifier predistortion architecture based on memory polynomial model , 7th International Conference on Information, Communications and Signal Processing, IEEE Conference Publications,Page(s): 1 – 5.

[44] Kenney, J.S., Woo, W., Ding, L., Raich, R., Ku, H. and Zhou, G.T. (2001) The impact of memory effects on predistortion linearization of RF power amplifiers. *ISMOT*, Montreal, Canada, 2001.

[45] Ding, L., Zhou, G., Morgan, D., Ma, Z., Kenny, J., Kim, J. and Giardina, C. (2004) A Robust Digital Baseband Predistorter Constructed Using Memory Polynomials. *IEEE Transactions on Communications, 52(1).*

[46] Diaz, H. and Desrochers, A.A. (1988) Modeling of nonlinear discrete-time systems from input-output data. *Automatica, 24(5)*: 629–641.

[47] Amrit Kaur, Pradip Sircar, Adrish Banerjee, "Interpolation of Lost Frames of a Video Stream using Object based Motion Estimation and Compensation", Proceedings of the IEEE INDICON 2008, December 11th-13th 2008.

[48] Ljung, L. (1999) *System Identification: Theory for the User*, 2nded. New Jersey: Prentice-Hall.

[49] Nelles, O. (2000) *Nonlinear System Identification.* New York: Springer - Verlag.

[50] Ogunnaike, Babatunde and Harmon, R.W. (1994) *Process Dynamics, Modelling and Control,* New York: Oxford University Press.

[51] Sjoberg, J. (1995)Nonlinear black box modeling in system identification: A unified overview. *Automatica, 31*:1691– 1724.

[52] Sowlati, T., Greshishchev, Y. and Salama, A. Phase (1997) Correcting Feedback System for Class-E Power Amplifier. *IEEE Journal Solid-State Circuits*, *32*: 544–550.

[53] CliftonJ.C., AlbashaL., Lawrenson A., Eaton A.M.,(2005), Novel multimode J-pHEMT front-end architecture with power-control scheme for maximum efficiency, IEEE Transactions on Microwave Theory and Techniques, 53(6) , Page(s): 2251 - 2258

[54] Angheon, K., Young, Y.W., Junghwan, M. and Bumman., K. (2008) A New Wideband Employing Feedback Linearization. *IEEE Transactions on Microwave Theory and Techniques, 56(*2*)*: 385–392.

[55] SunghoJeon, JunghyunKim, Jaekwon Lee, Ho Jin Ryu, Young-Woo Suh, Ha-Kyun Mok, Man-Sik Kim,(2010) An adaptive digital predistorter with iterative noise cancelation, IEEE International Symposium on Broadband Multimedia Systems and Broadcasting (BMSB), Page(s): 1 – 5.

[56] PingCao, SongbaiHe, Fei You,(2011), A feedback predistortion technique for power amplifiers based on amplitude and phase signal processing, IEEE 3rd International Conference on Communication Software and Networks (ICCSN), Page(s): 284 – 287.

[57] JinhyunSo, SunghoChoi, Seung Hyeok Ahn, Eui-Rim Jeong, Lee Y.H.,(2012). Digital predistortion based on envelope feedback, IEEE International Conference on Acoustics, Speech and Signal Processing (ICASSP), Page(s): 3169 – 3172.

[58] Sung Won Chung, Holloway J.W., Dawson J.L.,(2007).Open-Loop Digital Predistortion Using Cartesian Feedback for Adaptive RF Power Amplifier Linearization , IEEE/MTT-S International Microwave Symposium, Page(s): 1449 – 1452.

[59] Lie Donald Y.C., Popp J.D., Wang F., Kimball, D., Larson L.E.,(2007).Linearization of Highly-Efficient Monolithic Class E SiGe Power Amplifiers with Envelope-Tracking (ET) and Envelope-Elimination-and-Restoration (EER) at 900MHz System-on-Chip, IEEE Workshop on Circuits and Systems, Page(s): 1 - 4 .

[60] LiWang; WenhuaChen; Zhijun Zhang; Zhenghe Feng,(2010).A novel class-E power amplifier with an asymmetrical spurline filter and its linearization, IEEE International Conference on Wireless Information Technology and Systems (ICWITS), Page(s): 1 – 4.

[61] Design and implementation of a secure access system to information resources for IEEE802.11 wireless networks, J. Mejía N., S. Vidal B., J. López-Bonilla, Proc. of Elect ronics, Robotics & Automotive Mechanics Conf.-CERMA'06, IEEE 26-29 Sept., Cuernavaca-Morelos, Mexico 1 (2006) 58-63.

[62] Bennett, T.J. and Clements, R.F. (1974) Feedforward an alternative approach to Amplifier linearization.*Radio and Electrical Engineering*, 44(5): 257-262.

[63] Bauman, R.M. (1983) Adaptive Feed-forward System. U.S.P. 4,389,618.

[64] Chapman, R.H. and Turney, W.J. (1991) Feedforward distortion cancellation circuit. U.S.P. 5,051,704.

[65] Narahashi, S. and Nojima, T. (1991) Extremely low-distortion multi-carrier amplifier self-adjusting feedforward amplifier. *Proceedings of IEEE International Communications Conference.* pp. 1485-1490.

[66] Oberman, M.G. and Long, J.F. (1991) Feedforward distortion minimization circuit. U.S.P. 5,077,532.

[67] Kumar, S. and Wells, G. (1991) Memory controlled feedforward linearizer suitable for MMIC implementation.*Inst. Electrical Engineering Proceedings*. 138(1): pp. 9-12.

[68] Stapleton, S.P., Kandola, G.S. and Cavers, J.K. (1992) Simulation and Analysis of an Adaptive Predistorter Utilizing a Complex Spectral Convolution.*IEEE Transactions on Vehicular Technology*, *41(4)*: 1-8.

[69] Hilborn, D., Stapleton, S.P. and Cavers, J.K. (1994) An adaptive direct conversion transmitter. *IEEE Transactions on Vehicular Technology*, *43(2)*: 223-233.

[70] Wilson, J.F. (1994) The TERA system and its requirements for linear amplification, *IEE Colloquium on Linear RF Amplifiers and Transmitters*, *Digest no. 1994/089*, *4*: 1-7.

[71] Bruno, M.; Cousseau, J;Ghadam, A.S.H. ;Valkama, M.(2010)High Linearity RF Amplifier Design,IEEE conference in Argentina School of Micro-Nanoelectronics Technology and Applications(EAMTA),Page(s):97-102.

[72] LevyY., KaramG., Sari H.,(1995). Adaptation of a digital predistortion technique based on intersymbol interpolation IEEE Global Telecommunications Conference, 1995. GLOBECOM '95.,Volume: 1, Page(s): 145 – 150.

[73] Sills J.A., Sperlich R.,(2002). Adaptive power amplifier linearization by digital pre-distortion using genetic algorithms, IEEE Radio and Wireless Conference, RAWCON 2002, Page(s): 229 – 232.

[74] Katheleen J. Muhonen, Mohsen Kavehrad, Rajeev Krishnamoorthy,(2000).Look-Up Table Techniques for Adaptive Digital predistortion : A Development and Comparison,IEEE Transactions on Vehicular Technology, 49(5).

[75] QunWu, JinghuiQiu, Jiahui Fu, Yuemei Qin, Haoyang Xing,(2002).High precision RF linearizer by digital predistortion technique, Conference on Precision Electromagnetic Measurements, 2002, Page(s): 76 – 77.

[76] Larson L., KimballD., Asbeck P., Draxler P., Junxiong Deng; Ming Li,(2006). Digital predistortion techniques for linearized power amplifiers, Asia-Pacific Microwave Conference, Page(s): 1048 – 1051.

[77] Chih-HungLin, Hsin-HungChen, Yung-Yi Wang, Jiunn-Tsair Chen,(2006).Dynamically optimum lookup-table spacing for power amplifier predistortion linearization, IEEE Transactions on Microwave Theory and Techniques, 54(5) , Page(s): 2118 – 2127.

[78] HusseinM.A., WangY., Feuvrie B., Toutain S., Peyresoubes G.,(2008).Piecewise Complex Circular Approximation of the Inverse Characteristics of Power Amplifiers for Digital Predistortion Techniques.The Third International Conference on Digital Telecommunications,Page(s): 59 - 63 .

[79] Angheon Kim,YoungYunWoo, Junghwan Moon, Bumman Kim,(2008).A New Wideband Adaptive Digital Predistortion Technique Employing Feedback Linearization, IEEE Transactions on Microwave Theory and Techniques, 56(2), Page(s): 385 – 392.

[80] ShibingHu, JiqingLu, (2011).A novel digital predistortion technique for Ultra Wide-Band radar pulse-compression signals, Second International Conference on Mechanic Automation and Control Engineering (MACE), Page(s): 5508 – 5511.

[81] ZhengGao, PingGui, (2012).A look-up-table digital predistortion technique for high-voltage power amplifiers in ultrasonic applications, IEEE Transactions on Ultrasonics, Ferroelectrics and Frequency Control, 59(7), Page(s): 1550 – 1557.

[82] Tabrizi M.M., Masoumi, N. (2011).CMOS Linear High Performance Push Amplifier for WiMAX PowerAmplifier, 7th International Conference on Wireless Communications, Networking and Mobile Computing (WiCOM), Page(s): 1 – 4.

[83] Cavers, J.K. (1990) Amplifier Linearization Using a Digital Predistorter with Fast Adaptation and Low Memory Requirements. *IEEE Transactions on Vehicular Technology*, *39(4)*: 374-382.

[84] Bosch, W. and Gatti, G. (1989) Measurement and Simulation of Memory Effects in Predistortion Linearizers. *IEEE Transactions on Microwave Theory and Techniques, 37(12)*.

[85] Ku, H. and Kenney, J. (2003) Behavioral Modeling of Nonlinear RF Power Amplifiers Considering Memory Effects. *IEEE Transactions on Microwave Theory and Techniques, 51(12)*.

[86] Vuolevi, J., Rahkonen, T. and Manninen, J. (2001) Measurement Techniques for Characterizing Memory Effects in RF Power Amplifiers. *IEEE Transactions on Microwave Theory and Techniques*, *49(8)*.

[87] Faulkner, M., Mattsson, T. and Yates, W. (1991) Automatic Adjustment of Quadrature Modulators. *Electronic Letters*, *27(3)*: 214-216.

[88] Faulkner, M. and Mattsson, T. (1992) Spectral Sensitivity of Power Amplifiers to Quadrature Modulator Misalignment. *IEEE Transactions on Vehicular Technology, 41(4)*.

[89] Stapleton, S.P. and Costescu, F.C. (1992) An Adaptive Predistorter for a Power Amplifier Based on Adjacent Channel Emissions.*IEEE Transactions on Vehicular Technology,41(1)*: 49-56.

[90] Wright, A. and Durtler, W. (1992) Experimental Performance of an Adaptive Digital Linearized Power Amplifier. *IEEE MTT International Microwave Symposium Digest*, 1992.

[91] Cavers, J.K. and Liao, M. (1993) Adaptive Compensation for Imbalance and Offset Losses in Direct Conversion Transceivers. *IEEE Transactions on Vehicular Technology, 42(4)*.

[92] Voudouris, K. and Noras, J. (1993) Effects of Amplitude, Phase, and Frequency Imperfection on the Performance of a Direct Conversion Receiver (DCR) for Personal Communication Systems. *IEEE Microwave and Guided Wave Letters,* 3.

[93] Sundstroem, L., Faulkner, M. and Johansson, M. (1995) Effects of reconstruction filters in Digital Predistortion Linearizers for RF Power Amplifiers. *IEEE Transactions on Vehicular Technology, 44(1)*: 131–139.

[94] Cavers, J.K. (1996) The Effect of Quadrature Modulator and Demodulator Errors on Adaptive Digital Predistorters. *Mobile Technology for the Human Race, IEEE 46th Vehicular Technology Conference*, 1996. vol. 2.

[95] Goeckler, R. (1996) *Use Adaptive Digital Predistortion to Simulate a Linearization System*. RF Design.

[96] Cavers, J.K. (1997)The Effect of Quadrature Modulator and Demodulator Errors on Adaptive Digital Predistorters for Amplifier Linearization. *IEEE Transactions on Vehicular Technology*, *46*: 456–466.

[97] Cavers, J.K. (1997) New Methods for Adaptation of Quadrature Modulators and Demodulators in Amplifier Linearization Circuits. *IEEE Transactions on Vehicular Technology, 46(3)*: 707–716.

[98] Jeckeln, E.G., Ghannouchi, F.M. and Sawan, M. (1998) An L Band Adaptive Digital Predistorter for Power Amplifiers using Direct I-Q modem. *IEEE MTT International Microwave Symposium Digest*, 1998. pp. 719–722.

[99] Ren, Q. and Wolff, I. (1997) Improvement of Digital Mapping Predistorters for Linearising Transmitters. *IEEE Transactions on Microwave Theory and Techniques*.

[100] Zavosh, F. (1999) Digital Predistortion Techniques for RF Power Amplifiers with CDMA Applications. *Microwave Journal*.

[101] Singh, A. and Srinivasan S., Digital Signal Processing Implementations Using DSP Microprocessors, brooks/ Cole - Thompson Learning, Belmont, California, 2005.

[102] Kenington, P.B, Cope, M., Bennett, R. and Bishop, J. (2001) A GSM – EDGE High Power Amplifier Utilising Digital Linearization. *IEEE MTT International Microwave Symposium Digest*, 2001.

[103] Sills, J. (2001) Application of Digital Predistortion to BWA. *Intersil Wireless Networking Products System Engineering*, Scottsdale, August 20, 2001.

[104] Stapleton, S. (2001) *Amplifier Linearization Using Adaptive Digital Predistortion*. Applied Microwave & Wireless.

[105] Davis, M. (2002) *A Software Linearization Method for Enhancing Noise Loaded Performance of Amplifiers.* American Institute of Aeronautics and Astronautics.

[106] Hammi, O., Boumaiza, S., Jaidane, M. and Ghannouchi, F. (2003) Baseband Digital Predistortion using Subband Filtering Technique. *IEEE MTT International Microwave Symposium Digest*, 2003.

[107] Jeckeln, E.G., Shih, H., Martony, E. and Eron, M. (2003) Method for Modeling Amplitude and Bandwidth Dependent Distortion in Nonlinear RF Devices. *IEEE MTT International Microwave Symposium Digest*, 2003.

[108] Boumaiza, S. and Ghannouchi, F.M. (2003) Thermal Memory Effects Modeling and Compensation in RF Power Amplifiers and Predistortion Linearizers. *IEEE Transactions on Microwave Theory and Techniques, 51(12)*.

[109] Varahman, P., Atlasbaf, Z. and Heydarian, N.V (2005) Adaptive digital predistortion for power amplifiers used in CDMA applications. *Asia-Pacific Conference on Applied Electromagnetics*,2005. pp. 4.

[110] Raithwaite, R.N. and Santa, A. (2008) Wide bandwidth adaptive digital predistortion of power amplifiers using reduced order memory correction. *International Microwave Symposium Digest*, 2008. pp. 1517–1520.

[111] Tarasov, K.N., McDonald, E.J. and Grayver, E. (2008) Power amplifier digital predistortion - fixed or adaptive? *IEEE Conference on Military Communications*, 2008. pp. 1-7.

[112] Wang, A., Yu, J. and Gao, Z. (2009) Research of adaptive digital predistortion technology for wideband OFDM Power Amplifier. *5th International Conference on Wireless Communications, Networking and Mobile Computing*, 2009. pp. 1-4.

[113] Jiang, H. and Wilford, P.A. (2010) *IEEE Transactions on Signal Processing archive, 58(8)*.

[114] Rawat, M., Rawat, K. and Ghannouchi, F.M. (2010) Adaptive Digital Predistortion of Wireless Power Amplifiers/Transmitters Using Dynamic Real-Valued Focused Time-Delay Line Neural Networks. *IEEE Transactions on Microwave Theory and Techniques, 58(1)*: 95-104.

[115] Amandeep Singh Sappal, Dr. Manjit Singh Patterh and Dr. Sanjay Sharma, “Fast Complex Memory Polynomial Based Adaptive Digital Pre-Distorter”, International Journal of Electronics, Vol. 98, No. 7, Page,923–931, July 2011.

[116] Amandeep Singh Sappal, Dr. Manjit Singh Patterh and Dr. Sanjay Sharma, “Digital Pre-distortion of Power Amplifiers using look-Up Table Method with Memory Effects” ICGST-PDCS, Volume 8, Issue 1, Page 39-44,December 2008.

[117] Singla R, Sharma SK, "Low Complexity Look Up Table based Adaptive Digital Pre-distorter with Low Memory Requirements",EURASIP Journal on Wireless Communication and Networking 2012, 2012:43(9 February, 2012)

[118] Cripps, S.C., "Advanced Techniques in Power Amplifier Design", Artech House, 2002

[119] David, S., et al., "Thermal Transients in Microwave Active Devices and Their Influence on IM Distortion,"IEEE MTT Symposium Digest, Phoenix, AZ, Vol. 1,2001, pp. 431–434.

[120] Parker, A., and J. Rathnell, "Contribution of Self Heating to Intermodulation in FETs," IEEE International Microwave Symposium, Fort Worth, TX, WE4C-5, 2003.

[121] Hyunchul, K. and Kenney, J.S. Behavioral Modeling of Nonlinear RF Power Amplifiers Considering Memory Effects. IEEE Transactions on Microwave Theory and Techniques, 51(12), 2003

[122] Navid Lashkarian, Chris Dick, FPGA implementation of digital predistortion linearizers for wideband power Amplifiers; Signal Processing Division, Xilinx Inc., San Jose, USA,.2007.

[123] Lei Ding, Digital Predistortion of Power Amplifiers for Wireless Applications; 2004.

[124] Cripps, S.C.: RF power 2005. Microwave Journal 48 (2005), 22-36.

[125] Euns, C.; Powers, E.J.: A new Volterra predistorter based on the indirect learning architecture. IEEE Tran. On Signal Processing 45 (1997), 223-227.

[126] Kim, J.; Konstantinou, K.: Digital predistortion of wideband signals based on power amplifier model with memory. Electronics Letters 37 (2001), 1417-1418.

[127] Nazim Ceylan Linearization of power amplifiers by means of digital predistortion– Erlangen 2005.

[128] García, P.; Ortega, A.; de Mingo, J.; Valdovinos, A.: Nonlinear distortion cancellation sings LINC transmitters in OFDM systems. IEEE Trans. On Broadcasting 51 (2005), 84-93.

[129] H.S. Black, Translating System, U.S. Patent 1686792, issued Oct. 29, and U.S. Patent 2102671, issued Dec.1937.

[130] H. Ku, M. D. McKinley, and J. S. Kenney, “Extraction of accurate behavior models for power amplifiers with memory effects using two-tone measurements,” in IEEE MTT-S Int. Microwave Symp. Dig., Seattle, WA, June 2002, pp. 139–142.

[131] Altera, Digital Predistortion Reference Design, Altera corporation, 2006.

[132] Shawn P. Stapleton, Agilent EEs of EDA Digital Predistortion of Power Amplifiers in cooperation with: Agilent Technologies, Inc, 2005.

[133] Ding, L.; Raich, R.; Zhou, G.T.: A Hammerstein predistortion linearization design based on the indirect learning. IEEE Int. Conf. on Acoustics, Speech, Signal Proc. 3 (2002), 2689-2692.

[134] Jung, W.J.; Kim, W.R.; Kim, K.M.; Lee, K.B.: Digital predistorter using multiple lookup tables. IEEE Electronics Letters 39 (2003), 1386-1388.

[135] Baudoin, G.; Jardin, P.: Adaptive polynomial pre-distortion for linearization of power amplifiers in wireless communications and WLAN. Int. Conf. on Trends in Comm. EUROCON 2001 1, 157-160.

[136] Marsalek, R.; Jardin, P.; Baudoin, G.: From post-distortion to pre-distortion for power amplifiers linearization. IEEE Electronics Letters 7 (2003), 308-310.

[137] M Faulkner, M Johansson, Adaptive linearization using predistortion experimental results. IEEE Trans Veh Technol **43**, 323–332 (1994).

[138] J. S. Kenney, and A. Leke, "Wireless Report: Power amplifier spectral regrowth for digital cellular and PIIS applications," Microwave Journal, pp. 74-92, Oct. 1995.

[139] L. Ding, G. T. Zhou et al.,”A Robust Digital Baseband Predistorter Constructed Using Memory Polynomials,” IEEE Trans. on Comm., Vol.52, No.1, January 2004.

[140] S.P. Stapleton and F.C. Costescu, An Adaptive Predistorter for a Power Amplifier Based on Adjacent Channel Emissions, IEEE Trans. Veh. Tech., vol. 41, no. 1, Feb. 1992.

[141] Raytheon Systems Co. L.P. Strickland, C.P. Yates, J. Patel, F.G. Muir, L.H. Goree and C.J. Briden, "CE Phase III final Review," July 2001.

[142] P.B. Kenington and D.W. Bennett, "Linear Distortion Correction using a Feedforward System," IEEE Transactions on Vehicular Technology, Vol. 45, No. 1, pp. 474-480, February 1996.

[143] Lei Ding, Zhou G.T., Morgan D.R., Zhengxiang Ma, Kenney J.S., Jaehyeong Kim, Giardina C.R., "A robust digital baseband predistorter constructedusing memory polynomials," Communications, IEEE Transactions on, Jan.2004, Volume: 52, Issue: 1, page 159-165.

[144] Lei Ding, "Digital Predistortion of Power Amplifiers for Wireless Applications," PhD Thesis, March 2004.

[145] Roland Sperlich, "Adaptive Power Amplifier Linearization by Digital Pre-Distortion with Narrowband Feedback using Genetic Algorithms," PhDThesis, 2005.

[146] Helaoui, M. Boumaiza, S. Ghazel, A. Ghannouchi, F.M., "Power and efficiency enhancement of 3G multicarrier amplifiers using digital signal processing with experimental validation," Microwave Theory andTechniques, IEEE Transactions on, June 2006, Volume: 54, Issue: 4, Part 1,page 1396-1404.

[147] Roland Sperlich, "Adaptive Power Amplifier Linearization by Digital Pre-Distortion with Narrowband Feedback using Genetic Algorithms", PhD Thesis, 2005.

[148] Helaoui, M.Boumaiza, S.Ghazel, A.Ghannouchi, F.M., "Power and efficiency enhancement of 3G multicarrier amplifiers using digital signal processing with experimental

validation", Microwave Theory and Techniques, IEEE Transactions on, June 2006, Volume: 54, Issue: 4, Part 1, page 1396-1404.

[149] H. A.Suraweera, K. R. Panta, M. Feramez and J. Armstrong, "OFDM peak-to-average power reduction scheme with spectral masking," Proc. Symp. on Communication Systems, Networks and Digital Signal Processing, pp.164-167, July 2004.

[150] Zhao, Chunming; Baxley, Robert J.; Zhou, G. Tong; Boppana, Deepak; Kenney, J. Stevenson, "Constrained Clipping for Crest Factor Reduction in Multiple-user OFDM", Radio and Wireless Symposium, 2007 IEEE Volume , Issue , 9-11 Jan. 2007 Page(s):341-344.

[151] Olli Vaananen, "Digital Modulators with Crest Factor Reduction Techniques", PhD Thesis, 2006.

[152] Singh, S.P., Subrat Kar and Jain, V.K., "Review of physical layer related issues in WDM networks: Part-II", IETE Journal of Research, Vol. 50, No.4, pp.269-280, July-August 2004.

[153] Boumaiza, Slim, "Advanced Memory Polynomial Linearization Techniques," IMS2009 Workshop WMC (Boston, MA), June 2009.

[154] Amplifier Pre-Distortion Linearization and Modeling Using X-Parameters, Agilent EEsof EDA. Available at: www.home.agilent.com/agilent/redirector.jspx.

LIST OF PUBLICATIONS

- Ruchi Pasricha, S Kumar "Digital Predistortion of Power Amplifiers using Look-Up Table with Memory Effects for LTE Wireless Systems", EURASIP Journal on Wireless Communications and Networking 2012, 2012:330, doi:10.1186/1687-1499-2012-330, Published: 31 October 2012 . SCI Indexed Impact Factor: 0.87 (Springer)
- Ruchi Pasricha, S Kumar "Modeling of a Power Amplifier for Digital Pre-distortion Applications using Simplified Complex Memory Polynomial", Applied Mathematics and Information Sciences, June 2012. (Accepted) SCI Indexed Impact Factor: 0.642.(Natural Science Corporation).
- Ruchi Pasricha, Sanjay Sharma "Low Complexity look up table based Adaptive Digital Predistorter with low memory requirements" EURASIP Journal on Wireless Communication and Networking 2012, 2012:43, doi: 10.1186/1687-1499-2012-43 pp 1-8 , Published: 9 February,2012. SCI Indexed Impact Factor: 0.87 (Springer).
- Ruchi Pasricha, Sanjay Sharma "An FPGA Design of Fixed point Kalman filter" ICGST journal of DSP, Vol. 9/1, pp 1-9, June 2009.
- Ruchi Pasricha, Sanjay Sharma " Linearization of Volterra series model Power amplifier for wideband WCDMA base stations" IJST ,Vol. 2/3,pp 32-33,March 2009.
- Ruchi Pasricha, Sanjay Sharma "Digital Predistortion in WCDMA Power amplifier using Embedded Processor" ICGST journal of PDSC, Vol. 8/1, pp 1-10, December 2008.