LINEARIZATION OF RF POWER AMPLIFIERS WITH MEMORYLESS BASEBAND PREDISTORTION METHOD

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ABSTRACT

LINEARIZATION OF RF POWER AMPLIFIERS WITH MEMORYLESS BASEBAND PREDISTORTION METHOD

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In modern wireless communication systems, advanced modulation techniques are used to support more users by handling high data rates and to increase the utilization efficiency of the limited RF spectrum. These techniques are sensitive to the nonlinear distortions due to their high peak to average power ratios. Main source of nonlinear distortion in transmitter topologies are power amplifiers that determine the overall efficiency and linearity of the transmitter. To increase linearity without sacrificing efficiency, power amplifier linearization techniques may be a choice. Baseband predistortion technique is known to be one of the optimum methods due to its relatively low complexity and its convenience for adaptation. In this thesis, different memoryless baseband signal predistortion methods are investigated and analyzed by simulations. Look-Up Table(LUT) and Polynomial approaches are compared and LUT approach is found to be better in performance. Parameters, like indexing, training sequences and training duration are evaluated. An open loop testbench is built with a real amplifier and a different LUT predistortion method that is based on amplifier modeling is offered. It is evaluated by using two tone test and adjacent channel power suppression with 8PSK data. Also, some Look-Up Table parameters are re-investigated with the proposed method. The performances of the proposed method in different amplifier classes are observed. Along with these studies, a list of prerequisites for design of a predistortion system is determined.

Keywords: Predistortion, Linearization, Power Amplifier, Look-Up Table

ÖZ

RF GÜÇ YÜKSELTEÇLERİN HAFIZASIZ TEMELBANT ÖNBOZUNUM YÖNTEMİ İLE DOĞRUSALLAŞTIRILMASI

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Modern kablosuz iletişim sistemlerinde, sınırlı RF spektrumu kullanımının verimliliğini artırmak ve daha hızlı verileri destekleyerek kullanıcı sayısını artırmak için gelişmiş kiplenim teknikleri kullanılmaktadır. Bu teknikler, yüksek tepe-ortalama güç oranlarından dolayı doğrusal olmayan bozunumlara karşı hassastırlar. Gönderici yapılarındaki doğrusal olmayan bozunumların temel kaynağı, göndericinin toplam verimlilik ve doğrusallığını belirleyen güç yükselteçlerdir. Verimlilikten ödün vermeden doğrusallığı artırmak için, güç yükselteçleri doğrusallaştırma teknikleri bir seçenek olabilmektedir. Temelbant önbozunum yöntemi, karmaşıklık ve adaptasyona uygunluk açısından en optimum metodlardan biri olarak bilinmektedir. Bu tezde, hafıza etkisi olmayan farklı temelbant işaret önbozunum metotları araştırılmış ve benzetimlerle incelenmiştir. Başvuru çizelgeli ve polinomsal yaklaşımlar kıyaslanmış ve başvuru çizelgeli yaklaşımın daha iyi performansa sahip olduğu gözlemlenmiştir. Indeksleme, alıştırma dizileri ve alıştırma süreleri gibi parametreler değerlendirilmiştir. Gerçek yükselteç içeren açık döngü test düzeneği kurulmuş ve yükselteç modellemesine dayalı farklı bir başvuru çizelgeli önbozunum metodu önerilmiştir. Bu yaklaşım iki ton testi ve 8PSK data ile yapılan yan kanal gücü bastırma miktarları ile değerlendirilmiştir. Ayrıca, bazı başvuru çizelge parame-

treleri burada da incelenmiştir. Yaklaşımın farklı yükselteç sınıflarındaki performansı da gözlemlenmiştir. Bu çalışmaların yanısıra, bir önbozunum sisteminin tasarımı için gerekli olan önkoşullar belirlenmiştir.

Anahtar Kelimeler: Önbozunum, Doğrusallaştırma, Güç Yükselteç, Başvuru Çizelgesi

to the memory of my father

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CHAPTER 1

INTRODUCTION

1.1 Motivation

In modern wireless communication systems, as the information to be transferred becomes denser, high data rates are needed. Correspondingly, as the number of users increases more and more, spectral efficiency of the limited RF spectrum is becoming very critical. High data rates can be obtained by using digital modulation techniques such as Wideband Code Division Multiple Access (WCDMA) and Orthogonal Frequency Division Multiplexing (OFDM). However, these modulation techniques create non-constant envelope signals, i.e. they have very large peak to average power ratios.

Power amplifiers are important parts of radio transmitters. They amplify the signal to a certain power level and drive the antenna to make the transmission. Nevertheless, power amplifiers are inherently nonlinear. Spectral regrowth is an unavoidable result, when a non-constant envelope signal is applied to the power amplifier. Spectral regrowth means adjacent channel interference and as a result, efficiency of the utilization of RF frequency spectrum gets worse. Another consequence is the deterioration of communication quality as the symbols are exposed to these nonlinearities and they are damaged even before the channel. Therefore, as the usage of digital modulation techniques gain popularity, linearity specifications of amplifiers become more and more strict.

In order not to be ruined by these effects and satisfy the linearity specifications, operating the power amplifier at back-off can be one possible choice. However, recent modulation techniques needs larger peak to average power ratios, therefore, backing-off results very inefficient operation. As the main stage that is responsible for large power conversions is the power amplifier, inefficiency of such a block results in very high amount of power to be lost. This power means loss of money and in this technology era, battery life of a cell phone and electricity or cooling needs of a base station are determined with the efficiency of the amplifiers. Furthermore, as the power loss comes out as heat; it increases temperature to very high levels and may damage the device.

Another possible choice is the linearization of power amplifier unless there is chance to change the modulation. There are several methods in the literature and Feedback, Feedforward, Envelope Elimination and Restoration are some of them and they are studied in various works. Predistortion is another possible method and it tries to eliminate the nonlinearities by applying the inverse of them. Baseband predistortion focuses on obtaining the inverse nonlinearities of the amplifier by using the advantages of digital domain, and as its correction capability, correction bandwidth and relative cost is at optimum, it is more attractive.

1.2 Objective

The objectives of this thesis are;

- Understanding the sources of nonlinearities
- Learning the characterization and measurement of linearity
- Understanding the baseband predistortion approaches
- Extracting the difficulties and bottlenecks of a baseband predistortion system
- Gaining experience on algorithms through simulations
- Gaining experience on algorithms through a ‘Hardware in the Loop’ testbench

The main goal of this thesis is to implement a predistorter line-up and run it. In doing so algorithmic approaches are examined and method is investigated deeply. New approaches have been made and tried.

1.3 Outline

In this thesis, digital predistortion linearization method is investigated. In doing so, the studies are started by examining the causes of nonlinearities. Nonlinearities are defined and the consequences are stated. To characterize a power amplifier in linearity point of view, some measures of quantities are needed and recalled. The nonlinearity measurement can be done in several ways and all of these headings are mentioned in Chapter 1. Discussions about Nonlinear Modeling and Efficiency are done in this chapter. Linear Amplification is the main aim and the possible ways of such an amplification is discussed. Linearization Techniques are briefly mentioned and Predistortion is introduced in this part also. Some literature survey is again in this chapter.

In the 2^{nd} Chapter, Theory of Baseband Predistortion is introduced. Baseband predistortion can be realized in several ways. The algorithm can be constructed through the decisions on these ways so a classification is needed and made by examining various works in the literature. In this thesis, as a starting point Memoryless Signal Predistortion Algorithms are interested and evaluated, so the algorithms are briefly mentioned in this chapter. At the end of this chapter, some parameters and line-up component specifications that one should be aware of when design of a predistortion system is aimed are discussed. These are very important headings as each of them affect the predistortion performance more or less.

Simulations done to evaluate predistortion performance are presented in Chapter 3. A realistic amplifier model is stated and two of the Memoryless Signal Predistortion Algorithms mentioned in the 2^{nd} is tested in MATLAB simulation environment. At each algorithm, some other parameters, like Look-Up Table indexing, and data formats used for training, are evaluated. At the end of the chapter, a comparison is made between the two approaches.

In Chapter 4, instead of using an amplifier model, baseband predistortion is tested on a real amplifier. For this purpose an amplifier line-up operating at 100 MHz is designed. Some information about the behavior of this amplifier is presented in this chapter. A new algorithm is tried to be obtained in this chapter as the testbench used in this chapter is an open loop. So the third approach mentioned in the 2^{nd} chapter is used. The experiments made on the testbench are explained in three subsections. In the first subsection, parameters like indexing, Look-Up Table construction approach, and effect of filtering are evaluated. In the second one, Data Formats that are used to model the amplifier and evaluate the predistortion performance are tested. In the last subsection, predistortion performance when different amplifier classes are used is presented.

In Chapter 5, the results obtained throughout the thesis are summarized and works that can be done in the future are stated.

CHAPTER 2

POWER AMPLIFIER NONLINEARITIES and LINEARIZATION

In telecommunication applications, incident power to the receivers should be above the sensitivity level of the system. In order to send a message to far distances, RF signals must have sufficient power to overcome, at least, free space path loss. Hence, Power Amplifier with sufficiently high output power is the last stage of an RF transmitter chain before the antenna along with its harmonic elimination filter. See Figure 2.1

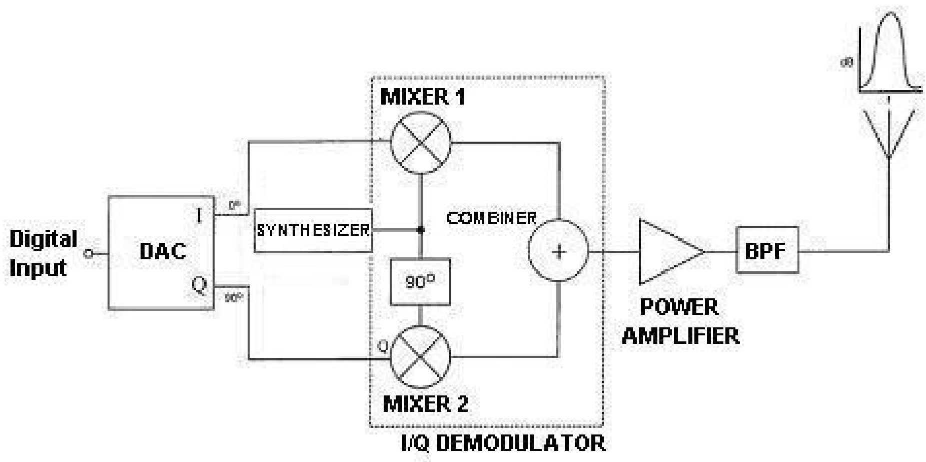


Figure 2.1: RF Transmitter Line-Up

Typically, signals prepared by DSP and RF Modulator units are input to the power amplifiers with a power in the range of miliwatts, and the power amplifiers increase this level to tens or even hundreds of watts. While doing so, DC power is drawn from the supply and converted to RF, thanks to transistor technology.

2.1 Causes for Nonlinearity

Power amplifiers are inherently nonlinear in nature and there are various reasons for this behavior. Probably, the first reason is the transfer characteristics of transistors. Amplification is done owing to the transconductance parameter of a transistor. The slope of the transfer function is the transconductance of the transistor, as shown in Figure 2.2. When transistor's gate voltage is varied, drain current varies correspondingly and amplification is done.

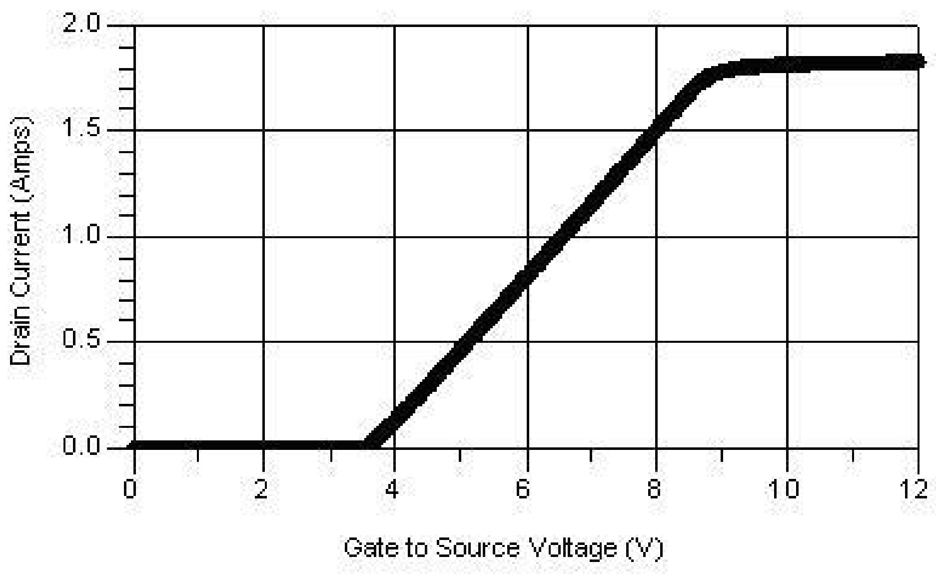


Figure 2.2: Drain Current vs. Gate-Source Voltage Graph of an NMOS Transistor

Since transistor is a real life component, ideal behavior is an unreasonable expectation. Any disorientation in its fabrication (See Figure 2.3) like doping, channel definition, unsymmetrical structure may result in an undefined, non-constant or partially constant transconductance. Therefore, not every input is amplified by the same gain factor.

In the first place, amplifiers have limited power capacity, in that, as its input is increased continuously, after some point its output cannot follow the change in the input and saturates. This can be seen from Fig. 2.2, end of the transconductance line causes saturation.

For MOSFET transistors there is another unwanted behavior which is 'Channel Length Modulation' (BJTs have 'Early Effect' in accordance). As seen on Figure 2.4, as drain voltage increases, drain current in saturation region does not stay constant as hoped. Therefore, drain

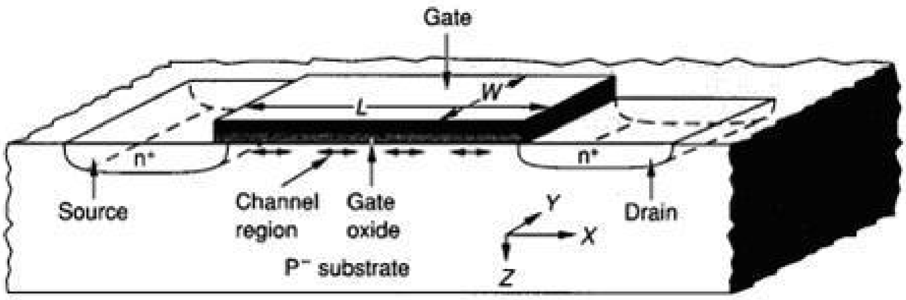


Figure 2.3: Internal Structure of an NMOS Transistor[1]

current variation is exposed to a new nonlinearity source.

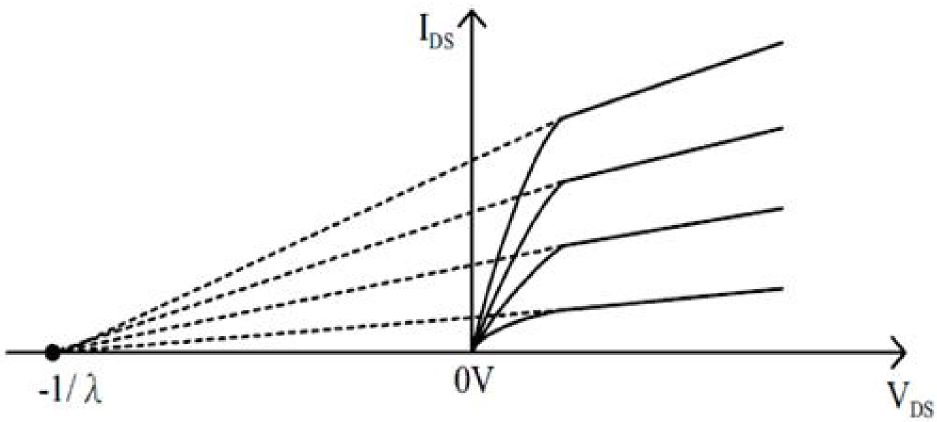


Figure 2.4: Drain Current versus Drain Voltage Graph of an NMOS Transistor[2]

Thirdly, in order to build an amplifier, some external components are needed. These are resistors, capacitors and inductors whose behavior may change with respect to the amount of voltage or current they handle. Therefore, overall characteristics of the amplifier may also be nonlinear because of the components around.

On the other hand, physical structure of the transistor includes some capacitances and these capacitances result in limited operational bandwidth and certain phase change. This causes the amplifier to behave different for different input frequencies. If a wideband signal is applied to the amplifier, the components in wideband signal will not be amplified by same amplifier characteristics. So, nonlinearity will be observed at the output.

2.2 Nonlinearities

All of these causes mentioned above make amplifier responsible for degradation of communication quality. In literature, the nonlinear behaviors of power amplifiers are investigated under certain titles. These can be listed as;

- AM-AM (Amplitude Modulation to Amplitude Modulation) Nonlinearity
- AM-PM (Amplitude Modulation to Phase Modulation) Nonlinearity
- Memory Effects

AM-AM Nonlinearity

Transistors are fabricated with different substrates like Silicon (Si), Gallium Arsenide (GaAs), Gallium Nitride (GaN) etc., each of which has certain power handling capability. For a certain substrate, transistors are manufactured with different sizes to make them handle some power. Gate length and induced channel width are the main parameters that determine power handling capacity.

However, for a certain substrate, gate length and induced channel width; as transistor is driven higher, output power cannot follow the change in input power and saturates. Along with the other minor problems discussed in the previous section, this behavior is the main responsible for the erroneous communication and spectral regrowth which will be explained in Sec. 2.3.

AM-AM nonlinearity can be investigated from Figure 2.5, which is taken from datasheet of a real transistor. In the figure, the output power versus input power and gain versus input power curves are illustrated. The amplifier's gain starts to fall as input power is increased, which is called 'Gain Compression'.

AM-AM distortion is a major problem for non-constant envelope signals. As these signals have information on their amplitudes, saturation of the amplifier means loss of information which degrades the quality of transmission.

In Figure 2.6, a sample input-output relation of a saturated amplifier is shown. Input is sinusoidal, however, at the output, the signal is no more a sinusoidal.

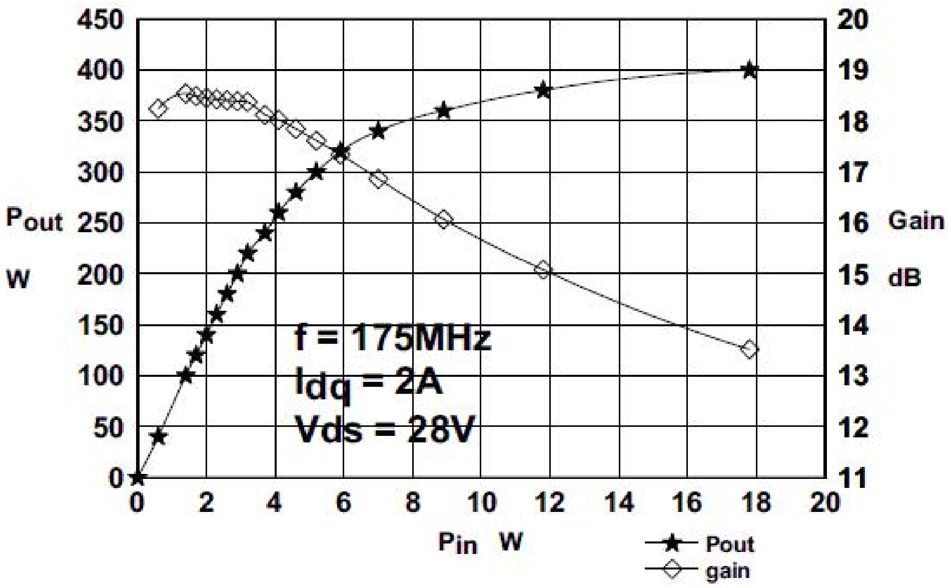


Figure 2.5: Output vs. Input Power for SEMELAB D1029UK Transistor[3]

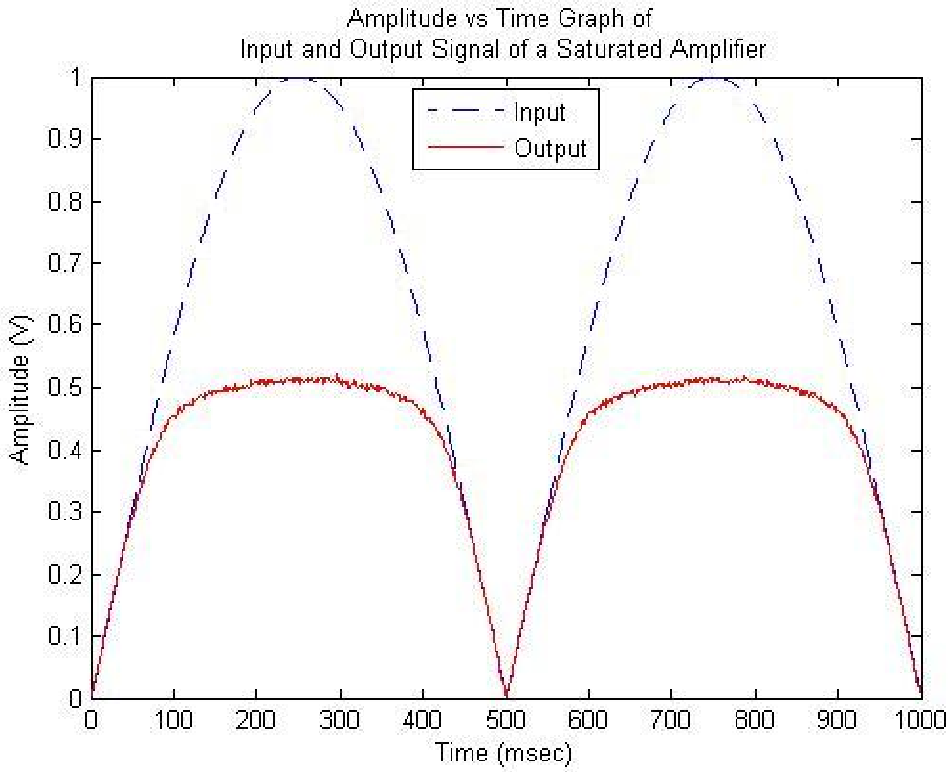


Figure 2.6: Amplitude of Input and Output Signal of a Saturated Amplifier(Norm.)

As mentioned before, AM-AM distortion also causes spreading of the input signal's frequency spectra, called spectral regrowth. In Figure 2.7, the resulting frequency spectrum of a 10 dB amplifier can be seen. Frequency spectrum is widened, and in addition to the main channel, neighboring channels which are called adjacent channels pop up.

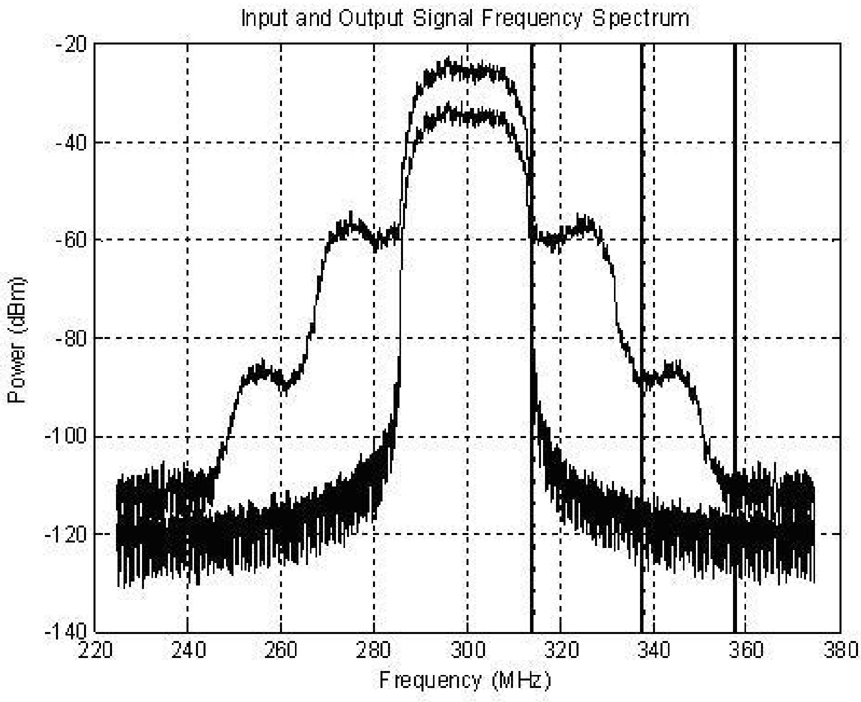


Figure 2.7: Output and Input Power Spectral Densities of an Amplifier(QAM)

AM-PM Nonlinearity

Another main nonlinearity source is the AM-PM distortion. AM-PM can be described as follows; there is a constant phase shift between the output and input of the amplifier in small signal operation. As the output power of the amplifier increases, phase shift starts to change. Therefore, if the input signal is not constant envelope, then, phase modulation is observed on the output signal.

One of the reasons of amplifier's AM-PM response is the internal capacitances of the transistor. See Figure 2.8, which is a High Frequency Model of a MOSFET.

Capacitances exist between various internal terminals of the transistor and all of those capacitances contribute to phase response in the operational bandwidth of the transistor. The major

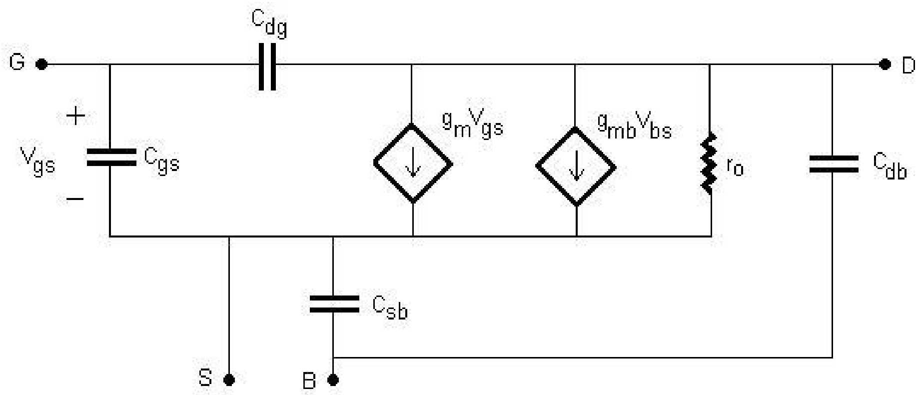


Figure 2.8: High Frequency Model of a MOSFET

responsibility for AM-PM nonlinearity is on C_{gs} capacitance. When input is driven by large signals, C_{gs} capacitance starts to change and therefore, output phase changes correspondingly. In [4], the role of nonlinear C_{gs} capacitor on IMD performance and phase distortion onset for applications in the saturated region was shown. The result of such a distortion is asymmetrical spectrum in adjacent channels as shown in Figure 2.9.

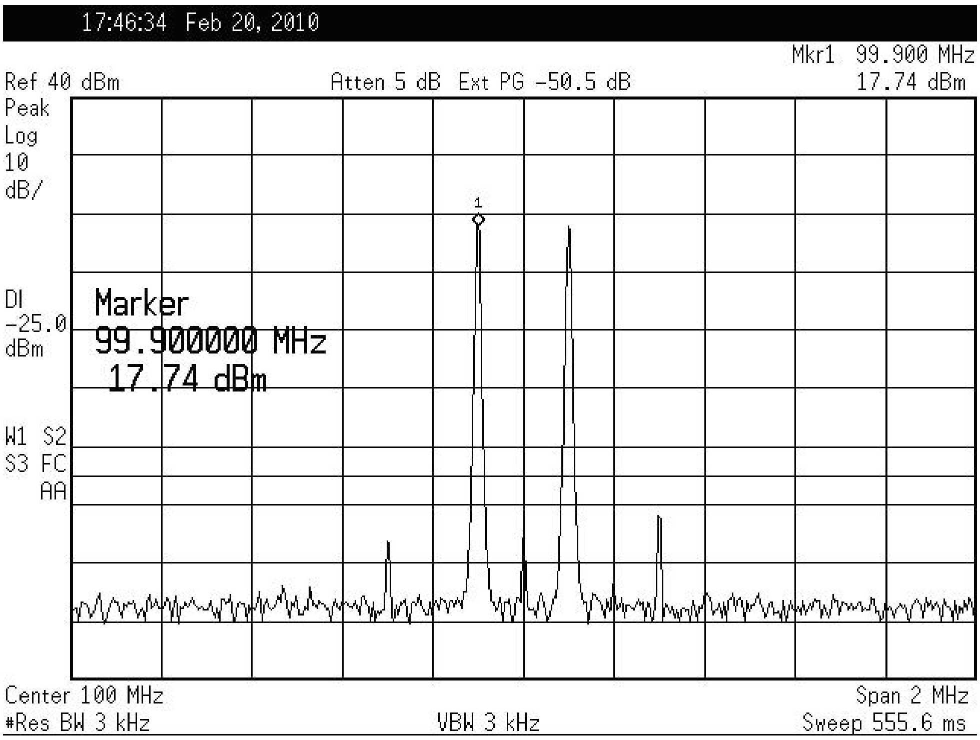


Figure 2.9: AM-PM Effect of an Amplifier(Asymmetrical IMDs)

In [5], authors claimed that the phase component dominates the gain component in its contribution to overall distortion; for conventional devices operated below saturation, and by non-uniform doping, phase nonlinearity of buried-layer GaAs MESFET's can be reduced.

Memory Effects

Memory effects are defined in simple terms as the causes that change the response of the amplifier in time. Transistor's internal structure, bias networks, input and output matching circuits can be named among the sources of memory effects[7].

The effects may occur in short term or long term. Short term is the durations that are in the order of RF signal periods. Long term effects are much higher than that and in the order of period of envelope frequency.

'Short Term Memory Effects' are basically due to the internal inductances and capacitances of the transistor, which store electric or magnetic energy. A capacitor that behaves in a certain way for some RF frequency by storing and pumping energy to the circuit, will not behave exactly the same way for some other RF frequency in the modulation bandwidth. Any wideband signal can be thought as a combination of many sinusoids, and amplifier will not show the same response to each sinusoid in the band.

On the other hand, input and output matching networks are also causes for short term effects. These matching networks are designed with capacitors, inductors and also transformers which are made with ferrites. And ferrites are well-known with their hysteresis effects.

'Long Term Memory Effects' are due to the change in behavior of the amplifier, when its envelope varies. It can be defined as 'Change of Baseband Frequency Response of the Amplifier'.

Long Term Memory Effects can be due to;

- Supply Voltage Variations
- Bias Circuitry Response
- Thermal Effects
- Trapping Effects

In order to amplify a signal, DC supply is needed for both biasing and current feeding. If supply voltage varies in time, this causes the amplifier response to change in time. On the other hand, the supply should be isolated from RF. This is done by RF Chokes which are generally implemented with inductors. Correspondingly, bypass capacitors are also used for DC isolation and coupling capacitors are used for rectification of DC supply and charge pumping. All of these components are some kinds of filters and have frequency response. The difference in the frequency response of bias circuits in the modulation bandwidth of the signal in hand results in the change of the transfer characteristics of the amplifier dynamically. This is named as bias circuitry response.

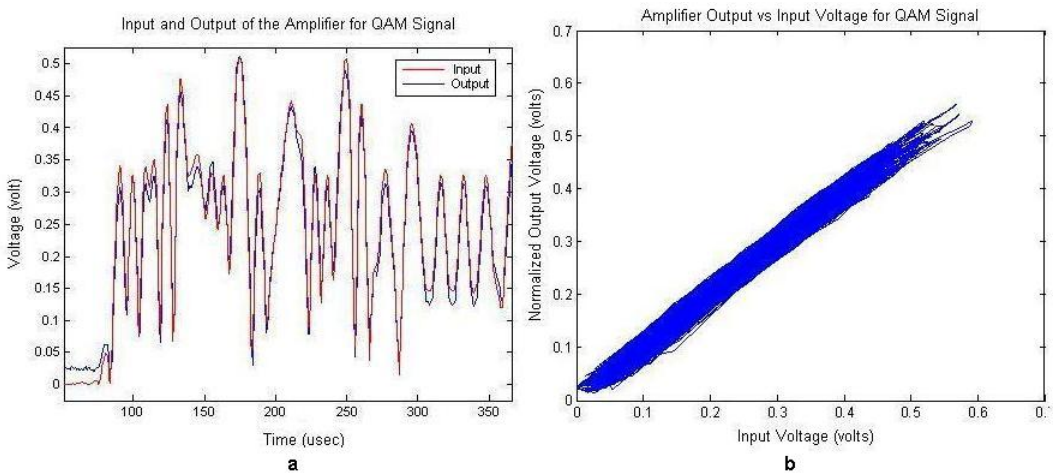


Figure 2.10: Input and Output of the Amplifier(QAM)(a)Time Domain b)Relative)

Thermal effects can be described as follows:

If the modulation of the RF signal is a constant envelope one, then the channel of the transistor will heat up and down uniformly. However, if a non-constant envelope signal is applied, then the dynamics differ. If a peak voltage occurs, transistor draws large amount of current which increases heat dissipation and results a rise in temperature. After the peak voltage, transistor cannot return to the equilibrium temperature fast enough and some of the parameters are changed in chain. For example, the gain is no more equal to the equilibrium gain. Therefore, the following signal will be amplified by a different gain response, which means output of the transistor is now dependent on the past input values.

Thermal effects are more effective below 100 kHz envelope frequencies as its time constants are on the order of microseconds to milliseconds[7]. Thermal effects show itself as spreading of the output vs. input voltage graphs and this can easily be seen in Figure 2.10a, which is the same figure as Figure 2.10b, but plotted relatively.

Thermal effects are also functions of bias point hence a function of class of operation[7].

The imperfections in manufacturing and material itself results in trapping effects. Local potentials and temperature at the channel and channel-oxide interface makes electrons and holes to be trapped and released. This action changes the charge density in the channel and results in change of behavior in time[7].

Trapping effects occur effectively at kilohertz through megahertz frequencies. LDMOS transistors are the ones that suffer less from trapping phenomena. However, GaAs and GaN FETs display more frequently[7].

2.3 Consequences of Nonlinearity

In previous sections, the causes for nonlinear amplification are discussed. In this section, the consequences of this behavior will be mentioned.

Harmonic Generation

AM-AM response of the amplifier results in signals created by the amplifier at frequencies other than the applied frequency. If w is the carrier frequency, signals at $2w$, $3w$ etc. will show up, at the output of the amplifier. These are called harmonics, which are one of the major problems in communication and should be kept below a certain level. This level is specified by standards of the communication method used. If at the output of the amplifier, harmonic suppression is not achieved, they can easily be suppressed by harmonic elimination filters.

Intermodulation Distortion

The non-linear behavior of the amplifier not only results in appearance of signals at harmonic frequencies but also generates signals from a combination of the input multitone. In other words, if more than one carrier frequency is applied to the input, at the output, the combination of the input tones will show up. If amplifier model is assumed to be;

$$y(t) = a_1 x(t) + a_3 x^3(t) \tag{2.1}$$

where a_1 and a_3 are polynomial coefficients, $x(t)$ is the input and $y(t)$ is the output signal and if two tone signal, at w_1 and w_2 frequencies, with Δw seperation is applied to the amplifier;

$$x(t) = V\cos(w_1 t) + V\cos(w_2 t), \tag{2.2}$$

then output of the amplifier becomes;

$$\begin{aligned} y(t) = \left(a_1 V + \frac{9a_3V^3}{4}\right)\cos(w_1 t) + \left(a_1 V + \frac{9a_3V^3}{4}\right)\cos(w_2 t) + \\ \left(\frac{a_3V^3}{4}\right)\cos((2w_1 - w_2)t) + \left(\frac{a_3V^3}{4}\right)\cos((2w_2 - w_1)t) \\ + HighFrequencyTerms, \end{aligned} \tag{2.3}$$

and as seen from the output, there exists tones other than fundamentals and these tones are only Δw away from the fundamentals. See Fig. 2.11.

Furthermore, if amplifier is more nonlinear than just 3^{rd} order, signals at $\pm nw_1 \pm mw_2$ frequencies can be observed. These products are called Intermodulation Distortion (IMD) products. If $|n| + |m|$ is even, then the products fall either around DC or at harmonic frequencies. Therefore, filtering them is easy. However, if $|n| + |m|$ is odd, then the IMD products may fall very close to modulation bandwidth as seen in Eq. 2.3, and filtering them cannot be an option. Among odd orders, third order is very crucial, as it is the highest in power and closest to the modulation bandwidth and most of the time it defines the quality of communication. Frequency spectrum of odd order nonlinearity products can be seen in Figure 2.12

Some of IMD products even fall into the modulation bandwidth and even ideal filters cannot filter them.

Spectral Regrowth

Spectral Regrowth is the same nonlinearity result as IMD products. As mentioned in previous section, two tone input results in appearance of tones other than inputs at output frequency

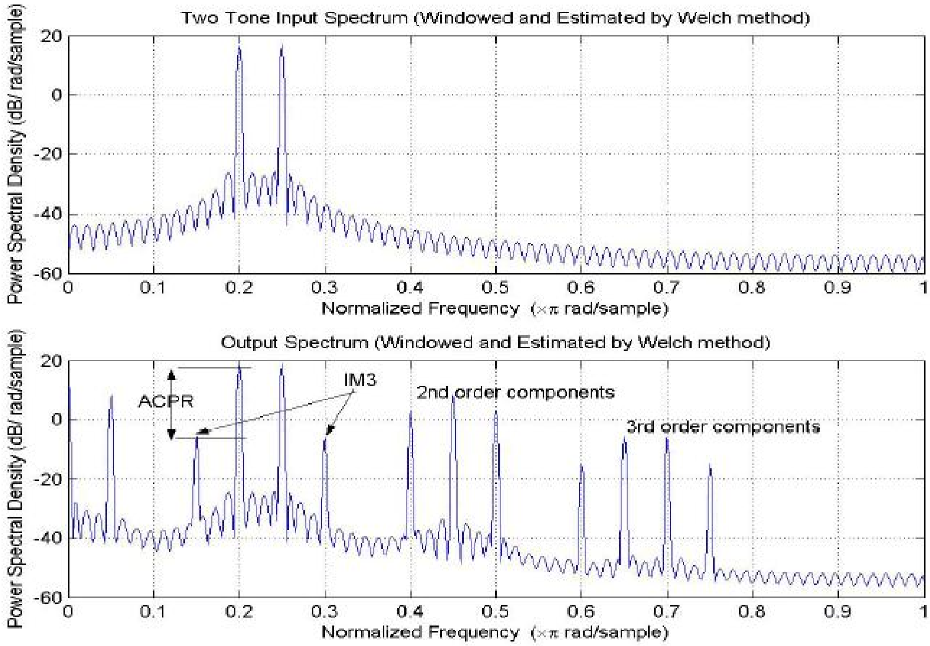


Figure 2.11: Two Tone Response for 3^{rd} Order Nonlinearity[8]

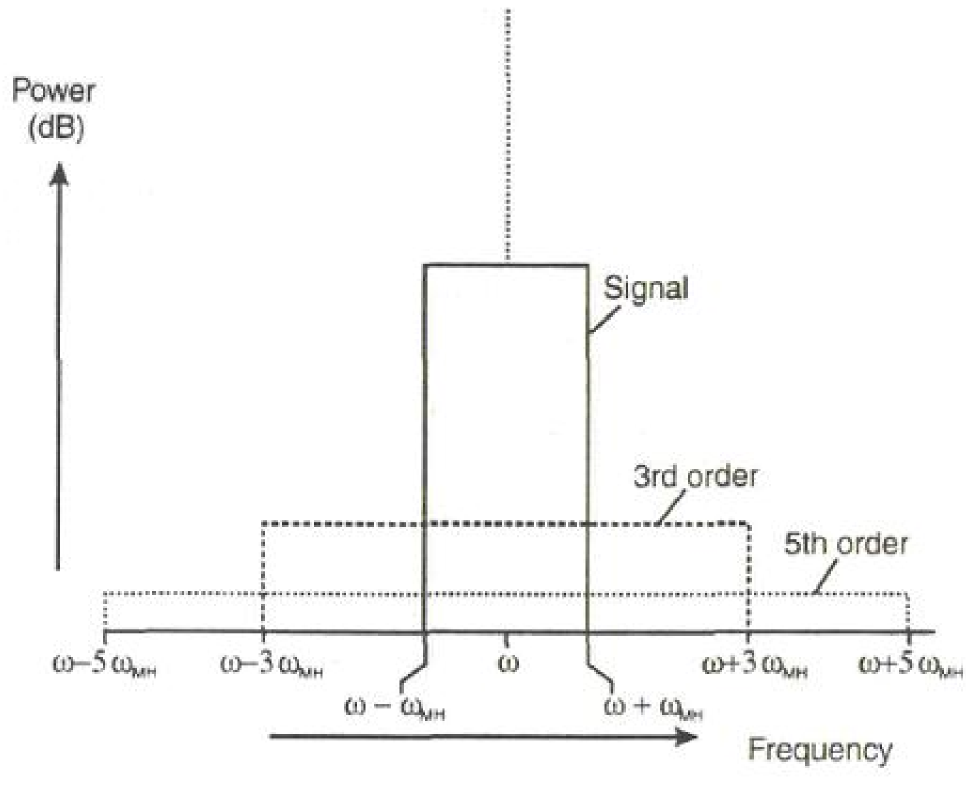


Figure 2.12: Spectral Regrowth for Different Order of Nonlinearity

spectrum. If more than two tones or a wideband signal is applied, signals close to input modulation bandwidth appear at the output spectrum. These unwanted signals make modulation bandwidth wider. As amplifier goes into saturation higher, the frequency spectrum grows higher and becomes wider. This effect is called Spectral Regrowth. See Figure 2.7.

In digital modulation techniques, bandwidth efficiency is the main advantage. In other words, for a certain bandwidth, bit rate can be increased by making signal non-constant envelope. However, spectral regrowth degrades bandwidth efficiency.

In fact, spectral regrowth term is used for modulated signals and IMD is used for multitone excitation.

Cross Modulation

Cross modulation is the transfer of modulation from an undesired carrier to a desired carrier and it is caused by the odd terms of the series expansion for the output current in terms of the input voltage[9]. Modulation on one of the carriers will be coupled to other carriers and modulate them unintentionally. This results in degradation of communication performance. For example, for the following input;

$$x(t) = V_1 \cos(w_1 t) + (1 + m(t)) \cos(w_2 t), \tag{2.4}$$

if amplifier model in Eq. 2.1 is used, then, at the amplifier output;

$$y(t) = \left(a_1 V_1 + \frac{3}{4} a_3 V_1^3 + \frac{3}{2} a_3 V_1 (1 + m(t))^2 \right) \cos(w_1 t) \tag{2.5}$$

will be observed at w_1 frequency. Third coefficient shows that an unmodulated carrier(w_1) is affected by a modulated carrier(w_2)[10].

Desensitization

Similar to cross modulation, desensitization is another nonlinearity problem that multi-carrier systems face. Desensitization occurs due to a single large interfering signal. This is called blocking and the interferer is called blocker[11]. If two tones with different amplitudes are applied to the amplifier, as the big one draws the amplifier into saturation, the smaller one is exposed to smaller gain.

If the input is

$$x(t) = V_1 \cos(w_1 t) + V_2 \cos(w_2 t) \tag{2.6}$$

with $V_2 \gg V_1$, output at w_1 frequency may become;

$$y(t) = a_1 V_1 \cos(w_1 * t) + \frac{3}{2} a_3 V_1 V_2^2 \cos(w_1 * t) + ... \tag{2.7}$$

Then the apparent gain of the w_1 component;

$$a_1' = a_1 \left(1 + \frac{3a_3}{2a_1} V_2^2\right) \tag{2.8}$$

and due to $V_2 \gg V_1$ condition and as a_3 has an opposite sign of a_1 because of compression characteristics, the amplifier is desensitized to small carrier at w_1.

2.4 Nonlinearity Characterization Parameters

In order to evaluate AM-AM response, there are some parameters that enable one to evaluate the linearity of amplifier. These are 1 dB Compression Point (P_{1dB}), Third Order Intercept Point (IP_3).

1dB Compression Point (P_{1dB})

AM-AM responses are monotonically increasing functions in general. As input is getting larger and larger, output goes into saturation and clips at some point. This effect can be seen also in Figure 2.5. For small signals, gain stays constant. As input gets larger, gain deviates from its small signal value more and more.

At a certain point, gain of the amplifier decreases 1 dB from its small signal value. This point is called 1 dB compression point of the amplifier. See in Figure 2.13. It can be input referred (IP_{1dB}) or output referred (OP_{1dB}), however, output referred is generally preferred. Its unit is dBm.

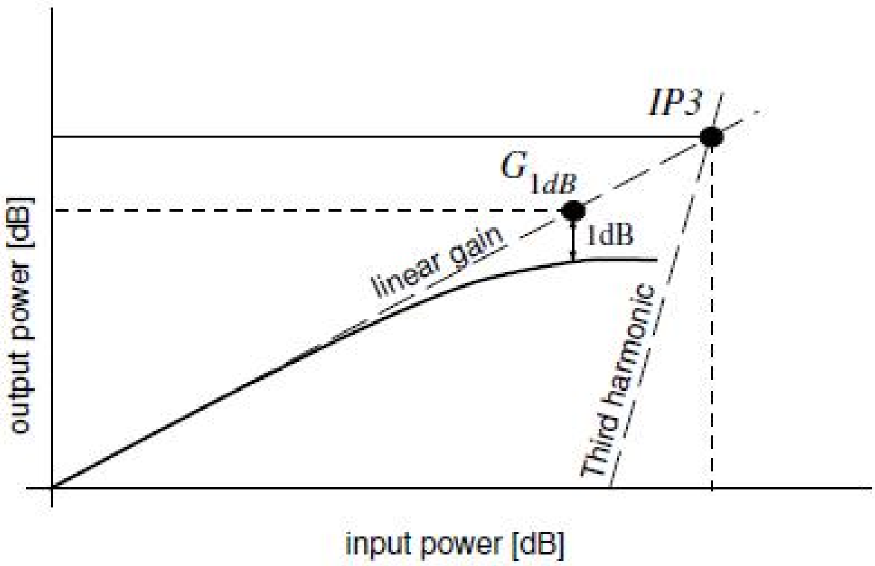


Figure 2.13: P_{1dB} and IP_3 illustration[12]

In order to measure OP_{1dB}, a single tone input signal should be applied and sweeped until amplifier goes into saturation.

On the other hand, there are some amplifier topologies, especially high efficiency classes like AB, which has AM-AM responses different than mentioned above. When input increases, gain does not compress, does not stay constant but increases for certain region. After some point, near saturation, it starts to decrease. For these responses, definition of OP_{1dB} is hard to be made.

Third Order Intercept Point (IP_3)

Intermodulation Distortion products are introduced in previous section. Amplifiers are also characterized by their IMD product performances. As expected, the more the amplifier goes into saturation, the higher the IMD products at the output of the amplifier exist.

The physical dynamics of the amplifier, that create third order IMD products, cause these products to increase three times as much as fundamental tones do. This can be seen from Eq. 2.3, where IMD terms have V^3 coefficients, while fundamentals have V and V^3 coefficients but V is dominating because most of the time $|a_1| \gg |a_3|$.

As input to the amplifier is increased, both fundamental and third order IMD will increase. As the amplifier go into saturation, both of them goes into saturation. However, there exists a hypothetical power that fundamental power and third order IMD power becomes equal if there would not be saturation. This point is called Third Order Intercept Point (IP_3). See Figure 2.13. This point can also be input referred (IIP_3) or output referred (OIP_3), which is more preferred.

The fifth order intercept point (OIP_5), and the seventh order intercept point (OIP_7) can also be defined in the same manner and show some degree of nonlinearity.

Both OP_{1dB} and OIP_3 show the nonlinear characteristics of the amplifier to a certain degree. As they are measured with single tone and two tones, saturation characteristics of the amplifier can be determined only at the frequencies at which tests are done. Therefore, amplifier responses to wideband signals may become different then expected.

2.5 Measures of Nonlinearity

The quality of the communication depends on the quality of the output signal. As amplifier is a non-linear block, output signal should be investigated in order to decide how linear the amplification is or how much distortion is added to the output signal.

In order to do that, there must be some measures of quantity that we rate the output signal. Commonly used four of them are;

- Bit Error Rate or Symbol Error Rate (BER-SER)
- Error Vector Magnitude (EVM)
- Adjacent Channel Power Ratio (ACPR)
- Intermodulation Levels(IM)

Bit Error Rate - Symbol Error Rate

Bit Error Rate(BER) or Symbol Error Rate(SER) are two ways in evaluation of communication system performance. Both are closely related to the constellation of the modulation, and as power amplifier nonlinearity is the main cause for constellation spreading, measurements of BER or SER are possible ways for rating linearity and performance.

SER is defined as;

$$SER = \lim_{N \to \infty} \frac{n(N)}{N}, \tag{2.9}$$

where N is the total number of symbols received and $n(N)$ is the number of erroneous symbols[12].

Error Vector Magnitude

Error Vector Magnitude is a similar conceptual parameter as BER, however, it is more meaningful from an amplifier point of view as it shows how much spreading occurs in the constellation. On the contrary, BER or SER are less quantized versions of EVM as they just state the Bit or Symbol wrong or correct. EVM shows degradation of the performance, even the bit is correct.

EVM is mostly defined in rms form as[12];

$$EVM_{rms} = \frac{\sqrt{\frac{1}{N_s} \sum_{k=1}^{N_s} |\varepsilon(k)|^2}}{\sqrt{\frac{1}{N_s} \sum_{k=1}^{N_s} |A(k)|^2}}, \tag{2.10}$$

where N_s is the number of constellation points used for EVM calculation, and,

$$\varepsilon(k) = \frac{V(k)W^{-k} - C_0}{C_1} - A(k), \tag{2.11}$$

where A(k) is an ideal transmitted constellation point, V(k) received constellation point at instant k, and constants C_0, C_1, and W compensate the constellation offset, constellation complex attenuation and frequency offset, respectively[12].

The maximal authorized values of EVM are different for each standards and vary from 8.5%(GSM) to 23.5%(IS95)[12].

Adjacent Channel Power Ratio

As mentioned in Sec. 2.3, the nonlinear behavior of the power amplifier results in the spectrum to be wider. Other than input bandwidth, the near channels also begin to contain considerable amount of power at the output spectrum. The amount of power in the near channels with respect to the main channel power is defined as a linearity measure and is called "Adjacent Channel Power Ratio". It is measured in dB. The channel adjacent to the adjacent channel can also be called Alternate Channel.

Adjacent channel power measurement is a common way of measuring linearity of an amplifier. As the measurement needs amplifier only as a Device Under Test, it rates the amplifier response only. However, BER-SER or EVM measurements quantify the system performance as a whole. They include modulator and demodulator, mixer and local oscillator effects along with the amplifier.

Intermodulation Levels

Two-tone test is the basic method to measure the Spectral Regrowth distortion. As the amplifier AM-AM nonlinearity causes the signals at $\pm nw_1 \pm mw_2$ frequencies to be observed when two tones are applied, the levels of these products with respect to the fundamental tones are straightforward measures of linearity and are called Intermodılation Levels. This is similar to the ACPR measurement; however, this time as there are tones only, the measurement becomes very easy.

For $|n| + |m| = 3$, the product is called Third Order Intermodulation(IM_3). It is the difference of one of the fundamental tones and 3^{rd} order intermodulation distortion product at the output of the amplifier. Its unit is dBc.

Other than third order, higher order intermodulations are also defined likewise. Fig. 2.14 illustrates the calculation of IM_3, IM_5, IM_7 and IM_9 intermodulation levels.

On the other hand, as the two tones have constant spacing in between, this measurement cannot measure the memory effects, so real performance to wideband modulation may be

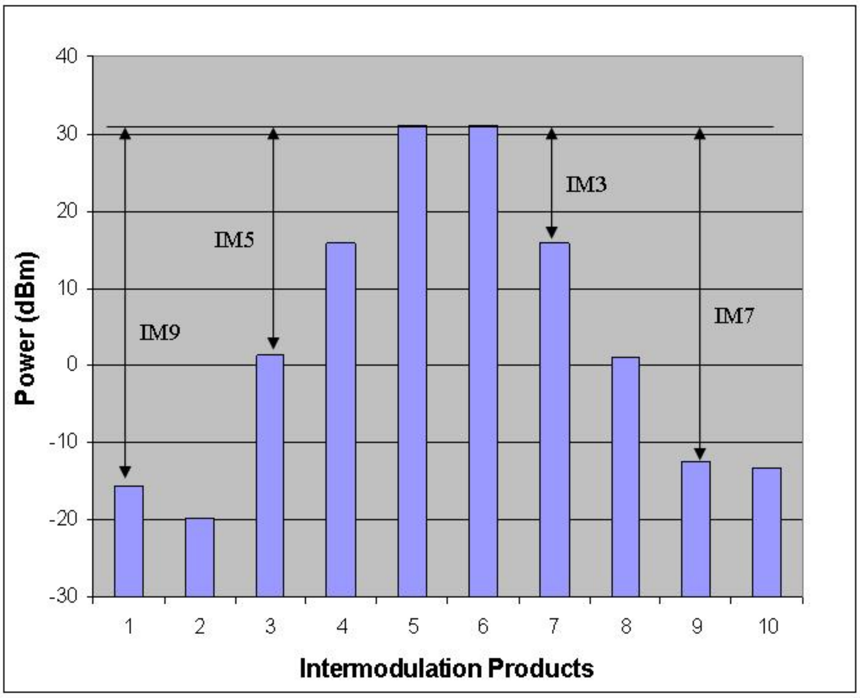


Figure 2.14: Intermodulation Levels

different. Two tone tests should be done for different spacings to get more realistic comments on amplifier nonlinearity.

2.6 Nonlinear Modeling

Power amplifiers are non-linear in nature. In order to linearize them, it is necessary to model the way they behave. Modeling is a quantitative problem and measured data is used for the aim. For modeling, basically, the response is measured and a proper function or combination of functions is fitted to the response.

As mentioned previously, there are AM-AM, AM-PM and Memory Effect nonlinearity problems of power amplifiers and modeling all of these nonlinearities as a black box is called behavioral modeling.

AM-AM and AM-PM responses of amplifiers are main tools to model the amplifier and to estimate the output of the amplifier to any kind of signal. Obtaining these responses is rather easy with respect to memory effects behavior. On the other hand, as the responses are measured by applying single tone and as amplifiers behave different even in the modulation bandwidth, performance is limited to that end. Therefore, AM-AM and AM-PM modeling, which is called Memoryless Modeling, is the basic, efficient, however, limited modeling approach, and most of the time, can be enough for the application in hand.

Along with AM-AM and AM-PM, modeling the memory effects is called Modeling with Memory. Modeling with Memory is, in general, very complicated issue. Functions that are used to model the amplifier is rather sophisticated and parameters are usually hard to find and more sensitive to any change. Therefore, in this respect, modeling with memory should be chosen when Memoryless Modeling cannot satisfy needs.

Memoryless Modeling

There have been many attempts to model the behavior of an amplifier. In Memoryless Modeling, it is assumed that output amplitude of the amplifier depends only on input amplitude. Also, input to output phase shift depends solely on input amplitude. Therefore, the resulting approximation functions, in Memoryless Modeling, are functions of one variable, which is input amplitude.

$$|v_{out}(t)| = G_{amp}(|v_{in}(t)|)\,|v_{in}(t)| \tag{2.12}$$

$$\angle(v_{out}(t)) = \phi_{amp}(|v_{in}(t)|) + \angle(v_{in}(t)) \tag{2.13}$$

In literature, there are several approaches to model amplifier. Some of these models can be listed as[10];

- Saleh Model
- Rapp Model
- Arctan Model
- Nth Order Polynomial Model

Saleh Model models AM-AM and AM-PM responses as[13];

$$G(|x|) = \left\{ \begin{array}{l} \frac{\alpha_a |x|}{1+\beta_a |x|^2} if\ |x| \le \frac{1}{\sqrt{\beta_a}} \\ \frac{\alpha_a}{2\sqrt{\beta_a}} if\ |x| > \frac{1}{\sqrt{\beta_a}} \end{array} \right\}, \tag{2.14}$$

$$\theta(|x|) = \frac{\alpha_p\, |x|^2}{1 + \beta_p\, |x|^2}. \tag{2.15}$$

As seen from AM-AM response, or gain response, up to $\frac{1}{\sqrt{\beta_a}}$ input level, the output compresses as input increases and after that point output saturates. α_a is the small signal gain and β_a determines the amount of compression. AM-PM response has a similar formula and $\frac{\alpha_p}{\beta_p}$ gives the maximum phase shift as input goes to infinity.

Rapp model approximates only the AM-AM response of an amplifier and has a formula

$$G\,|x| = \frac{\nu\,|x|}{\left[1 + \left(\frac{\nu |x|}{A_0}\right)^{2p}\right]^{\frac{1}{2p}}}. \tag{2.16}$$

The Rapp model has three parameters ν, A_0 and p : The parameter ν is the linear small-signal gain. A_0 is the limiting output amplitude. p is the smoothness factor. It controls the smoothness of the transition of the AM/AM response from the linear region to the limiting region[14].

Arctangent model offers the same formula for both AM-AM and AM-PM responses[8];

$$G\,(|x|) = |\gamma_1 \arctan(\zeta_1\,|x|) + \gamma_2 \arctan(\zeta_2\,|x|)|\,, \tag{2.17}$$

and

$$\theta\,(|x|) = \angle\,(\gamma_1 \arctan(\zeta_1\,|x|) + \gamma_2 \arctan(\zeta_2\,|x|))\,, \tag{2.18}$$

where γ_1 and γ_2 are complex numbers and ζ_1 and ζ_1 are real numbers and they are found by curvefitting[15].

Taylor Theorem implies any nonlinear behavior can be modeled as a Taylor series polynomial. As infinite number of polynomials may be needed for convergence to original nonlinearity, there is always a trade of between how much accuracy the modeling needs and how many coefficients the system can handle. It is known from amplifier behavior that the order of nonlinearity may become very huge. For example, if the output of an amplifier is investigated, 10th to 15th harmonics can be seen. However, at the same time, it can be seen that harmonics of the amplifier always decreases, which makes neglecting higher order terms after some degree possible.

The most important nonlinearities that amplifier designers face are the harmonics and IMD products. Harmonics can easily be eliminated with a low pass or band pass filter and most important ones are 2nd and 3rd harmonics due to their closeness to fundamental and due to their power level. On the other hand, among IMD products, third order and fifth order ones are more important due to their closeness and power level. Considering these, modeling the amplifier up to fifth order seems to be reasonable.

$$y(t) = a_1 x(t) + a_2 x^2(t) + a_3 x^3(t) + a_4 x^4(t) + a_5 x^5(t) \tag{2.19}$$

On the other hand, depending on the nonlinearity under investigation, some more terms may be further neglected. It is known that even order terms have influences on harmonics and bias point. Therefore, if harmonics are not one of the interests, and if Memoryless Modeling is the aim then even order terms can be neglected.

$$y(t) = a_1 x(t) + a_3 x^3(t) + a_5 x^5(t) \tag{2.20}$$

If 3^{rd} order IMD products are the only interests and if the effect of fifth order term on 3^{rd} order IMD can be neglected, then fifth order term can be omitted.

$$y(t) = a_1 x(t) + a_3 x^3(t) \tag{2.21}$$

Modeling with Memory

Memory Effects of the amplifiers are not apparent when input signal is narrowband. When wideband signals are used, like WCDMA, these effects become effective. Therefore, for wideband signals, memoryless models of the amplifier cannot give correct expectations. Modeling with memory is more complicated but gives more satisfactory results for these signals.

As mentioned before, memory effects are both time dependent and frequency dependent effects. In order to include memory effects in modeling, AM-AM and AM-PM responses should include these time and frequency dependence. In literature, there are several approaches. Some of the most important ones are[16];

- Volterra Series Model
- Wiener Model
- Hammerstein Model
- Wiener-Hammerstein Model
- Memory Polynomial Model

Volterra series is the generic method among all, however, as number of coefficients increase exponentially when nonlinearity order increases, this approach becomes unattractive[16]. Wiener, Hammerstein, Wiener-Hammerstein and Memory Polynomial Models are special cases of Volterra Series Model. Memory Polynomial Model will be explained below. For more information about the other models refer to [16].

Memory Polynomial Model

Memory Polynomial Model models the amplifier AM-AM and AM-PM responses with memory effects at the same time. It is a special case of Volterra series and the model is[12];

$$y(n) = \sum_{k=1}^{K} \sum_{l=1}^{L} f_{kl} x(n - ln_0) \left| x(n - ln_0) \right|^{k-1}, \tag{2.22}$$

with K, L is the degree of nonlinearity and memory length of modeled amplifier, respectively, f_{kl}'s are complex coefficients that are intended to be determined and n_0 is the elementary delay of memory.

2.7 Efficiency

2.7.1 Power Efficiency

Conversion is a lossy work to do. While converting DC to RF, some of the power is dissipated in the RF transistor. As the output power is huge, dissipation is also huge and more critical. The issue, in general, is to reduce the dissipated power, which we call increasing efficiency. Efficiency can be described as;

$$\eta = \frac{P_{out}}{P_{total}} = \frac{P_{RF_{out}}}{P_{DC_{drawn}}} \tag{2.23}$$

If efficiency of amplification is not good enough, this results in both loss of money and emission of heat.

Heat results in an increase in transistor's junction temperature. So, transistor may burn and device cannot operate if the heat sink of the amplifier is insufficient. Therefore, junction temperature of a transistor should always be kept below from a certain temperature. This can be done by heat sinks or forced cooling or by increasing efficiency of the amplifier, which is possible by considering amplifier classes.

Class Selection

In order to handle the inefficiency of amplifiers, throughout the power amplifier history, many approaches of amplifier topologies have been offered. Most common ones are A, B, AB, C. Lately announced ones are D, J, etc. All of them have their own advantages and disadvantages.

Class A amplifiers are the most inefficient amplifiers of all. Ideal efficiency is 50%, which is highly theoretical and obviously the disadvantage of this class. On the other hand, Class A amplifiers, due to their high bias points, do not distort the signal and make linear amplification.

Class B is a middle stage efficient amplifier. Ideal efficiency is 78,5%. Nevertheless, as no bias drawn, positive cycles of the RF are amplified and signal is distorted, which means nonlinear amplification. There is an intermediate class which is called Class AB. It takes the advantages of Class A and Class B amplifiers as the name implies. Efficiency and Class Selection relation can be seen in Figure 2.15.

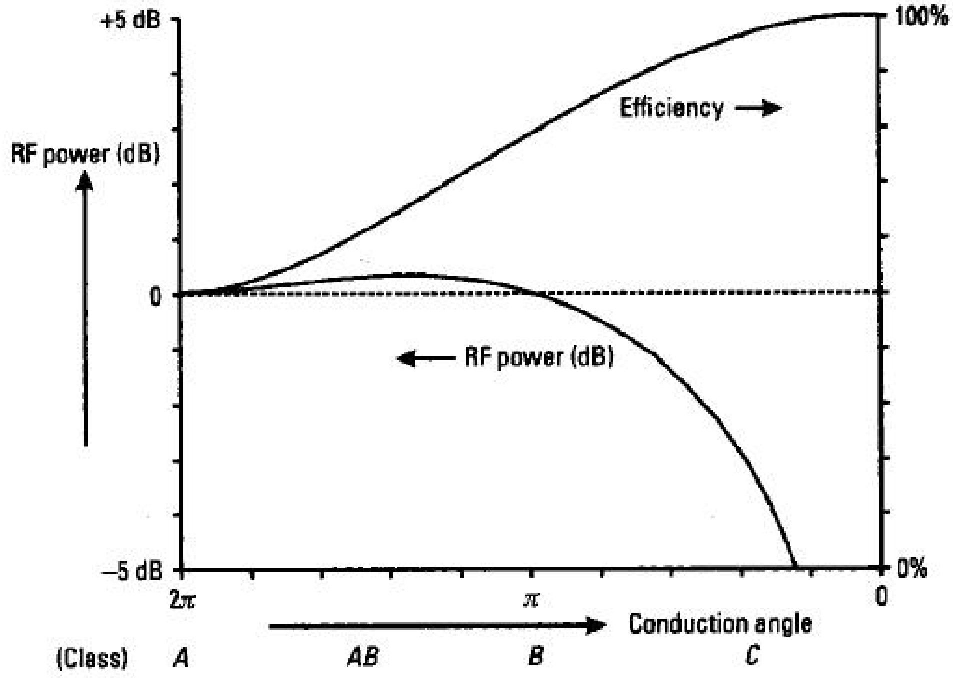


Figure 2.15: The Efficiency and Output Power vs. PA Class of Operation[17]

Classes of amplifiers are introduced according to their bias conditions and conduction angle. However, recently announced ones offer some signal logic in it. For example, Class D is based on the idea that "If amplifier dissipates power when both drain voltage and current are nonzero, then put some intentional harmonic to the input in order to lessen the duration of simultaneous nonzero drain voltage and current".

In Class B, C or even in Class D, in order to increase efficiency of the amplifier, bias conditions are diminished and clearness of the input signal is changed, which results nonlinear amplification. Therefore, efficiency seems to be a trade for the linearity in amplification.

Operation at Back-Off

For linear amplification "Operation at Back-off" may be another solution. Operation at Back-off can be defined as taking less average output power from an amplifier which can give more. So Back-off is a different name for inefficient operation.

Amplifiers are designed to give certain amount of power to predefined loads. In Figure 2.17, D1030 transistor is used and an amplifier which can give 350 Watts of power output is designed. As digitally modulated signals have high peak to average power ratios(PAPRs) or Crest Factors, linear amplification of those signals needs care. In order to amplify these signals linearly, even peak voltages should be linearly amplified, which means peak voltages must be amplified up to OP_{1dB} at most, considering the gain compression. OP_{1dB} is 320 Watts for amplifier in Figure 2.17. This means the average output power of the amplifier will be situated at much lower than OP_{1dB} level. Let's say at 160 Watts for the example, which is 3 dB (Crest Factor of Input) below OP_{1dB}. As the average power is at that low level, efficiency is lowered relative to an operation at OP_{1dB} average, 320 Watts average. Then operation is said to be at "Crest Factor" dB backed-off from its optimum efficiency power output. Back-off level and efficiency comparison for a Class AB amplifier can be seen in Figure 2.16.

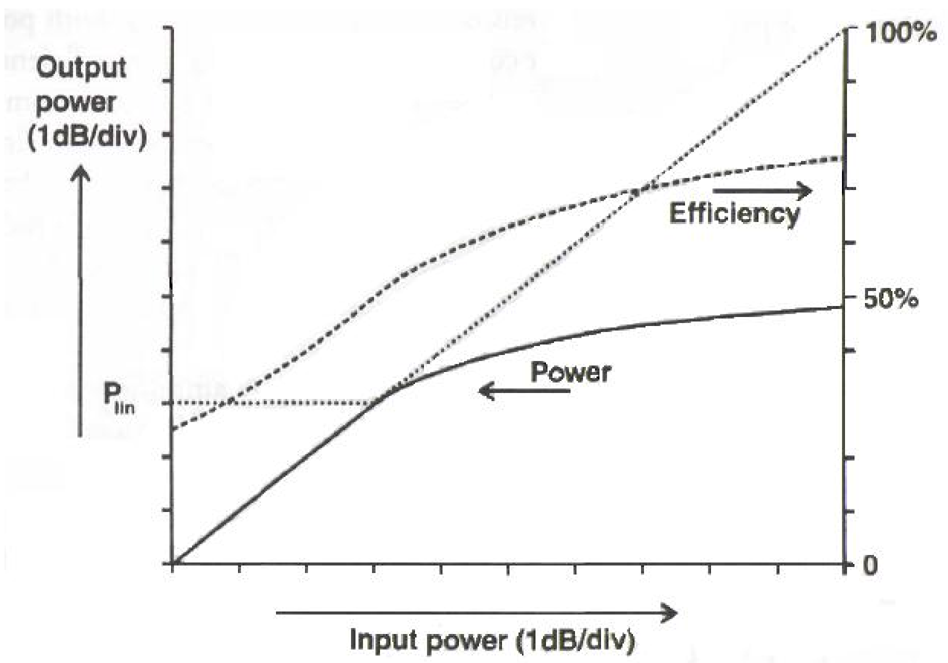


Figure 2.16: Back-off Level vs Efficiency Graph[17]

2.7.2 Cost Efficiency

Another important definition of efficiency is how many watts of power output can be gained per dollar given. In other words, how efficient the transistor is used.

If the aim is to design of an amplifier that is capable of emitting certain amount of output power, then there must be some specifications about how much linear the output signal should be.

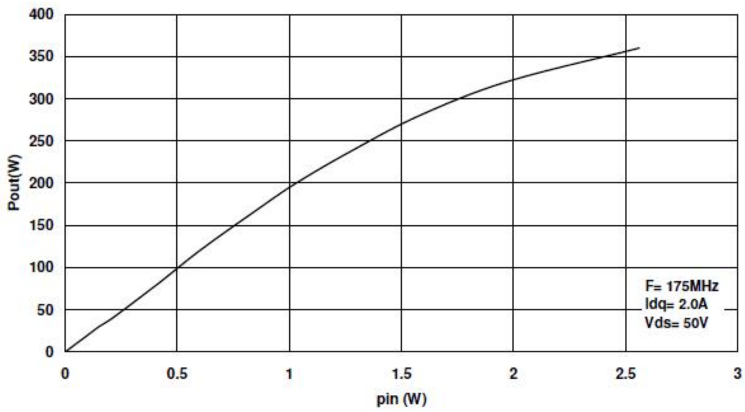


Figure 2.17: Output Power versus Input Power for SEMELAB D1030[18]

As seen in Fig. 2.17, for larger output power, the transistor saturates more and more. For above example, the transistor can emit 320 Watts output, but it is operated at 160 Watts as it is linear up to there. Let's say the aim is to design 320 Watts amplifier. There are three options;

1. First option is to buy two of these transistors and design two amplifiers, which have above characteristics, to get 160W linear power from each and combine them. In this way, 320 Watts of power is taken from two transistors.
2. Second option is to buy a bigger transistor that can give 320W linear power so pay more.
3. Third option is to buy one transistor and design an amplifier, which has above characteristics, to get 320W output. In this way, 320 Watts of power is taken from one transistor. However, at this time, the output is more non-linear than the first two choices.

That is the point where Cost Efficiency vs Linearity appears. There is a way to go on the second option and make it linear.

2.8 Linear Amplification

Up to this point, nonlinear natures of amplifiers are introduced and efficiency problem is mentioned. It is obvious that efficiency of an amplifier is inherently a trade-off to linear amplification.

For an RF designer, the aim is to satisfy the needs of the two ends. The current drawn from the supply should not exceed a certain value for safe operation and at the same time, amplification should be linear enough to meet the ACPR or Two-Tone Test specifications.

In general, linearity specification determines the design and back-off operation is chosen. Then, the dimensions of the amplifier are determined according to the heat that will be dissipated. However, in this fast developing technology era, products are becoming smaller and smaller, therefore, dimensions become also one of the most important criteria for a design.

On the other hand, cost efficiency is another important factor in a design and usage of two transistors instead of one is a loss. For linear amplification there are two ways to follow.

- Crest Factor Reduction
- Linearization

2.8.1 Crest Factor Reduction

As the conflict of efficiency and linear amplification appears in high crest factor modulations due to the need for large back-off, then the first thing to do is to reduce peak to average powers of the modulations. In this way, backing off will be reduced and inefficient operation will be diminished.

There are some possible ways to do that which are[12];

- Clipping
- Coding
- Tone Reservation and Tone Injection

- Partial Transmit Sequence
- Interleaved PC-OFDM
- Constant Peak Power Circuit

However, they, generally, result in data rate loss or they increase complexity. For detailed information, refer to [12].

2.8.2 Linearization

If crest factor reduction techniques are not suitable for operation, as it changes modulation properties, "Linearization" can be a good solution. By linearization, power amplifier linearization is meant. By using external components and structures, it is possible to design a unit, which operates with power amplifier and does necessary operations to make amplifier amplify linearly. This is possible with a number of techniques.

2.9 Linearization Techniques

2.9.1 Feedback Linearization

Among all linearization techniques, feedback linearization is one of the most effective techniques conceptually. Feedback achieves linearization by causing amplifier output to follow amplifier input[10]. A basic form of feedback can be seen in Fig. 2.18.

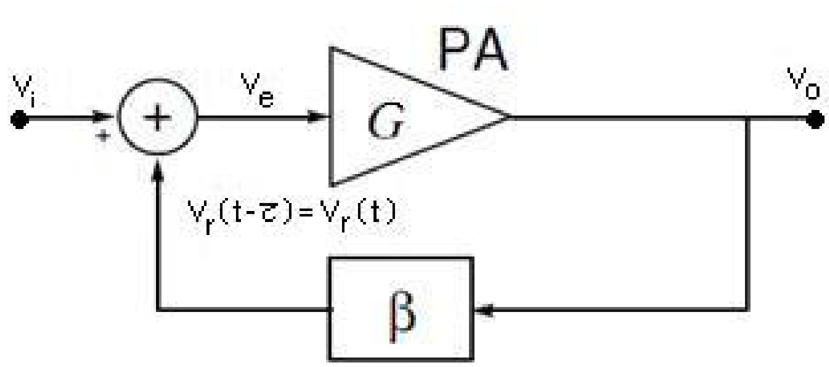


Figure 2.18: Feedback Linearization[10]

If loop delay τ is neglected, then;

$$v_o = \frac{G}{1+G\beta}v_i = G_{eff}v_i, \tag{2.24}$$

$$G_{eff} = \frac{G}{1+G\beta}. \tag{2.25}$$

As seen in Eq. (2.25), gain of amplifier is reduced by $1+G\beta$ factor which is the first drawback of this method. If we differentiate two sides of Eq. (2.25);

$$\frac{\partial G_{eff}}{G_{eff}} = \frac{1}{1+G\beta}\frac{\partial G}{G} \tag{2.26}$$

A differential change in G effects G_{eff} only by a factor of $\frac{1}{1+G\beta}$.

This last equation shows the underlying linearization principle of feedback systems. Gain is reduced by a factor; however, gain compression of amplifier, which is the main responsible for distortion, is reduced by the same factor.

Second drawback of feedback systems is that loop delay τ is neglected in Eq. (2.24). Therefore, it is assumed that the system is non-causal. However, the loop delay exists and determines the maximum modulation bandwidth that the system can linearize.

There are different topologies that feedback linearization can be applied. RF Feedback and Modulation Feedback are the main approaches.

In RF Feedback, output RF signal is fed back by using a feedback network and subtracted from input as in Fig. 2.18. The feedback network used can be active or passive. An amplifier can be used as an active feedback network or resistors and transformers can be used as passive feedback networks[19]. RF Feedback is relatively easy to implement, however, due to non-causality assumption stability should be satisfied. To ensure stable operation, the bandwidth of the input signal must be smaller than a certain value determined by the delay introduced[10].

In Modulation Feedback, complex envelope of the RF is fed back to the input by down-conversion and demodulation. Therefore, the upconverters are included into forward path.

Thus, the feedback can compensate also for the nonlinearity of upconverters[12]. See Figure 2.19. Modulation feedback can be implemented by three basic approaches which are; Envelope, Polar and Cartesian Feedback approaches.

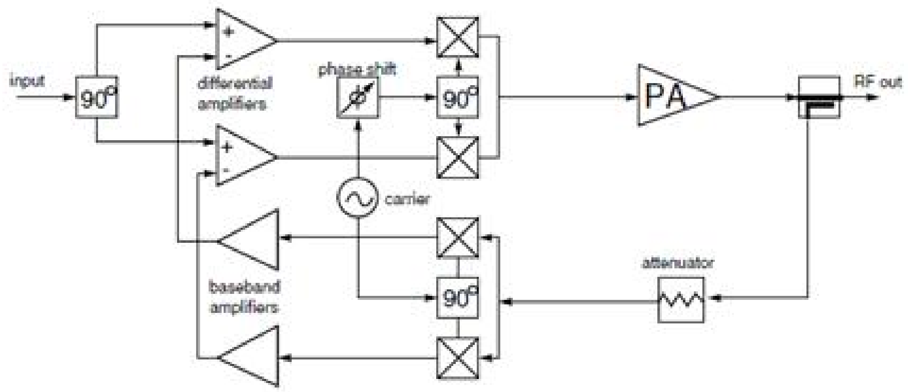


Figure 2.19: Modulation Feedback Linearization[12]

Envelope feedback is the one where only the envelope of the output is fed back to the input and hence it linearizes only AM-AM nonlinearities. Polar feedback method feeds complex envelope back to input and so it is capable of linearize both AM-AM and AM-PM nonlinearities. Cartesian feedback method uses I and Q signals to linearize AM-AM and AM-PM nonlinearities as in Figure 2.19.

2.9.2 Feedforward Linearization

In feedforward linearizer, there are two main loops, which are carrier cancellation loop and error cancellation loop.

In the carrier cancellation loop, the idea is to amplify the signal without considering distortion and take a sample of it. On the parallel branch, input is supplied to a subtractor with a proper delay in order to subtract it from the signal sampled from main branch. The aim is to get only the distortion products in this branch, so this loop is called carrier cancellation. See Figure 2.20.

In the second loop, the aim is to amplify the distortion products obtained in the previous loop with an error amplifier up to a proper level. This level is determined by the distortion products that main amplifier produces. Then, the amplified distortions are subtracted from main branch

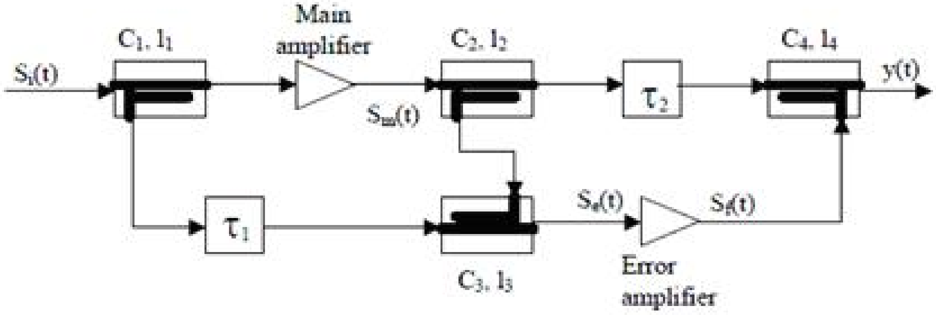


Figure 2.20: Feedforward Linearization[20]

signal and a clear output signal is obtained at the output.

The system has lots of parameters to be optimized for the best efficiency and a given linearity performance[20]. Feedforward linearization has better performance compared to feedback linearization as it has no causality problems. Therefore, modulation bandwidth is not limited by feedback delay but is very sensitive to calibration. Linearity performance of the linearizer depends on how well amplitude, delay and phase matching are maintained within these two loops[20].

The main advantages of feedforward linearization are[12]:

- The performance is independent from the sources of nonlinearity.
- It doesn't reduce the PA gain as feedback linearization does.
- The basic feedforward system is unconditionally stable.
- The bandwidth of signals can be relatively large.

The disadvantages of feedforward are:

- Changes of components' characteristics (due to aging, temperature, . . .) are not compensated in basic feedforward system, so an adaptive compensation system has to be added.
- It is sensitive to the matching of circuit elements over the working bandwidth.
- More components are needed than in the case of feedback linearization.

2.9.3 Envelope Elimination and Restortation

Envelope Elimination and Restoration Linearization Method depends on the principle that any narrow-band signal can be separated into simultaneous amplitude (envelope) and phase modulation components. Using this principle, the input RF signal is resolved into amplitude modulation and phase modulation components and they are combined back after amplification[19].

In order to separate it, at first, RF signal is splitted into two, and one of them is passed through a limiter to make it only phase modulated.

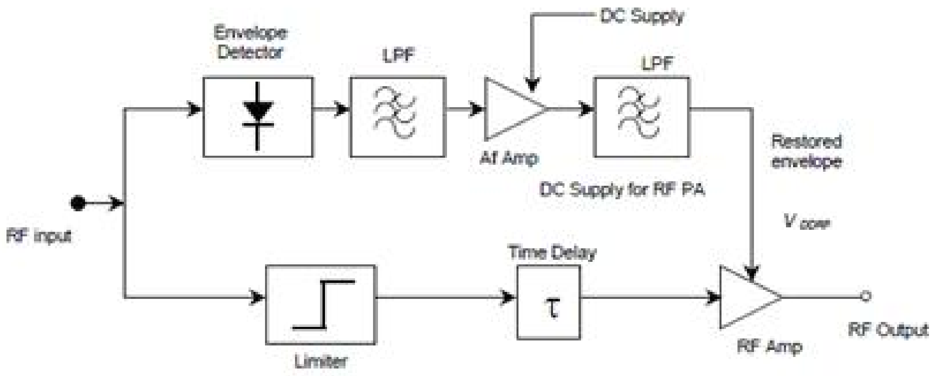


Figure 2.21: Envelope Elimination and Restoration Linearization[20]

Second RF signal is used to get the envelope, i.e., amplitude modulation, of the signal. The envelope is then amplified by ultra efficient, linear audio frequency amplifiers to modulate the supply voltage of the RF amplifier that is used to amplify phase modulated signal. See Figure 2.21.

Amplifiers used in this technique can have 100% efficiency theoretically leading to a very efficient linearizer. However, at high power applications this technique may require power regulators practically difficult to implement and at low envelope levels RF power transistors may cut-off causing distortion[20].

2.9.4 Linear Amplifiation using Nonlinear Components

This method has a similar approach with EER method. The idea is to split the incoming signal into two constant envelope signals and amplify them separately. Due to the constant envelopes high efficiency but nonlinear amplifiers can be used. In this way, high efficiencies can be achieved easily. See Figure 2.22.

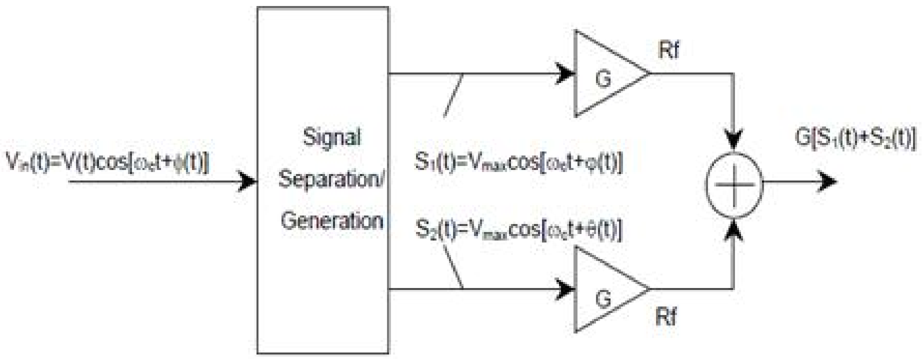


Figure 2.22: Linear Amplification using Nonlinear Components[20]

The main problems with this technique are the strict cancellation requirements for the gain and phase matching of the two RF paths; loss of efficiency during the cancellation process and the thermal tracking problem which would be overcome by integrating both of the amplifier chains in the same module[20].

2.9.5 Predistortion

Predistortion, as the name implies, distorts the incoming signal so that the overall characteristics of predistorter and power amplifier becomes undistorted. In AM-AM nonlinearity point of view, predistortion makes expansion on the input signal and compensates the compression characteristics of the amplifier. Therefore, a predistortion linearizer tries to obtain exact inverse nonlinear characteristics of the power amplifier.

Predistorter can be thought as an eyeglass which tries to create inverse distortion that eyes will create.

Nonlinearities of amplifiers are mentioned in Sec.2.2. AM-AM distortion is the main responsible for all results of nonlinearity and any predistortion algorithm should compensate for that behavior. So gain characteristic of amplifier, which is compressing in general, should be compensated by expansion characteristics of predistortion. See Fig. 2.23. AM-PM nonlinearity,

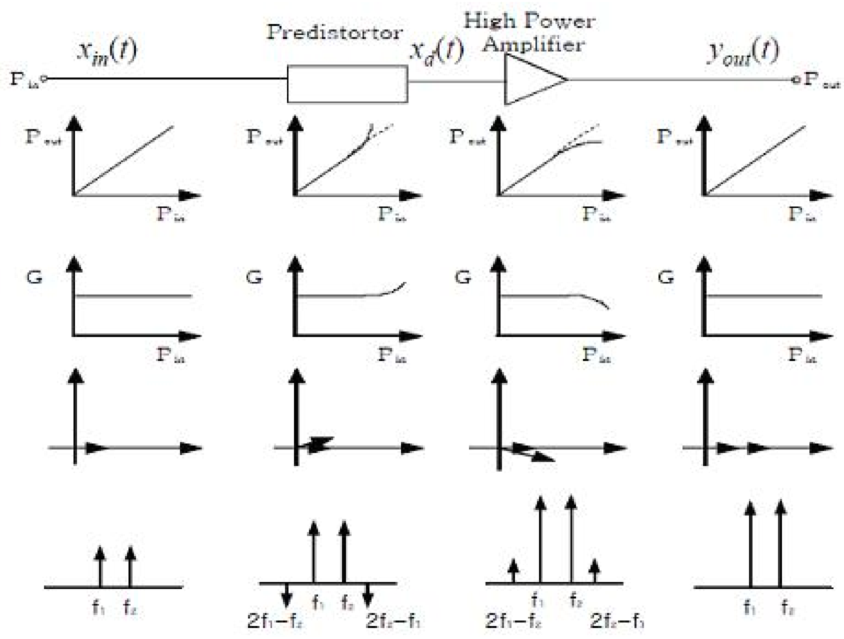


Figure 2.23: Predistortion Linearization[21]

which appears when amplifier goes higher points in saturation region, results asymmetrical IMDs. A predistortion algorithm should also include inverse of that behavior to complete memoryless part of predistortion. If input signal bandwidth is high or amplifier is not designed well from memory effect point of view (like choice of transistor, handling with heat, bias circuitry response), then AM-AM and AM-PM compensation will not be enough as amplifier shows high degree of memory effects. In this case, predistortion algorithm should also include compensation of those effects. Because of this reason, according to memory content, predistortion is classified as;

- Memoryless Predistortion
- Predistortion with Memory

Furthermore, as the power amplifier characteristics may vary due to the temperature changes, aging of the devices or amplifier output and antenna matching, it is desirable to allow the adaptability of the predistorter system[12]. Therefore, predistortion can also be classified as;

- Adaptive Predistortion
- Non-adaptive Predistortion

Predistortion can be done either in analog domain or in digital domain. In this respect, predistortion can be investigated under two titles;

- Analog Predistortion
- Digital Predistortion

2.9.5.1 Analog Predistortion

Analog predistortion implies that all the predistortion operations are done in analog domain. Analog predistortion can be further divided into two kinds;

- RF Predistortion
- Analog IF Predistortion

In RF predistortion, the nonlinear predistorting element or network operates at the final carrier frequency. In Analog IF predistortion, the predistorting element or network operates at a convenient intermediate frequency[22].

Both of the approaches have the same idea that the inverse distortion characteristics of the amplifier should be applied to the input signal by using analog hardware. As mentioned in above sections, intermodulation distortions of amplifier come from odd order polynomial functions that amplifier responses have. The most effective one is third order or cubic nonlinearity. Because obtaining the exact inverse of amplifier characteristic with analog components is hard to achieve, only cubic nonlinearity is tried to be linearized, in general.

There are various ways to implement the inverse cubic nonlinearity in the cubic predistorter. It is generally implemented by using a single diode or a pair of anti-parallel diodes or by using FET transistors[10].

A basic line-up of RF Cubic Predistortion can be seen in Figure 2.24. Incoming signal is splitted into two. Lower branch tries to obtain cubic expansion characteristics and then, by using a variable attenuator and a phase shifter, tuning for cancellation can be done. In the upper branch, input is time delayed for compensation of the delay in the lower branch. They are added and applied to the nonlinear amplifier.

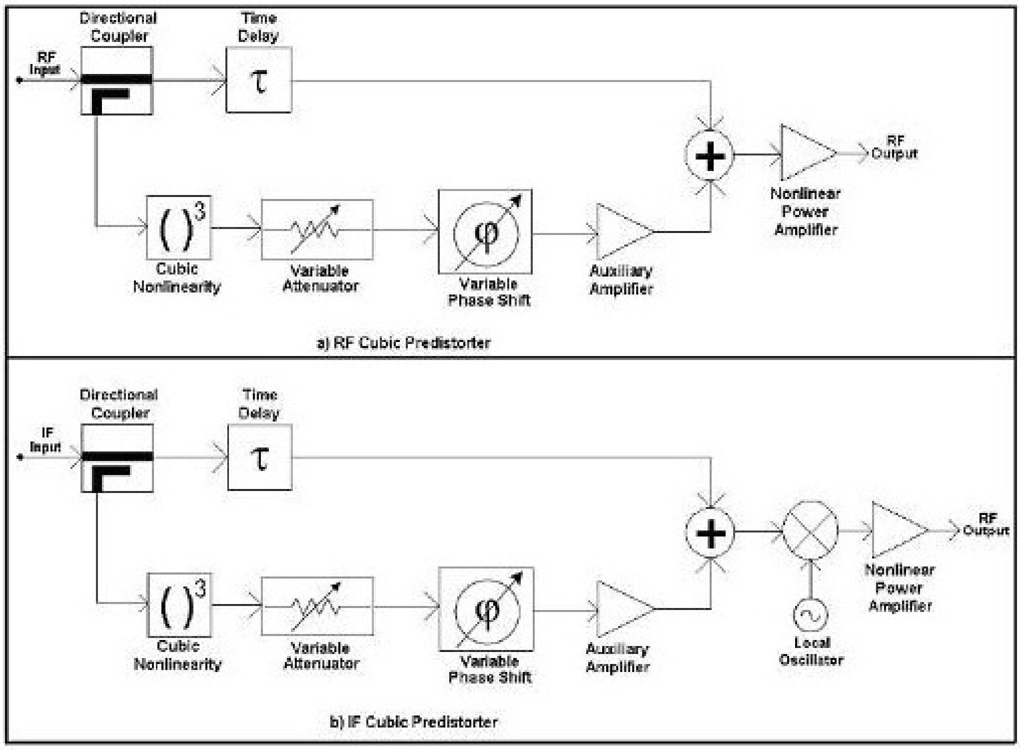


Figure 2.24: RF and IF Predistortion Linearization Line-Ups[12]

Analog IF Predistorter has the same general idea as RF Predistorter; however, it makes the operation at a certain intermediate frequency. The use of the Analog IF predistorters working on intermediate frequency allows to apply the same predistorter for different carrier frequencies[12]. Therefore, the components used up to mixer always operate at a single frequency. In this way, the nonlinear characteristic of these components that depends on operation frequency is not a problem for Analog IF predistorter, unlike RF predistorter.

On the other hand, due to high frequency of operation of RF Predistortion, it is difficult to make it adaptive[12]. In that sense, adaptability of Analog IF Predistortion can be achieved more easily if intermediate frequency is chosen to be lower than RF frequency.

The fundamental advantages of RF or Analog IF predistortion are their ability to linearize the entire bandwidth of an amplifier. Because of this reason, they are ideal to be used in wideband multicarrier systems such as satellite amplifiers or base-station applications[22].

2.9.5.2 Digital Predistortion

Digital predistortion, on the other hand, tries to obtain the inverse characteristics by using the advantages of digital domain. Operations that are done in digital domain for predistortion are much easier than those in analog domain. Consequently, digital predistortion has better performance, in general.

On the other hand, the main drawback of the digital baseband predistortion concept is probably the high power consumption of additional elements - especially fast A/D and D/A converters. The heart of digital baseband predistortion - a DSP, an FPGA or an ASIC, consumes power too, but as it is already present in most of the digital communication systems, one can usually neglect the increase in power consumption[12].

Digital predistortion can be divided into two categories;

- Digital IF Predistortion
- Baseband Predistortion

Although, digital predistortion name is generally used instead of baseband predistortion, as IF predistortion can also be done digitally, baseband predistortion name is more appropriate for I/Q predistortion.

Before explaining Digital Predistortion approaches, it is necessary to talk about transmitter structures. All linearization methods affect or determine the transmitter structure somehow. For digital predistortion the situation is not different. In order to apply digital predistortion, transmitter structure should be chosen appropriately. Complex Envelope Signal Predistortion Transmitter structures can be realized by constructing two different RF upconversion topologies which are[7];

- Analog Quadrature Modulation Architecture
- Digital Quadrature Modulation Architecture.

In Analog Quadrature Modulation(AQM), digitally created I and Q signals are converted to analog with separate DACs. Then, both of the signals are upconverted to IF or RF (dependıng

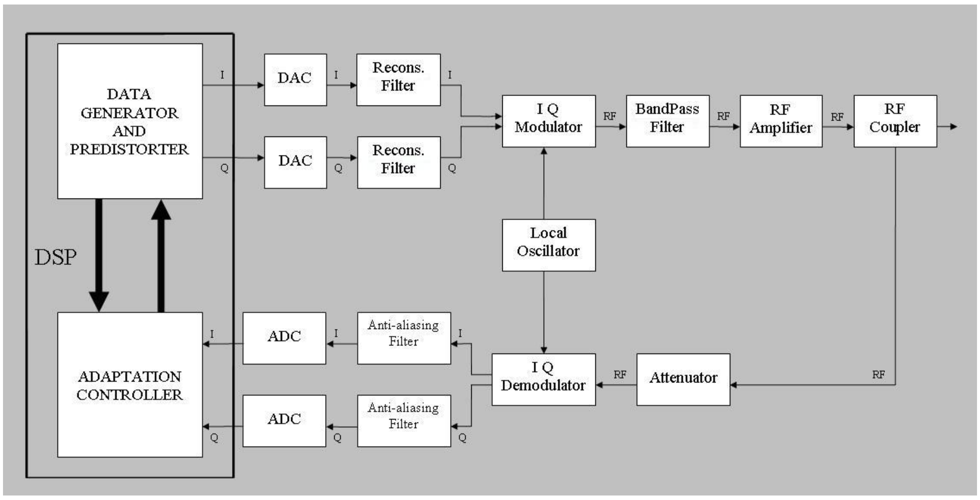


Figure 2.25: Analog Quadrature Modulation Architecture

on the existence of a second upconverter) by IQ Modulator integrated circuits and added together. For adaptation purpose, a sample from the output should be taken and downconverted to baseband I and Q signals by IQ demodulators. ADCs are used to digitize the signals. See Figure 2.25.

On the other hand, in Digital Quadrature Modulation(DQM), it is the IF signal which is created in digital domain from I and Q signals using direct digital synthesizers. IF signal is converted to analog with a single ultra high speed DAC. Then, upconversion to RF is realized by mixers. Adaptation is done accordingly with mixers and high speed ADCs. I and Q signals are obtained from digital IF by downconversion in FPGAs. See Figure 2.26.

With DQM topology, predistortion of the complex envelope input signal can be realized by digital IF or baseband predistortion. On the other hand, if AQM topology is used, then predistortion must be realized by baseband predistortion only.

All three possible ways of digital predistortion (Baseband with AQM, Baseband with DQM, Digital IF with DQM) have advantages and disadvantages.

In Digital IF Predistortion, single DAC is used so I/Q imbalance is avoided, however, as IF signal is at high frequency, the single DAC should be very high speed. At the same time, ADC in the adaptation branch, should also be a high speed one, even if this also avoids any I/Q imbalance. The need for high speed converters can be diminished by using dual conversion topology, where two local oscillators and two mixers are needed, and this time

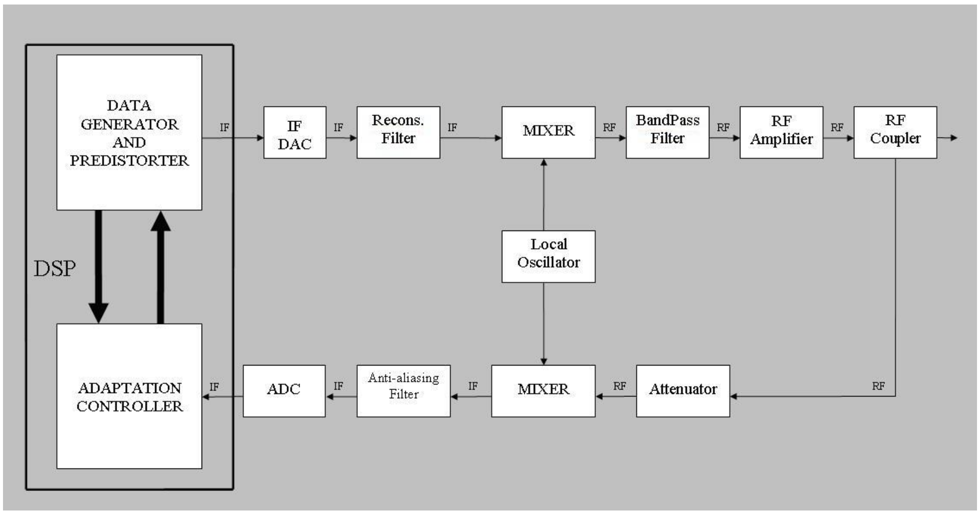


Figure 2.26: Digital Quadrature Modulation Architecture

their nonlinearities would appear.

In Baseband Predistortion, need for dual independent DACs and ADCs inserts other distortions, and this makes predistortion more difficult[7]. However, Digital IF Predistortion has reduced bandwidth with respect to Baseband Predistortion for equal sampling rates[7].

As the main topic of this thesis is Baseband Predistortion, a more detailed structural explanation will be done in Chap. 3.

2.9.5.3 Literature

Digital predistortion dates back to 1989, when Nagata offered an adaptive predistortion linearizer. From experimental results, of nearly -60 dB out-of-band emission and 33% power efficiency achievement was obtained in that work[23]. Then, Cavers proposed a gain based predistorter to lessen the need for memory and decrease the time of convergence that Nagata faced[24].

At the same time, a global optimization algorithm was presented which uses polynomial predistortion and postdistortion[25]. This approach uses polynomial functions to predistort I and Q signals. Another proposed technique was based on the use of a predistortion circuit, whose AM-AM and AM-PM responses are separately implemented as polynomial approximations of the respective responses of the ideal linearizer[26].

After the introduction of this linearization technique, the researches concentrated on improvements. The misalignment caused by the modulation and upconversion circuits as well as temperature and loading changes was analyzed by Faulkner[27]. Effects of imperfections in quadrature modulators and demodulators -gain and phase imbalance and dc offset- on amplifier linearization circuits were investigated in [28].

Optimum table spacing for Look-Up Table Predistorters and performance analyzes were introduced by Cavers[29]. Indexing by investigating the gradient of amplifier AM-AM and AM-PM characteristics was proposed[30].

For polynomial predistortion, a set of orthogonal polynomial basis functions for predistorter modeling was offered. The orthogonal polynomials alleviate the numerical instability problem associated with the conventional polynomials and generally yield better predistortion linearization performance[31].

In [32], polynomial and look-up table predistortion was compared and, considering the practical implementation conditions, LUT method was proven to be more powerful. Although polynomial functions have lower performance than LUT method and require high computational complexity, it is possible to use data pre-processing to improve polynomial predistortion performance[33]

A Multilevel-LUT-based, adaptive, digital, baseband predistortion architecture for RF power amplifier linearization is offered in [34]. Compared with the conventional LUT and polynomial-based predistorters, the proposed algorithm enhanced the dynamic behavior of the treatment while preserving the inherent advantages of a LUT-based approach, including the hardware efficiency and high compensation accuracy[34].

Lately, it is shown that the use of linear interpolation alone significantly reduces the minimum LUT size that is needed to meet the spectral performance required, and the optimal spacing of a linearly interpolated LUT predistorter is derived and is shown to provide a greater performance impact than in the case of the uninterpolated LUT in [35].

Recently, as wideband spectrally efficient modulation schemes appeared, and amplifier memory effects become important. In order to linearize power amplifiers with memory, researches concentrated on Memory Predistortion Algorithms. Memory Polynomial is a possible way and firstly announced in [36].

Other than memory polynomials, Volterra series representation is another approach for memory predistortion algorithms. Actually, memory polynomial is a basic version of Volterra Series. Wiener, Hammerstein, Wiener-Hammerstein and parallel Wiener structures are also special cases of the Volterra series and are used for memory predistortion[37].

In [38], different approaches of predistortion with memory effects are discussed and a digital predistortion solution that enables closed-loop wideband linearization is presented with excellent linearization capabilities.

CHAPTER 3

BASEBAND PREDISTORTION

3.1 Theory

In predistortion, the aim is to obtain the inverse characteristics of the amplifier. Baseband predistortion tries to get these characteristics by modeling the amplifier nonlinearity through the usage of baseband information.

Baseband information, which is I and Q signals, forms the envelope of complex modulations. This can be seen from quadrature modulated signal equations; Eq. (3.1) or equivalently Eq. (3.2).

$$V_{in}(t) = I(t)\cos(wt) + Q(t)\sin(wt) \tag{3.1}$$

$$V_{in}(t) = (\sqrt{I^2(t) + Q^2(t)})\cos(wt + \arctan(\frac{Q(t)}{I(t)})), \tag{3.2}$$

or equivalently,

$$z_{in} = (\sqrt{I^2 + Q^2})\angle\arctan(\frac{Q}{I}) = r_{in}\angle\theta_{in} \tag{3.3}$$

where I and Q signals are the signals created according to complex modulation scheme used and w is the IF frequency that modulator uses, V_{in} is time domain signal and z_{in} is its phasor. r_{in} and θ_{in} are its magnitude and phase components.

From Eqs. (2.12) and (2.13), it is known that, if Eq. (3.3) is used as an input to the amplifier, then at the output;

$$z_{out} = G_{amp}(r_{in})\, r_{in} \angle \left(\theta_{in} + \phi_{amp}(r_{in})\right), \tag{3.4}$$

will be obtained.

If the input envelope versus the output envelope is plotted, then the AM-AM response of the amplifier is obtained. See Fig. 2.10. Hence, it can be said that baseband response is the AM-AM response of the amplifier. Similarly, if the phase difference of input and output signals is plotted with respect to the envelope of the input signal, the AM-PM response of the amplifier will be obtained.

By using this phenomenon, I/Q signals created in signal processors, which can be DSPs (Digital Signal Processors), FPGAs (Field Programmable Logic Arrays) can be used to obtain AM-AM and AM-PM nonlinearities of the amplifier.

If a predistorter with gain and phase distortion functions of G_{pd} and ϕ_{pd} is used before the amplifier as in Fig. 3.1, instead of Eq. (3.4);

$$z_{out} = G_{amp}\left(G_{pd}(r_{in})\, r_{in}\right) r_{in} \angle \left(\theta_{in} + \phi_{pd}(r_{in}) + \phi_{amp}\left(G_{pd}(r_{in})\, r_{in}\right)\right), \tag{3.5}$$

will be obtained.

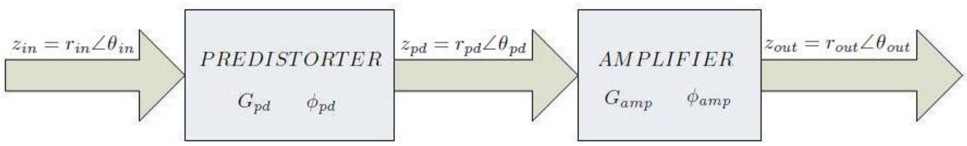


Figure 3.1: Predistortion Operation

Hence, correct G_{pd} and ϕ_{pd} functions are the functions that satisfy the following equations;

$$G_{amp}\left(G_{pd}(r_{in})\right) = G_0, \tag{3.6}$$

$$\phi_{pd}\left(r_{in}\right) + \phi_{amp}\left(G_{pd}\left(r_{in}\right) r_{in}\right) = 0, \tag{3.7}$$

where G_0 is small signal gain of the amplifier.

Figure 3.2 illustrates the operation of predistortion on AM-AM nonlinearity for a single point. If input $r_{in_{lim}}$ is applied to the amplifier when there is no predistortion, then, compressed output, $r_{out_{lim}}$, will be observed. However, if a predistorter that maps $r_{in_{lim}}$ to $r_{in_{sat}}$ is used, then, amplifier will have an output $r_{out_{sat}}$.

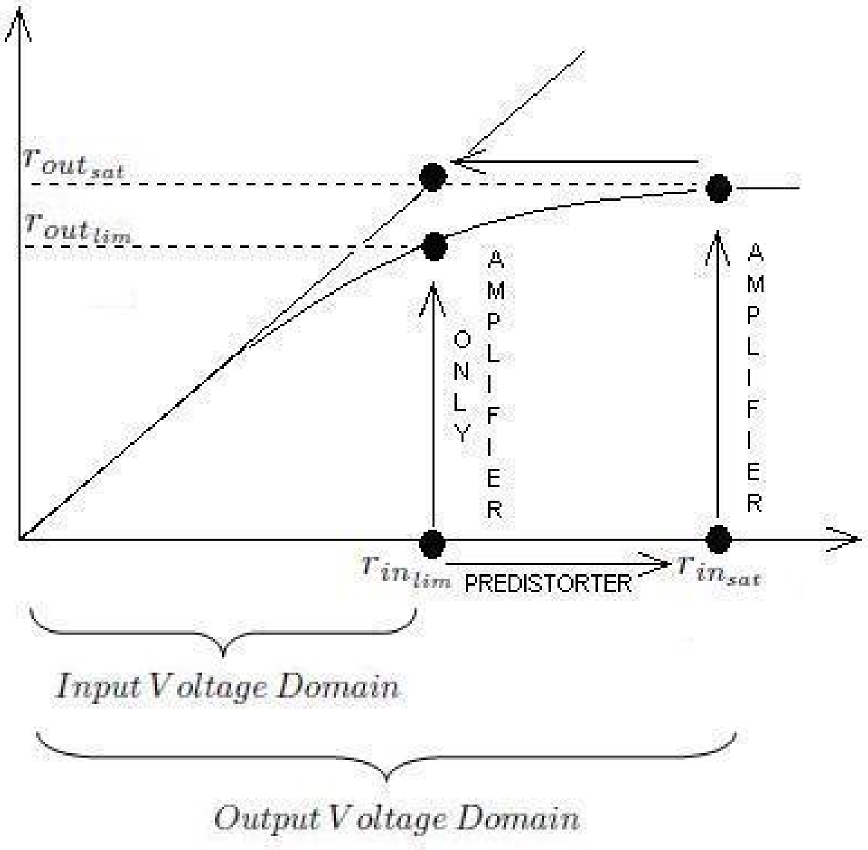


Figure 3.2: Predistortion Theory

Therefore, predistorter and amplifier combination is said to make linear amplification, $r_{in_{lim}}$ to $r_{out_{sat}}$. Predistortion for AM-PM nonlinearity is similar and easier, as there is only sign change.

It should be pointed out that, as output of the amplifier cannot be more than $r_{out_{sat}}$ point, as it saturates there, output of predistorter and amplifier combination, also, cannot be more than $r_{out_{sat}}$. Hence, there is a limit for predistorter input, shown as $r_{in_{lim}}$, and if more than this limit is applied to predistorter, predistorter can only map it to $r_{in_{sat}}$ and clipping occurs at the output of amplifier, which is a nonlinearity that no method can linearize.

Therefore, there exists an Input Voltage Domain, an interval, in which predistortion function is defined. Outside this domain predistortion has nothing to do. Accordingly, predistortion function has also an Output Voltage Domain, an interval in which output is defined, see in Fig. 3.2.

Determination of predistortion function can be realized through various approaches which will be mentioned later in this section.

3.2 Classification

In the literature, there are various kinds of baseband predistortion approaches. In order to gather these methods in a more comprehensive manner, they are classified by different criterions in Fig. 3.3. Position in DSP, memory effect content, adaptability, and formation of predistortion function headings should be decided to realize a predistortion.

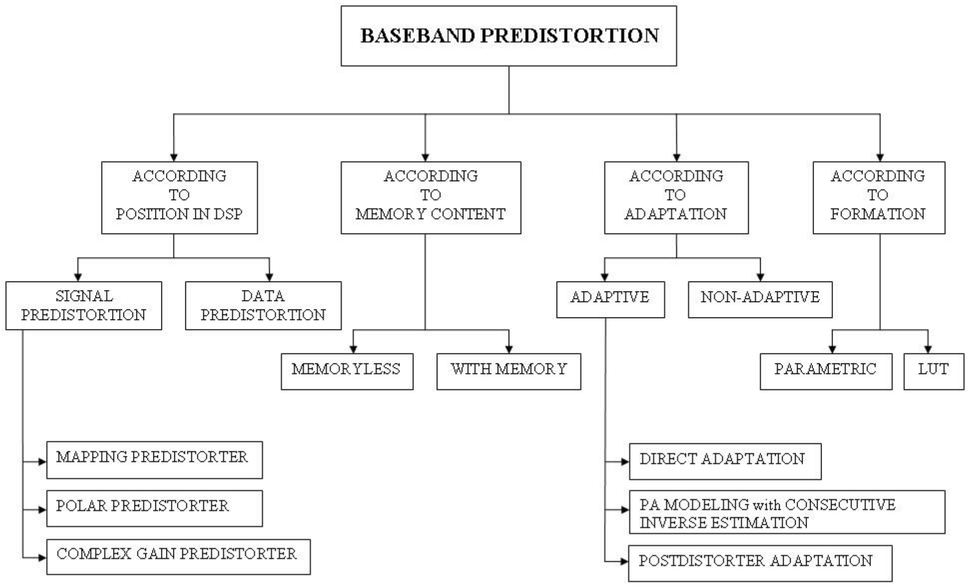


Figure 3.3: Classification of Baseband Predistortion

First criterion is the position of predistorter inside the digital signal processors;

- Data Predistortion
- Signal Predistortion

Data Predistorters tries to compensate for the deformation of constellation diagram. They are simpler, but cannot normally eliminate the adjacent channel emissions[12]. The main drawback of data predistorters is their dependence on the signal constellation and the pulse shaping filter and they do not work well if the processing produces almost continuous input signal levels[16].

Signal Predistorters are independent of modulation format. They, generally, work on the signal, after modulation and baseband pulse shaping. Because of the large dynamics of signal amplitudes, their adaptation is slow in comparison with Data predistorters[12].

Signal Predistorters can be realized by using three structures[39]:

- Mapping Structure
- Polar Structure
- Complex Gain Structure.

Mapping Structure is initially used by Nagata[23]. In this work, 2D Look-up Table, indexed by I and Q, is used to obtain complex predistortion coefficients of complex signals. Due to large two-dimensional look-up table, this method is rather inefficient to implement because of slow adaptation time and high memory requirements. See Fig. 3.4.

Considering the fact that AM/AM and AM/PM responses of the memoryless power amplifier depend only on the input amplitude, two-dimensional look-up table in the mapping predistorter can actually be replaced by two one-dimensional LUTs. This replacement is basically the polar representation of Cartesian domain and is called Polar Structure Signal Predistortion. Polar structure is proposed by Faulkner et al[40]. See Fig. 3.9.

Complex Gain structure employs a complex gain table instead of the gain and phase tables in the polar predistorter and is offered by Cavers[41]. See Fig. 3.5.

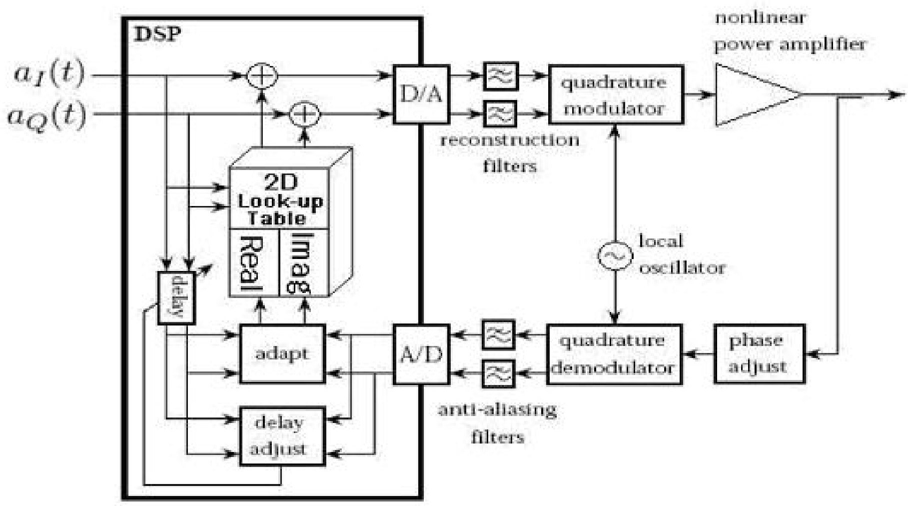


Figure 3.4: Mapping Structure Signal Predistortion Topology[39]

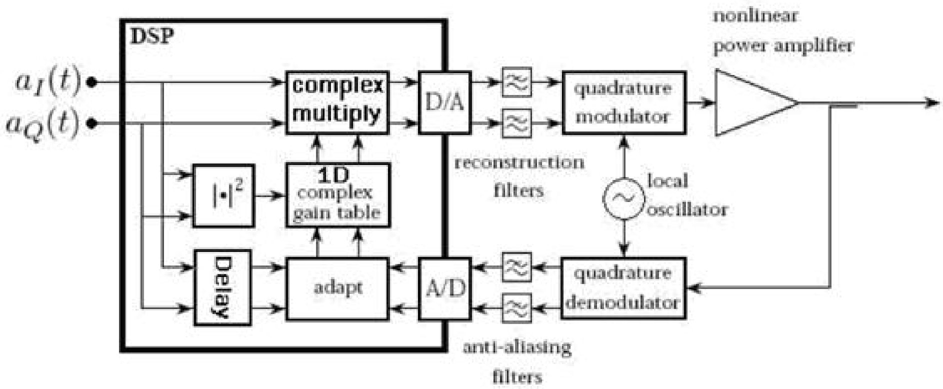


Figure 3.5: Complex Gain Structure Signal Predistortion Topology[39]

According to the capability of compensation of amplifier memory effects, predistortion may need to compensate for the effects of memory. As mentioned in Sec. 2.9.5, baseband predistortion can also be divided into two groups; Memoryless Predistortion and Predistortion with Memory.

In general, for wideband or high power applications, the power amplifier exhibits memory effects, for which memoryless predistorters can achieve only limited linearization performance[16]. Therefore, Predistortion with Memory may become necessary, even though, it is a difficult way of predistortion. To accomplish this task, modeling the amplifier memory effects and compensation of them is necessary and it is generally chosen if Memoryless Predistortion cannot satisfy the needs.

According to the formation of inverse amplifier characteristics, baseband predistortion can be separated into two groups;

- Look-Up Table Predistortion
- Parametric Predistortion.

As amplifier AM-AM and AM- PM characteristics are nonlinear, the inverse characteristics, which are predistorter characteristics, are also nonlinear. If the nonlinear predistorter characteristics are modeled by piecewise linear functions, then this is a Look-Up Table Predistortion. LUT entries holds the boundary values of these piecewise linear functions. On the other hand, if the nonlinear predistorter characteristics can be modeled as nonlinear functions such as Polynomial, Spline, Volterra series etc., then this is called Parametric Predistortion.

LUT can model any inverse characteristic; however, success of parametric modeling is a matter of number of coefficients and fitting accuracy. LUT Predistortion has better performance in Error Vector Magnitude(EVM) measurements; however, discontinuities occurring in LUTs causes adjacent channel power and broadband noise to increase[7].

The advantage of baseband predistortion lies in the possibility of its simple extension to adaptive form[12]. Quadrature Modulation Architectures are suitable for application of adaptation loops as mentioned in Sec. 2.9.5. According to the methods of adaptation, baseband predistortion can be classified into three main approaches[12].

- Direct predistorter adaptation
- PA modeling with consecutive inverse estimation(Extraction)
- Predistortion using postdistorter adaptation (Indirect learning)

In direct predistorter adaptation, the predistorter is adjusted in order to minimize the error between the attenuated output of the amplifier r'_{out} and the original signal r_{in}[12]. This method is a kind of trial and error approach.

$$\frac{r_{out}}{G_0} = G_{amp}\Big(G_{pd}\left(r_{in}\right)\Big) \tag{3.8}$$

$$G_{pd}(r_{in}) = G'_{amp}(\frac{r_{out}}{G_0}) \tag{3.9}$$

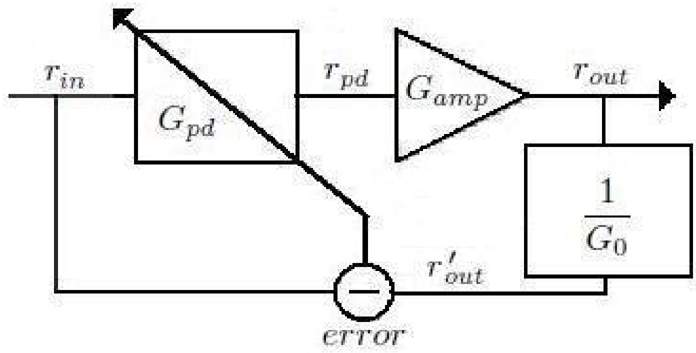


Figure 3.6: Direct Adaptation Topology[12]

As G_{amp} is a nonlinear function, the inverse function has to be derived by classical iterative optimization techniques[12].

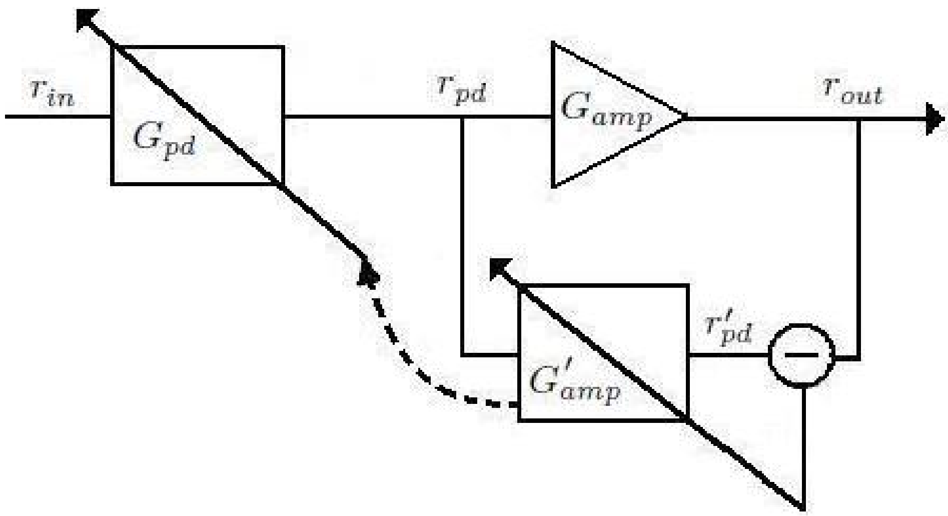


Figure 3.7: PA Modeling with Consecutive Inverse Estimation(Extraction)[12]

The second architecture, shown in Fig. 3.7, consists of two steps(in some implementations more). First, the PA input r_{pd} and output r_{out} are used to calculate the amplifier characteristics. This forward amplifier model is then used to calculate the amplifier inverse, which is G_{pd}[12].

The third possible way to adapt the predistorter is based on the direct identification of the PA

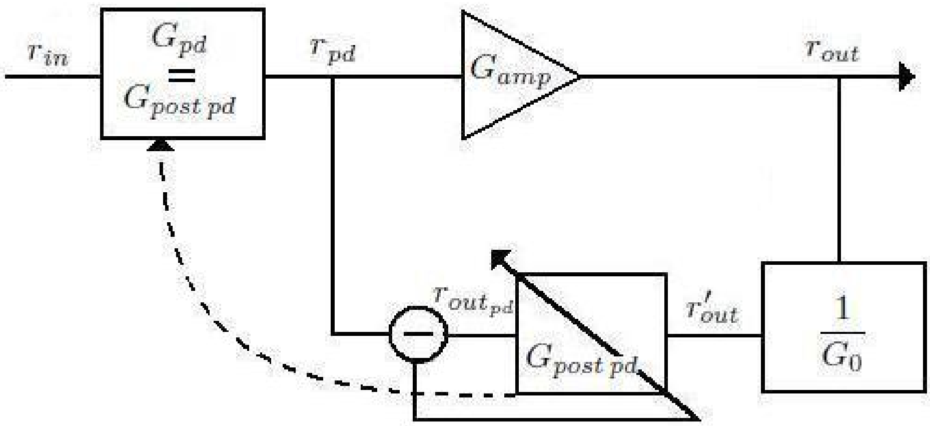


Figure 3.8: Predistortion using Postdistorter Adaptation (Indirect Learning)[12]

inverse using the input and output of the PA[12].

$$r_{out_{pd}} = G_{postpd}\left(\frac{G_{amp}\left(r_{pd}\right)}{G_0}\right) = r_{pd} \tag{3.10}$$

$$G'_{postpd}(r_{pd}) = \frac{G_{amp}\left(r_{pd}\right)}{G_0} \tag{3.11}$$

$$G_{postpd}(r_{pd}) = \frac{G'_{amp}\left(r_{pd}\right)}{G_0} \tag{3.12}$$

Predistorter characteristic G_{pd} is exactly equivalent to G_{postpd}. As in direct method, G_{amp} is a nonlinear function, and the inverse function G'_{amp} has to be derived by classical iterative optimization techniques.

3.3 Memoryless Signal Predistortion Algorithms

In this section, different kinds of predistortion approaches will be mentioned. Actually, in literature, nearly all of the cross combinations of above mentioned predistortion structures can be found. The ones introduced below are the kinds that are used throughout this thesis. Each approach will be introduced under *usage* and *adaptation* titles. Usage mentions how predistorter processes the input data, and adaptation states how predistorter adapts itself.

3.3.1 Polar Polynomial Predistortion with Direct Adaptation

In polar polynomial predistorters, both amplitude and phase corrections are implemented by using polynomials[10]. Both Gain and Phase Polynomials are polynomials with real coefficients. Gain polynomial models the expansion characteristics, which compensates for AM-AM nonlinearity of the amplifier. Phase polynomial models phase shift necessary for compensation of AM-PM nonlinearity of the amplifier. The degrees of each polynomial are independent from each other and depend on how much nonlinear the AM-AM and AM-PM responses are.

Usage

As mentioned previously, both polynomials are functions of input amplitude only, as AM-AM and AM-PM are. If the input signal to the polar polynomial predistorter is denoted by r_{in}, gain of the polar polynomial predistorter is denoted by G_{pd} and phase shift of the polar polynomial predistorter is denoted by ϕ_{pd}. G_{pd} and ϕ_{pd} are given by[10];

$$G_{pd}\left(r_{in}\right) = \sum_{k=0}^{N} a_k r_{in}^k = \left(\overline{A}\right)^T \overline{R_a} \tag{3.13}$$

$$\phi_{pd}\left(r_{in}\right) = \sum_{k=0}^{M} p_k r_{in}^k = \left(\overline{P}\right)^T \overline{R_p} \tag{3.14}$$

where $r_{in} = |z_{in}|$, $\overline{A} = [a_0 a_1 a_2 ... a_N]^T$ is the gain polynomial coefficient matrix, $\overline{P} = [p_0 p_1 p_2 ... p_M]^T$ is the phase polynomial coefficient matrix, $R_a = \left[1 r_{in} r_{in}^2 ... r_{in}^N\right]^T$, $R_p = [1 r_{in} r_{in}^2 ... r_{in}^M]^T$ are independent variable matrices and N and M are the orders of the magnitude and phase polynomials, which are not necessarily the same.

Predistortion coefficients found by above formulas are used to distort incoming signal as stated in Eq. (3.5).

Adaptation

In literature there are different adaptation algorithms applied to polar polynomial predistorters. The most common ones are LMS(least mean squares) and RLS(recursive least squares) algorithms[10].

An example of an LMS algorithm applied to a polar polynomial predistorter can be found in[42]. In that paper, the LMS algorithm was applied to adaptation to minimize the mean square error between the desired amplifier output and the observed amplifier output in the presence of a polar polynomial predistorter located before the amplifier. This resulted in the following update rule for gain and phase polynomials[42];

$$\overline{A}(n+1) = \overline{A}(n) + \mu_a \overline{R}_a(n)\, e_a(n), \tag{3.15}$$

$$\overline{P}(n+1) = \overline{P}(n) + \mu_p \overline{R}_p(n)\, e_p(n), \tag{3.16}$$

where n is the iteration index, μ_a and μ_p are small positive convergence constants, and, $e_a(n)$ and $e_p(n)$ are errors from the previous iteration and can be defined as;

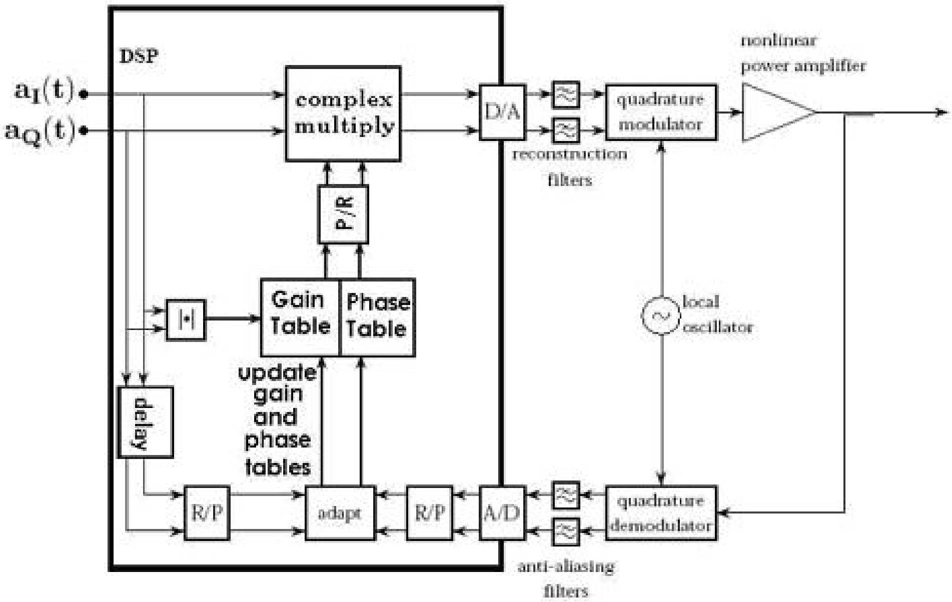
$$e_a(n) = \left[G_0 r_{in}(n) - \left| G_{amp}\left(r_{in}(n) \left(\overline{A}(n) \right)^T \overline{R}_a(n) \right) \right| r_{in}(n) \right], \tag{3.17}$$

$$e_p(n) = \left[-\arg\left(G_{amp}\left(r_{in}(n) \left(\overline{A}(n) \right)^T \overline{R}_a(n) \right) \right) - \left(\overline{P}(n)^T \overline{R_p}(n) \right) \right], \tag{3.18}$$

where $r_{in}(n)$ is the magnitude of input signal at the nth iteration, G_0 is the small signal gain of the amplifier.

3.3.2 Polar Look-Up Table Predistortion with Direct Adaptation

A polar Look-up- Table Predistorter can be seen in Fig 3.9. In this approach, AM-AM and AM-PM nonlinearities of amplifier are compensated by corresponding gain and phase values of predistorter as in polar polynomial case. These values are listed in two 1-D tables which are indexed according to amplitude or power of the input signal or any other mapping functions. Indexing methods, including amplitude and power, and their evaluation was discussed in[29]. An example of a LUT indexed by input amplitude can be seen in Tab. 3.1.



R/P : rectangular coordinates to polar coordinates

P/R : polar coordinates to rectangular coordinates

Figure 3.9: Polar Look-Up Table Predistortion with Direct Adaptation Topology[10]

Table 3.1: Look-Up Table with Amplitude Indexing[8]

$z_{in} = r_{in}\angle\theta_{in} \longrightarrow$

$Input\,Amplitude$	$Gain\,Coefficient$	$Phase\,Coefficient$
r_1	G_{pd_1}	ϕ_{pd_1}
r_2	G_{pd_2}	ϕ_{pd_2}
r_3	G_{pd_3}	ϕ_{pd_3}
$\cdots$	$\cdots$	$\cdots$
$\cdots$	$\cdots$	$\cdots$
r_i	G_{pd_i}	ϕ_{pd_i}
$\cdots$	$\cdots$	$\cdots$
r_N	G_{pd_N}	ϕ_{pd_N}

$\xrightarrow{z_{pd} = G_{pd_i} r_{in}\angle(\theta_{in} + \phi_{pd_i})}$

$==$

$z_{pd} = r_{pd}\angle\theta_{pd}$

Usage

Each value on the table corresponds to certain input magnitude or power, depending on indexing method. If input amplitude is exactly equal to one of the row, then that gain and phase value is used directly, such that;

$$z_{pd} = r_{pd}\angle\theta_{pd} = G_{pd_i} r_{in}\angle(\theta_{in} + \phi_{pd_i}). \tag{3.19}$$

On the other hand, if input amplitude belongs to somewhere between two rows, then linear or any other interpolation method is used to get correct gain and phase predistortion values;

$$G_{pd_j} = G_{pd_i} + \left(\frac{G_{pd_{i+1}} - G_{pd_i}}{r_{i+1} - r_i} \right) (r_{in} - r_i), \tag{3.20}$$

$$\phi_{pd_j} = \phi_{pd_i} + \left(\frac{\phi_{pd_{i+1}} - \phi_{pd_i}}{r_{i+1} - r_i} \right) (r_{in} - r_i), \tag{3.21}$$

where G_{pd_j} and ϕ_{pd_j} are gain and phase LUT coefficients calculated for input between i^{th} and $(i+1)^{th}$ levels. Then, these coefficients are used to predistort the signal as in Eq. (3.19).

Adaptation

For adaptation purpose, feedback path is used. After delay calibration and normalization calculations, each data sent is compared with the data sampled from the feedback. As in polynomial case, the aim is to decrease the error. Error is defined as the difference between the desired output and the observed output;

$$e_G(n) = G_0 r_{in}(n) - G_{amp}\left(G_{pd_n} r_{in}(n)\right), \tag{3.22}$$

$$e_\phi(n) = \theta_{in} - \left[\theta_{in} + \phi_{pd_n} + \phi_{amp}\left(G_{pd_n} r_{in}(n)\right)\right], \tag{3.23}$$

where θ shows phase of the signal, however, ϕ shows the AM-PM function of the corresponding nonlinear block.

At each iteration, the error, calculated in Eqs. (3.22) and (3.23), is used to update gain and phase predistortion values. If the data belongs to certain row in LUT, then adaptation is done according to following equations[8];

$$G_{pd_{n+1,i}} = G_{pd_{n,i}} + \mu_G e_G, \tag{3.24}$$

$$\phi_{pd_{n+1,i}} = \phi_{pd_{n,i}} + \mu_\phi e_\phi, \tag{3.25}$$

where μ_G and μ_ϕ are small positive convergence constants, n is the iteration index and i is entry level of LUT.

On the other hand, if the input amplitude is in between the two rows, then update is applied to both of the entries. Of course, the entry closer to the input value gets the larger share of error contribution. Hence share of error contributions can be calculated as[8];

$$\Delta_{n,i} = \frac{r_{i+1} - r_{in}}{r_{i+1} - r_i}, \tag{3.26}$$

$$\Delta_{n,i+1} = \frac{r_{in} - r_i}{r_{i+1} - r_i}, \tag{3.27}$$

where $\Delta_{n,i}$ is the share for lower neighbour entry and $\Delta_{n,i+1}$ is the share for the upper one.

Then, update can be done by the following equations[8];

$$G_{pd_{n+1,i}} = G_{pd_{n,i}} + \mu_G \, \Delta_{n,i} \, e_G, \tag{3.28}$$

$$G_{pd_{n+1,i+1}} = G_{pd_{n,i+1}} + \mu_G \, \Delta_{n,i+1} \, e_G, \tag{3.29}$$

$$\phi_{pd_{n+1,i}} = \phi_{pd_{n,i}} + \mu_\phi \, \Delta_{n,i} \, e_\phi, \tag{3.30}$$

$$\phi_{pd_{n+1,i+1}} = \phi_{pd_{n,i+1}} + \mu_\phi \, \Delta_{n,i+1} \, e_\phi, \tag{3.31}$$

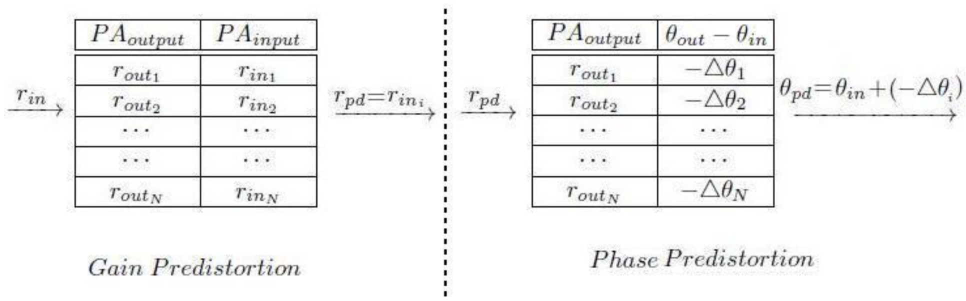
where n is the iteration index and i is entry level of LUT, again.

3.3.3 Polar Look-Up Table Predistortion by PA Modeling with Consecutive Inverse Estimation Adaptation

This approach is based on modeling the amplifier as the name implies. As amplifier's real time response can be obtained, inverse of the characteristic can be calculated. Jeckeln[43] offers such a method called adaptive predistortion with real time PA modeling. This method, basically, measures input and output of the amplifier, converts them in polar form to get AM-AM and AM-PM characteristics of the amplifier. After that, input magnitude, output magnitude and phase shift information are put in order. Interchanging the magnitude information and changing the sign of phase information gives the predistorter characteristics.

Let r_{in} and r_{out} be the normalized input and output magnitudes of the amplifier and $\Delta\theta_{out}$ is the phase difference of input(θ_{in}) and output(θ_{out}) of the amplifier, and these are measured and listed in PA modeling time interval.

Table 3.2: Usage of Jeckeln's Method[43]



As seen in Table 3.2, for a new input value, algorithm searches for PA_{output}, which is equal to the new input, and finds corresponding PA_{input} to make it predistorter output. Phase predistortion is done exactly the same way, but the output obtained in Gain Predistortion used as input. In this way, necessary PA_{output} can be obtained for linear amplification. Nevertheless, as the algorithm is based on searching, and as this searching is made in a nonlinear table, this procedure is rather slow.

Other than Jeckeln algorithm, Han's 3-step algorithm is offered for initialization and update of the LUT contents[44].

At the first step, AM/AM and AM/PM characteristics of the amplifier are estimated using transmitting signals. Since the transmitting signals are known, the gain and phase shift properties can be obtained by comparing and averaging the known and demodulated signals from the amplifier output[44]. A table is done from the gathered information as in Jeckeln's Method.

Second step calculates the inverse function of the amplifier from the table estimated in the first step with the assumption of amplifier gain is one. Then, for each entry in predistorter LUT, corresponding input amplitude is searched in output table. The nearest data set in the table is taken and inverse of the amplifier gain is written on that entry as LUT gain value. At the same time, negative of the phase shift on that data set is written on the same LUT entry.

From that point on (3^{rd} step), adaptation is done by comparing amplifier output and amplifier model output with the help of mean-square minimization of the error. The updates of the PA model gain and phase functions are performed using steepest descent algorithm formulas;

$$G_{amp_i}(n+1) = G_{amp_i}(n) + 0.5\mu_{G_{amp}}\left[-\nabla_{G_{amp}}(n)\right], \tag{3.32}$$

$$\phi_{amp_i}(n+1) = \phi_{amp_i}(n) + 0.5\mu_{\phi_{amp}}\left[-\nabla_{\phi_{amp}}(n)\right], \tag{3.33}$$

with gradients;

$$\nabla_{G_{amp}}(n) = \frac{\delta J_{G_{amp}}}{\delta G_{amp_i}}, \tag{3.34}$$

$$\nabla_{\phi_{amp}}(n) = \frac{\delta J_{\phi_{amp}}}{\delta \phi_{amp_i}}, \tag{3.35}$$

where J being the mean-square criterion function $J_{G_{amp}} = E\left\{\left|e_{G_{amp}}\right|^2\right\}$ and $J_{\phi_{amp}} = E\left\{\left|e_{\phi_{amp}}\right|^2\right\}$ with $e_{G_{amp}}$, $e_{\phi_{amp}}$ being the amplitude and phase error. After the adaptation, the predistorter characteristic is updated performing the second step again[12].

3.4 Parameters and Specifications

As mentioned previously, one of the possible complex envelope signal transmission line-up can be seen in Fig. 2.25. In this section, basic information about mission of each component will be introduced and problems encountered during usage will be mentioned. However, at first, some important parameters will be discussed.

3.4.1 Predistortion Bandwidth

As the name implies, predistortion distorts the incoming signal using a nonlinear function which is, ideally, the inverse of the amplifier. Depending on the order of amplifier nonlinearity, predistortion nonlinearity can be assumed as a high order polynomial. Therefore, output of the predistorter has higher adjacent channel powers with respect to incoming signal. Indeed, the adjacent channels include distortion terms which are needed to linearize the amplifier. Hence, the output frequency spectrum is a key parameter that determines line-up component specifications.

For a certain data rate and certain complex baseband input signal bandwidth, the speed of signal processors, conversion rate of DACs and ADCs, and pass band of reconstruction filters are variables that depend on bandwidth of predistorted signal. Depending on amplifier nonlinearity, predistorter output spectrum may expand up to 4th to 5th adjacent channels. However, linearization targets determine the adequate spectrum to achieve the desired linearity. For example, if the 1st and the 2nd adjacent channels are needed to be linearized, then, 5 to 7 times the bandwidth of the incoming signal is sufficient as a system parameter.

In general, as the predistortion bandwidth needed increases, i.e. above 1 MHz, then Analog Quadrature Modulation topology may be prefered for a PA with severe nonlinear distortion, since the Digital Quadrature Modulation topology cannot practically achieve as wide a predistortion bandwidth as for a given maximum practical clock rate[7].

3.4.2 Delay Synchronization

There exists a feedback delay in baseband predistortion systems. As predistortion coefficients are calculated through the comparison of input and output signals of amplifier, the correct input and output pairs must be processed. The time interval between the creation of modulation signals and the reception of them from the feedback path needs to be determined for correct formation of predistortion parameters.

Therefore, adaptation of baseband predistortion is highly dependent on the quality of delay synchronization. This delay may, similar to the PA nonlinearity, vary with time due to aging or temperature[12]. There are some approaches to calculated this delay.

Nagata's Iteration Method

Nagata offers an iterative way to calculate the delay[23]. He used the following formula;

$$d_{i+1} = d_i - T\, sign\left[\left\{q\left(t_i\right) - q\left(t_i - d_i\right)\right\} q'\left(d_i\right)\right], \tag{3.36}$$

where d_i denotes delay at time t_i, T is time step and $q(t_i)$ and $q(t_i - d_i)$ represents feedback signal and modulating signal, respectively. $q'(d_i)$ is the time derivative of the modulating signal at time d_i.

Correlation Method

Authors of a US patent[54] claim the simple correlation method for the delay estimation in PA;

$$C(m) = \frac{1}{N}\sum_{n=0}^{N-1} z_{pd}(n)\, z_{out}^{*}(n+m), \tag{3.37}$$

where z_{pd} and z_{out} are phasors of predistorter and amplifier output signals, respectively, m is the delay variable, N is the length of signals to correlate. If there is a delay of d samples, the sequence $C(m)$ must reach its maximum at m = d[12].

In [57], correlation method is used to measure loop delay by applying a two-step procedure. At the first step, using correlation, integer delay is calculated. At the second step a Delay Locked Loop, introduced below, and an interpolator is used to calculate fractional part of the delay. In [34] and [58], the same approach is used, however, instead of signals itself, amplitude-difference function is used in Eq. (3.37).

For the above two approaches, there is no need for synchronization time to detect delay, because the method is applied with the real modulated signals on the fly. However, there are some standards like TETRA which gives users time to calculate parameters of linearization at the beginning of each transmission burst. For those kinds of modulations, as the time can be used with arbitrary data, special signals can be used to make delay synchronization more accurate or faster.

For example, correlation method can be applied with signals that have special autocorrelation characteristics. Barker Sequences are one of such special sequences. Their peak autocorrelation, which is the value obtained when there is no delay in the system, is far greater than any other autocorrelation of the sequence. Therefore, usage of this sequence for delay estimation increases quality of the result.

Ramp Method

As mentioned above, if a synchronization time is defined then ramp function can also be used to detect delay. In [56], searching for the same levels in the ramp input and its response is offered to be an alternative way to detect the delay. However, this method is dependent on the quality of normalization as the input and output sequences are exposed to linear scalings or nonlinear distortions. If indexes corresponding to this point are denoted as i_{lin}, k_{lin} in input ramp and ramp response, respectively, the delay introduced by the amplifier is simply;

$$delay = k_{lin} - i_{lin}, \tag{3.38}$$

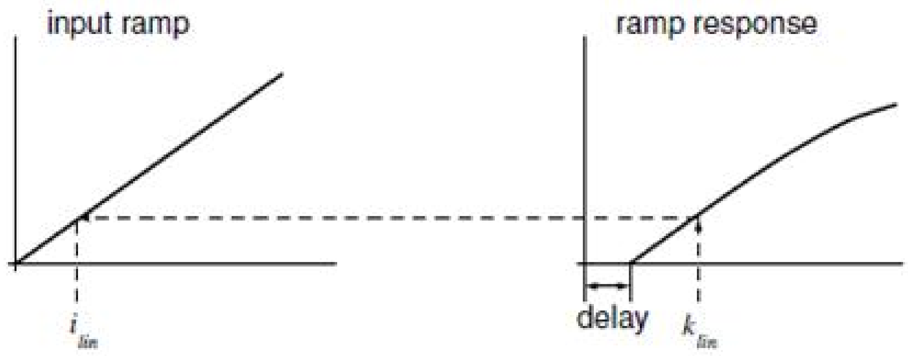


Figure 3.10: Ramp Function Delay Synchronization Approach[56]

Delay Locked Loop (DLL) Method

In [55], a delay control loop circuit for predistortion systems is presented and applied to polynomial IF predistortion. It is called Delay Locked Loop(DLL). It needs extra circuit for detection as shown in Fig. 3.11.

The output of the discriminator is filtered by a loop filter and used as the control voltage of a voltage controlled oscillator. The phase of the VCXO controls the sampling time of the A/D converter in order to eliminate the delay between forward and feedback path of the predistorter[12].

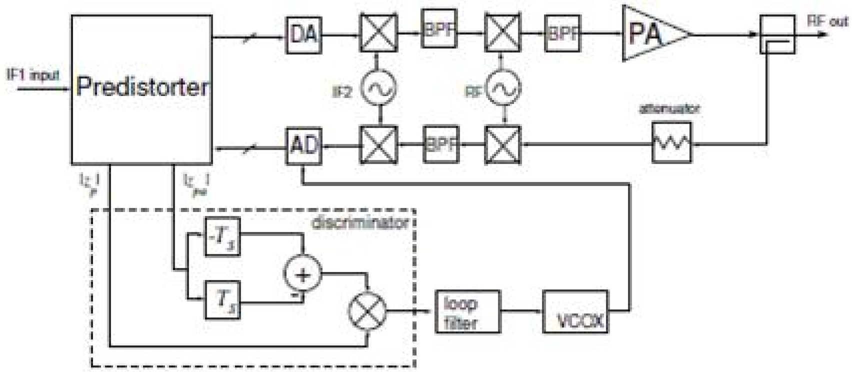


Figure 3.11: Delay Locked Loop Approach for Delay Synchronization[12]

3.4.3 Line-Up Component Specifications

Signal Processor

Signal processors are the hearts of digital radios. More than one signal processors are being used to handle different functions inside a radio. The systems are so complex that one signal processor deals with the operation, while another one copes with modulation issues. However, in a basic digital transmitter, one signal processor makes all coordination and modulation operations at the same time.

Predistortion is a method that needs extra computational power to make mathematical operations. For a basic predistortion algorithm, finding correct index, calculation of coefficients and making use of them need hundreds or thousands of flops inside a processor. When it comes to adaptation, the computational complexity of calculations are much higher. So, predistortion linearization needs extra care in determination of processor to be used.

As a processor, Digital Signal Processors (DSPs) can be used. They have built-in arithmetic logic units, multipliers and interfaces. However, they do operations one by one, in that, at each cycle they do one operation or a bunch of operations. This limits the speed of operation.

Another possible choice for signal processing is Field Programmable Gate Arrays (FPGAs). There are millions of functional units in an FPGA and they can be programmed via VHDL or Verilog hardware description languages to operate in a predefined way. FPGAs operate in a way different from DSPs, in that, functional units within any FPGA can operate in parallel if they are programmed so. Therefore, FPGAs can handle much more operations per second than DSPs. For example, at one clock cycle a DSP can do one multiplication, while an FPGA can do even hundreds of it. As a result, FPGAs are used instead of DSPs when there are operations that must be completed at the same time, or when number of operations is huge.

Application Specific Integrated Circuits (ASICs) are components which are produced by a manufacturer to do certain operation, for example, Predistortion. They can be faster and low power than DSPs and FPGAs, they are usually cheaper, however, they have high initial cost and generally make a design less flexible.

A DSP can be used to complete all of the duties in digital domain. However, there are some real life problems related to DSP, these affect quality of predistortion.

- Fixed Point Operation

Baseband Predistortion is a linearization method that has several numbers of underlying mathematical operations due to its digital structure. These mathematical operations are done in DSPs' multipliers and accumulators, whose structure is determined by the resolution of the DSP architecture, called Word Lengths. If larger variables are used instead of native resolution of the DSP, algorithms slows down significantly.

There are Floating point DSPs, which have much more efficient real number processing capabilities and which are capable of doing operations on those numbers easily. However, they are expensive and they consume more power.

- Speed of Operation

Detecting I and Q signals, finding corresponding predistortion coefficients and application of these coefficients need a time interval. Depending on the sampling rate of the modulation, this time interval may decrease and necessity of faster processors may occur. For high data rate systems, it is better to use an FPGA and make all of mathematical operations as fast as possible.

On the other hand, as mentioned in Sec. 3.4.1, the output signal of the predistorter has a higher bandwidth than the source signal, which dictates the minimum oversampling rate. But, at the same time, any increase on the oversampling rate increases the power consumption in the DSP, which is not desirable[47].

Digital to Analog Converter and Analog to Digital Converter

- Zero Order Hold Effect

At an ideal DAC output, the signal is kept constant between two clock pulses. This behavior leads to $\frac{sin(x)}{x}$ filtering of the signal in frequency domain. Pass band insertion loss of $\frac{sin(x)}{x}$ may degrade the performance of predistortion depending on oversampling rate. Inverse $\frac{sin(x)}{x}$ filters can be used to pre-compensate this insertion loss variation to achieve flat amplitude response from DC to near Nyquist[7]. This compensation needs to be done in signal processors which is another work load for them and which, in return, degrades the speed of operation. On the other hand, there exist some DACs that compansate for this effect.

- Resolution

Along with fixed point operation of signal processors, DAC and ADC resolutions are other important parameters that determine quantization errors. Resolutions of DAC and ADC result in quantization noise to occur on the signal, which result incorrect measurements and degrade predistortion performance.

In [34], the effects of DAC and ADC resolutions on mean square error are obtained as in Fig. 3.12. It is also revealed that, the DAC resolution is more critical than the ADC. Because the DAC outputs drive the PA directly, and the quantization noise passes through without attenuation, while the ADC outputs are used to update the LUT contents and the quantization noise effect can be mitigated by averaging.

In [12], it is claimed that the performance of the predistorter in the case of 13 bit quantization, which models commercial 14 bits, is close to the unquantized case. Also Potter[7] states 14 or 16 bits are required in a typical predistortion system.

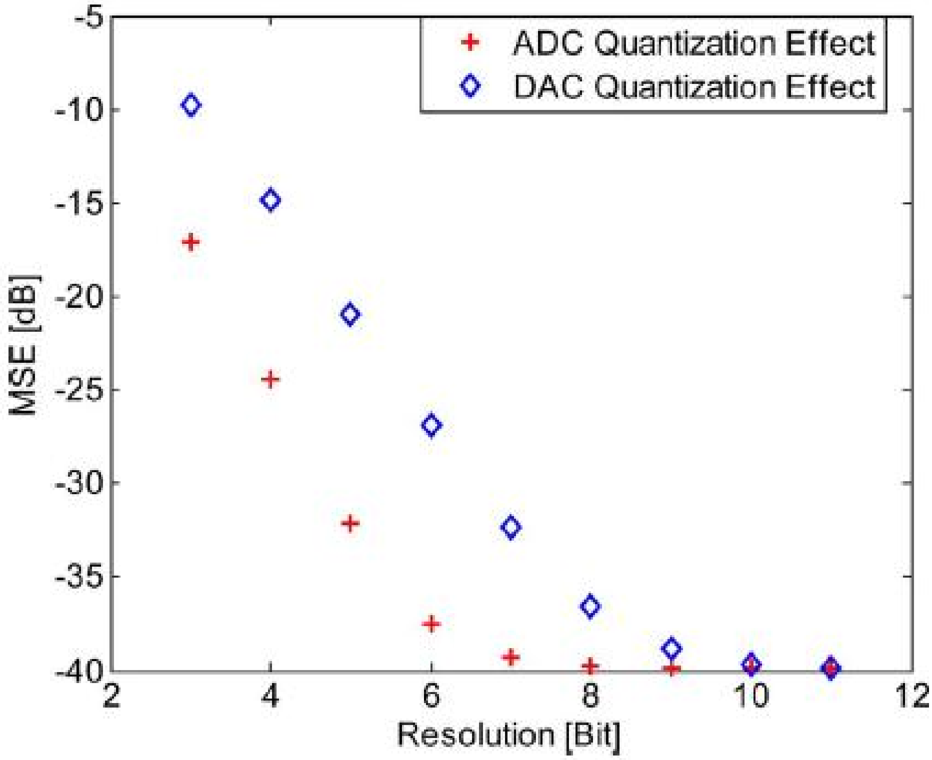


Figure 3.12: Quantization Effects of ADCs and DACs[34]

- Speed

As mentioned in Sec. 3.4.1, predistorted signal has a wider bandwidth than input signal. Therefore, the speed of the DAC is determined by predistorted signal's bandwidth, not by input signal's. For example, if input signal bandwidth is BW, and if predistortion is intended to linearize amplifier so that the $1st$ and the $2nd$ adjacent channels are suppressed, amplifier can be assumed to have, at least, 5^{th} order nonlinearity. The inverse of this function has infinite number of terms, however, at least 5^{th} to 7^{th} orders should be taken into account. So the system is designed so that 5^{th} to 7^{th} order nonlinearities exist at the predistorter output. $7BW$-bandwidth puts sampling frequency of DACs at $7BW$ Hz for an ideal reconstruction filter, in order to avoid aliasing. However, considering the realizable reconstruction filters, more than $10BW$ Hz can be applicable.

ADC specification is similar to DACs. Their speed is determined by predistorted signal bandwidth and data rate of the modulation.

Reconstruction Filter and Anti-Aliasing Filter

Reconstruction filters are used in transmitter chain after DA converters because DA converter output frequency spectrum includes image terms situated at multiples of sampling frequency. The passband of the filter must cover the predistorted signal bandwidth as mentioned in Sec. 3.4.1. Passband insertion loss of the filter must be flat and passband phase response must be linear (linear phase) so that filter causes constant group delay.

However, this is not an ideal filter and the analog implementation of the baseband reconstruction filter is relatively of low order and finite bandwidth due to is complexity and cost, therefore it has some passband ripple and phase delay as illustrated in Fig. 3.13. This type of filters have a near constant group delay and negligible attenuation for low frequencies in the passband. Closer to the cut-off frequency, the attenuation is significant and/or the group delay is not constant[47]. Therefore, frequencies close to the edge are affected by non-ideal filter characteristics, and degrade the predistortion performance.

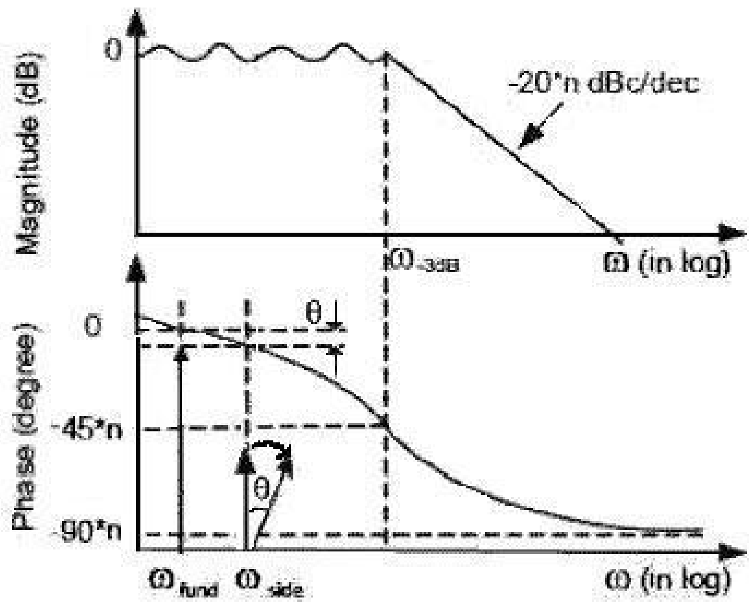


Figure 3.13: Magnitude and Phase Responses of Nth-Order Analog Filter[46]

One solution to reduce the impact of these filters is to use a high sampling rate reducing the demands on the filter and thus the memory effects introduced[47]. However, usage of high sampling rate is not an easy way as mentioned in the DAC part. For detailed filter effect and sampling rate analysis refer to [47].

Lately, some integrated circuit companies announced the usage of integrated circuit high order linear phase low-pass filters.

Other than reconstruction filters, some bandpass filters may also be used in transmitter chain for different purposes. For example, mixer's spuriouses and harmonics can be filtered by a band pass filter with sufficient insertion loss characteristics. Another band pass or low pass filter may be added to filter amplifier harmonics. All of these filters cause similar performance degradation problems in a predistortion system.

Anti-aliasing filters are used before ADCs to avoid aliasing. Their specifications are similar to reconstruction filters. Bandwidth of the distorted signals at the output of the amplifier must be included in the pass band of anti-aliasing filters to detect errors on the signals and characterize the amplifier. Flatness of pass band gain and linearity of phase response are also important as they cause distortion.

Modulator and Demodulator

In Analog Quadrature Modulation scheme, analog modulators are used. See Fig. 2.25. They take I and Q signals and multiply them with IF signals which have 90 degree phase difference.

However, in practice, the quadrature carriers in the analog modulator do not have exactly the same amplitude and exact phase difference of 90 degrees. These effects are called gain and phase imbalances and can cause cross-talk between I and Q channels. In addition, leakage of the carrier to the transmitted signal manifests itself in the demodulated received signal as a dc offset[48]. For narrowband signals modulator imbalances can be assumed as frequency independent. For wideband inputs, the gain and phase imbalances exhibit frequency dependent behavior[48].

The adaptation of digital predistorter is highly sensitive to the imperfections of the quadrature modulators. If the modulators' imperfections exceed requested limits, some method for their compensation as for example the adaptive LMS compensation can be used[12]. Modeling and compensation of modulator gain phase imbalance are studied in [49], [50], [51], [52].

Demodulator is the dual of modulator. It downconverts IF to baseband I/Q signals. In order to do this, it uses two IF signals having 90 degree phase shift. Due to the same reasons that modulators have, demodulators cause distortions on I and Q signals.

In [53], experiments have shown that even if there is no spectral spreading, there are always linear distortions in linearized systems if the demodulator errors exist. As a result, BER of the system will be deteriorated. Digital demodulation topology is not affected from demodulation errors.

On the other hand, Digital Quadrature Modulation scheme (Fig. 2.26) does not suffer from these problems, as these processes are done in digital domain.

Local Oscillator

Local oscillators are used to provide IF frequency for upconversion and downconversion. Local oscillators are characterized by their *PhaseNoise* parameters, which show how much the frequency of the oscillator shifts.

If local oscillators of modulator and demodulator are different in a predistortion system, then, phase noises of local oscillators will degrade performance. So, in general, single local oscillator is used in predistortion systems to overcome phase noise problem.

The problem arises in adaptation algorithm. For adaptation purpose, I and Q signals are fed back and compared with the ones created in processor. If I_{sent} and Q_{sent} are signals sent at time $t = t_1$, then, modulator output is;

$$V_{out_{mod}}(t_1) = I_{sent}(t_1)cos(wt_1 + \phi(t_1)) + Q_{sent}(t_1)sin(wt_1 + \phi(t_1)). \tag{3.39}$$

where w is the carrier frequency and $\phi(t_1)$ is phase noise at $t = t_1$.

If τ seconds are passed until this signal is fed back to demodulator, then, at time $t_1 + \tau = t_2$, the LO input of demodulator will become;

$$V_{LO_{demod}}(t_2) = cos(wt_2 + \phi(t_2)). \tag{3.40}$$

If Eq. (3.40) is used, then at demodulator output;

$$I_{taken}(t_2) = \frac{I_{sent}(t_1)}{2} cos\left(w\tau + \phi(t_1) - \phi(t_2)\right), \tag{3.41}$$

which shows the problem that phase noise causes.

The basic solution is to compensate for the delay of the feedback signal by adding τ seconds delay to LO input of the demodulator so that modulator and demodulator have the same LO frequency at the time they use LO. If demodulator LO source is delayed by τ seconds;

$$V_{LO_{demod}}(t_2) = cos(wt_1 + \phi(t_1)). \tag{3.42}$$

Then, demodulator outputs, I_{taken} and Q_{taken}, will become;

$$I_{taken}(t_2) = \frac{I_{sent}(t_1)}{2}, \tag{3.43}$$

$$Q_{taken}(t_2) = \frac{Q_{sent}(t_1)}{2}, \tag{3.44}$$

which is the ideal case.

CHAPTER 4

SIMULATIONS

In this chapter, simulations that are done in order to understand the theory of predistortion and to see the limitations in basic Memoryless Signal Predistortion Algorithms will be presented. MATLAB is used as the simulation environment. Firstly, the amplifier model used throughout the simulations will be given. The two predistortion approaches, introduced in Sec. 3.3, will be studied and different parameters that affect the performance will be investigated. At the end, a performance comparison will be given.

4.1 Amplifier Model

Amplifier Modeling is mentioned in Sec. 2.6. In this work, a fifth order polynomial model is used to model AM-AM nonlinearity of the amplifier in baseband. As the approaches are memoryless and modeling the intermodulation products is the main interest, odd order terms are used only.

In order to make the model more realistic, polynomial coefficients are associated with Nonlinearity Characterization Parameters, mentioned in Sec. 2.4. Therefore, as a baseband amplifier model;

$$V_{out}(t) = a_1 V_{in}(t) + a_3 V_{in}^3(t) + a_5 V_{in}^5(t), \tag{4.1}$$

is used. The baseband coefficients a_1, a_3 and a_5 are;

$$a_1 = 10^{G_0/20}, \tag{4.2}$$

$$a_3 = \frac{-1}{2R} 10^{\frac{3G_0}{20} - \frac{IP_3}{10}}, \tag{4.3}$$

$$a_5 = \frac{1}{4R^2} 10^{\frac{G_0}{4} - \frac{IP_5}{5}}, \tag{4.4}$$

where R is the reference impedance of the circuit. The calculation details of baseband coefficients a_1, a_3 and a_5 can be found in [59].

In order to model AM-PM nonlinearity, a realistic AM-PM model is needed and Saleh Model is used, which is given in Eq. (2.15) and repeated here for convenience;

$$\phi_{amp}(|V_{in}(t)|) = \frac{\alpha_p \, |V_{in}(t)|^2}{1 + \beta_p \, |V_{in}(t)|^2}. \tag{4.5}$$

For these models, the AM-AM and AM-PM characteristics can be seen in Fig. 4.1 and Fig. 4.3 with corresponding parameters used. The two figures are obtained from the same parameters but plotted with respect to power and voltage, respectively. Succeding figures will be plotted with respect to voltages.

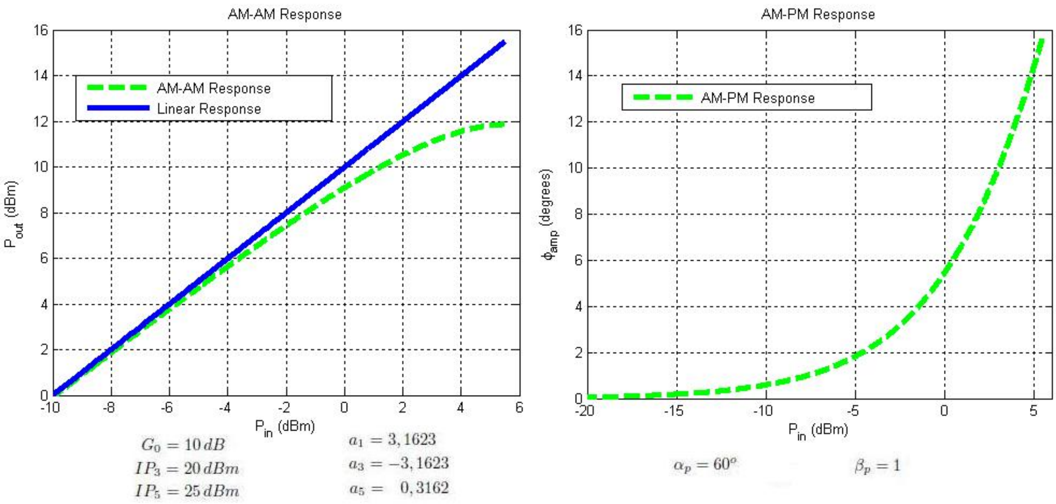


Figure 4.1: AM-AM and AM-PM Responses of the Model(Power)

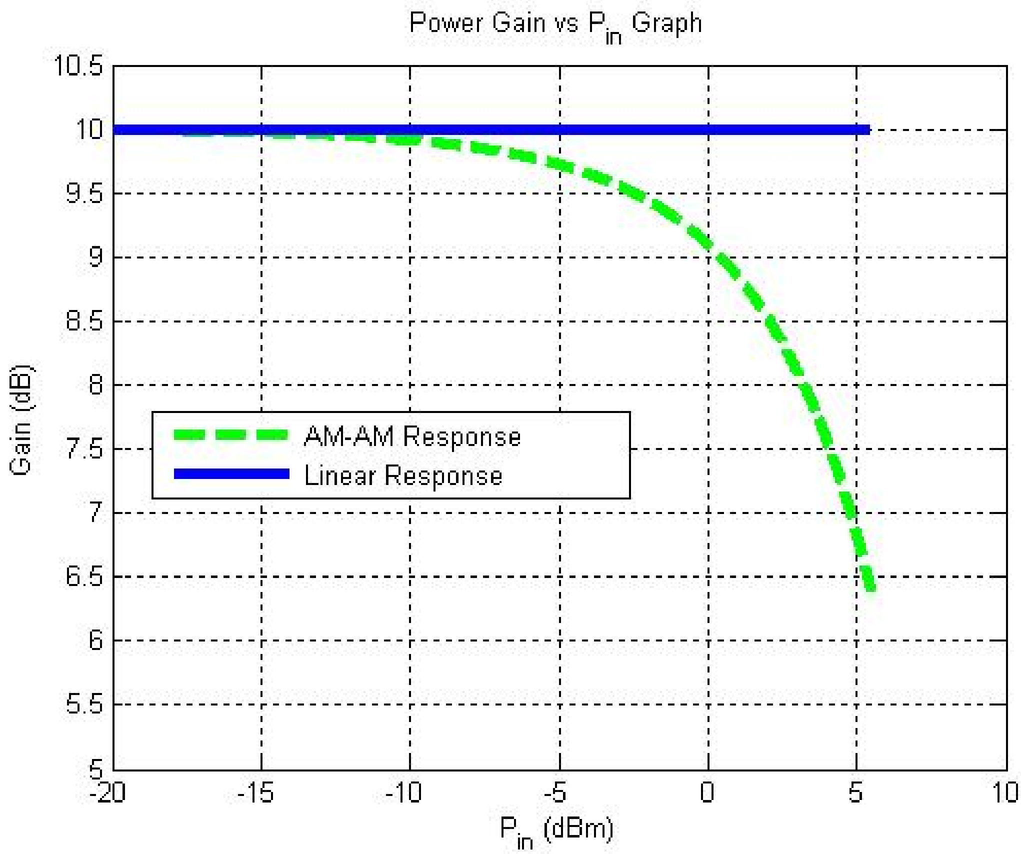


Figure 4.2: AM-AM Response of the Model (Power Gain)

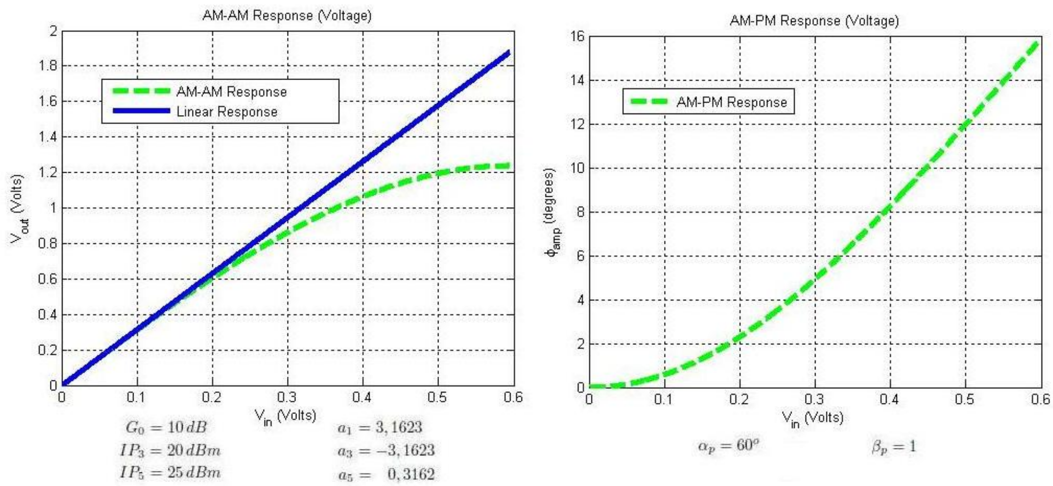


Figure 4.3: AM-AM and AM-PM Responses of the Model(Voltage)

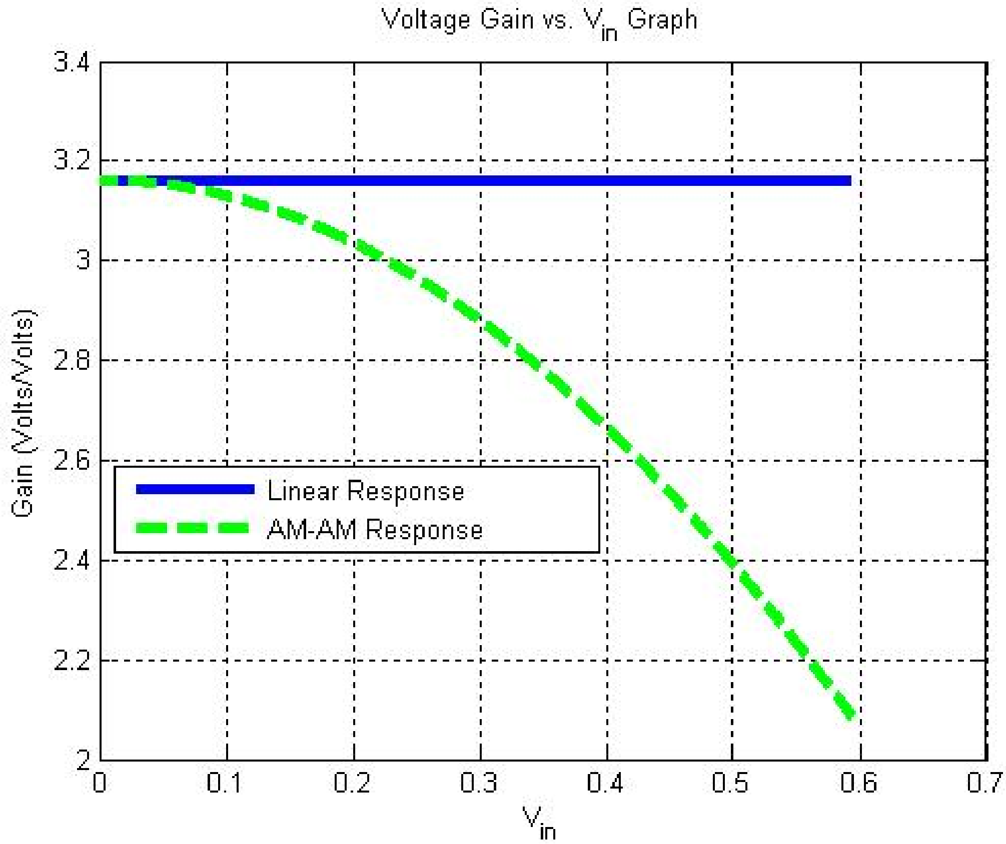


Figure 4.4: AM-AM Response of the Model (Voltage Gain)

4.2 Approaches

In this section, Polynomial and LUT Predistortion approaches will be evaluated. For evaluation, key parameters and effects of each one will be mentioned, while the others are remained at their optimum values. Throughout the simulations wideband modulation is preferred to two tones. Therefore, after training of predistortion parameters, an $8PSK$ data with $3,13dB$ Crest Ratio and 16640 sample points are used as input to evaluate the adjacent channel improvement.

4.2.1 Polar Polynomial Predistortion with Direct Adaptation

Effects of Polynomial Order

The effects of polynomial order are investigated. For this aim different polynomial structures are used as Gain Polynomial of Polar Polynomial Predistorter. Phase predistortion polynomial is determined as 2^{nd} order polynomial and remains the same for different Gain Polynomials. The formulas and adaptation algorithms used can be seen in Secs. (3.13) and (3.15). Tab. 4.1 shows the results obtained after adaptation with 10000 random input data.

Table 4.1: Effects of Gain Polynomial Order on Predistortion Performance

Gain Polynomial Used	Main Channel Power (dBm)	Adjacent Channel Power (dBm)	Alternate Channel Power (dBm)	IM3 (dB)	IM5 (dB)
No Predistortion	7,64	-26,76	-56,34	34,40	63,98
3^{rd}Odd Order	8,54	-39,36	-43,57	47,90	52,11
5^{th}Odd Order	8,50	-40,71	-44,49	49,21	52,99
7^{th}Odd Order	8,43	-41,87	-44,71	50,30	53,14
9^{th} Odd Order	8,45	-41,38	-44,28	49,83	52,73
11^{th}Odd Order	8,44	-41,70	-44,95	50,14	53,39
5^{th}Order	8,41	-39,64	-40,22	48,05	48,63
7^{th}Order	8,61	-38,68	-39,69	47,29	48,30
11^{th}Order	8,61	-38,84	-39,76	47,45	48,37

Results show that predistortion by using this approach linearizes the adjacent channel by approximately $15dB$, however, it also distorts alternate channel by $10dB$. 7^{th} only odd order polynomial gives the best result as expected, because the model is 5^{th} order. 5^{th} odd order predistorter gives close results. Adding even order terms degrades performance. 7^{th} odd order polynomial predistorter is used from this point on.

It is obvious from above results that the adaptation algorithm is not good enough to find correct coefficients and predistorter function cannot converge to exact inverse function.

Secondly, Phase Polynomial is evaluated by using 7^{th} odd order polynomial as the Gain Polynomial. Results can be seen in Tab. 4.2.

2^{nd} order polynomial seems to model the inverse of AM-PM nonlinearity of the amplifier best and $15dB$ linearization in adjacent channel achieved, while alternate channel is distorted by nearly $12dB$.

Table 4.2: Effects of Phase Polynomial Order on Predistortion Performance

Phase Polynomial Used	Main Channel Power (dBm)	Adjacent Channel Power (dBm)	Alternate Channel Power (dBm)	IM3 (dB)	IM5 (dB)
No Predistortion	7,60	-27,12	-57,14	34,72	64,74
1^{st} Order	8,48	-37,24	-43,39	45,72	51,87
2^{nd} Order	8,43	-41,35	-44,94	49,78	53,37
3^{rd} Order	8,59	-39,88	-44,26	48,47	52,85
4^{th} Order	8,48	-41,22	-44,59	49,70	53,07
5^{th} Order	8,55	-40,03	-43,95	48,58	52,50

Effects of Training Sequence

In order to calculate polynomial predistortion coefficients, input and output data are compared. If there is a time interval for training, then any kind of data can be used. For polynomial adaptation ramp and random data are used for training. Results can be seen in Tab. 4.3.

Table 4.3: Effects of Training Sequenceon Predistortion Performance

Training Sequence Used	Main Channel Power (dBm)	Adjacent Channel Power (dBm)	Alternate Channel Power (dBm)	IM3 (dB)	IM5 (dB)
No Predistortion	7,68	-26,93	-56,78	34,61	64,46
Ramp 100000 pts	7,85	-28,73	-59,09	36,58	66,94
Ramp 1000000 pts	8,55	-38,66	-42,93	47,21	51,48
Random 10000 pts	8,53	-41,04	-44,30	49,57	52,83

Adjacent channel powers show that Ramp data with 100000 points cannot train predistortion coefficients, however, Random data with 10000 points achieve $15dB$ linearization in adjacent channel. Ramp data can reach to the performance of Random data with 1000000 points at least. Again, the alternate channel is distorted by $12dB$. The conclusion from this experiment is that random data converges faster for polynomial predistortion.

Effects of Number of Samples used for Training

As Random data is found to be the best trainer, number of points needed to achieve satisfactory results is investigated. The results obtained after adaptation with random input data are shown in Tab. 4.4.

From Tab. 4.4, it can be concluded that in order to train polynomial coefficients at least 10000 points should be used. If 50000 points are used then, $15dB$ linearization is achieved and $12dB$ distortion occured in alternate channel.

Table 4.4: Effects of Number of Samples used for Training on Predistortion Perf.

Number of Samples Used	Main Channel Power (dBm)	Adjacent Channel Power (dBm)	Alternate Channel Power (dBm)	IM3 (dB)	IM5 (dB)
No Predistortion	7,58	-27,42	-57,45	35,00	65,03
100 points	8,43	-26,54	-59,96	34,97	68,39
1000 points	8,40	-28,04	-56,87	36,44	65,27
5000 points	8,39	-34,98	-48,29	43,37	56,68
10000 points	8,27	-40,30	-43,38	48,57	51,65
50000 points	8,39	-42,19	-45,01	50,58	53,40
100000 points	8,45	-40,80	-43,69	49,25	52,14

4.2.2 Polar Look Up Table Predistortion with Direct Adaptation

Effects of LUT Indexing and LUT Size

The first questions to be answered in Look-Up Table Predistorter are how indexing or mapping will be done in the algorithm and how many entries there will be in the Look-Up Table. These two parameters are closely related, so they must be investigated together. In Sec. 4.2.2, indexing was done according to amplitude of the incoming signal. It can also be done according to the square of that, which corresponds to power. Then formulas will include r^2 instead of r. On the other hand, LUT size is important because in LUT predistortion inverse characteristics of amplifier is modeled with piecewise linear functions and the number of these functions is determined by LUT size. The larger the number is, the better the approximation will be achieved up to some point. Tabs. 4.5 and 4.6 shows the results obtained after adaptation with 5000 ramp data (explained in next topic).

Table 4.5: Results of Amplitude Indexed LUT Predistortion

Size of LUT	Main Channel Power (dBm)	Adjacent Channel Power (dBm)	Alternate Channel Power (dBm)	IM3 (dB)	IM5 (dB)
No Pred.	7,61	-27,07	-57,20	34,68	64,81
8	8,17	-26,96	-28,76	35,13	36,93
16	8,31	-39,87	-40,32	48,18	48,63
32	8,39	-49,72	-50,58	58,11	58,97
64	8,36	-57,63	-60,75	65,99	69,11
128	8,40	-60,54	-64,45	68,94	72,85
256	8,41	-62,68	-68,69	71,09	77,10
512	8,47	-63,56	-69,87	72,03	78,34
1024	8,46	-64,13	-69,98	72,59	78,44
100000	8,48	-64,19	-70,34	72,67	78,82

Results in Tables 4.5 and 4.6 are plotted in Fig. 4.5 for better comparison.

Table 4.6: Results of Power Indexed LUT Predistortion

Size of LUT	Main Channel Power (dBm)	Adjacent Channel Power (dBm)	Alternate Channel Power (dBm)	IM3 (dB)	IM5 (dB)
No Pred.	7,61	-27,07	-57,20	34.68	64,81
8	8,37	-36,64	-37,31	45.01	45.68
16	8,34	-51,55	-52,10	59.89	60.44
32	8,34	-60,21	-62,03	68.55	70.37
64	8,45	-61,17	-64,46	69.62	72.91
128	8,46	-62,16	-67,05	70.62	75.51
256	8,38	-63,58	-69,75	71.96	78.13
512	8,34	-64,20	-69,78	72.54	78.12
1024	8,48	-64,38	-70,00	72.86	78.48
100000	8,49	-64,50	-70,50	72.99	78.99

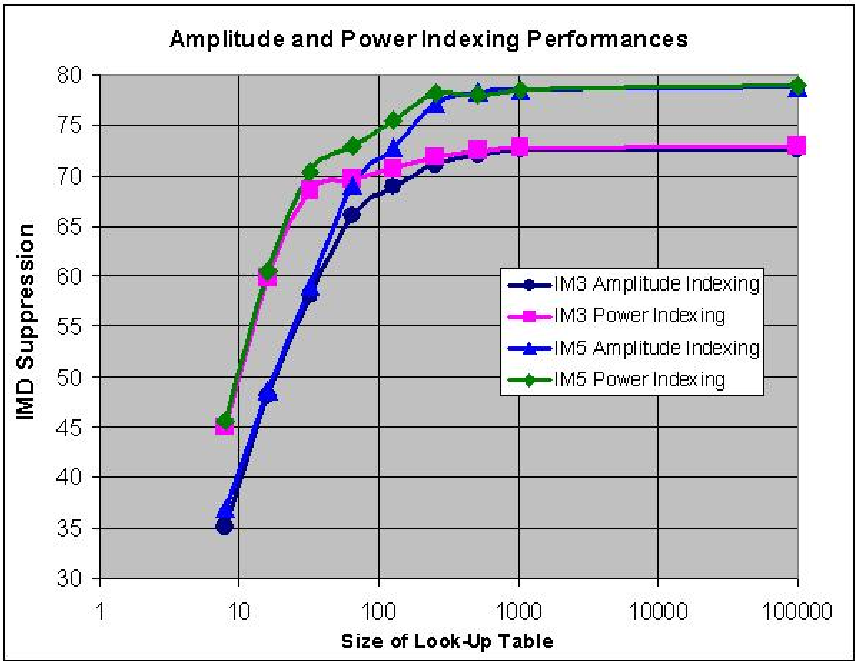


Figure 4.5: Effects of Indexing on LUT Predistortion Performance

There are two important observations obtained from Fig. 4.5. Firstly, for LUT sizes smaller than 64, Power Indexing gives 10 to 12 dB better performance. This proves the theory that power indexing should be better as it focuses the saturation region, which is the main responsible for the nonlinearity. Secondly, it can be seen from the figure that after LUT size of 64, the performances of indexing methods become closer and LUT sizes larger than 256 gives nearly the same results. From above discussions, Power Indexing with LUT size of 512 is selected.

Effects of Training Sequence and Number of Samples used for Training

In polynomial predistortion, random data had better performance. In LUT predistortion, by including physical assumptions into adaptation algorithm, training algorithms that have better performance than just applying random data can be obtained. One of such algorithms is developed in this thesis.

Modified Ramp Algorithm

In Sec. 3.1 the theory of predistortion is introduced. Fig. 3.2 illustrates the algorithm. For LUT predistortion, there exist very important points on AM-AM graph, which are $r_{in_{lim}}$ and $r_{in_{sat}}$. These are the end points of Input Voltage Domain and Ouput Voltage Domain of predistorter. As the last entry of LUT predistorter must map $r_{in_{lim}}$ to $r_{in_{sat}}$ value, then, it is obvious that last entry of LUT must be $\frac{r_{in_{sat}}}{r_{in_{lim}}}$.

Second physical assumption is that amplifier AM-AM response is continuous, or, there exists no discontinuity on amplifier's compression function. This assumption guarantees the continuity of predistorter's expansion function.

Using these two information LUT entries can be found easily. At first, last entry of LUT is tried to be found by continuously applying same input data and using the formulas given in Sec. 4.2.2. When the error defined decreases to some predetermined value, the second entry from the end is tried to be found. As predistortion function is accepted as continuous, then this second entry should be very close to last entry. So last entry is copied to one below and input data related to that entry is applied continuously until error decreases enough. This procedure continues until first LUT entry is done. As initial values are very realistic, number of samples needed is very few.

Random data and Modified Ramp data are used to train predistortion parameter and results can be seen from Table 4.7.

It is obvious from the results that offered approach has $28dB$ better performance in adjacent channel, $34dB$ in alternate channel.(5000pts) For an approximately equal performance, random data should be applied by 20 times more, which means 20 times longer in time.

Table 4.7: Results of Random and Modified Ramp Sequence LUT Adaptation

Number of Points Used	Main Channel Power (dBm)	Adjacent Channel Power (dBm)	Alternate Channel Power (dBm)	IM3 (dB)	IM5 (dB)
No Pred.	7,66	-26,79	-56,51	34,45	64,17
Random 1000 pts	8,11	-24,04	-24,86	32,15	32,97
Random 5000 pts	8,38	-36,11	-36,16	44,49	44,54
Random 10000 pts	8,41	-43,87	-43,64	52,28	52,05
Random 50000 pts	8,40	-46,96	-47,52	55,36	55,92
Random 100000 pts	8,42	-63,73	-66,40	72,15	74,82
Mod. Ramp 4997 pts	8,33	-64,28	-69,81	72,61	78,13

4.3 Comparison of Two Approaches

The best performances achieved from the polynomial and LUT approaches mentioned above can be compared.

Polynomial Predistortion is realized by 7^{th} odd order gain polynomial and 2^{nd} order phase polynomial. Inverse functions obtained from adaptation with 10000 random data are as follows;

$$r_{pd}(t) = 0.9859r_{in}(t) + 1.7466r_{in}^3(t) + 0.3154r_{in}^5(t) + 0.048r_{in}^7(t), \quad (4.6)$$

$$\phi_{pd}(t) = 0.284 - 5.4916r_{in}(t) - 39.1974r_{in}^2(t). \quad (4.7)$$

For illustration, normalized AM-AM, Gain and AM-PM responses of amplifier with and without predistortion and predistorter are plotted and can be seen in Figs. 4.6, 4.7 and 4.8.

From Figs. 4.6 and 4.7, it is seen that polynomial function found in training cannot compensate for amplifier nonlinearity. Its expansion characteristic is just not right. The reason for this is that, the adaptation algorithm in Sec. 3.3.1 is designed for adaptation with point by point procedure, however, the polynomial tries to model whole range at the same time. Calculation of predistortion coefficients by considering a number of points at the same time will make the performance better. One important point in these figures is that all of them are plotted upto about $0.4V$ V_{in} value, not $0.6V$ as in Fig. 4.3, because $0.4V$ is the end point of Input Voltage Domain of predistorter.

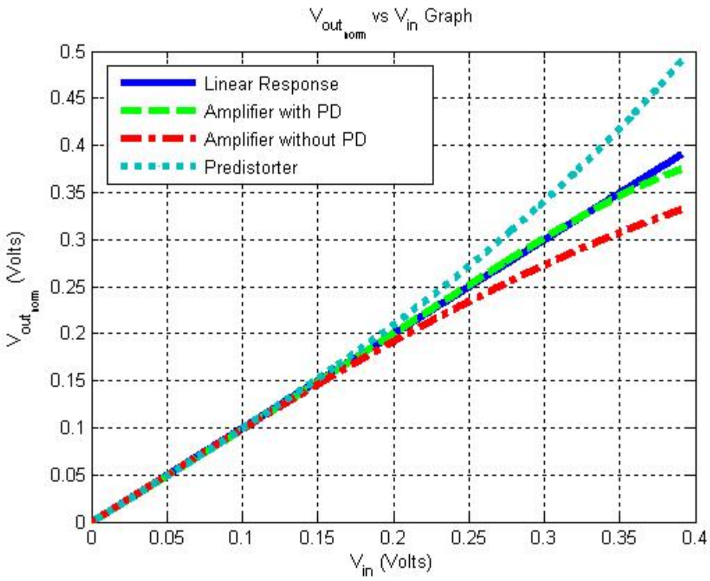


Figure 4.6: AM-AM Responses of Amplifier and Polynomial Predistorter

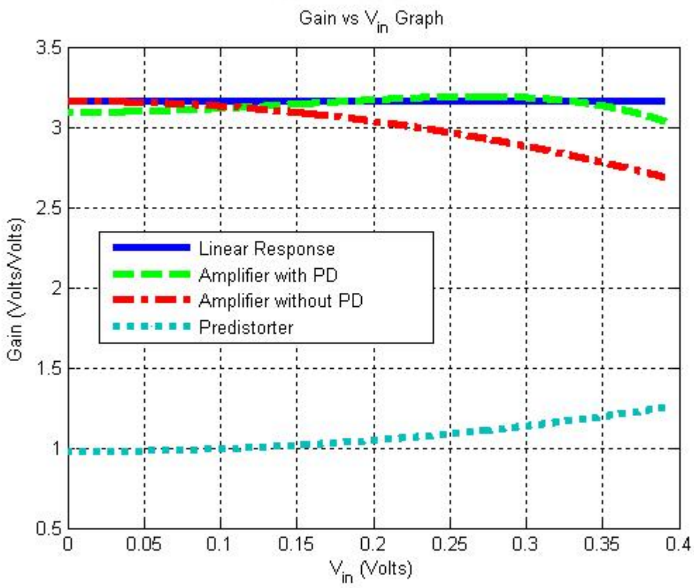


Figure 4.7: Gain Responses of Amplifier and Polynomial Predistorter

Phase Predistortion decreases the overall phase shift to lower than 1 degrees, however, as gain predistortion is not good enough phase polynomial cannot be evaluated.

By using these functions, the input and output power spectral density responses to $8PSK$ input mentioned at the beginning are plotted in Figs. 4.9 and 4.10.

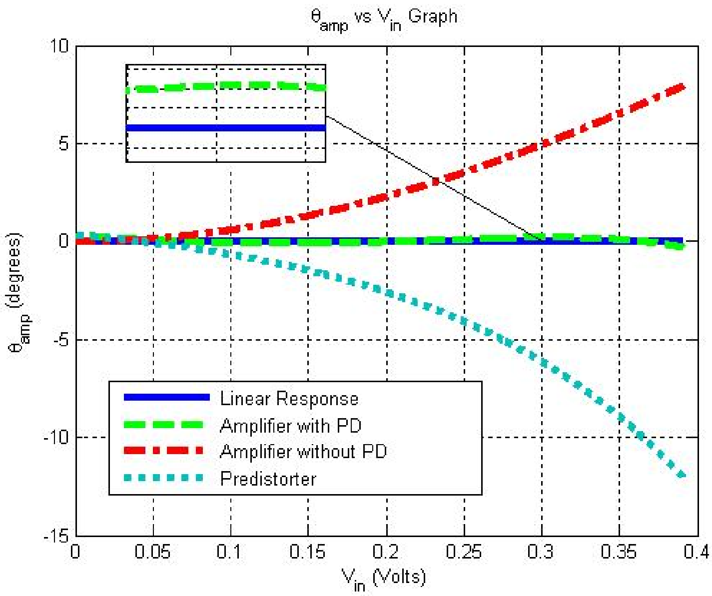


Figure 4.8: AM-PM Responses of Amplifier and Polynomial Predistorter

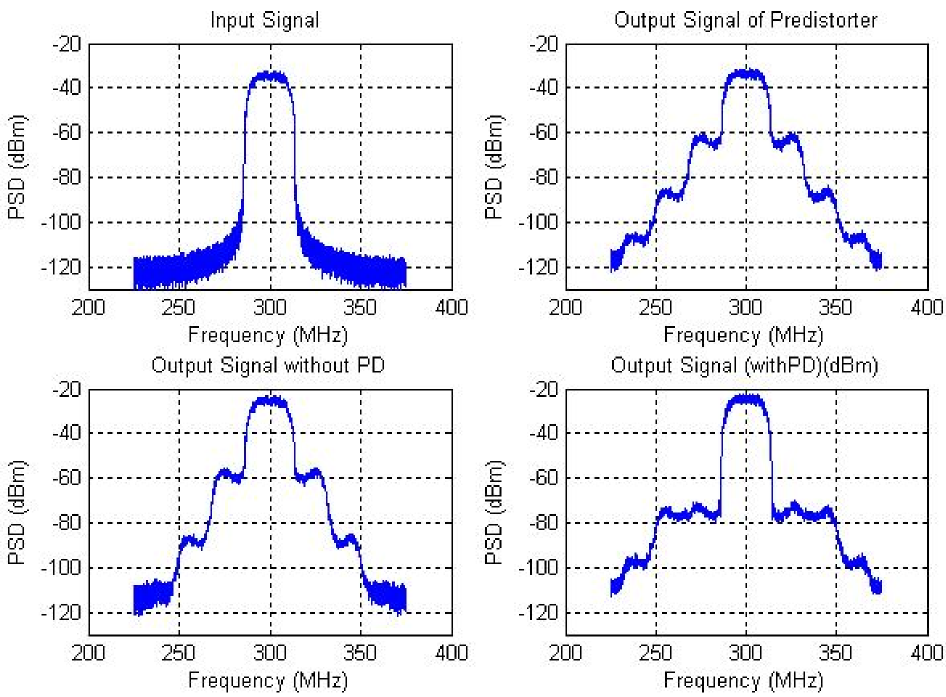


Figure 4.9: Power Spectral Densities of Input and Output(Polynomial)

Improvement of $15dB$ in adjacent channel and degradation of $12dB$ in alternate channel can be seen from the figures. One important observation in Fig. 4.9 that should be emphasized

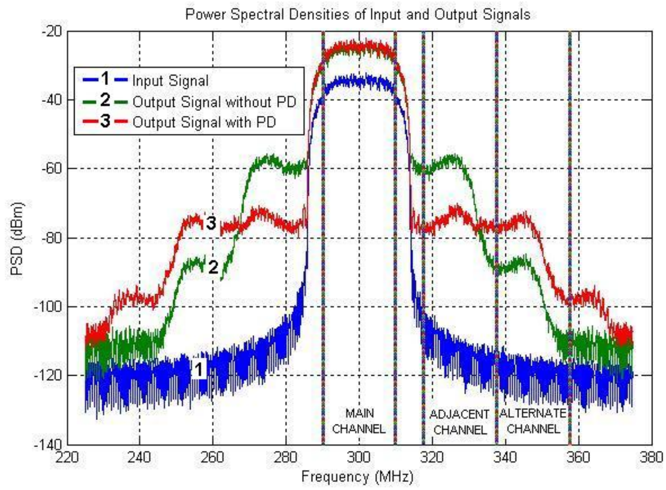


Figure 4.10: Power Spectral Densities of Input and Output(Polynomial)(SinglePlot)

is the PSD of predistorter output. As gain predistortion polynomial is 7^{th} order, the spectrum includes distortions up to 2^{nd} alternate channel. This proves the discussions made in Sec. 3.4.1.

LUT Predistortion is realized by power indexed LUT with a size of 512. The entries of gain and phase predistortion tables are obtained from adaptation with Modified Ramp data with an approximately 5000 points. The characteristics found are plotted for illustration as in polynomial case.

AM-AM and Gain Responses in Figs. 4.11 and 4.12 shows how well the LUT predistorter models the inverse nonlinearity.

Phase LUT is also very successful in modeling inverse nonlinearity as shown in Fig. 4.13. The combination of amplifier and predistorter has nearly zero phase distortion.

By using these look-up tables, the input and output power spectral density responses to $8PSK$ input are plotted in Figs. 4.14 and 4.15.

As seen from figures, the performance is satisfactory. With a correct training, LUT is far better than polynomial case as it achieves $38dB$ linearization compared to $15dB$ of polynomial in adjacent channel. In alternate channel $14dB$ linearization is achieved via LUT predistortion while polynomial distorts there by $12dB$. There are two observations from PSD plots. The predistorter bandwidth spreads again up to 7 times the input signal bandwidth as mentioned

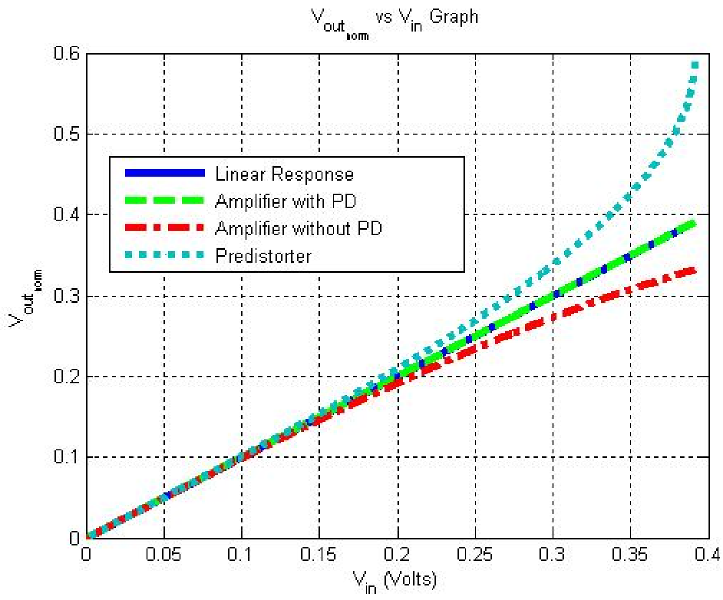


Figure 4.11: AM-AM Responses of Amplifier and LUT Predistorter

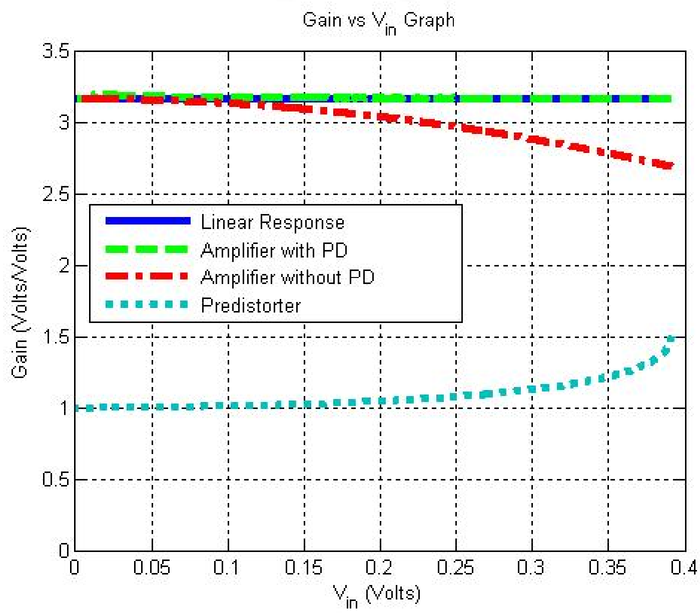


Figure 4.12: Gain Responses of Amplifier and LUT Predistorter

in Sec. 3.4.1. Secondly, the final performance in Fig. 4.15 shows that $15dB$ increase in noise level occurs, while the gain is $10dB$. In 2^{nd} alternate channel, the rise in spectrum is due to predistortion. So, there is still $5dB$ to linearize.

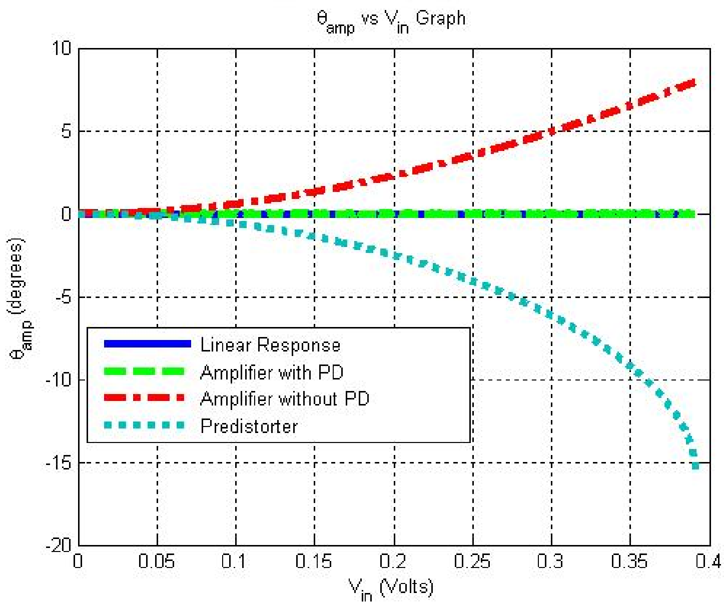


Figure 4.13: AM-PM Responses of Amplifier and LUT Predistorter

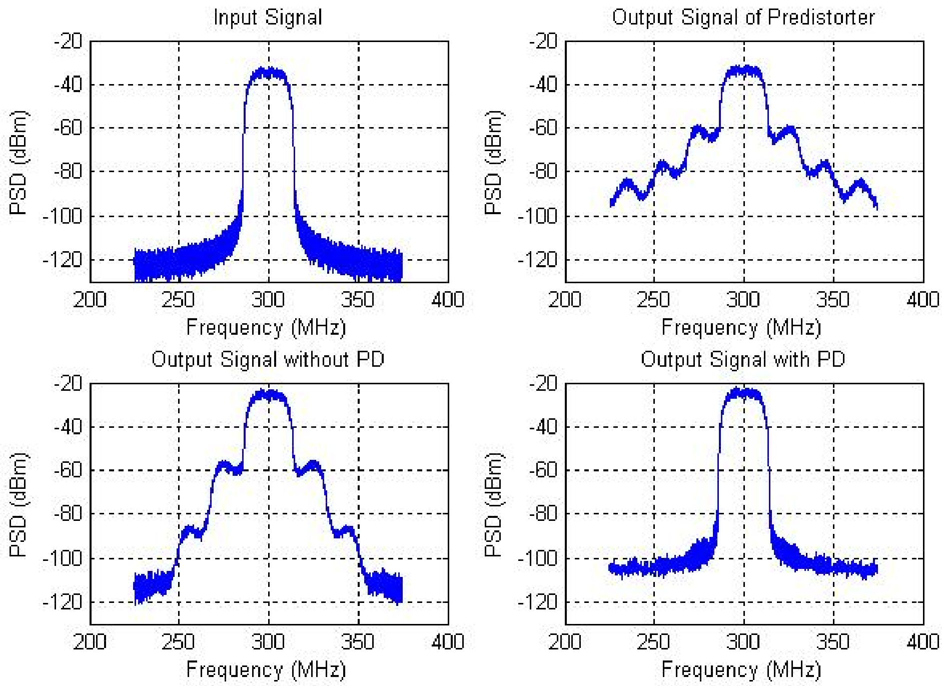


Figure 4.14: Power Spectral Densities of Input and Output(LUT)

Finally, it should also be noted that predistortion can linearize the output power that corresponds to Input Voltage Domain of the predistorter which equals to $\frac{r_{out_{sat}}}{a_1}$ which is approxi-

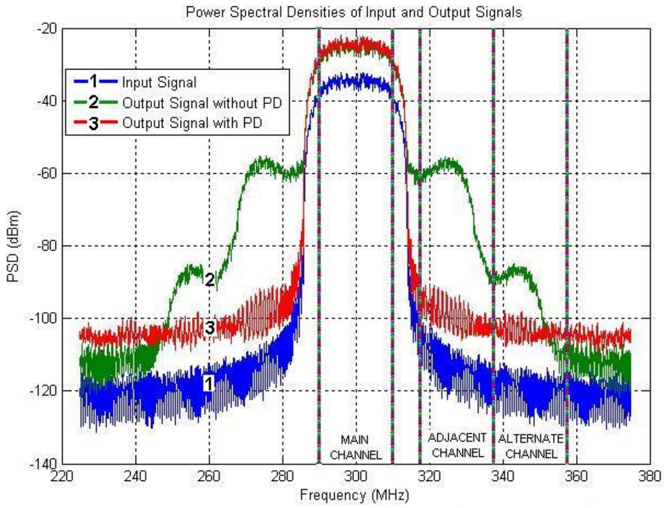


Figure 4.15: Power Spectral Densities of Input and Output(LUT)(SinglePlot)

mately $0.4V$ or $1.5dBm$ in this case. If gain response of amplifier is investigated from Figs. 4.2 and 4.4, it is seen that this point is nearly 1.5dB compression point for this amplifier model.

CHAPTER 5

HARDWARE IN THE LOOP

From the experiences gained in the previous chapter, Look-Up Table Predistortion approach is decided to be used. In order to evaluate LUT predistortion algorithms on real amplifiers rather than simulations, an amplifier is designed. To create and collect data, a testbench is built. Information on the testbench and results obtained from this testbench will be given in this section.

5.1 Testbench

Up to this point MATLAB codes are used to evaluate predistortion. Predistortion parameters are trained with random or ramp data and predistortion performance is evaluated with 8PSK data.

From this point on, predistortion performance on real amplifier will be evaluated. For this purpose an amplifier is designed on PCB and a testbench is built. The testbench used in this section can be seen in Fig. 5.1.

With this testbench, Analog Quadrature Modulation topology is tried to be realized. For this reason Signal Generator(SG) is used for modulation. SG in the testbench has arbitrary waveform generation option, and it is used to create RF signal from I/Q signals calculated in MATLAB. This signal is applied to amplifier and its output is attenuated and delivered to both the Vector Signal Analyzer(VSA) and Spectrum Analyzer(SA). VSA is used for demodulation purpose. It measures and saves I and Q signals. SA is used for evaluation of predistortion by measuring ACPR.

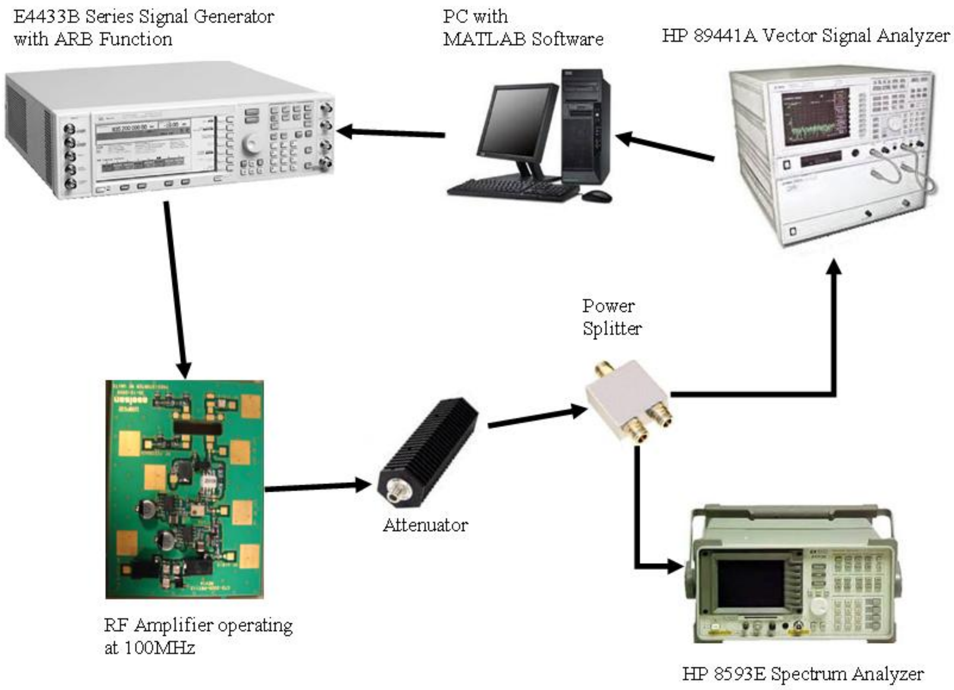


Figure 5.1: Testbench Used to Evaluate Predistortion

5.1.1 Designed Amplifier

An advantage of baseband predistortion is its independence from RF frequency, because the process is realized on I and Q signals. Hence, the amplifier on which the test will be made can operate at any frequency. For this reason, the carrier frequency is selected randomly as 100 MHz in VHF range.

The amplifier is composed of three blocks. Two of them are IC amplifiers with 13 dB gain, 22 dBm P_{1dB} and 41 dBm IP_3. The last stage is designed for operation at 100 MHz. Semelab's D2019UK transistor is used. Biasing and stability circuits are added. The last stage is biased to different levels in order to change the class of operation throughout the tests. However, for 280 mA biasing, it has 24 dB gain, 35 dBm P_{1dB} and 43 dBm IP_3.

5.1.2 Usage of Testbench

There are three phases for realization of baseband predistortion on the testbench offered. These are PA Modeling, Predistortion and Evaluation phases.

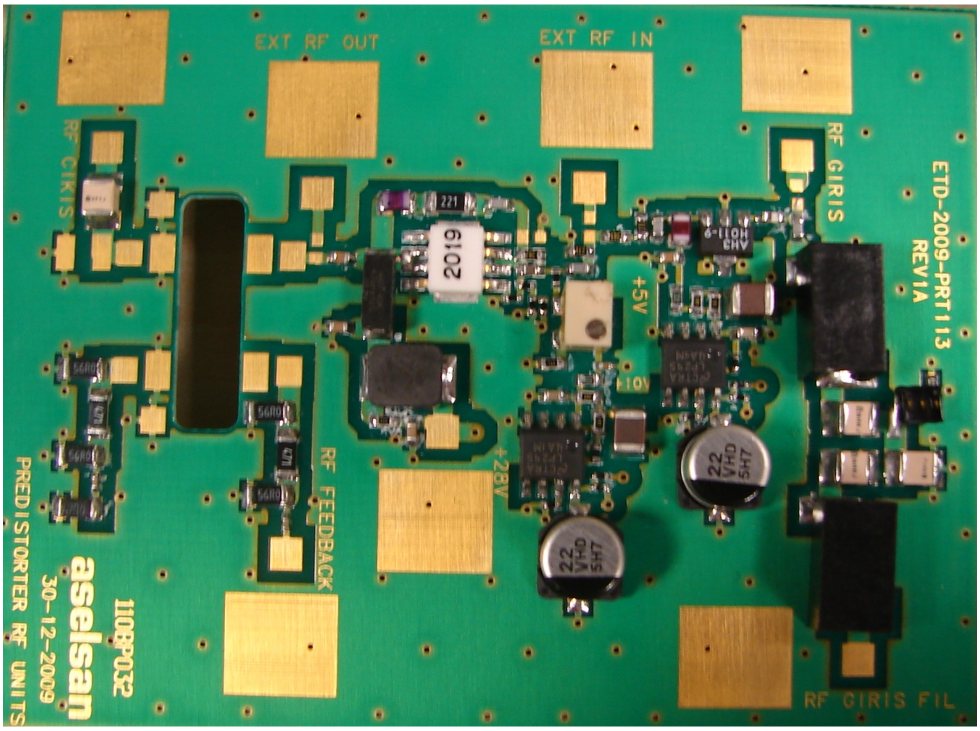


Figure 5.2: Designed Amplifier on PCB

PA Modeling Phase

In PA Modeling phase the aim is to obtain AM-AM and AM-PM responses of amplifier. In Chap. 4, training is done by continuously applying data and by the usage of direct adaptation algorithms, which needs a closed loop. However, in this line-up the loop is not closed. Therefore, PA Modeling with Consecutive Inverse Estimation adaptation method, mentioned in Sec. 3.2, is more suitable.

In PA Modeling phase, signals with varying envelopes can be used as an input signal, however, the envelope should sweep the whole input range of amplifier. Ramp, triangular, sinusoidal or modulated signals can be used as input. However, the bandwidth of the envelope signal should not be very wideband. In fact, as the bandwidth of the modeling data gets larger, memory effects of amplifier become effective. From now on, this input signal is named as *ModelingData*.

In PA Modeling phase, the procedure to obtain the nonlinearities can be listed as follows:

1. Modeling data, which is composed of baseband I and Q signals, is created in MATLAB.

2. Modeling data is loaded to SG's internal memory by using a software provided by the vendor of SG.

3. After necessary adjustments done on SG (like sampling freq., center freq. and recons. filter) amplifier is driven by the RF signal output of SG.

4. The output power of the amplifier is set to saturation power by using power meter.

5. This output is attenuated and divided into two to feed VSA and SA.

6. In VSA, RF signal is demodulated and I and Q signals are sampled by the same sampling frequency used in SG.

7. Demodulated signals are stored into a floppy disk and loaded to MATLAB.

8. In MATLAB, PA Modeling is performed by comparing sent and taken data.

In MATLAB, PA is modeled by using the equations given in Sec. 3.1. However, the data taken from VSA has a certain delay and loop gain, which must be compensated for correct modeling. For this reason, as the first step, delay between data sent and data taken is calculated with correlation method by using amplitudes of the signals. See Sec. 3.4.2.

After compensation of the delay, the loop gain is calculated by averaging the gain of the signals in small signal region. As the data sent is normalized to unity, samples, whose magnitudes are in 0.05 to 0.2 ranges, are used to calculate loop gain. After the compensation of the loop gain AM-AM Model of amplifier is found. For AM-PM response, the difference of phases of input and output is calculated.

The model obtained by following the above procedure can be seen in Figs. 5.3 and 5.4.

Predistortion Phase

In Predistortion Phase, the aim is to fill gain and phase predistortion look-up tables. This is realized by using an algorithm developed independent from but similar to the one mentioned in [44]. There is a procedure to obtain gain and phase LUTs and to explain it easily a sample predistortion table is given in Tab. 5.1. The procedure to obtain LUT parameter can be listed as follows;

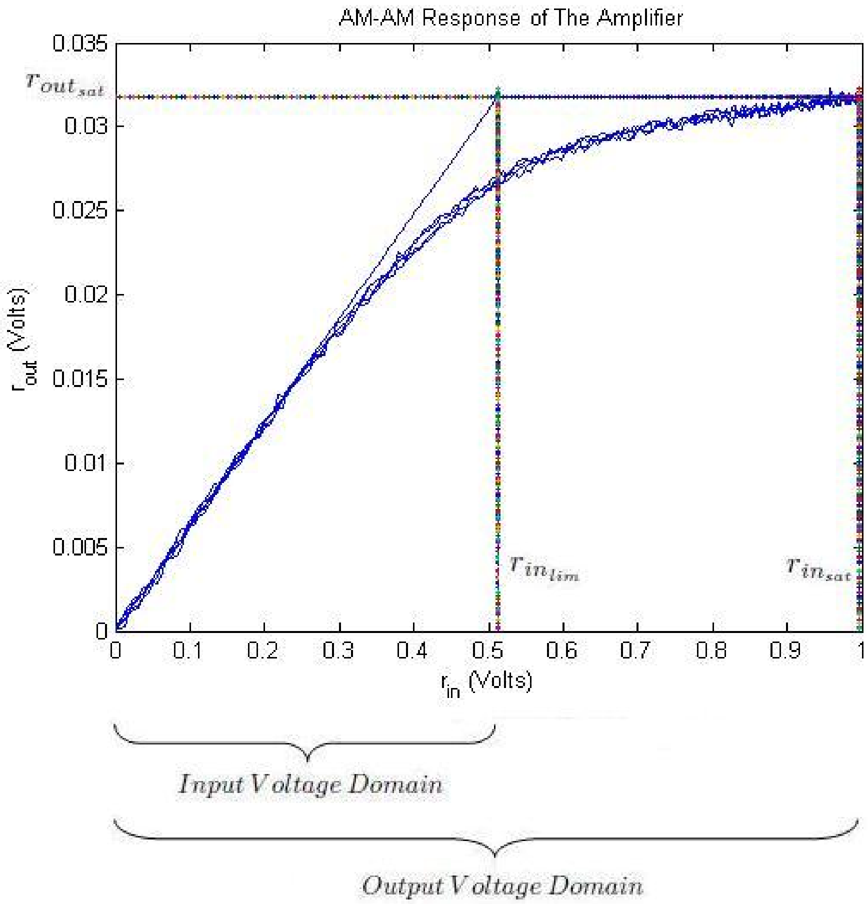


Figure 5.3: AM-AM Response of the Amplifier Obtained in MATLAB

1. AM-AM and AM-PM responses obtained in the previous phase are listed and named as; r_{sent}, r_{taken} and θ_{diff}, which are amplitude vector of input signal, amplitude vector of output signal divided by small signal gain and phase difference vector of output and input signals, respectively.

2. Differential voltage of LUT entries is determined and is named as r_{step}.

3. Gain and Phase LUTs are initialized with all ones and all zeros, respectively.

4. To calculate predistorter's gain, Tab. 5.1 should be investigated carefully. In the table, data sent and data measured, which are obtained in PA Modeling Phase, are listed in the first two columns. These are input and output of amplifier with a small signal gain of 10. The 3^{rd} column r_{taken} is the normalized output. r_{taken} can also be interpreted as ideal input values for linear amplification.

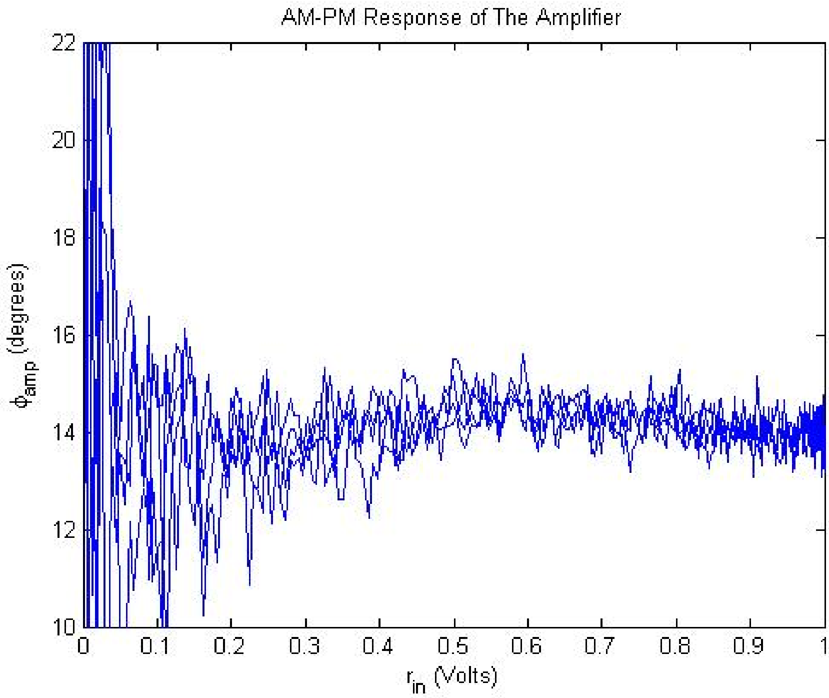


Figure 5.4: AM-PM Response of the Amplifier Obtained in MATLAB

Table 5.1: Sample Predistortion

$r_{sent} = r_{in}$	$r_{meas} = r_{out}$	$r_{taken} = r_{out}/a_1$	θ_{diff}	*Predistorter Gain*	*Predistorter Phase*
1.0	10.0	1.0	0	$1.0/1.0 = 1.00$	0
2.0	20.0	2.0	1	$2.0/2.0 = 1.00$	-1
3.0	29.0	2.9	2	$3.0/2.9 = 1.03$	-2
4.0	38.0	3.8	3	$4.0/3.8 = 1.05$	-3
5.0	46.0	4.6	4	$5.0/4.6 = 1.09$	-4
6.0	54.0	5.4	6	$6.0/5.4 = 1.11$	-6
7.0	61.0	6.1	9	$7.0/6.1 = 1.15$	-9
8.0	68.0	6.8	13	$8.0/6.8 = 1.18$	-13
9.0	74.0	7.4	18	$9.0/7.4 = 1.22$	-18
10.0	80.0	8.0	24	$10.0/8.0 = 1.25$	-24

Let's focus on the bottom row. Input is 10V, output is 80V and normalized output is 8V. To observe 80V at the output of the amplifier, linear amplification imposes 8V input to be applied. However, amplifier output becomes 68 when 8V is applied. On the other hand, amplifier output becomes 80V when its input is 10V, so we should, somehow, convert the incoming 8V to 10V. So the predistorter will make the necessary mapping. From now on, it will map input of 8V to 10V so that amplifier output will become 80V. This mapping is realized by multiplication and predistorter gain should be $10/8 = 1.25$. Therefore the formula can be stated as follows;

$$G_{pd}(n) = \frac{r_{sent}(n)}{r_{taken}(n)} \tag{5.1}$$

It is worth noting that r_{taken} values correspond to Input Voltage Domain of predistorter and r_{sent} values correspond to Output Voltage Domain of predistorter. So r_{taken} value should be used to determine which LUT entry should be updated.

5. From r_{taken} value, corresponding LUT entry is found by dividing it to r_{step} and rounding down.

6. Throughout the usage of each value in the input-output lists, r_{taken} value in hand, generally, does not belong to some specific entry, as it is a real number. So it should affect both the upper and lower entries. If, somehow, it belongs to a specific entry, then the algorithm will affect only that entry. For the time being, let's say it is somewhere in between. There is one gain information and two entries to be updated, therefore, the previous r_{taken} value can be used for the aim, assuming that input data is continuous. Then, the correct gain values that belong to LUT entries can be obtained as follows;

$$G_{pd}(i) = G_{pd}(n) + \left[\frac{G_{pd}(n) - G_{pd}(n-1)}{r_{taken}(n) - r_{taken}(n-1)} \left(r(i) - r_{taken}(n)\right)\right], \tag{5.2}$$

$$G_{pd}(i+1) = G_{pd}(n) + \left[\frac{G_{pd}(n) - G_{pd}(n-1)}{r_{taken}(n) - r_{taken}(n-1)} \left(r(i+1) - r_{taken}(n)\right)\right], \tag{5.3}$$

where n denotes the order of values in the lists, i denotes the table entries. These values are used to update the LUT entry. In order to equally weight the values obtained for a certain entry, averaging is used. To do that, a variable that counts the number of values used to calculate the average up to that instant is held. The update formulas are;

$$LUT_G(i) = \frac{N_i * LUT_G(i) + G_{pd}(i)}{N_i + 1}, \tag{5.4}$$

$$LUT_G(i+1) = \frac{N_{i+1} * LUT_G(i+1) + G_{pd}(i+1)}{N_{i+1} + 1}, \tag{5.5}$$

For phase LUT, the approach is similar. When 80V occurs at the output of the amplifier, then, it is known that amplifier causes 24 degrees phase shift, therefore, if 8V is applied to predistorter it will map the magnitude to 10V and it should shift the phase by -24 degrees to make total shift 0 degrees. Therefore; the formula is;

$$\phi_{pd}(n) = -\theta_{diff}(n), \tag{5.6}$$

Considering that the r_{taken} value corresponds to a point between two indexes, the phase LUT entries can be obtained from the following formulas;

$$\phi_{pd}(i) = -\theta_{diff}(n) - \left[\frac{\theta_{diff}(n) - \theta_{diff}(n-1)}{r_{taken}(n) - r_{taken}(n-1)}(r(i) - r_{taken}(n))\right], \tag{5.7}$$

$$\phi_{pd}(i+1) = -\theta_{diff}(n) - \left[\frac{\theta_{diff}(n) - \theta_{diff}(n-1)}{r_{taken}(n) - r_{taken}(n-1)}(r(i+1) - r_{taken}(n))\right], \tag{5.8}$$

$$LUT_{\phi}(i) = \frac{M_i * LUT_{\phi}(i) + \phi_{pd}(i)}{M_i + 1}, \tag{5.9}$$

$$LUT_{\phi}(i+1) = \frac{M_{i+1} * LUT_{\phi}(i+1) + \phi_{pd}(i+1)}{M_{i+1} + 1}, \tag{5.10}$$

where N and M are arrays of numbers to calculate averages on each entry.

Evaluation Phase

In Evaluation Phase, the most important work is the application of predistortion. The procedure can be described as follows;

1. Data to be transmitted is created and scaled so that it does not exceed Input Voltage Domain of predistorter.
2. According to the indexing method amplitude or power information is calculated.
3. Using the information, correct LUT entry is found. If the data belongs to a point between two entries, the coefficient is found by linear interpolation and applied as mentioned in Sec. 4.2.2.

4. From SA, the output of the amplifier is measured for both predistorted and unpredistorted cases.

In Evaluation Phase, beside Modeling data, different data can be used. For example, a sinusoidal can be used as Modeling Data and 8PSK signal can be used in Evaluation. From now on, the data used in evaluation is called *EvaluationData*.

5.2 Experiments

In order to evaluate performance of predistortion offered in previous part, several tests are done. In the first place, the method is tried to be optimized, so the parameters like indexing, LUT size, LUT construction and filtering are investigated and results that provides optimum performance are determined. Secondly, data formats used in modeling and evaluation are compared and their relations to memory are analyzed. Thirdly, the effect of amplifier classes are examined.

5.2.1 Parameter Performances

Indexing and LUT Size

Amplitude and power indexing are possible ways of indexing. With different LUT sizes, indexing methods give different performances. Results related to these two parameters are obtained under the following conditions and can be seen in Fig. 5.5.

- Class A amplifier with 36 dBm output power is used.
- Two-Tone Data is used as Modeling and Evaluation Data.
- LUT construction with linear interpolation, explained in 5.1.2, is performed.
- Filtering applied to I/Q signals taken (in MATLAB).

Interestingly, power indexing works well even with LUT size of 4, which is possibly due to linear interpolation and amplifier's smooth response. This result verifies the observation

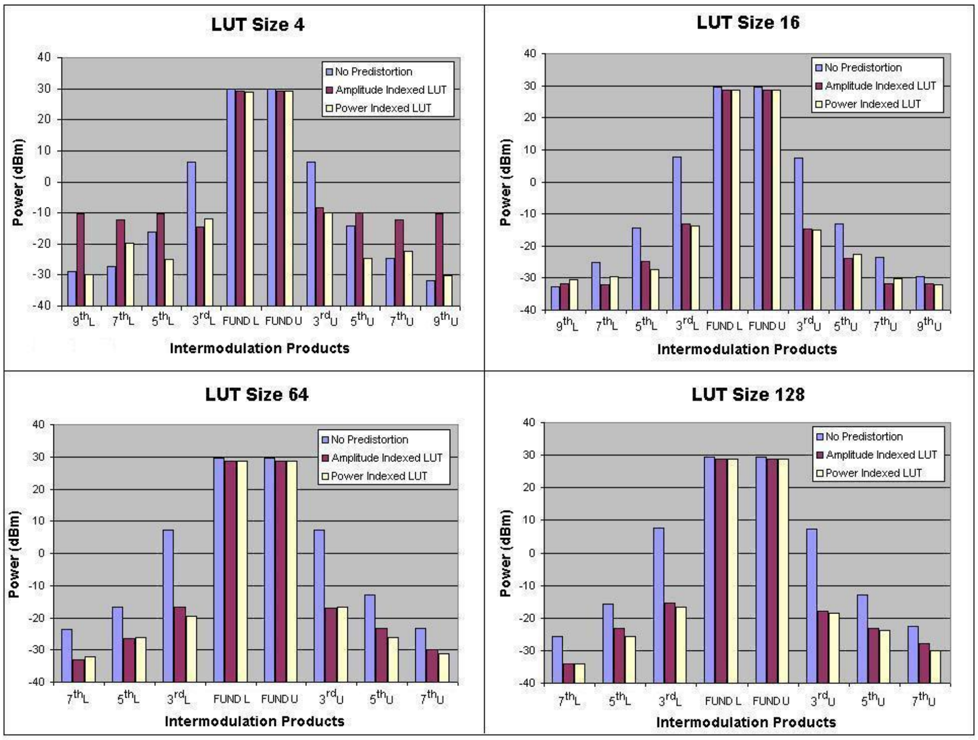


Figure 5.5: Comparison of Different Indexing Methods and LUT Sizes

obtained in Sec. 4.3, which is the fact that power indexing works better for smaller LUT sizes. For LUT sizes, of 16 and 64, satisfactory results are achieved. For these sizes amplitude and power indexing give approximately similar results. For LUT size of 128 the intermodulation levels achieved are listed for numerical observation.

Table 5.2: Intermodulation Levels for 128 LUT Size

Size of LUT	$IM7_L$(dB)	$IM5_L$(dB)	$IM3_L$(dB)	$IM3_U$(dB)	$IM5_U$(dB)	$IM7_U$(dB)
No Predistortion	55,25	45,12	21,96	22,19	42,38	52,23
Ampl. Ind. Pred.	62,82	51,99	44,27	46,62	52,04	56,59
Power Ind. Pred.	62,79	54,57	45,53	47,33	52,64	58,74

It is seen that with 128 LUT size and power indexing, nearly 24 dB improvement can be achieved in 3^{rd} order intermodulation for Class A amplifier with 36 dBm output power. For 5^{th} and 7^{th} orders 10 dB and 7 dB improvements are observed, respectively.

LUT Construction

In Sec. 5.1.2, the method used for LUT construction is mentioned. In this approach LUT entries are calculated by linear interpolation of two data. This can be done without interpolation. For that approach, data affects only the nearest LUT entry. The performances of both approaches are examined and can be seen in Fig. 5.6.

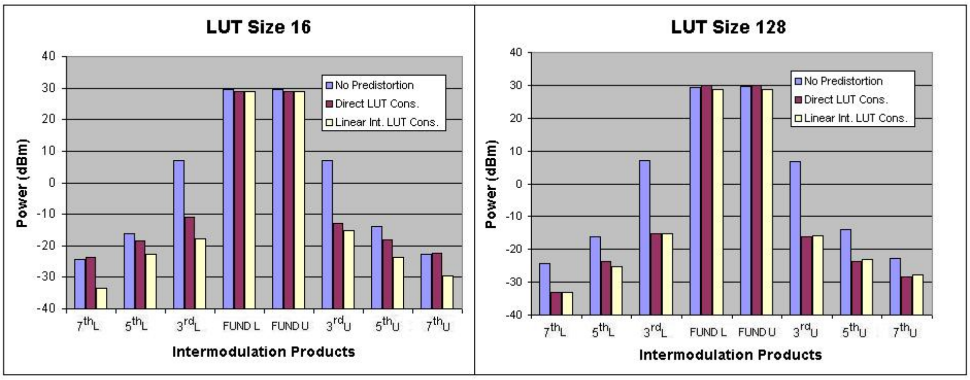


Figure 5.6: Comparison of Different LUT Construction Methods

It is seen that linear interpolation makes predistortion 7 dB better for lower LUT sizes. For LUT size of 128 the perfromances are the same.

Necessity of Anti-Aliasing Filters

As the testbench does not include anti-aliasing filters, data taken includes high frequency noises in the sampling frequency band. This causes the LUT entries to be filled with errors so the performance of predistortion is degraded. Digital FIR filter is used in MATLAB to avoid this problem. An important parameter is the phase response of the filter. The response should be linear in the pass band in order not to add any group delay distortion as mentioned in Sec. 3.4.3. The filter characteristics and evaluation of filtering can be examined in Figs. 5.7 and 5.8.

Effect of filtering can be seen in both LUT sizes. It improves the performance of predistortion up to 4 dB. This proves the necessity of anti-aliasing filters, as they increase SNR of data taken.

Necessity of Phase Predistortion

Phase distortion is an important nonlinearity and needs to be corrected. The algorithm handles it by measuring AM-PM behavior of the amplifier. The importance and necessity of phase predistortion can be seen in Fig. 5.9.

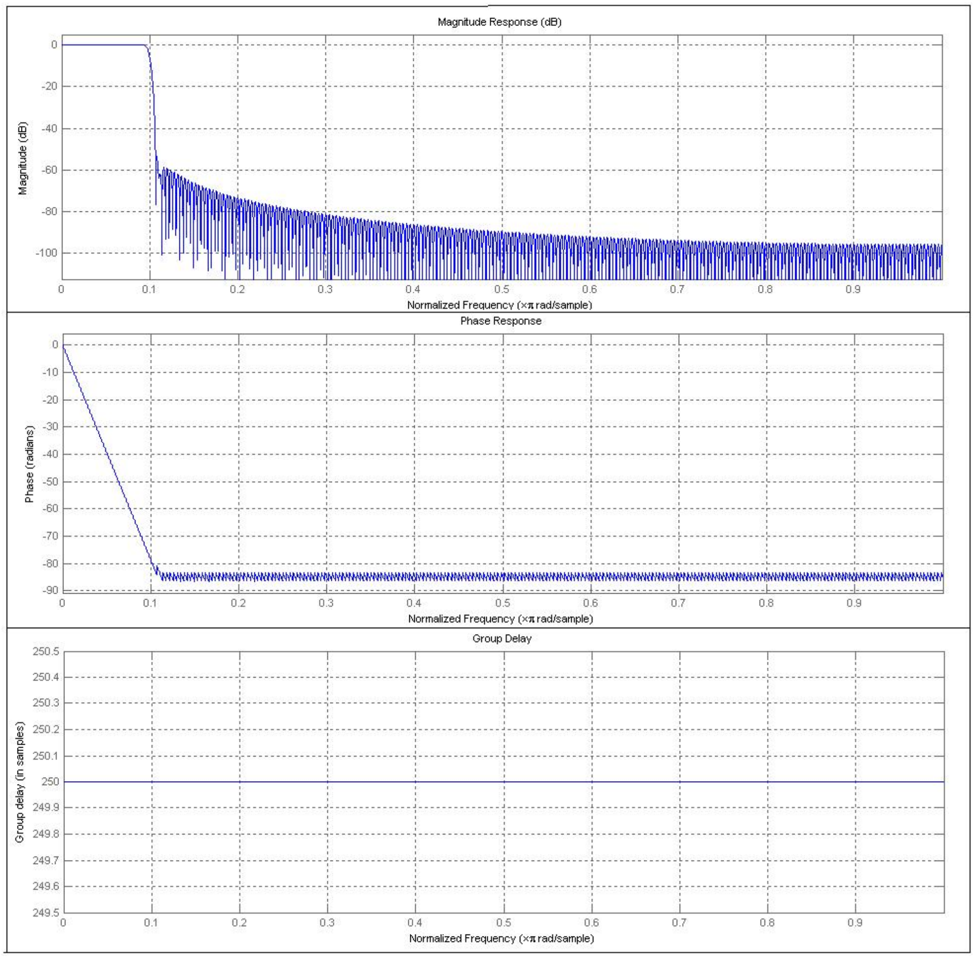


Figure 5.7: Digital Filter Responses

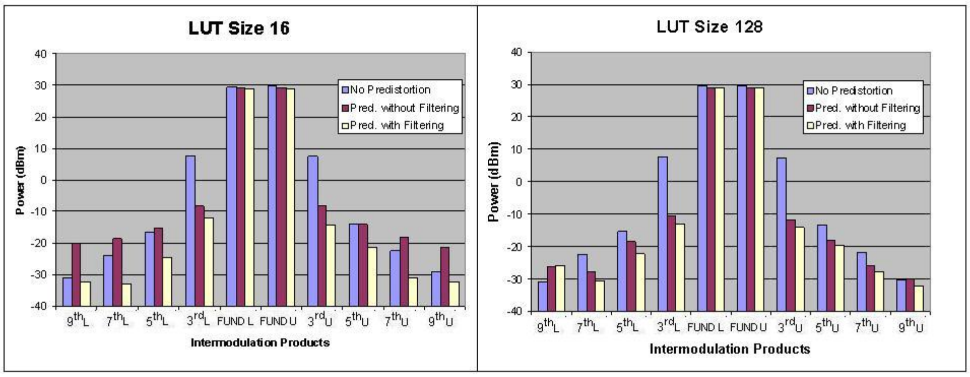


Figure 5.8: Evaluation of Digital Filtering

If only Gain predistortion is applied 7 dB to 12 dB worse performance is obtained and 5 dB asymmetry occurs in 3^{rd} order intermodulation products, which is expected as mentioned in

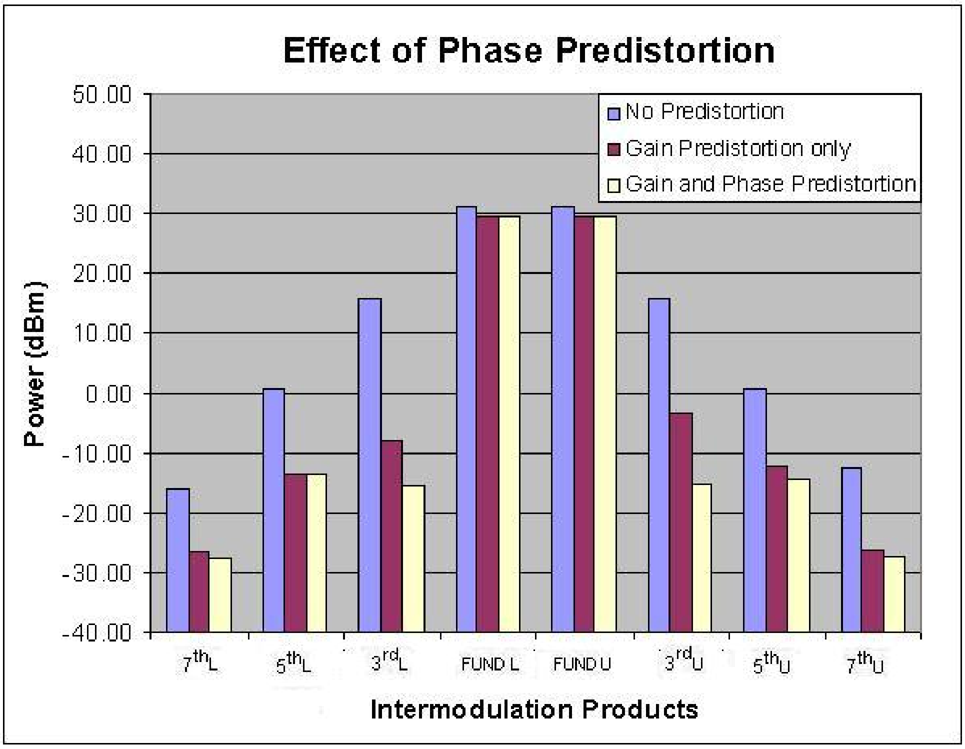


Figure 5.9: Effect of Phase Predistortion

Sec. 2.2. This time Class A amplifier with 37 dBm output power is used.

5.2.2 Data Formats and Memory

In this part effect of Modeling Data on predistortion performance will be examined. As mentioned before, various kinds of data can be used as the Modeling Data, as long as its envelope sweeps the dynamic range of the amplifier. Sinusoidal signals and modulated signals are among them.

5.2.2.1 Two Tone Data

A sinusoidal signal has important characteristics when used in quadrature modulation schemes. A single tone sinusoidal signal in baseband becomes two tone when it is modulated. See App. A for related calculations. Luckily, the performance of predistortion can also be measured with two tone signals effectively. Therefore, a sinusoidal signal in baseband can be used to model the amplifier in PA Modeling Phase and the same data can also be used to evaluate predistortion performances in Evaluation Phase.

At first, two tone signal with 2kHz spacing (1kHz in baseband) is used as Modeling Data. Amplifier Responses are measured by the method mentioned in 5.1.2. The AM-AM and AM-PM responses of the amplifier, operated as Class A with 37 dBm output power, and calculated LUTs can be seen in Fig. 5.10.

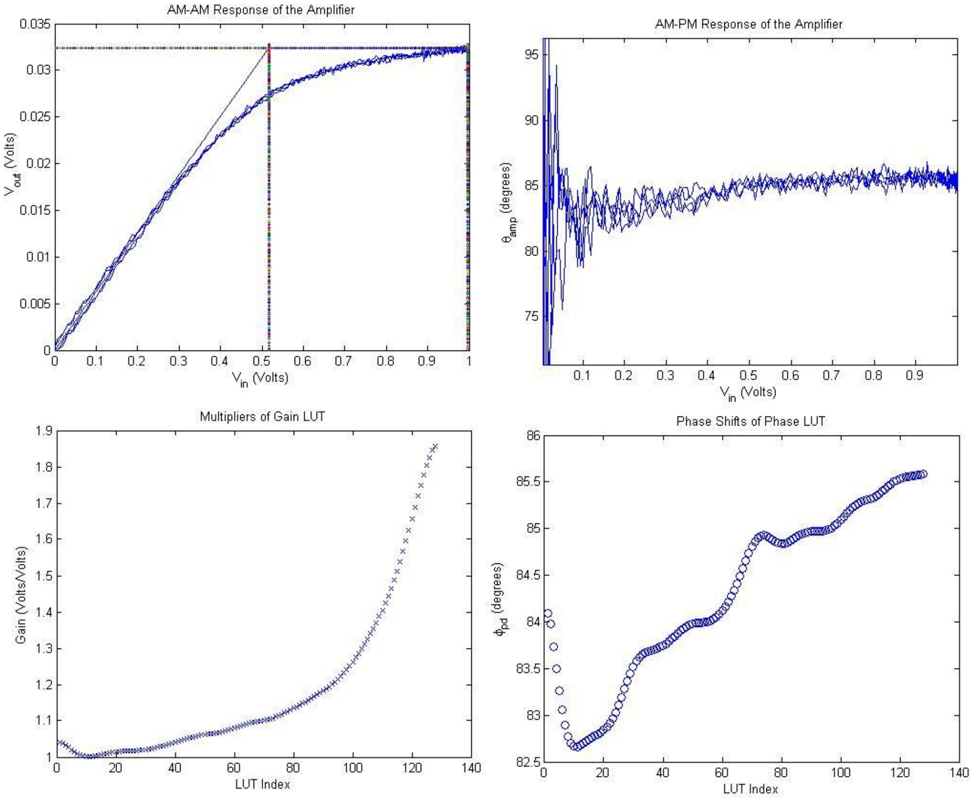


Figure 5.10: Amplifier Responses and Predistortion Tables for 2kHz Spacing

The performance of the predistortion is investigated with 2kHz again. The resultant performance is in Fig. 5.11.

It is seen that, nearly 30 dB improvement in IM_3, 19 dB improvement in IM_5, 10 dB improvement in IM_7, 12 dB improvement in IM_9 can be achieved.

If the validity of the predistortion tables used in Fig. 5.11 is investigated for different spacings, predistortion performance for wideband signals can be predicted. For that reason, data with different spacings are used for Evaluation. Fig. 5.12 shows the improvements for different spacings.

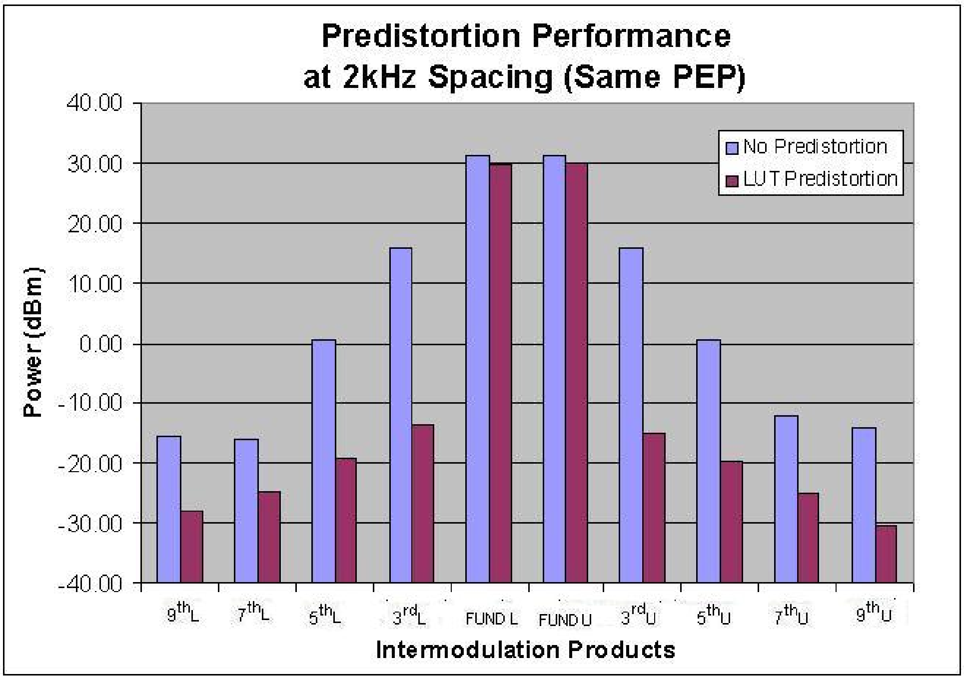


Figure 5.11: Predistortion Performance of 2kHz LUTs for 2kHz Spacing

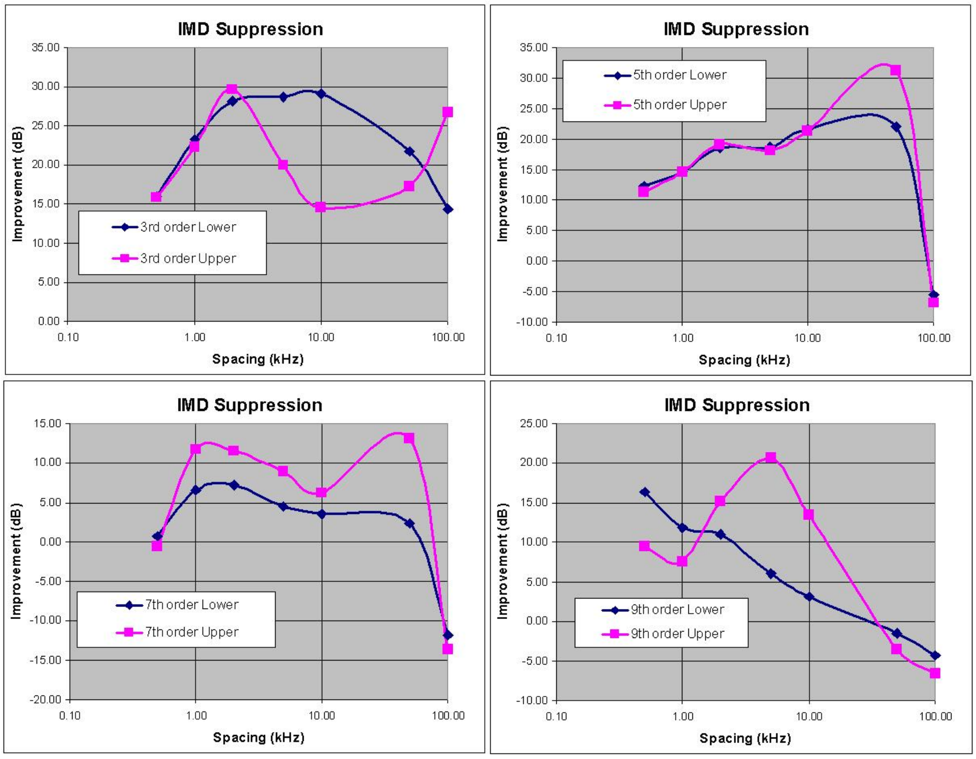


Figure 5.12: Predistrotion Performance of 2kHz LUTs for Different Spacings

It can be seen that, up to 100kHz spacings, 3^{rd} order IMDs are suppressed by, at least, 15 dB. Higher order intermodulation performances show that LUTs obtained by 2kHz spacing is effective up to 50kHz spacings.

Performances on linearization of low spacings degrade because the signal generator used creates two tone signal with low spacings very distorted. For example, 0.5kHz spacing IM_3 of output of SG is nearly 30 dBc, which is the level that the predistorter can achieve.

After investigation of LUT predistortion performance on various tone spacings, linearization of wideband data is examined. For this aim, 8PSK data is used as in Chap. 4. The data is created in MATLAB environment and I and Q components are loaded to SG. 34kHz and 85kHz bandwidths are used for evaluation. For the same peak power, channel powers of predistorted and unpredistorted signals are measured by the Spectrum Analyzer. The results of both bandwidths are nearly the same and only results of 85kHz are given. The power levels can be seen in Tab. 5.3 and Fig. 5.13.

Table 5.3: Predistortion Performance of 2kHz LUT with 8PSK Data

	IM7 L (dBc)	IM5 L (dBc)	IM3 L (dBc)	Main Channel Power (dBm)	IM3 U (dBc)	IM5 U (dBc)	IM7 U (dBc)
Without Pred.	61.0	50.6	24.0	33.5	25.5	51.3	62.4
With Pred.	60.6	59.4	50.1	31.3	50.2	59.3	61.4

For better visualization amplitudes of predistorted and unpredistorted data are shown in Fig. 5.14.

5.2.2.2 Wideband Data

For modulation techniques that have no synchronization time to model amplifier, it seems that the modulation itself can be used as Modeling Data. For that purpose, 8PSK data is used, and amplifier is characterized. The amplifier responses and predistortion tables can be seen in Fig. 5.15.

The linearization results obtained by the usage of 8PSK data as Modeling data can be seen in Tab. 5.4 and Fig. 5.16.

From the Tab. 5.4, the performance of 8PSK data used for Modeling does not work as good as two tone data. This is because of the memory effects of the amplifier that appear when

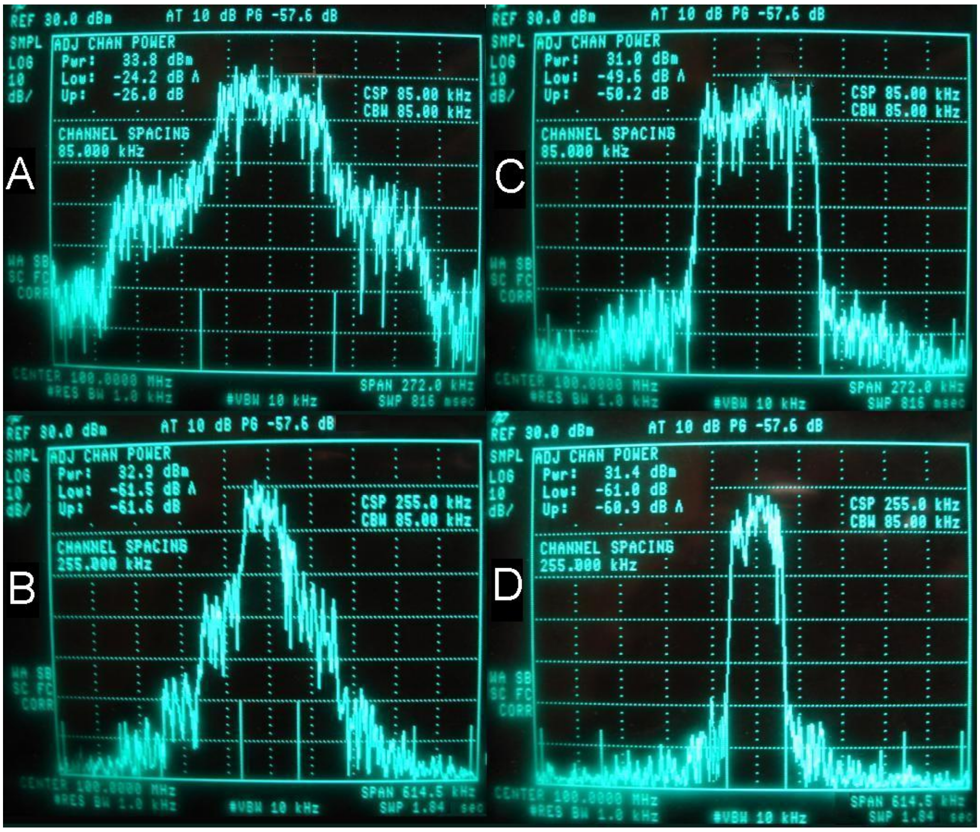


Figure 5.13: Predistortion Performance of 2kHz LUT with 8PSK Data A)Unpredistorted case adjacent channel view B)Unpredistorted case wideband view C)Predistorted case adjacent channel view D)Predistorted case wideband view

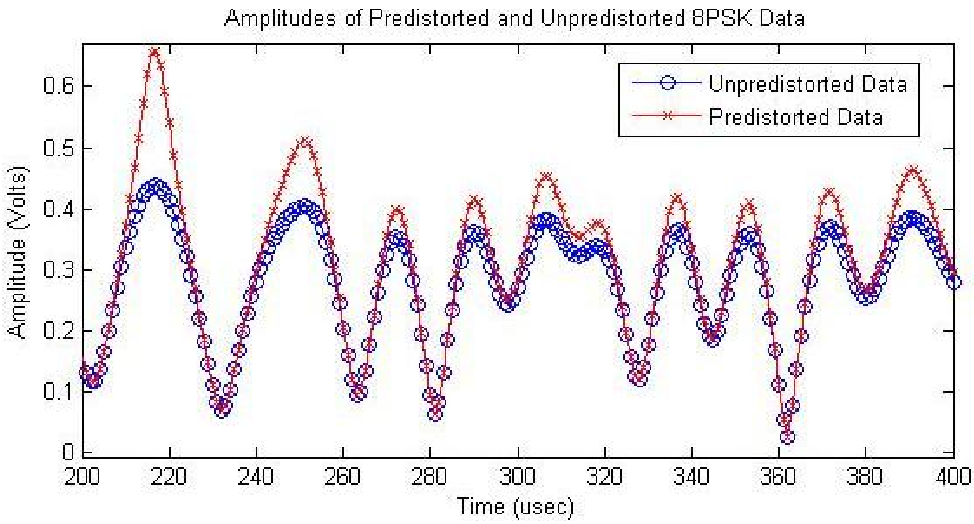


Figure 5.14: Amplitudes of Predistorted and Unpredistorted 8PSK Data

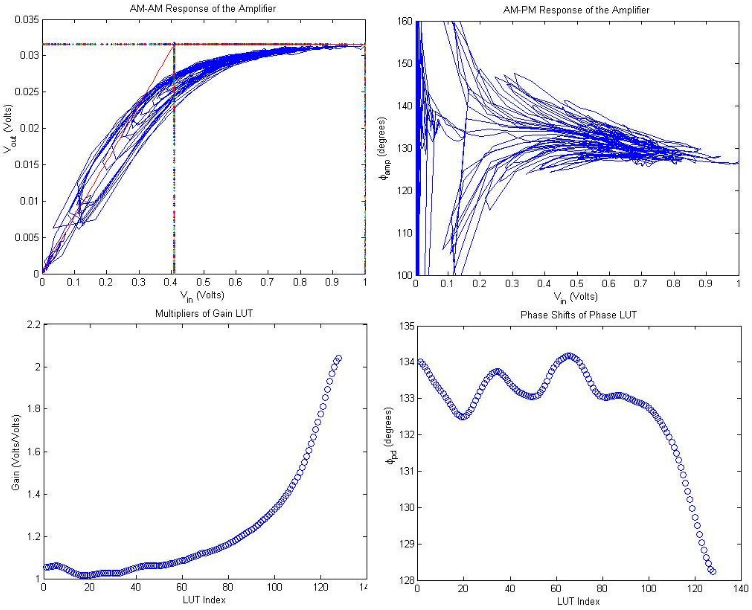


Figure 5.15: AM-AM and AM-PM Responses of the Amplifier and Predistortion Tables obtained by 8PSK Data

Table 5.4: Predistortion Performance of 8PSK LUT with 8PSK Data

	IM7 L (dBc)	IM5 L (dBc)	IM3 L (dBc)	Main Channel Power (dBm)	IM3 U (dBc)	IM5 U (dBc)	IM7 U (dBc)
Without Pred.	61.5	50.4	24.2	33.8	26	51.5	61.6
With Pred.	43.0	41.3	40.5	31.3	40.5	41.9	43.9

wideband data is used. In Fig. 5.15, AM-AM and AM-PM Responses are spreaded, so important parameters like small signal gain, loop delay, Input Voltage Domain of the predistorter cannot be calculated correctly. This results in miscalculation of LUT entries. Nevertheless, the 3^{rd} order intermodulation products are suppressed by nearly 15 dB, while 5^{th} and 7^{th} are degraded by 10 dB to 15 dB.

5.2.3 Amplifier Classes

Effects of amplifier classes on efficiency and linearity are mentioned in Sec. 2.7. As AM-AM responses of amplifier classes change from classical monotonic Class A response, it is important for a predistorter to obtain inverse of those characteristics. In this section, predistortion performance on different amplifier classes will be stated. At the end, a predistorted class C amplifier will be compared with unpredistorted so called “Linear” Class A amplifier.

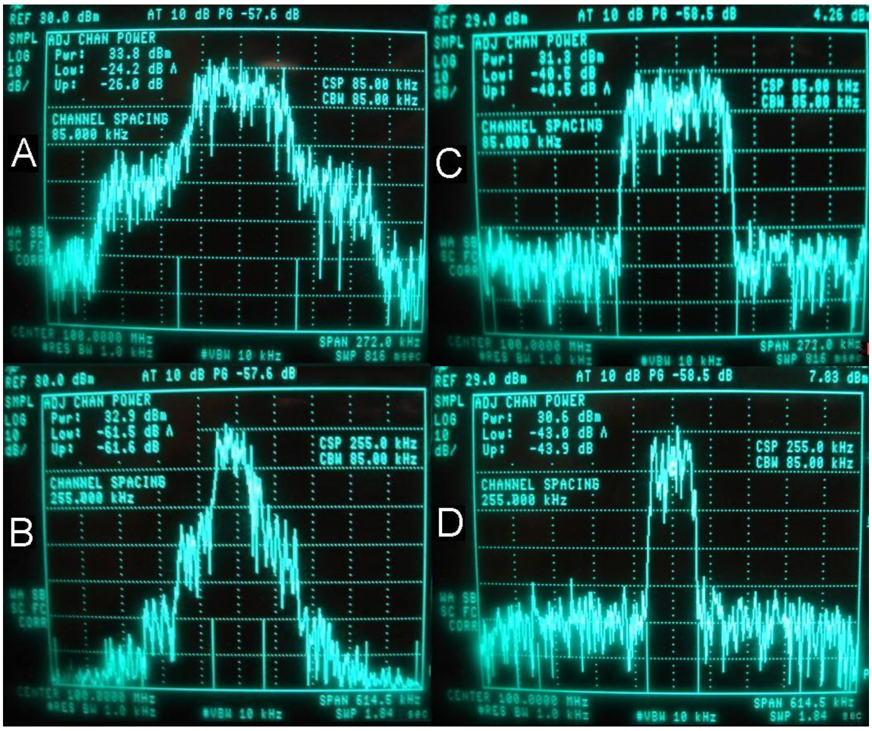


Figure 5.16: Predistortion Performances of 8PSK LUT with 8PSK Data A)Unpredistorted case adjacent channel view B)Unpredistorted case wideband view C)Predistorted case adjacent channel view D)Predistorted case wideband view

Amplifier classes are defined in terms of output power, output impedance and bias point. For example, for a Class A amplifier if output power is 2 Watts and output impedance is 50 Ohms then quiescent current drawn should be at least 282 mA, for 28V drain voltage. However, in predistortion methods as the main purpose is to push amplifier into high saturation regions to get the most, the amplifier will no longer behave like Class A from linearity point of view.

For an amplifier, biased as Class A, the responses of the amplifier, the LUTs of the predistorter and predistortion performances can be seen in Figs. 5.17 and 5.18.

The linearization results in Fig. 5.18 are taken for the same peak power, 5 Watts (37 dBm). The transistor datasheet gives 2.5 Watts rated power, however, more power is taken by pushing it into saturation. This increases the cost efficiency of the design, but, mean time before failures should also be considered. From the results, nearly 30 dB improvement in IM_3, 20 dB improvement in IM_5, 12 dB improvement in IM_7, 13 dB improvement in IM_9 can be achieved.

For Class AB amplifier biased at 50 mA, the responses can be seen in Fig. 5.19.

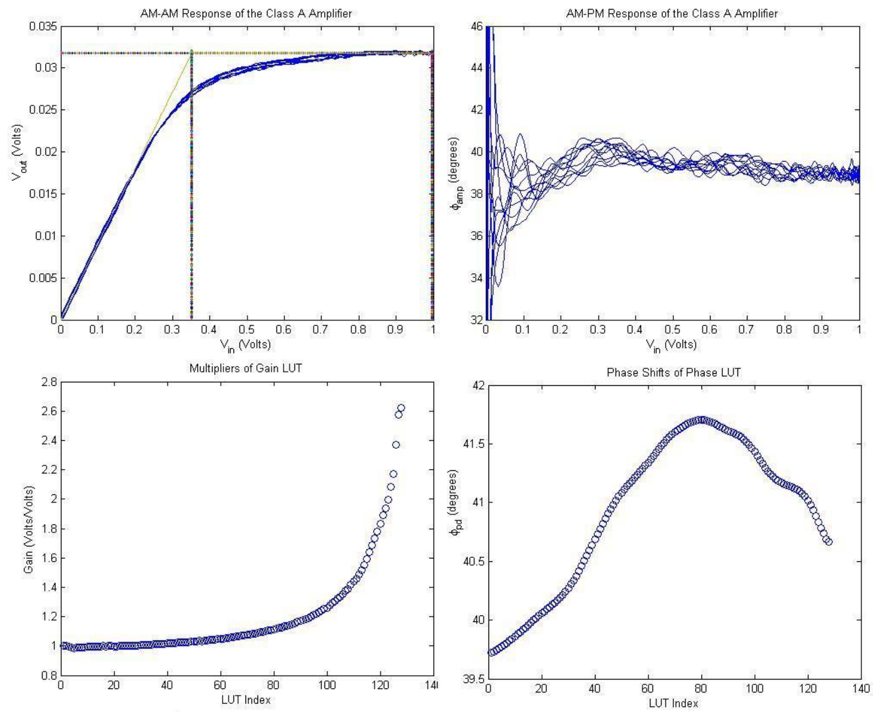


Figure 5.17: Class A Amplifier Responses, Predistortion Tables and Performances

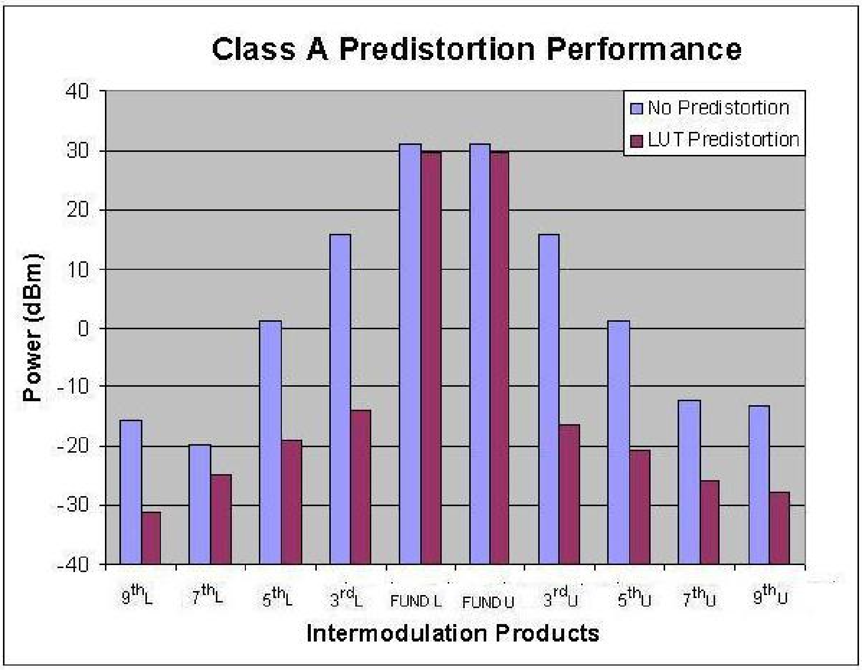


Figure 5.18: Predistortion Performances on Class A Amplifier

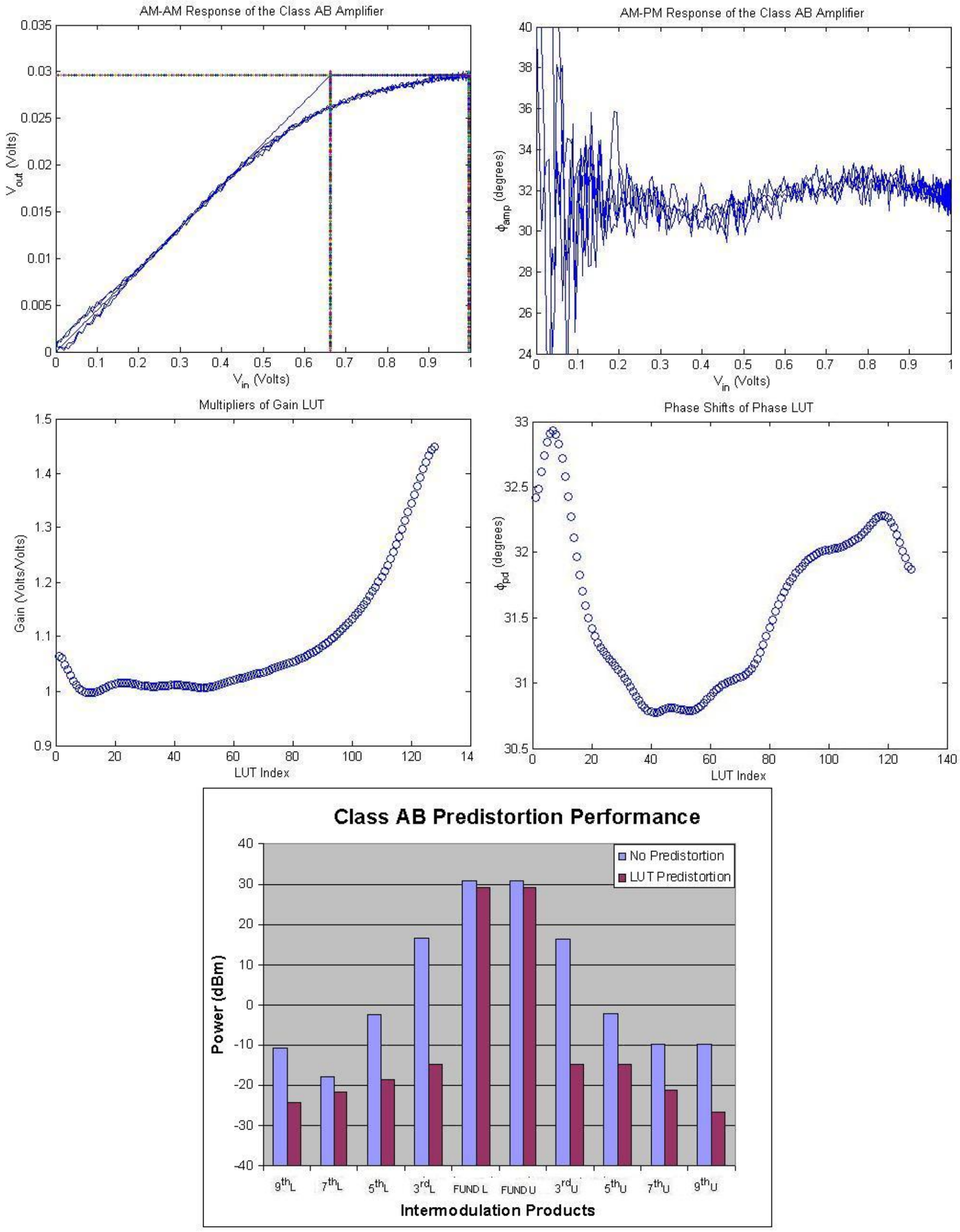


Figure 5.19: Class AB Amplifier Responses, Predistortion Tables and Performances

The results in Fig. 5.19 are obtained at 4.6 Watts (36.62 dBm) peak power. As seen from the AM-AM characteristics of Class AB amplifier, the gain is nearly constant in small signal region as in Class A case, therefore, predistorter coefficients are one in this range. The predistortion performance for Class AB amplifier can be summarized as; 30 dB improvement in IM_3, 15 dB improvement in IM_5, 10 dB improvement in IM_7, 15 dB improvement in IM_9.

For Class B amplifier biased at 0 mA, the responses can be seen in Fig. 5.20.

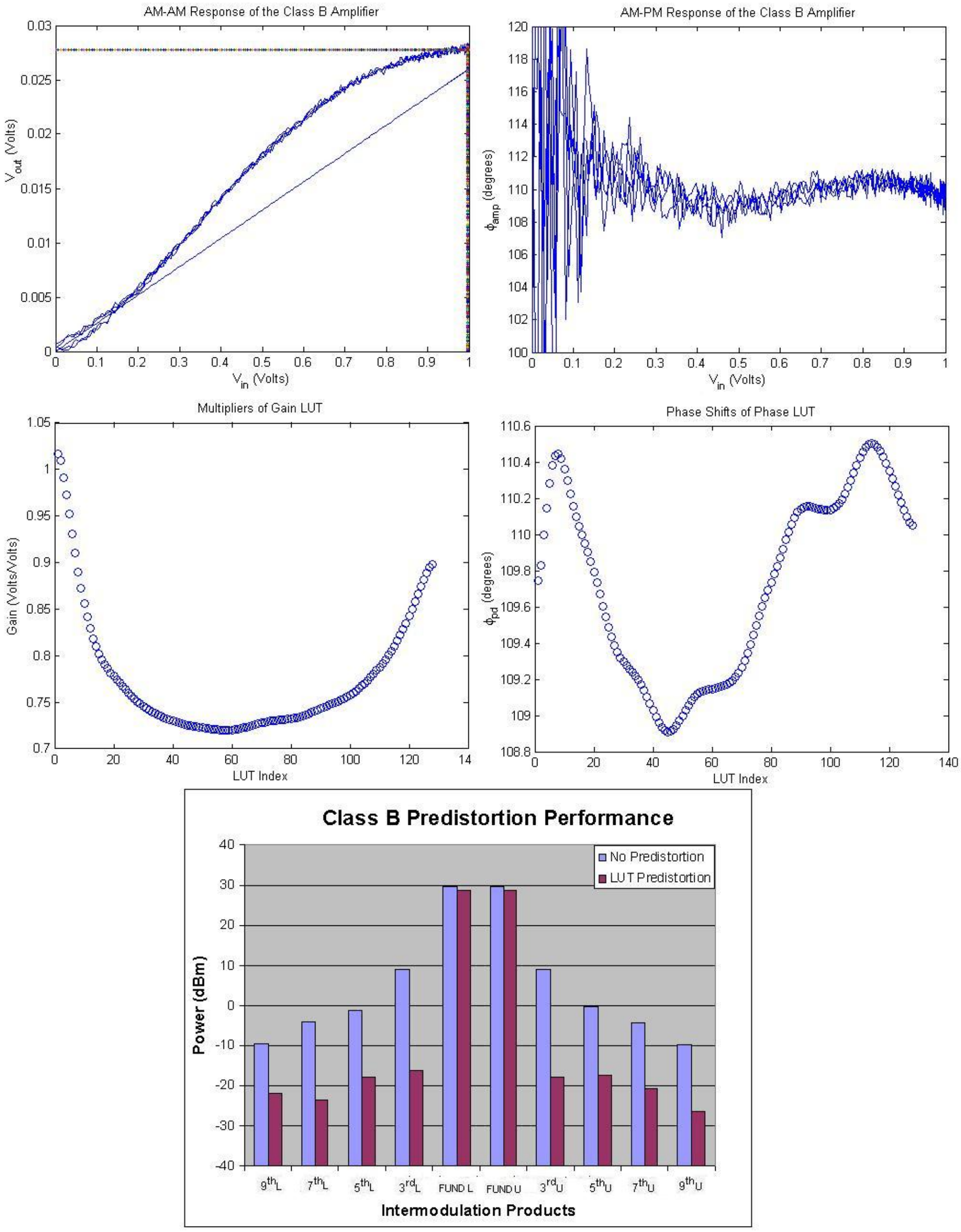


Figure 5.20: Class B Amplifier Responses, Predistortion Tables and Performances

The results in Fig. 5.20 are taken at 4.17 Watts (36.2 dBm). AM-AM response of Class B amplifier is different from that of Class A. It expands after small signal region and then saturates, the predistortion algorithm offered is capable of obtaining the inverse of amplifier response. See Fig. 5.20. The low values of IMDs at the output of the Class B amplifier with respect to Class A case (in unpredistorted cases) is due to the fact that gain of Class B amplifier decreased and the driver cannot drive Class B into higher saturation regions. Nonetheless, the predistorted response achieves the same IMD levels(-15 dBm) as Class A case.

For Class C operation, the amplifier is biased so that conduction angle is nearly 150 degrees, the responses can be seen in Fig. 5.21.

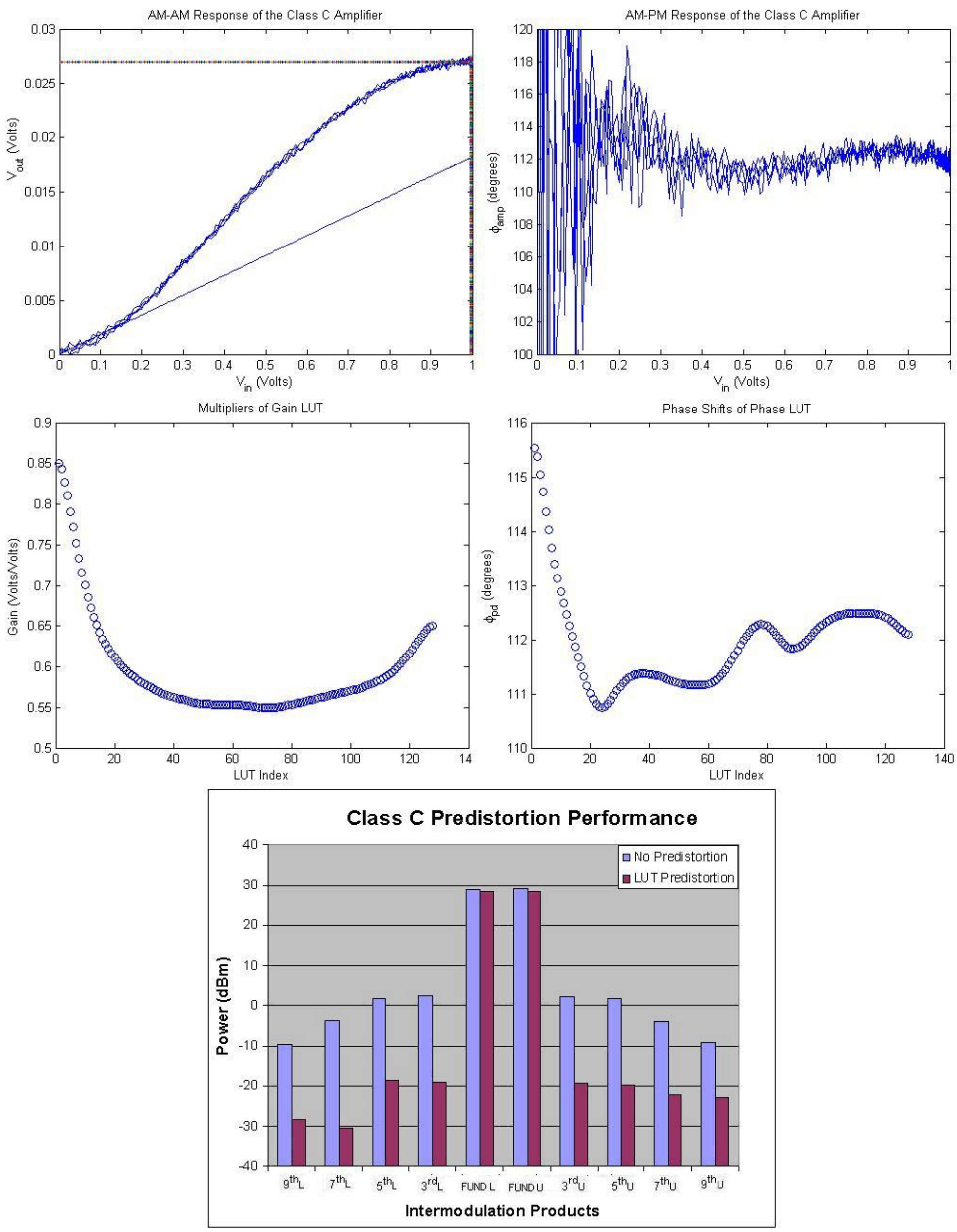


Figure 5.21: Class C Amplifier Responses, Predistortion Tables and Performances

The linearization results in Fig. 5.21 are taken at 4 Watts (36 dBm) peak power. In Class C amplifier the AM-AM response is similar to Class B case and LUT predistorter gets its inverse. The IMD levels of unpredistorted output of Class C is better than that of Class A, because Class C cannot be driven into high saturation by driver amplifier as gain of Class C is very low.

From all the predistortion performances on different amplifier classes, it can be concluded that LUT predistortion method offered in Sec. 5.1.2 can model and obtain any nonlinear characteristics as long as it is continuous. The IMD levels obtained in predistortion results of each class are nearly equal and can be seen in Fig. 5.22.

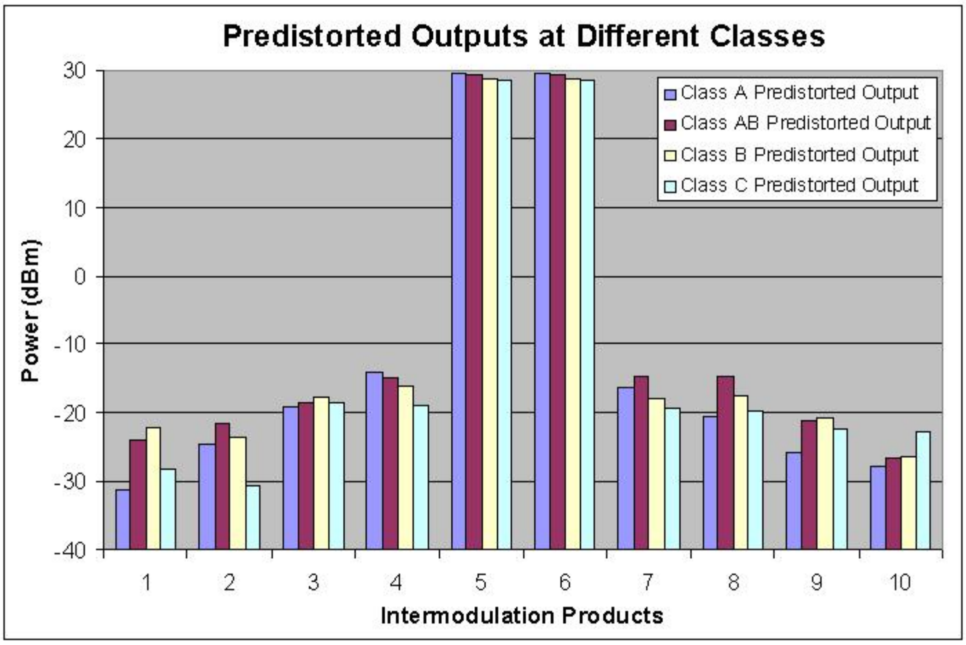


Figure 5.22: Predistorted Outputs at Different Amplifier Classes

As mentioned in Sec. 2.7, there is a way to make amplifier both efficient and linear. At 36 dBm peak power, 162 mA current is drawn form 28V supply in the Class C amplifier with predistortion. By this way, 47 dBc IM_3 and 48 dBc IM_5 are achieved. For the same output power and Class A amplifier, without predistortion, 301 mA current is drawn from supply and 22 dBc IM_3 and 42 dBc IM_5 are observed.

If the Class A amplifier is 3 dB backed-off to make its output 2 Watts (33 dBm) peak, then 282 mA, bias current is drawn, and 36 dBc IM_3 and 52 dBc IM_5 are obtained. Again, Class C with predistortion is more linear, more efficient and has more output power than Class A, which proves the linear and efficient operation.

CHAPTER 6

CONCLUSION AND FUTURE WORK

Newer digital modulation techniques enable high data rates with good spectral efficiency by putting the information on amplitude of the signal. EDGE, OFDM and WCDMA are some of the amplitude modulated signals that are sensitive to linearity of the transmitter. The main nonlinear block in a transmitter chain is the power amplifier. AM-AM and AM-PM responses of power amplifiers cause the adjacent channels to pop up. Firstly, this degrades the spectral efficiency of the modulation which is the reason why that modulation is used. Secondly, allocation in adjacent channels leads to sensitivity deterioration of the cellular radios communicating at those allocated channels. Another result is the degradation of communication quality as the symbols are exposed to these nonlinearities and they are damaged even before the channel. All of these consequences diminish the communication quality. Therefore, as the usage of these techniques are becoming popular, linearity specifications of amplifiers becoming more and more strict.

One way of linear amplification is to operate the amplifier at large back-off, nevertheless, peak to average power ratios of newer modulations can be as much as 12 dB so the efficiency drops to very low values. Crest Factor Reduction Techniques are other possible ways, however, they result in data rate loss or they increase complexity. In order to make a linear amplification without sacrificing efficiency or data rate, Linearization Methods are offered.

Among several techniques, Baseband Predistortion is more attractive due to its correction capability, correction bandwidth and relative cost. Correction capability of predistortion can be up to 20-30 dB for memoryless applications and 50 dB for with memory applications. Correction bandwidth is also a matter of memory content and successful results can be obtained for bandwidth ranges on the order of MHz. Also, as the transmitter architectures include receiver paths, they can be used for predistortion adaptation issues, which reduces the cost for such an auxiliary system.

The main purpose of this thesis was to get an insight of a system that can easily be used inside of the wireless communication structures. This has been accomplished in two ways throughout the thesis. In the first place, the concept of nonlinearity and linearization has been investigated. Secondly, some simulation and hardware in the loop circuits have been offered. So, as a first step, the causes for nonlinearity have been briefly mentioned. The nonlinear behaviours of power amplifiers have been introduced and consequences of these behaviors have been stated. To evaluate the linearity of an amplifier and also to evaluate the predistortion performance some measures of quantities were needed and have been introduced. The behavioral modeling of amplifier is an important concept as it helps one to understand how predistortion should act to model the inverse of the amplifier.

Efficiency is an important parameter as it determines the cost and safety of the system. As non-constant envelope signals needs good linearity, amplifiers are pushed to inefficient operation modes and this increases the need for high power dissipation on the amplifier and the cooling needs, which, in turn, increases the volume and cost of the product. On the other hand, as linearity imposes the usage of more than one transistor to get certain amount of power, cost effectiveness becomes another issue.

Linear amplification problem can be solved by either Crest Factor Reduction Techniques or Linearization Methods. The former results in reduction of data rates and increases complexity of the system. Various linearization methods are offered in the literature. Each of them have certain capability, while they have worse performances in a certain aspect.

Baseband predistortion is a kind of predistortion. It is done in digital domain by using I and Q signals. Theory and classification have been explained and illustrated at the beginning. Classification has proved that there are various possible ways, each of which is a topic of different papers. In this work, the memoryless approaches have been examined as a first step to predistortion concept. The importance of memory content has been taken for granted from previous works, however, the limits of memoryless approaches has been wanted to be measured. Some parameters and line-up component specifications have been deeply investigated and a list of prerequisites for design of such a system has been obtained.

In order to evaluate the effectiveness of a baseband predistortion system, MATLAB simulation environment is used. A realistic amplifier model has been constructed. For baseband predistortion, the inverse of the amplifier nonlinearities should have been obtained so the formation of inverse functions have been realized by Look-Up Tables and Polynomial Functions. Each of them has been optimized through investigation of various system parameters.

For Polynomial Functions effects of polynomial order has been investigated. The theory leads us to think that, if a power amplifier of N^{th} order AM-AM nonlinearity is at hand, then $(N+2)^{th}$ order of inverse nonlinearity is needed at least as each order of nonlinearity contributes to lower order of nonlinearity also. Therefore, as the amplifier model was a 5^{th} order polynomial, then, 7^{th} order predistortion should be used. The simulation results have shown that 7^{th} order gives the optimum results. Optimum training sequence has been analyzed and Ramp and Random input sequences have been compared. Due to the fact that the algorithm optimizes the polynomial coefficients for each single data, random data has been expected to be the optimum training sequence and simulations has proved that. Number of samples used for training has also been analyzed.

For Look-Up Table predistortion LUT indexing and LUT sizes have been investigated. The theory leads to think that as power amplifier compression causes the distortion, power indexing should be better than amplitude indexing. However, some other publications that defend the opposite have been made. The results of the simulations have supported the theory, such that, for lower LUT sizes power indexing has given 10 to 12 dB better performances. For higher LUT sizes the performances of amplitude and power indexing have converged to each other. Interestingly, even with a LUT size of 8, 10 dB improvement in 3^{rd} order intermodulation has been obtained by power indexing. For training of LUT entries a novel method has been offered. The method is based on the physical assumptions. As long as the amplifier responses are continuous the method works well. On the other hand, any discontinuity does not damage the method but lengthens the training duration. By this way, 20 times faster convergence time to obtain optimum LUT entries has been achieved with respect to the usage of random data.

In order to compare Polynomial and LUT predistortion methods, an 8PSK data with 3,13dB Crest Ratio and 16640 sample points is used as input to evaluate the adjacent channel improvement. The results has shown that LUT approach has had better results as it has achieved

38 dB linearization compared to 15 dB of polynomial predistortion in adjacent channel. In alternate channel 14 dB linearization has been achieved via LUT predistorton while polynomial predistortion distorts there by 12 dB.

Predistortion on amplifier model has given satisfactory results, however, the performances on a real amplifier have also been investigated. A testbench has been constructed and a different algorithm has been used. The reason is that the previous methods have been applicable to closed loops because of their direct adaptation. However, the testbench constructed has been an open loop. Therefore, a different adaptation method, PA modeling with consequtive inverse estimation adaptation method has been used. In this approach, at the first place, the amplifier has been modeled by applying a special input data called Modeling Data. Then, amplifier has been modeled from the data taken. Inverse of it has been obtained by logical inferences and LUT entries have been obtained. By this way, a different kind of LUT predistortion has been tried on a real amplifier.

Some experiments have been done on this method. At the first hand, the method has been optimized, so the parameters like indexing, LUT size, LUT construction and filtering have been investigated and results that provide optimum performance have been determined. Secondly, data formats used in modeling and evaluation have been compared and their relations to memory have been analyzed. Thirdly, the effect of amplifier classes have been examined.

At the beginning, LUT indexing and LUT sizes have been examined. In conjunction with the results obtained in simulation power indexing has given better results for lower LUT sizes and for higher LUT sizes performances have converged. It has been seen that with 128 LUT size and power indexing, nearly 24 dB improvement can be achieved in 3^{rd} order intermodulation. For 5^{th} and 7^{th} orders 10 dB and 7 dB improvements are observed, respectively. On the other hand, it has been shown that, if linear interpolation is used for LUT construction, 7 dB improvement can be achieved for lower LUT sizes like 16. For higher LUT sizes like 128 the performances have become the same. Also the importance of anti-aliasing filters have been shown. Up to 4 dB improvement has been achieved through the usage of anti-aliasing filters. Phase predistortion is a part of memoryless predistortion and is needed. The importance of phase predistortion has been shown by neglecting it and 7 dB to 12 dB worse performance has been achieved.

In the second place, data formats used for PA modeling and evaluation have been investigated. Two tone data and wideband data have been used for modeling. As memory effects of amplifier has been appeared when wideband data is used, the two tone data used for modeling has given better performance. With two tone data nearly 30 dB improvement in IM_3, 19 dB improvement in IM_5, 10 dB improvement in IM_7, 12 dB improvement in IM_9 can be achieved. This has been the best performance, for different spacings at least 15 dB improvement in IM_3 is achieved. The performance has been verified by applying 8PSK data with 85kHz bandwidth. The adjacent channel emission has been diminished by 25 dB and in the alternate channel it has been 10 dB. If wideband data is used for modeling, then for the same data the emission improvement has dropped down to 16 dB in adjacent channel and alternate channel emission has been increased.

Predistortion performance on different amplifier classes have been investigated. High efficient classes like AB, B or C have very nonlinear AM-AM responses. The method has proved its capability. At each class the IMD levels have been suppressed to noise floor. The importance of predistortion has been seen in class A and class C amplifier results. At 36 dBm peak power, 162 mA current is drawn form 28V supply in the Class C amplifier with predistortion. By this way, 47 dBc IM_3 and 48 dBc IM_5 are achieved. For the same output power and Class A amplifier, without predistortion, 301 mA current is drawn from supply and 22 dBc IM_3 and 42 dBc IM_5 are observed. Even 3 dB backed-off class A amplifier cannot reach predistorted class C performance.

In conclusion, in this thesis power amplifier nonlinearities and its linearization have been presented. Baseband predistortion techniques have been introduced and a Polynomial and LUT baseband predistortion approaches have been simulated. It has been shown that LUT predistortion has better performance. Furthermore, a hardware in the loop testbench has been built and satisfactory results for a memoryless predistortion algorithm have been reached.

Possible future research activities can be summarized as follows. The algorithm offered in Chap.5 can be realized in a closed loop system. For closed loop operation Analog Quadrature Modulation topology can be constructed with a proper selection of signal processor. Each of the system parameter that belongs to line-up components can be tested through that structure. Validity of the approach in the operational bandwidth can be tested. This will also lead to consider the effects of the memory. As the bandwidth of the input signal become wider, memory effect of the amplifier will become apparent. Then, predistortion with memory can be followed. Also, the causes for memory effects can be analyzed and appropriate predistortion techniques can be applied. The amplifier designed and the testbench constructed in this thesis are being used in a different master thesis work. Polynomial approach and predistortion with memory will be applied and performances will be compared.

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APPENDIX A

TWO TONE DATA CREATION WITH QUADRATURE MODULATION

If the baseband signal is determined to be sinusoidal then in quadrature modulation schemes this can be realized by the following equations;

$$I_{in}(t) = cos\,(w_{BB}t)\,, Q_{in}(t) = 0, \tag{A.1}$$

where w_{BB} is the frequency of baseband single tone.

If these I and Q signals are modulated via qaudrature modulator, then, at the output;

$$V_{mod}(t) = cos\,(w_{BB}t)\,cos\,(w_c t)\,, \tag{A.2}$$

will be obtained. w_c is the carrier frequency. This equation can be rearranged as follows;

$$V_{mod}(t) = \frac{1}{2}(cos\,((w_c - w_{BB})t) + cos\,((w_c + w_{BB})t)), \tag{A.3}$$

which is a two tone signal with $2w_{BB}$ spacing.

APPENDIX B

USAGE OF TEST EQUIPMENTS

B.1 ESG4433B Arbitrary Waveform Generator

This Arbitrary Waveform Generator(AWG) is used in creating Quadrature Modulated RF signal from I and Q signals created according to the modulation of interest. In MATLAB, one can create certain I and Q data. Then, by using these I and Q data, an array called $Complex = I + jQ$ is created. In order to scale these values to the correct ranges of the DACs of AWG, a MATLAB code ($arbsave(Complex, 0, 0, 1)$) should be run. It is provided by the manufacturer of AWG. Then, $i.bin$ and $q.bin$ are created. By using AWG communication program ($ESG\_ARB.exe$), one can download this files to the instrument via RS232 or HPIB interface.

What the instrument does is to take I and Q files, convert this digital data to analog with DACs at the sampling frequency entered (up to 40 MHz) and filter these analog signals by the internal reconstruction filters (250kHz,2.5MHz,8MHz or through) and make quadrature modulation ($I * coswt + Q * sinwt$) at the RF frequency entered. It can amplify the signal up to 15 dBm peak envelope power. However, 50dBc IMD can be achieved up to 0 dBm output power.

B.2 HP89441A Vector Signal Analyzer

This Vector Signal Analyzer(VSA) is used to obtain I and Q data from the output of the amplifier. In order to do that, attenuated RF output (below -10dBm level is suitable) is applied to the VSA. Its RF unit downconverts to IF and sends the signal to IF section. In IF section, the demodulation process is carried out.

If one selects Vector Mode of the instrument from Instrument Mode section and in Measurement Data section 'Main Time Ch1' is chosen, then by selecting I or Q in Data Format section, one can observe I and Q demodulated time domain waveforms. In order to take the signals at a certain sampling frequency one should consider the following formula;

$$SamplingFrequency = 1.28 * Span \tag{B.1}$$

Also, to specify the time length, main length should be arranged, if limit is reached, number of frequency points should be increased from ResBW/Window section. This data can be downloaded to floopy disk by using Save/Recall—Save Trace—Into file commands. Saved data is in *.dat* format. To change it *.mat* file, which can be read by MATLAB, *sdftoml.exe* program should be used. Then, by using load command of MATLAB, saved I and Q waveforms can be examined.