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Adaptive Predistortion Linearization of Nonlinear Power Amplifiers

**by
Bakri Aboukarr, B.Eng.**

**A thesis submitted to the
Faculty of Graduate Studies and Research
in partial fulfillment of the requirements
for the degree of
Master of Engineering**

**Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
Carleton University**

September, 1990

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**The undersigned recommend to the Faculty of
Graduate Studies and Research acceptance of the thesis**

**“Adaptive Predistortion Linearization
of Nonlinear Power Amplifiers”**

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in partial fulfillment of the requirements
for the degree of Master of Engineering**

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Abstract

The radio spectrum is a scarce resource, which necessitates the use of highly efficient modulation schemes in digital transmission systems. Although, theoretically such modulation methods achieve high spectrum efficiency, they tend to suffer from the nonlinearities of high power amplifiers and as a result they spread the spectrum of the filtered signal, causing some adjacent channel interference. This thesis addresses the problem of adaptive linearization of high power amplifiers. The technique considered here is independent of the modulation scheme and is suitable for narrow band signals. In practice, it is applied at base band. In this technique, the low pass equivalent envelope of the modulated signal is distorted prior to amplification (i.e., predistorted), such that the amplified signal is a linear version of the modulated signal. The predistortion is achieved by mapping the modulated signal to a distorted signal using table lookup functions. The mapping is done using the polar representation of the low pass complex envelope of the modulated signal since the amplifier's nonlinearities are a function of the amplitude. The adaptation of the table lookup functions are based on the Least Mean Squared (LMS) algorithm. The adaptation technique is analyzed and tested on a Quadrature Phase Shift Keying (QPSK) modulation system using computer simulation. The results obtained indicate that the technique provides excellent performance. In addition, the technique enjoys the advantages of implementation simplicity, low memory and processing requirements and fast convergence.

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List of Acronyms

QPSK	- Quadrature phase shift keying
HPA	- High power amplifier
IF	- Intermediate frequency
RF	- Radio frequency
LMS	- Least mean square
MSE	- Mean squared error
LINC	- Linear amplification with nonlinear components
AM-AM	- Amplitude modulation to amplitude modulation
AM-PM	- Amplitude modulation to phase modulation
DC	- Direct current
dB	- Decibels
D/A	- Digital to analog
A/D	- Analog to digital

List of Symbols

V_{in}	- Input signal
A	- Amplitude of input signal
V_{out}	- Output signal
w	- Output of amplifier (complex), sampling rate
v	- Input of amplifier (complex)
P_{max}	- Maximum output power of amplifier
ϕ	- Phase shift introduced by nonlinear system
ϕ_{outBP}	- Resultant phase shift from bandpass output
J_i	- Bessel function of order i
a_n	- Coefficients of a Taylor series
A_i	- Magnitude of signal i
ϕ_i	- Phase of signal i
$Q()$	- The AM-AM characteristic function of the predistorter
$G()$	- The AM-AM characteristic function of the amplifier
$R()$	- The overall AM-AM characteristic function of the predistorter-amplifier system
g	- The overall constant gain of the predistorter-amplifier system
$G^{-1}()$	- The inverse function of $G()$
$P()$	- The AM-PM characteristic function of the predistorter
$F()$	- The AM-PM characteristic function of the amplifier
$S()$	- The overall AM-PM characteristic function of the predistorter-amplifier system
C	- The overall constant phase change introduced by the predistorter-amplifier system

e_m	- Quantization error in the magnitude of the amplified signal
$\sigma_{e_m}^2()$	- Variance of the error e_m
e_p	- Quantization error in the phase of the amplified signal
$\sigma_{e_p}^2()$	- Variance of error e_p
D	- Distance between two successive quantization levels
γ_i	- Slope around a region on one of the characteristic curves
α	- Adaptation coefficient for the magnitude lookup table
β	- Adaptation coefficient for the phase lookup table
$\epsilon()$	- Adaptation error in the magnitude lookup table
T	- Convergence time
N	- Number of iteration required for an entry to converge
L	- Length of table
θ_e	- Phase error due to changing carrier frequency

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Chapter 1

Introduction

1.1 Problem Description

Since the radio spectrum is a scarce resource, it must be carefully allocated and managed. To prevent significant interference between users of the radio spectrum, stringent restrictions on spectral emission are specified and enforced by regulatory agencies [1]. These restrictions motivate the users to seek highly efficient transmission schemes. Spectrally efficient modulation schemes often involve amplitude as well as phase modulation. Even constant envelope signals, such as quadrature phase shift keying (QPSK), when filtered, produce amplitude fluctuations. Although, theoretically such modulation methods achieve high spectrum efficiency, they easily suffer from the nonlinearities of high power amplifiers (HPA) and tend to spread the spectrum of the filtered signal. Therefore, linear amplifiers are, in general, required to maintain the spectral efficiency advantage of these schemes. But linear amplifiers have low dc-to-RF power conversion efficiency, while nonlinear amplifiers have high dc-to-RF power conversion efficiency. Linearization of HPA is one way to preserve high dc-to-RF power conversion efficiency and at the same time achieve a high degree of spectrum efficiency of the modulated signal.

Several HPA linearization techniques exist and may be applied at different stages of processing the transmitted signal. Depending on the technique used, linearization of the

HPA can take place at baseband [2, 3], at IF stage [4], or at RF [5]. In this thesis, we are only concerned with baseband methods.

Several linearization techniques have been developed previously and reported in literature. Examples of these techniques include predistortion, postdistortion, negative feedback, feedforward, linear amplification with nonlinear components (LINC) and power back-off [6]. These techniques have their advantages and their drawbacks depending on the frequency of operation, stability, and feasibility. A brief summary of these methods is given in chapter 2.

Among baseband linearization techniques, the predistortion method has attracted a wide interest recently and appears to be very promising for practical realization [2, 3].

This thesis introduces an adaptive technique for linearization of HPA based on the polar representation of the modulated signal and the least mean squared (LMS) algorithm. The adaptation technique is analyzed and tested on QPSK modulation system using computer simulation. The results obtained indicate that the technique provides excellent performance. In addition, the technique enjoys the advantages of implementation simplicity, low memory and processing requirement and fast convergence.

1.2 Classification of Nonlinear Systems

In this section, a brief description of different classes of nonlinear systems is presented. A detailed description of these classes can be found in [6].

Nonlinear systems are classified as:

(1) Memoryless systems:

The output signal of these systems depends only upon the value of the instant input and not upon past values of the input signal. Also the output signal does not contain any phase nonlinearities.

For example if the input signal is:

$$V_{in} = A \cos(\omega t + \theta), \quad (1.1)$$

then the output signal can be found using the power series representation as:

$$V_{out} = \sum_{i=0}^{\infty} a_i V_{in}^i, \quad (1.2)$$

and the bandpass output signal can be written as:

$$V_{out_{BP}} = G(A) \cos(\omega t + \theta), \quad (1.3)$$

where $G(A)$ is the AM-AM nonlinear distortion function associated with this memoryless system.

In general the output signal does not only contain AM-AM distortion, but it also contains a number of harmonics and intermodulation products not associated with input signal's frequencies.

(2) Systems with memory:

The output signal of this class of nonlinearity depends upon the value of the instant input and upon past values of the input signal. Such systems introduce a frequency dependent AM-AM and AM-PM nonlinear distortion.

For example if the input signal is:

$$V_{in} = A \cos(\omega t + \theta), \quad (1.4)$$

then the bandpass output signal can be written as:

$$V_{out_{BP}} = G(A, \omega) \cos(\omega t + \theta + F(A, \omega)), \quad (1.5)$$

where $G(A, \omega)$ and $F(A, \omega)$ are the frequency dependent AM-AM and AM-PM nonlinear distortion functions associated with system respectively.

The Volterra series is the most popular representation of nonlinear systems with memory. In general, most high power amplifiers can be classified as nonlinear systems with memory due to the presence of active components in any electronic device (i.e., capacitors, inductors etc.).

(3) Systems effectively without memory:

This class of nonlinearity is not fully memoryless for the presence of AM-PM distortion nonlinearity, and is a special case of the systems with memory. For a narrow band signal, it is assumed that the AM-AM and AM-PM nonlinear distortion functions are fixed over that range of frequency, i.e., flat frequency response.

For example if the input signal is:

$$V_{in} = A \cos(\omega t + \theta), \quad (1.6)$$

then the bandpass output signal can be written as:

$$V_{out_{BP}} = G(A) \cos(\omega t + \theta + F(A)), \quad (1.7)$$

where $G(A)$ and $F(A)$ are the frequency independent AM-AM and AM-PM nonlinear distortion functions respectively.

Using the quadrature envelope representation of the output signal, such system can be represented as a sum of two fully memoryless systems. Now resolving the output signal into quadrature envelope components $P(A)$ and $Q(A)$ where,

$$P(A) = G(A) \cos(F(A)) \quad (1.8)$$

and

$$Q(A) = G(A) \sin(F(A)), \quad (1.9)$$

the output signal can be written as:

$$V_{outBP} = P(A) \cos(\omega t + \theta) - Q(A) \sin(\omega t + \theta). \quad (1.10)$$

This class of nonlinearity is mostly used for representation of high power amplifiers when operated over a narrow band of frequencies.

1.3 Characteristics of High Power Amplifiers

1.3.1 A Mathematical Model

High power amplifiers are classified as nonlinear systems effectively without memory for a narrow band signals. For wide band signals they are classified as nonlinear systems with memory. If an input signal with constant power and variable frequency was applied to the HPA, the observed output power would be constant over small changes of frequency (narrow band), but it would change over large changes of frequency (wide band).

In this thesis we consider only narrow band signals. Consequently, we model the HPA as a nonlinear system “effectively” without memory. For computational and simulation purposes, the following complex gain HPA characterization is used:

$$w = \underbrace{\frac{2v\sqrt{P_{max}}}{1 + |v|^2}}_{AM-AM} \exp \left\{ \underbrace{j\phi_0 \frac{2|v|^2}{1 + |v|^2}}_{AM-PM} \right\} \quad (1.11)$$

The characterization of equation 1.11 is an accurate representation for TWT amplifiers [5]. For convenience, the maximum possible HPA power, P_{max} , is assumed to occur at an input power of 1W. This scaling of v simplifies matters without affecting the generality of the results. Thus, $|w|^2 = P_{max}$ when $|v|^2 = 1$, and the phase shift for that maximum power is ϕ_0 . Figures 1.1 and 1.2 illustrate the AM-AM and AM-PM curves of equation 1.11. The AM-AM curve shows that the effect of compression is clearly evident at high input levels, as is the linear gain at low input levels, while the AM-PM curve shows that this phase change is very small at low input levels and increases rapidly at high input levels.

The effects of AM-AM and AM-PM conversion will be studied in the following sections.

1.3.2 AM-to-PM Conversion

This section is a summary of the work done by Taraschuk [6] which explains the effects of phase nonlinearities introduced by HPAs.

To simplify the analysis of AM-PM conversion, let us assume that the system contains only phase nonlinearities. Thus if the input signal is:

$$V_{in} = A \cos(\omega_0 t), \quad (1.12)$$

Figure 1.1 Typical AM-AM characteristics curve

Figure 1.2 Typical AM-PM characteristics curve

then the output signal can be written as:

$$V_{out} = A \cos(\omega_0 t + \phi(V_{in})). \quad (1.13)$$

The nonlinear phase distortion function $\phi(V_{in})$ can be represented by the Taylor (power) series as:

$$\phi(V_{in}) = \sum_{n=0}^{\infty} a_n V_{in}^n, \quad (1.14)$$

but due to the infinite number of terms, the mathematical analysis will become complex. However, reducing the power series to the quadratic term only, some insight can be gained.

Therefore for the phase nonlinearity given by:

$$\phi(V_{in}) \approx a_2 V_{in}^2, \quad (1.15)$$

the output will become:

$$V_{out} \approx A \cos\left(\omega_0 t + a_2 (A \cos(\omega_0 t))^2\right). \quad (1.16)$$

Removing all frequencies not associated with ω_0 , V_{out} can be written as

$$V_{out} = A \cos(\omega_0 t + \phi_{out_{BP}}). \quad (1.17)$$

It was found [6] that:

$$\phi_{out_{BP}} = \arctan \left\{ \frac{\sin\left(\frac{a_2 A^2}{2}\right) J_0\left(\frac{a_2 A^2}{2}\right) - \cos\left(\frac{a_2 A^2}{2}\right) J_1\left(\frac{a_2 A^2}{2}\right)}{\cos\left(\frac{a_2 A^2}{2}\right) J_0\left(\frac{a_2 A^2}{2}\right) - \sin\left(\frac{a_2 A^2}{2}\right) J_1\left(\frac{a_2 A^2}{2}\right)} \right\}. \quad (1.18)$$

Where $J_0()$ and $J_1()$ are Bessel functions, and $\phi_{out_{BP}}$ referred to as the band pass frequency independent AM-PM nonlinear distortion function.

In general the AM-PM nonlinear distortion function is not only envelope-dependent, but is also dependent on the instantaneous input signal. However, for narrow band, as shown in the previous example, the nonlinear distortion function is envelope dependent.

1.3.3 Restoration of Filtered Sidelobes

In Frequency Division Multiple Access (FDMA) systems, each channel is allocated a specific bandwidth. Within that bandwidth, the frequency spectral density must fall below specific limits. As an example of these limitations, it is stated in some applications that the signal level should be at least 50dB below the main lobe of the adjacent channels in order to prevent interference.

To illustrate the restoration of filtered sidelobes, the simple case of AM-AM conversion is considered. Assume an input signal V_{in} that have a filtered spectrum $V(f)$ with bandwidth w , and let the output of the amplifier be represented by a Taylor series as follows:

$$V_{out}(t) = a_1 V_{in}(t) + a_2 V_{in}^2(t) + a_3 V_{in}^3(t) + \dots, \quad (1.19)$$

then the spectrum of V_{out} can be written as

$$V_{out}(f) = a_1 V_{in}(f) + a_2 V_{in}(f) * V_{in}(f) + a_3 V_{in}(f) * V_{in}(f) * V_{in}(f) + \dots, \quad (1.20)$$

where (*) is the convolution operator [7]. From convolution properties, $V_{in}(f) * V_{in}(f)$ will produce a spectrum with bandwidth $2w$ and $V_{in}(f) * V_{in}(f) * V_{in}(f)$ will produce a spectrum with bandwidth $3w$, etc.

Therefore, $V_{out}(f)$ appears as an unfiltered spectrum with a very large bandwidth made of a main lobe and a number of sidelobes. This effect looks like the regeneration of filtered sidelobes.

As an illustration, figures 1.3, 1.4 and 1.5 show the spectrum of an unfiltered QPSK signal, the filtered version and the amplified spectrum of the filtered signal, respectively. The unfiltered QPSK modulated signal does not meet the restrictions set by regulatory agencies because the sidelobes adjacent to the main lobe are approximately 9.5dB below the main lobe, therefore filtering is required. The filtering was done using an 8th order butterworth filter with cutoff frequency of 1Hz. But the spectrum of the amplified signal shows the regeneration of the filtered sidelobes, which made the adjacent sidelobes to the main lobe to increase by approximately 20dB.

1.4 Thesis Objectives

Nonlinearities present in typical high power amplifiers distort filtered modulated signals. The spectrum of the distorted signal has high sidelobes. This means that the transmitted signal will interfere with the adjacent channels.

This thesis considers the linearization of HPAs to prevent the regeneration of sidelobes in the transmitted signal. The linearization is performed by the intentional predistortion of the filtered signal prior to amplification.

First, a non-adaptive predistorter was designed and applied to a baseband QPSK filtered signal. Then, an adaptive predistorter was designed and also applied to baseband QPSK filtered signal. The implementation of both distorters was considered in terms of complexity, time-delay and feasibility.

1.5 Thesis Organization

The remaining chapters of this thesis are organized as follows.

Figure 1.3 Power spectrum of modulated QPSK signal

Figure 1.4 Power spectrum of filtered QPSK signal

Figure 1.5 Power spectrum of amplified filtered QPSK signal

In chapter 2, a brief survey of different linearization techniques is presented. In chapter 3, the proposed linearization method (predistortion) is discussed in details. First, a mathematical description of the predistortion process is presented. This is followed by adaptation of non-adaptive as well as adaptive predistortion techniques. Performance criteria and implementation considerations are also discussed for both techniques in chapter 3.

In chapter 4, the predistorter implementation and simulation results are presented and discussed for both non-adaptive and adaptive predistorter designs.

Finally, chapter 5, summarizes major the conclusions of this work and provides recommendations for future research.

Chapter 2

Linearization Techniques

2.1 Introduction

To eliminate the nonlinear distortion introduced by high power amplifiers, the linearization of high power amplifiers is required. Different linearization techniques have been studied and analyzed in the past. Each of these techniques have their advantages and their drawbacks. A brief description of some of these techniques is given in the following sections. An excellent survey of linearization techniques has been presented by Green [8] and Taraschuk [6].

2.2 Predistortion

Predistortion is the intentional distortion of the transmitted signal prior to amplification as shown in figure 2.1, i.e., wrapping the transmitted signal's constellation points prior to amplification, so that the overall predistorter-amplifier system gain is linear. The predistortion can be implemented at base band frequency, intermediate frequency or at radio frequency depending on the application. One of the major drawbacks of predistortion is the difficulty of cancelling the nonlinear distortion over a wide range of frequencies (wide band). However, this technique is capable of cancelling the nonlinear distortion over a narrow band. Another consideration is that aging and temperature changes will make it difficult for the predistorter to cancel the entire nonlinear distortion. Therefore, to overcome this problem, an adaptive predistorter is recommended.

2.3 Postdistortion

Postdistortion systems are similar to predistortion systems in concept, except the distortion is implemented after amplification as shown in figure 2.1.

Postdistortion is less attractive than predistortion because any associated signal losses are more severe at higher power.

2.4 Negative Feedback

Negative feedback is used to reduce the nonlinear distortion by subtracting the feedback amplifier's nonlinear distortion from the transmitted signal. This technique is not suitable when amplifying a high frequency signal because the feedback path delay would be very large compared to the modulated signal's periods. However, it provides low distortion levels when the amplifier is operated at low frequency signals. Finally, negative feedback reduces the amplifier's gain which is costly for high power amplifiers.

2.5 Feedforward

The feedforward technique does not have the problem of feedback path delay associated with the negative feedback technique, but is similar to it in concept. This technique eliminates the nonlinear power amplifier distortion by subtracting the amplified signal from a scaled version of the transmitted signal, then amplifying this difference (distortion) and subtracting it from the amplified signal as shown in figure 2.2. For perfect nonlinear distortion cancellation, the two amplifiers used in this method must have the exact nonlinear distortion functions. However, this is not possible because mobile transmitters are frequently turned on and off, i.e., the nonlinear distortion functions of

Predistortion System

Postdistortion System

Figure 2.1 Predistortion / Postdistortion system block diagrams

the two amplifiers will change frequently, thereby preventing proper cancellation of the nonlinearities [2].

2.6 Linear Amplification with Nonlinear Components (LINC)

The linear amplification with nonlinear components (LINC) technique requires two amplifiers similar to the feedforward technique. This technique works perfectly if the two amplifiers are memoryless and have no phase distortion. However, high power amplifiers are not considered fully memoryless because they introduce AM-PM conversion, and it is hardly realized that the two amplifiers will have the exact nonlinear distortion functions so that they would cancel each other's nonlinearities.

2.7 Power Back-off

The power back-off technique is not really a linearization technique because the amplifier is being operated around its linear region. The implementation of such system is the easiest, but it suffers from a high cost for higher power amplifiers than required and from high power consumption.

Figure 2.2 Feedforward distortion reduction block diagram

Chapter 3

Proposed Linearization Method for HPA

3.1 Mathematical Description

The predistortion linearization technique involves transformation of the signal prior to amplification, such that the output signal of the amplifier is a linear version of the predistorter input signal. For example, if an amplifier exhibits a convex type nonlinearity, linearization requires predistortion with a concave nonlinearity.

To examine this method analytically, the following notations are introduced. Let $x(t)$, $y(t)$ and $z(t)$ denote the signals at the predistorter input, amplifier input and amplifier output respectively. The three signals are shown in figure 3.1.

Each of the three signals can be expressed in a low pass equivalent form as follows:

$$x(t) = A_x e^{j\phi_x} \quad (3.1)$$

$$y(t) = A_y e^{j\phi_y} \quad (3.2)$$

$$z(t) = A_z e^{j\phi_z} \quad (3.3)$$

where A_i represents the amplitude of signal i and ϕ_i represents the phase of the signal.

For the magnitude analysis, we have

$$A_z = G(A_y)$$

Figure 3.1 A cascaded predistorter-amplifier to be analyzed

$$\begin{aligned}
 &= G(Q(A_x)) \\
 &= R(A_x),
 \end{aligned}
 \tag{3.4}$$

where $G()$ is the amplifier's magnitude characteristic function, $Q()$ is the predistorter's magnitude characteristic function, and $R(A_x)$ is the overall magnitude characteristic function of the cascaded predistorter-amplifier system [9].

For amplitude linearization, the following equation must be satisfied

$$R(A_x) = g.A_x, \tag{3.5}$$

where g is the required overall constant gain. From equations 3.4 and 3.5, it follows that:

$$g.A_x = G(Q(A_x)). \tag{3.6}$$

Defining G^{-1} to be the inverse function of G and assuming that G^{-1} exists, equation 3.6 can be rewritten as:

$$Q(A_x) = G^{-1}(g.A_x). \tag{3.7}$$

As an example, consider G to be a square function, i.e.,

$$G(A_y) = A_y^2 \tag{3.8}$$

if the required linear gain is 5, then

$$G^{-1}(A_x) = \sqrt{A_x} \tag{3.9}$$

and

$$\begin{aligned}
 Q(A_x) &= G^{-1}(g.A_x) \\
 &= \sqrt{5.A_x}.
 \end{aligned}
 \tag{3.10}$$

For the phase analysis, the phase difference $\phi_z - \phi_x$ can be expressed as follows:

$$\begin{aligned}\phi_z - \phi_x &= F(A_y) + P(A_x) \\ &= S(A_x),\end{aligned}\tag{3.11}$$

where $F()$ is the amplifier's phase characteristic function, $P()$ is the predistorter's phase characteristic function, and $S(A_x)$ is the overall phase characteristic function of the cascaded predistorter-amplifier system [9].

For phase linearization, the following condition must be satisfied

$$S(A_x) = C,\tag{3.12}$$

where C is a constant phase change. Equation 3.11 can be written as

$$\begin{aligned}P(A_x) &= C - F(A_y) \\ &= C - F(G^{-1}(g.A_x)).\end{aligned}\tag{3.13}$$

If $C = 0$, i.e., the overall phase change is zero degrees, then equation 3.11 becomes

$$P(A_x) = -F(G^{-1}(g.A_x)).\tag{3.14}$$

$$P(A_x) = -F(Q(A_x)),\tag{3.15}$$

To summarize, the above analysis shows that it is possible to obtain both the magnitude and phase characteristics of the predistorter from the magnitude and phase characteristics of the amplifier. If the amplifier's gain and phase characteristic functions are

$G(A_x)$ and $F(A_x)$ respectively, then the predistorter magnitude and phase characteristics are given by:

$$Q(A_x) = G^{-1}(g.A_x), \quad (3.16)$$

$$P(A_x) = C - F(Q(A_x)) \quad (3.17)$$

where C is the overall constant phase change and g is the overall constant gain of the predistorter-amplifier system.

Figure 3.2 shows the output power vs. input power for a typical TWT high power amplifier with its saturation point around an input power of 5 dB, while figure 3.3 shows its output phase in degrees vs. the input power. Notice that the phase shift introduced by the amplifier is zero for small values of input power, while there is a rapid change in the phase shift near the saturation point. Figures 3.4 and 3.5 show the magnitude and phase characteristics of the predistorter corresponding to that amplifier that makes the overall gain equal to 5 and the overall phase equal to zero degrees. Notice that, the concavity of the predistorter is opposite to that of the amplifier in both the magnitude and phase characteristics.

3.2 Non-adaptive Method

In the previous section, an analytical continuous (analog) predistorter's magnitude and phase characteristics were derived for a known amplifier's characteristics. In practice, the amplifier manufacturer does not provide the exact characteristics of the amplifier. In order to obtain the exact characteristics of the amplifier, a training sequence will be used.

Figure 3.2 Amplifier's AM-AM characteristics curve

Figure 3.3 Amplifier's AM-PM characteristics curve

Figure 3.4 Predistorter's AM-AM characteristics curve

Figure 3.5 Predistorter's AM-PM characteristics curve

Such a sequence compares the input signal values and the corresponding output signal values of the amplifier to obtain its exact characteristics.

Since the objective of this work is to obtain a digital implementation of the predistortion scheme, the approach followed will rely on linearization of a discrete version of the waveform. Assuming the input to the predistorter is quantized to n bits, then the predistorter can be visualized as two table lookup functions (one for the magnitude and one for the phase), each of length 2^n words.

The characteristic equations for the discrete predistorter are identical to the analog one derived in the previous section. The table lookup technique works as follows: A_x , the transmitted signal, is quantized to n bits (i.e. A_x is one of the 2^n possible inputs) and used as an address to the two tables that correspond to $Q(A_x)$ and $P(A_x)$.

3.3 Quantization Noise

Due to quantization, equations 3.5 and 3.12 become

$$R(A_x) = g \cdot A_x + e_m \quad (3.18)$$

$$S(A_x) = C + e_p, \quad (3.19)$$

where e_m and e_p are the quantization noise of the magnitude and phase respectively.

To derive the variance of e_m , let A_q be the quantization level for the range of inputs of A_x defined by

$$A_q - D/2 < A_x < A_q + D/2, \quad (3.20)$$

where D is the distance between two successive quantization levels. Then the output of the predistorter, A_y , for that range of inputs will have the following limits:

$$Q(A_q) - \gamma_1 D/2 < A_y < Q(A_q) + \gamma_1 D/2 \quad (3.21)$$

where γ_1 is the slope of the predistorter's magnitude curve around A_q , and the output of the amplifier, A_z , will be contained between the following limits:

$$G(Q(A_q)) - \gamma_1 \gamma_2 D/2 < A_z < G(Q(A_q)) + \gamma_1 \gamma_2 D/2, \quad (3.22)$$

where γ_2 is the slope of the amplifier's magnitude curve around $Q(A_q)$. However, since the overall gain of the cascaded predistorter-amplifier system is g , then $\gamma_1 \gamma_2 = g$ and the output of the amplifier will be defined between the limits:

$$G(Q(A_q)) - gD/2 < A_z < G(Q(A_q)) + gD/2. \quad (3.23)$$

Assuming that the input signal, A_x , is uniformly distributed over the interval

$$A_q - D/2 < A_x < A_q + D/2, \quad (3.24)$$

then the variance of e_m for that quantization level is

$$\sigma_{e_m}^2(A_x) = \frac{(gD)^2}{12} \quad (3.25)$$

[10] . Assuming further that the input signal is uniformly distributed over all entries of the table, then the average variance can be expressed as:

$$\begin{aligned} \langle \sigma_{e_m}^2 \rangle &= \frac{(gD)^2}{12} \\ &= \frac{g^2 A_{max}^2}{12L^2}, \end{aligned} \quad (3.26)$$

where A_{max} is the maximum input signal and L is the table length ($L = 2^n$).

Now consider the variance of e_p for the input range defined by the interval:

$$A_q - D/2 < A_x < A_q + D/2, \quad (3.27)$$

the output magnitude, A_y , of the predistorter will be contained in the interval:

$$Q(A_q) - \gamma_1 D/2 < A_y < Q(A_q) + \gamma_1 D/2 \quad (3.28)$$

and the range of the output phase, $P(A_x)$, will be given by:

$$P(A_q) - \gamma_3 D/2 < P(A_x) < P(A_q) + \gamma_3 D/2, \quad (3.29)$$

where γ_3 is the slope of the predistorter's phase curve around A_q , and the output phase of the amplifier, $F(Q(A_x))$, will be contained in the interval:

$$F(Q(A_q)) - \gamma_1 \gamma_4 D/2 < F(Q(A_x)) < F(Q(A_q)) + \gamma_1 \gamma_4 D/2, \quad (3.30)$$

where γ_4 is the slope of the amplifier's phase curve around $Q(A_q)$.

Note that

$$F(Q(A_q)) = -P(A_q) \quad (3.31)$$

and

$$\gamma_1 \gamma_4 = -\gamma_3 \quad (3.32)$$

for $C = 0$.

Therefore, the variance of e_p for the quantization level A_q is

$$\sigma_{e_p}^2(A_x) = \frac{(\gamma_3 D)^2}{12}. \quad (3.33)$$

This shows that the variance is proportional to γ_3^2 , the slope of the predistorter's phase curve. Consequently, the variance of e_p is maximum for large inputs, where the value of γ_3 is large. Assuming that the input signal is uniformly distributed over all entries of the table, the average variance will be given by:

$$\begin{aligned}
 \langle \sigma_{e_p}^2 \rangle &= \frac{1}{L} \sum_{i=1}^L \frac{(\gamma_3(A_i)D)^2}{12} \\
 &= \frac{\langle \gamma_3^2 \rangle D^2}{12} \\
 &= \frac{\langle \gamma_3^2 \rangle A_{max}^2}{12L^2}.
 \end{aligned} \tag{3.34}$$

From equations 3.26 and 3.34, both variances are inversely proportional to L^2 . Therefore, if the table length is doubled, which is equivalent to increasing n by one, then both variances will be reduced by a factor of 4 or 6dB. In the simulation runs to be presented later, $n = 8$ (or $L = 256$) was used, which produced a very small quantization noise.

3.4 Adaptive Method

The characteristics of most amplifiers are expected to vary with temperature, aging and frequency switching. Normally high power amplifiers can operate over a large range of frequencies (i.e. wide band), and by switching from one narrow band to another, the AM-AM and AM-PM characteristics of the amplifier will change. Therefore, it is adequate to have some form of adaptation to track such changes. The non-adaptive predistortion system discussed earlier uses a fixed lookup table functions. In this section,

an algorithm that will adapt the table entries to track the changes in the amplifier nonlinearities is derived. The algorithm does not require a prior knowledge of the amplifier's characteristics.

In the following subsections, the block diagram of cascaded amplifier-predistorter system shown in figure 3.6 will be used as a reference.

3.4.1 Adaptation Algorithm

Initially all entries in the predistorter table are set to unity gain and zero phase shift. Suppose that we apply a fixed complex input signal, $a_i = ae^{j0^0}$, to the predistorter input and observe the magnitude of the predistorter output $Q(i)$ in successive adaptation cycles. The following sequence of values will result:

$$Q(0) = a$$

$$Q(1) = a + \alpha(a - G(0)/g)$$

$$Q(2) = a + \alpha(a - G(0)/g) + \alpha(a - G(1)/g)$$

.

.

.

$$Q(i) = a + i\alpha a - \frac{1}{g} \sum_{j=0}^{i-1} G(j) \quad (3.35)$$

where g is the required gain and α is an adaptation constant. Similarly the successive values of the amplifier's magnitude output $G(i)$ are given by:

$$G(0) = f(a)$$

Figure 3.6 Block diagram of an adaptive predistorter-amplifier system

$$\begin{aligned}
G(1) &= f(a + \alpha(a - G(0)/g)) \\
G(2) &= f(a + \alpha(a - G(0)/g) + \alpha(a - G(1)/g)) \\
&\cdot \\
&\cdot \\
&\cdot \\
G(i) &= f\left(a + i\alpha a - \frac{\alpha}{g} \sum_{j=0}^{i-1} G(j)\right)
\end{aligned} \tag{3.36}$$

Assuming the amplifier is linear in the immediate vicinity of a with slope γ , the following approximation can be applied:

$$f(a + x) \approx G + \gamma x \tag{3.37}$$

given that

$$f(a) = G \tag{3.38}$$

and using the notation:

$$G(0) = G, \tag{3.39}$$

the following equations are obtained:

$$\begin{aligned}
G(0) &= G \\
G(1) &= G + \alpha\gamma(a - G/g) \\
G(2) &= (G - ag)(1 - \alpha\gamma/g)^2 + ag
\end{aligned}$$

$$G(i) = (G - ag)(1 - \alpha\gamma/g)^i + ag. \quad (3.40)$$

Equation 3.40, shows that if $G = ag$, i.e., the predistorter is already adapted to the amplifier, the output of the amplifier will always be ag (because the first term will always be zero). Also, if the term $(1 - \alpha\gamma/g)^i$ is a monotonically decreasing function of i , then $G(i)$ will approach ag with an increasing number of cycles.

Define $P(i)$ to be the predistorter phase output. To simplify the notation, we let $P = P(i)$. Assuming that the predistorter's magnitude table has already been adapted to the amplifier's magnitude characteristics, resulting in $G(i) \approx ag$ and $F(i) = F$, it is possible to obtain the following relationships:

$$\begin{aligned} P(i) &= P \\ P(i+1) &= P - \beta(P + F) \\ P(i+2) &= P(1 - \beta)^2 - \beta F(1 + (1 - \beta)) \\ &\vdots \\ P(i+n) &= P(1 - \beta)^n - \beta F \left(\sum_{j=0}^{n-1} (1 - \beta)^j \right). \end{aligned} \quad (3.41)$$

Note that

$$\sum_{j=0}^{\infty} (1 - \beta)^j = 1/\beta \quad \text{for } 0 \leq \beta \leq 2 \quad (3.42)$$

Therefore, equation 3.41 will become

$$P(\infty) = -F, \quad (3.43)$$

which is the required phase shift to cancel the phase shift introduced by the amplifier.

To summarize; for a fixed complex input a , equation 3.40 gives the magnitude of the amplifier response, and equation 3.41 gives the phase shift introduced by the predistorter. The following sections will provide a detailed examination of both equations in terms of convergence conditions and convergence time.

3.4.2 Convergence Conditions

A closer examination of equation 3.40 reveals that for

$$G(\infty) = ag, \quad (3.44)$$

it is necessary to have

$$0 < |1 - \alpha\gamma/g| < 1. \quad (3.45)$$

This inequality can be rearranged to give:

$$0 < \alpha\gamma/g < 2. \quad (3.46)$$

We may then state that for $0 < \alpha\gamma/g < 1$, $G(i)$ will monotonically converge to ag without any oscillation, i.e., the error

$$\begin{aligned} \epsilon(i) &= G(i) - G(\infty) \\ &= (G - ag)(1 - \alpha\gamma/g)^i \end{aligned} \quad (3.47)$$

will monotonically approach zero because the term $(1 - \alpha\gamma/g)$ varies in value between zero and unity.

For $1 \leq \alpha\gamma/g < 2$, the error $\epsilon(i)$ will oscillate but decrease in magnitude to zero because the term $(1 - \alpha\gamma/g)$ varies in value between zero and -1 .

Therefore, the condition for convergence depends on three parameters : α , γ , and g . Where α is the adaptation constant, γ is the slope of amplifier's magnitude characteristics curve around the input a , and g is the required gain from the predistorter-amplifier system.

The parameter γ is the slope of the amplifier around the input a . But the slope around the input a is not necessarily constant, and at the beginning of adaptation, we could be operating around one region while at the end of adaptation, we might be operating around another region on the amplifier's magnitude characteristics curve. Therefore, we will use γ as the maximum slope on the curve. Although, with careful design, one can divide the amplifier's magnitude characteristics curve into regions of different slopes and vary α for each different region to keep $\alpha\gamma/g$ constant.

To summarize, the value of $\alpha\gamma/g$ should be kept between zero and two to guarantee convergence, and the only parameter that can be controlled to achieve this is α . If γ is taken as the maximum slope, then α would be constant, but if the amplifier's magnitude characteristics curve is divided into regions of different slopes, then we could have different α for each slope. The next section will show what values of α should be used to minimize the convergence time.

Considering equation 3.41 with respect to phase adaptation, it is readily seen that in

order to guarantee convergence value of the term $|1 - \beta|$ should be limited as follows:

$$0 < |1 - \beta| < 1, \quad (3.48)$$

or

$$0 < \beta < 2. \quad (3.49)$$

Note that the convergence of the phase depends on the convergence of the magnitude, and that the error starts to decrease after the magnitude's error has approached zero. Consider now the terms $(1 - \beta)^n$ and $\sum_{j=0}^{n-1} (1 - \beta)^j$ in equation 3.41. For the term $(1 - \beta)^n$ to approach zero with minimum number of cycles, β should be close to unity. Also to make the term $\sum_{j=0}^{n-1} (1 - \beta)^j$ approach a constant value with minimum value of n , β should also approach unity. Choosing $\beta = 1$, equation 3.41 becomes

$$P(i + n) = -F \quad \text{for } n \geq 0 \quad (3.50)$$

which gives the exact opposite phase shift required to cancel the amplifier's phase nonlinearity. It can be concluded that the phase converges exactly along with the convergence of the magnitude.

3.4.3 Convergence Time

In order to make such an algorithm practical, it is necessary that all entries in the table converge very fast to the desired values.

As mentioned in the previous section, by choosing $\beta = 1$, the phase and magnitude will converge together, and only the magnitude affects the convergence time.

Consider the magnitude's error $\epsilon(i)$ given by equation 3.47:

$$\epsilon(i) = (G - ag)(1 - \alpha\gamma/g)^i, \quad (3.51)$$

for the error to approach rapidly zero, it is necessary to have:

$$\alpha\gamma/g \approx 1. \quad (3.52)$$

The condition represented by equation 3.52 is almost impossible to achieve in practice, since the amplifier is nonlinear and hence its magnitude characteristics curve does not have a constant slope γ . For simplicity, γ would be taken as the maximum slope of the amplifier's characteristics curve, and this leads to the approximation given by:

$$\alpha \approx g/\gamma, \quad (3.53)$$

which would minimize the convergence time for points around the maximum slope. Note, that the maximum slope of the amplifier's characteristics curve is around the region of small inputs (the linear region of the amplifier), while the slope is minimum around larger inputs (near saturation). Also for maximum efficiency, the complex envelope of filtered QPSK signal fluctuates in magnitude around its average, which is near saturation, resulting in a larger probability of transmitting samples near the saturation point compared to samples of smaller inputs. Therefore, if γ is taken as the maximum slope, samples near the saturation point would require more iterations to converge than samples of small inputs. In terms of convergence time, samples near the saturation point may converge before samples of small inputs. For example, for $\epsilon(i) < 10^{-5}$ and $\alpha\gamma/g = 0.9$, the required number of iterations, N , will be 5.

To calculate the convergence time for all entries in the table, assume that each entry converges with an approximately the same number of cycles N , the table has L (2^n) elements and a sampling rate of w hertz. Therefore, the time required for the entire

table to converge is

$$T = \frac{NL}{w}. \quad (3.54)$$

For example if $N = 30$ iterations, $L = 256$ entries and $w = 9600 * 8$ samples/sec (assuming 8 samples per baud and a baud value of 9.6k), then 100 msec is required for all entries in the table to converge.

3.5 Convergence Criteria

In the previous section, the convergence conditions and the convergence time for the proposed adaptation algorithm were presented. Implementation of the algorithm, however, requires the existence of an indicator for the convergence process. That is, a criterion for detecting convergence must be established.

The criterion should consider the difference between the actual content of the entries in the table and the desired content, i.e., the error. The sign of the error is irrelevant, hence an absolute value of the error or the squared error will be a good indicator. For analysis simplicity, the squared error will be used. Since we have a relatively large number of entries in the table, a running Mean Squared Error (MSE) average would be more practical. This can be obtained by taking the MSE for every K samples, the larger the K the more accurate the MSE will be.

Another criterion is to observe the spectrum of the amplifiers output, i.e., observing the error in the frequency domain. As proven earlier, a nonlinear amplifier restores the filtered sidelobes, but a linear amplifier keeps the spectrum unchanged. For example if a signal $x(t)$ with spectrum $X(f)$ is applied to a linear amplifier with gain g , then the

output of the amplifier will be

$$\begin{aligned}y(t) &= g.x(t) \\ Y(f) &= g.X(f).\end{aligned}\tag{3.55}$$

Therefore, by inspecting the spectrum of the amplifier's output every n samples and comparing it to the spectrum of the predistorter input, it can be shown that the restored sidelobes are being removed, and at some point the two normalized spectra will become identical.

The latter criterion, namely observing the error in frequency domain, is more powerful than the first one which is based on the MSE value in the sense that it provides gradual demonstration of the linearization process. It also shows that the predistorter is not adding any frequency components to the signal. However both criteria are being considered in this thesis.

3.6 Implementation Considerations

When a transmitter that includes an adaptive predistortion linearizer is implemented, one must consider every component that could affect its performance. To study such effects, a block diagram of a transmitter with the proposed linearizer is shown in figure 3.7.

The predistorter requires two inputs, the signal to be transmitted and a feedback signal from the amplifier output. The transmitted input signal should be subjected to a time delay which is equivalent to the delay encountered in the feedback path. This delay can be introduced through hardware delay circuits as shown in figure 3.7 [2]. Alternatively, the

Figure 3.7 Block diagram of a typical transmitter with an adaptive predistorter

delay circuit can be removed and the predistorter will buffer a sequence of the transmitted input signal and a sequence of the feedback signal. Then the correlation function of the two sequences will provide a good measure of the delay. Similarly, one can observe the slope change of both sequences and derive a measure of the delay encountered in the feedback path.

In practice, a problem is encountered when the carrier frequency shifts from w_0 to w_1 . This will introduce a phase rotation θ_e on the feedback signal. This phenomenon also occurs when the delay time in the feedback path changes. The phase rotation will trigger the adaptation algorithm. Then, the output spectrum will degrade until the update of the tables converges again. In order to avoid such phase rotation, a circuit which adjusts the phase of the modulator oscillator is necessary. This function can be implemented by using a conventional Phase Locked Loop (PLL) circuit to detect the phase error between two signals on the main path and the feedback path as shown in figure 3.7 [2].

In terms of memory size requirements, this system requires two tables, each around one kbyte; which is easily found in a single RAM chip.

The digital to analog (D/A) and analog to digital (A/D) converters are available in a wide variety of speed, dynamic range and accuracy [2]. A 16 bit converter is suitable for the purpose of this work.

Finally, if the other circuit components, especially in the feedback path, introduce any nonlinear distortions, the predistorter will detect errors and update the entries in the tables. But this update will not improve the spectrum of transmitted signal, because these nonlinearities were not introduced in the main path (including the amplifier) . To avoid

this problem, linear components should be used.

In summary, the above discussion indicates that it is feasible to implement the linearizer with existing components.

Chapter 4

Predistorter Implementation and Results

4.1 Introduction

The previous chapter has provided the background material for the simulation of the proposed linearization approach. A transmitter that includes a non-adaptive predistorter at baseband requires an upconverter circuit to generate the RF modulated carrier, while a transmitter that includes an adaptive predistorter requires an upconverter circuit to carry the signal to RF for transmission and a down converter circuit to bring it down as a feedback signal. In a properly designed system, the two converters do not introduce any significant distortion to the system. Hence, in the simulation reported here the effects of any frequency conversion of the signal are ignored. The simulation was carried out on the SUN network using a package called Block Oriented Systems Simulator (BOSS). The simulation system consists of functional blocks. Some of these blocks are primitive, i.e., perform simple function like addition, and larger functional blocks can be built from primitive blocks. Also it is possible to create your own functional blocks by writing a Fortran code and linking it to the package. A Fortran code was written for the predistorter and amplifier functional blocks that were used in the simulation.

4.2 Non-adaptive Predistorter Design

The non-adaptive predistorter is simply a table lookup complex function, i.e., the entries of the table are complex. For simulation purposes, the complex table can be

viewed as two tables; one table for the real part and the other for the imaginary part. Alternatively, the complex table can also be viewed as two tables; one table for the magnitude and the other for the phase. In either approach, the magnitude of the input signal determines what entry in the table should be accessed, i.e., the magnitude of the input signal acts as an address to the tables.

In the simulation, the latter approach was used in the implementation of the complex table. Therefore, depending on the magnitude of the input signal, the predistorter will generate a complex signal with new predistorted magnitude and phase value. The magnitude and phase tables of the predistorter are calculated as described in section 3.2.

Consider the amplifier's characteristics that were shown in figures 3.2 and 3.3. The predistorter characteristics which correspond to this amplifier can be evaluated, as derived earlier, using the relationships:

$$Q(A_x) = G^{-1}(g \cdot A_x) \quad (4.1)$$

$$P(A_x) = C - F(Q(A_x)), \quad (4.2)$$

where $Q(A_x)$ is the predistorted magnitude and $P(A_x)$ is phase change imposed by the predistorter for an input A_x , with A_x is quantized to n bits (i.e., 2^n addresses are contained in each table) and $Q(A_x)$ and $P(A_x)$ are quantized to m bits. Figures 4.1 and 4.2 show the magnitude and phase characteristics of the predistorter, using $g = 5$, $C = 0$, $n = 8$ bits (256 addresses in each table).

To study the effects of the predistorter, a simulation is performed on a QPSK signal. First, without the predistorter, and then with the predistorter. Figure 4.3 shows the block

Figure 4.1 Predistorter's AM-AM characteristics curve

Figure 4.2 Predistorter's AM-PM characteristics curve

diagram of a QPSK transmitter that was simulated. The simulation of this system without the predistorter gave the amplified signal whose magnitude spectrum is shown in figure 4.5. The high power sidelobes shown in figure 4.5 shows the extend to which the signal is distorted, particularly when compared to the filtered signal shown in figure 4.4. When the predistorter was used in the simulation, we obtain an undistorted amplified signal with very low power sidelobes as shown in figure 4.6.

In practice, the exact characteristics for each amplifier are not usually provided by its manufacturer. Even if some characteristics were initially known, the aging process of the amplifier is bound to cause changes in the value of the amplifier operating parameters, and the characteristics are normally given for a specific operating frequency. Therefore, a training sequence will be needed to determine the exact characteristics of the amplifier at the time it is being used and then calculate the predistorter table lookup functions:

4.3 Adaptive Predistorter Design

The adaptive predistortion system has the same elements in the forward path as the non-adaptive system, with the addition of a feedback path and the ability to change the entries of the lookup table functions. Since, the adaptation is taking place at baseband, the amplified signal must be down converted and fed back.

To study the adaptation process, a number of simulation runs were first carried out with the predistorter having its initial values of unity gain and zero phase change. Values of table entries, spectrum of the amplified signal and the running mean squared error were observed as they change over time. Later after the predistorter has been adapted to

Figure 4.3 Block diagram of transmitter with a non-adaptive predistorter

Figure 4.4 Magnitude spectrum of a filtered QPSK signal

Figure 4.5 Magnitude spectrum of the amplified signal without predistortion

Figure 4.6 Magnitude spectrum of amplified signal with predistortion

the amplifier, we will change the amplifier's characteristics (make it drift) and do some of the above observations again.

Figure 4.7 shows the block diagram of the simulated QPSK transmitter with an adaptive predistorter. To obtain meaningful comparisons, the characteristics of the amplifier used here are identical to those used in the non-adaptive predistorter case and which are given by figures 4.1 and 4.2.

Using $\alpha = 0.6$, we obtain $\alpha\gamma/g \approx 1.15$ which is in the linear region of the amplifier's magnitude characteristics curve. Selection of such value of α will minimize convergence time. Two entries of the predistorter tables were observed to study how they change from their initial value to the required value throughout the adaptation process.

The first entry corresponds to an output around the linear region of the amplifier's magnitude characteristics curve, where $\alpha\gamma/g \approx 1.08$. As explained in the previous chapter, for $\alpha\gamma/g > 1$ the error will oscillate until it reaches its minimum value. For a minimum error of 10^{-5} , and for $\alpha\gamma/g \approx 1.08$, a total number of 5 iterations are required ($N = 5$). Figures 4.8 and 4.9 show the value of this entry of the predistorter table throughout the simulation run. Note, that the error oscillates as expected and a total of $N = 4$ iterations were required for convergence, which is within the theoretical bound of $N = 5$ iterations.

The second entry corresponds to an output around the saturation region of the amplifier's magnitude characteristics curve, where $\alpha\gamma/g \approx 0.444$. As explained in the previous chapter, for $\alpha\gamma/g < 1$ the error will decrease monotonically until it reaches its minimum value. For a minimum error of 10^{-5} , and for $\alpha\gamma/g \approx 0.444$, N is calculated

Figure 4.7 Block diagram of the adaptive transmitter used in simulation

Figure 4.8 Magnitude of an entry, on the linear region, as it changes

Figure 4.9 Phase of an entry, on the linear region, as it changes

to have a maximum of 20 iterations. Figures 4.10 and 4.11 show the value of this entry of the predistorter table throughout the simulation run. Note also, that the error decreases monotonically as expected and that a total of $N = 6$ iterations were required for convergence, which is within the theoretical bound of $N = 20$ iterations.

The update of the second entry occurs more often than the first entry because the average of the transmitted signal is near saturation. Therefore, in terms of the time required for both entries to converge, the first entry required 35 msec of time and the second entry required 20 msec of time (for transmission rate of 9.6 kbaud and each baud consists of 8 samples). Note that, the magnitude and phase of each of the table entries required the same number of iterations since $\beta = 1$ was used in the simulation.

For the same value of $\alpha = 0.6$, a simulation was run and the spectrum of the amplified signal was observed. The spectrum of the amplified signal is initially the same as the one shown in figure 4.5, i.e., as if the predistorter was not present. After 6000 samples ($6000/8 = 750$ symbols), the spectrum became almost identical to the filtered signal spectrum (shown in figure 4.4) as shown in figure 4.12.

We examine now the change in the Mean Squared Error value throughout the adaptation process which takes place during each simulation run. For the same value of $\alpha = 0.6$, the MSE for both the magnitude and phase tables is taken every 10,000 samples (1250 symbols) and is shown in figures 4.13 and 4.14 respectively. The MSE curves decrease in value with some fluctuations. The fluctuations are due to some entries in the tables that are not being accessed that often. However, eventually all entries in both tables will converge and no fluctuations will appear in the MSE curves. Note that the fluctuations in the MSE curve of the phase table occur at the same position the fluctuation

Figure 4.10 Magnitude of an entry, on the saturation region, as it changes

Figure 4.11 Phase of an entry, on the saturation region, as it changes

Figure 4.12 Magnitude spectrum of amplified signal after the predistortion table has converged

in the MSE curve of the magnitude table. Also at the time the MSE curves approach their minimum, the spectrum of the amplified signal was almost identical to the filtered signal.

It is adequate to find a value of α that minimizes convergence time of the MSE value for both tables. Therefore, to find the optimum value of α , α can be varied and the MSE curve that gets to its minimum the fastest, will determine the optimum α . Figures 4.15 and 4.16 show the MSE for various values of α . It is evident from these curves, that the optimum value of α is between 0.5 and 0.6 and the value of $\alpha = 0.5$ is the closest to optimum.

Using $\alpha = 0.5$, the spectrum of the amplified signal was observed over time, and it was found that the spectrum of the amplified signal was almost identical to the spectrum of the filtered signal after approximately 5100 samples (638 symbols). In terms of time, it took approximately 66 msec, assuming 8 samples per baud and a baud value of 9.6k, while theoretically it should take approximately 100 msec ($T = \frac{NL}{w} = 100\text{msec}$ where $N = 30$ was assumed, $L = 256$ and $w = 76800$ samples/sec) of time for both tables to converge.

Finally, in order to test the adaptation algorithm against a change in the carrier frequency, after the predistorter tables have converged, a sudden change in the amplifier's characteristics functions was introduced. The modified amplifier's characteristics functions are shown in figures 4.17 and 4.18. The effect of this change can be seen by observing the MSE as shown in figures 4.19 and 4.20. The amplifier's characteristics functions are changed on the 24th interval, where the MSE value was minimum. Exactly on the next interval, the MSE value jumps and reconvergence starts again. It took

Figure 4.13 The MSE of the magnitude's table for $\alpha = 0.6$

Figure 4.14 The MSE of the phase's table for $\alpha = 0.6$

Figure 4.15 The MSE of the magnitude's table for different values of α 's

Figure 4.16 The MSE of the phase's table for different values of α 's

approximately 52 msec (4000 samples = 500 symbols) of time for the spectrum of the amplified signal to become identical to the spectrum of the filtered signal.

4.4 Comparison to Recent Research

Predistortion linearization techniques have attracted a wide interest recently. Two powerful predistorters were recently reported by Cavers [3] and Nagata [2].

The predistorter reported by Nagata uses a complex lookup table approach that maps every possible complex input signal to an entry in the table. The updating of the table is done in Cartesian form (i.e., the complex signal is represented as real and imaginary parts) using a linear adaptation scheme, and requires relatively small computational effort. This method has two basic drawbacks: (i) it requires a large amount of memory storage (about 2 Mwords) and, (ii) it requires a phase shifter in the feedback path for stability in the adaptation update. The large memory requirements lead to a very slow convergence in this method (about 10 sec, at 16 ksym/sec). In systems involving frequency conversion, switching to a new carrier requires readjustment of the phase shifter, and reconvergence of the table which require additional 10 seconds.

Most of the drawbacks discussed above are due to the large table size (memory requirements). The table size can be reduced by noting that the predistorter does not require the instantaneous input signal, instead it only needs the amplitude of the input signal to produce complementary nonlinearities to that of the amplifier's AM-AM and AM-PM. This observation was exploited in the approach reported here as well as in the technique developed by Cavers.

Figure 4.17 The modified amplifier's magnitude characteristics curve

Figure 4.18 The modified amplifier's phase characteristics curve

Figure 4.19 The MSE of the magnitude's table

Figure 4.20 The MSE of the phase's table

The predistorter reported by Cavers, uses a complex lookup table approach. But it maps a quantized amplitude of the input signal to an entry in the table. The update of the table is done in Cartesian form using a nonlinear adaptation scheme that eliminates the need of a phase shifter. This method requires a small amount of memory (about 100 complex words) and converges fast (few msec). The drawbacks of this method is its relatively large computational effort.

The predistorter described in this thesis has similar performance to the one described by Cavers in terms of its memory size and convergence time. The main difference between the predistorter reported here and the one developed by Cavers is in the realization details. Our predistorter is realized in polar form as two real lookup tables, one for the magnitude (AM-AM) and one for the phase (AM-PM). In terms of its computational effort, our method requires less computations for coefficient updates. However, since we are using polar rather than Cartesian tables, there may be an additional computation for Cartesian to polar conversion. It should be noted however, that Caver's method also requires the calculation of the input signal amplitude.

In general, the predistorter proposed in this thesis and the one proposed by Cavers are very similar in terms of memory requirements and adaptation speed.

Chapter 5

Conclusions and Recommendations

5.1 Conclusions

In this we have presented a powerful baseband adaptive predistortion linearizer, that is capable of linearizing an AM-AM and AM-PM characteristics functions. The linearization is independent of the modulation scheme and is suited for narrow band signals.

In chapter 3, we have given a mathematical description of the predistortion linearization technique for analog waveforms. Then the non-adaptive predistortion linearization technique for digitized waveforms were studied along with their quantization noise effects. It was found that the variances of both the amplitude and phase quantization noise are inversely proportional to the squared value of the table size (L^2), and hence doubling the table size will reduce the variance by a factor of 4 or 6 dB. Following this, the adaptive predistortion method was analytically analyzed. Its convergence conditions were derived and it was found that to guarantee convergence, it is necessary to have

$$0 < \alpha\gamma/g < 2, \quad (5.1)$$

where α is the adaptation coefficient, γ is the slope around a region on the amplifier's amplitude characteristics curve and g is the required constant gain. Also the convergence time was analyzed and the criteria for convergence time were described. The two criteria of convergence used in this thesis are the Mean Squared Error in time domain and the

observation of the spectrum of the amplified signal in the frequency domain. Finally, the implementation of a transmitter that includes the proposed adaptive predistortion linearizer was analyzed. It was found that it is feasible to implement the linearizer with existing components.

In chapter 4, we have suggested an implementation of both the non-adaptive predistorter and the adaptive predistorter along with their computer simulation. It was found that for a non-varying amplifier characteristics, the non-adaptive predistorter was able to cancel most of the nonlinearities introduced by the amplifier. Then the adaptive predistorter was tested, and it was found that it requires on the order of tenth of a second for the predistorter to adapt to a completely unknown amplifier. Also it was found that it requires few msec for the predistorter tables to reconverge upon a switch to a new channel.

In terms of memory storage requirements, both the non-adaptive and adaptive predistorters require less than 1k words to eliminate most nonlinear distortion in a typical class AB TWT power amplifier.

5.2 Recommendations for Future Research

The adaptation coefficient α depends on the slope of the amplifier's AM-AM characteristics curve in order to guarantee convergence of the entire table. In this thesis, the choice of α was based on the maximum slope of the amplifier's AM-AM characteristics curve. Therefore convergence of the entries near the maximum slope region will take place faster than other table entries. To make all entries of the table converge at the same rate, it is recommended that different adaptation coefficients be used for different regions

on the amplifier's AM-AM characteristics curve. It is expected that such an approach will minimize convergence time for the entire table.

It is also recommended that a transmitter with the proposed adaptive predistorter be implemented and tested on a real system.

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