İSTANBUL TECHNICAL UNIVERSITY ★ INSTITUTE OF SCIENCE AND TECHNOLOGY

ADAPTIVE FEEDFORWARD AMPLIFIER DESIGN

M.Sc. Thesis by
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Department : **Electronics and Communication Engineering**

Programme: **Electronics Engineering**

JUNE 2005

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Date of submission : **6 May 2005**

Date of defence examination : **3 January 2005**

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JUNE 2005

İSTANBUL TEKNİK ÜNİVERSİTESİ ★ FEN BİLİMLERİ ENSTİTÜSÜ

UYARLANIR İLERİ BESLEMELİ KUVVETLENDİRİCİ TASARIMI

YÜKSEK LİSANS TEZİ

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Tezin Enstitüye Verildiği Tarih : 6 Mayıs 2005
Tezin Savunulduğu Tarih : 3 Haziran 2005

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HAZİRAN 2005

FOREWORD

Communications has made great strides in the last 50 years, in quality of course but even more spectacularly, in the quantity of information exchanged. From the very few TV channels transmitted over the air to the hundreds of channels available via cable or satellite. This means that much more bandwidth is required. In modern systems, more complex modulations are being used to increase the bandwidth efficiency. These modulations require high fidelity transmitters using highly linear amplifiers.

In this thesis, firstly, I explained what the importance of linearity is for a communication system. Secondly, I gave the information about design requirements of power amplifiers. Then, I studied feedforward amplifiers after giving brief information about other linearization techniques and described the methods of adaptation of feedforward systems. Finally, I have simulated the results of study on a 5.8 GHz feedforward power amplifier.

I would like to thank to Prof. Dr. Osman Palamutçuoğulları, Dr. Bülent Yağcı and my family for their contributions to this master thesis study.

June 2005 Engin KURT

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LIST OF ABBREVIATIONS

PCS :Personal Communication Systems
IMT-2000 :International Mobile Telecommunications in the year 2000
UMTS :Universal Mobile Telecommunications Systems
UTRA :UMTS Terrestrial Radio Access
WCDMA :Wideband Code Division Multiple Access
CW :Cosine Wave
DC :Direct Current
IM :Intermodulation
IP3 :Third-Order Intercept Point
PA :Power Amplifier
SSB :Single Side Band
AM :Amplitude Modulation
PM :Phase Modulation
RF :Radio Frequency
BJT :Bipolar Junction Transistor
MOSFET :Metal Oxide Semiconductor Field Effect Transistor
LDMOS :Laterally Diffused Metal Oxide Semiconductor
VMOS :Vertical Metal Oxide Semiconductor
MESFET :Metal Semiconductor Field Effect Transistor
GaAs :Gallium Arsenide
PEP :Peak Envelope Power
VSWR :Voltage Standing Wave Ratio
LINC :Linear Amplification With Nonlinear Components
CALLUM :Combined Analog Locked Loop Universal Modulator
VCO :Voltage Controlled Oscillator
DSP :Digital Signal Processing
IMD :Intermodulation Distortion
LMS :Least Mean Square
ADS :Advanced Design System
MMIC :Monolithic Microwave Integrated Circuit

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ÖZET

20. yüzyılın ikinci yarısında, iletişim sistemleri muazzam gelişme gösterdi. Havadan iletilen çok az sayıdaki TV kanallarından kablolu TV yada uydu aracılığyla yayılan yüzlerce TV kanallarına. Hemen hemen sadece özel profesyonel kullanım için olan 'walkie talkie'ler üzerinde yapılan basit bilgi değiş tokuşundan herkes tarafından kullanılan modern cep telefonlarına. Basit verilerin üzerinde yazdırıldığı gürültülü ve yavaş teleks makinalarından faks makinesine ve elektronik postaya. Oyunun şimdiki adı kapasite ve hız.

Günümüzde sayısal işleme, iletilmek istenen bilgi üzerindeki hemen hemen tüm fazlalıkları sıyırıp atabilmeyi mümkün kılmaktadır ve böylelikle kapasiteyi arttırmaktadır. Bununla birlikte, bilindiği üzere iletişim kanalının en yüksek kapasitesi frekans, band genişliği ve işaret gürültü oranı tarafından sınırlanmaktadır. Modern sistemlerde, band genişliği verimi karmaşık modülasyonlar kullanılarak arttırılır. Bu tip modülasyonlar yüksek doğrusallıkta kuvvetlendiriciler kullanan yüksek doğruluk düzeyine sahip vericileri gerektirir.

Maalesef doğrusallık, genellikle çok zayıf elektriksel verimlilik, yüksek maliyet ve düşük güvenilirlik manasına gelmektedir. Bu sorun ilk olarak yüksek kapasiteli eşeksenli kablo telefon sistemleri üzerinde çalışan mühendisler tarafından ele alınmıştır. Geri besleme, ön bozma ve ileri besleme gibi doğrusallaştırma teknikleri bu zorlukların üstesinden gelmek için geliştirilmiştir.

Bu tez çalışmasında, ilk olarak doğrusallığın öneminin bir iletişim sistemi için ne olduğu açıklanmıştır. İkinci olarak, güç kuvvetlendiricilerinin tasarım isterleri hakkında bilgi verilmiştir. Daha sonra geri besleme, ön bozma ve ileri beslemeyi içeren doğrusallaştırma teknikleri gözden geçilmiştir ve uyarlanır ileri besleme sistemleri detaylı bir şekilde incelenmiştir. Analizin sonuçları 5.8 GHz'lik bir uygulama için uyarlanır ileri besleme kuvvetlendirici tasarımıyla benzetim olarak yapılmıştır.

ADAPTIVE FEEDFORWARD AMPLIFIER DESIGN

SUMMARY

Communication systems have gone forward greatly in the second half of the 20^{th} century. From the very few TV channels transmitted over the air to the hundreds of channels available via cable or satellites. From the simple information exchanged on ‘walkie talkies’ for almost only exclusive professional usage, to modern cellular telephones for everybody. From simple data being printed on noisy and slow teletype writers, to faxes and to electronic mail. The name of the game is now capacity and speed.

Today digital processing makes it possible to strip down almost all redundancy from the information transmitted and thereby increase capacity. However, it is known that the ultimate capacity of a communication channel is limited by frequency, bandwidth and signal-to-noise ratio. In modern systems, complex modulations are being used to increase the efficiency of bandwidth. These types of modulations require high fidelity transmitters using highly linear amplifiers.

Unfortunately, linearity usually means very poor electrical efficiency, high cost and low reliability. This problem has first been tackled by the engineers working on high capacity coaxial cable telephone systems. Linearization techniques such as feedback, predistortion and feedforward are developed to overcome these difficulties.

In this thesis, firstly, it is explained what the importance of linearity is for a communication system. Secondly, the information about design requirements of power amplifiers is given. Then, the linearization techniques including feedback, predistortion and feedforward overviewed and adaptive feedforward systems studied and analyzed in detail. The outcome of the analysis is simulated by designing of an adaptive feedforward amplifier for a 5.8 GHz application.

1. INTRODUCTION

With the evolution of existing and new standards for mobile communication systems and wireless multimedia services, the quantity and complexity of the signals to be transmitted form a single location is increasing. The demands placed on radio frequency power amplifiers, which are used in such systems, are subsequently increasing in terms of bandwidth, output power, efficiency and allowable level of output distortion. There is a growing need for amplifiers, which amplify all types of signals without adding significant distortion and capable of operating over a wide bandwidth and at potentially high levels of output power.

The primary goal of any radio system is to transmit and/or receive information. For broadcast radio and radiotelephony, information is usually in the form of speech; however, text, pictures and video are also being used for data transfer and wireless multimedia applications. For example, most analog first generation cellular systems is limited to speech, but the development of second generation digital systems is allowing both speech and limited data capabilities. Third generation systems and indeed developed second generation systems are being supported much higher rates of data transfer. One of the reasons that it is becoming more practical and cost effective to offer such services is that radio frequency power amplifiers, which are inherently nonlinear, can now be built to very high specifications and fulfill the requirements of a "linear amplifier". This has not been possible thus far using traditional techniques because the amplifiers generate distortion in the form of intermodulation and spectral regrownth. Power generated outside of the transmit channel causes interference in adjacent radio channels, while power generated in-band can cause errors in signal vectors and hence, degradation in demodulation accuracy.

Because of their versatility and flexibility, linear amplifiers are finding an increasing number of applications in cellular radio systems, personal communication systems

(PCS), international mobile telecommunications in the year 2000 systems (IMT-2000), and universal mobile telecommunications systems (UMTS). Linear amplifiers are capable of amplifying single-carrier and multicarrier signals, analog and digital signals, and constant envelope and non-constant envelope signals. Linear amplifiers are thus effectively transparent to the modulation format and number of carriers. Furthermore, depending upon the choice of linearization technique, linear amplifiers can operate with low levels of distortion over the wide bandwidths that are necessary to support high data rate services such as the Internet and wireless multimedia. For example, wideband linear amplifiers are integral part of the third generation system UTRA (UMTS terrestrial radio access), which is based upon wideband code division multiple access (WCDMA).

One of the primary goals of amplifier design is to produce an amplifier that has good efficiency and low distortion; however, in practice there is a trade-off between distortion performance and efficiency. For example, so-called Class A amplifiers have good distortion performance but low efficiency while so-called Class C amplifiers and to some extent Class B amplifiers are reasonably efficient but introduce significant distortion. As the power level increases, efficiency becomes more important.

A number of techniques, referred to as linearization techniques, have been developed that eliminate or reduce the amount of distortion added by an inherently nonlinear power amplifier. An easy to use linearization method for correcting distortion in amplifiers is to apply negative feedback, however such a technique is inherently bandwidth limited and is not suitable for wideband applications such WCDMA. An alternative approach, which is suitable for wideband applications, is to use Class AB amplifiers, which are more efficient than Class A, and apply feedforward linearization [1].

2. LINEARITY IN COMMUNICATION SYSTEMS

2.1 Distortion

Distortion is change in form of signal during transmission usually with impairment of quality.

All amplifiers possess this property of distorting the signals they are required to amplify. The existence of distortion and hence nonlinearity in audio amplifiers is very displeasing to the ear and high fidelity amplifiers have been designed and refined over the years to reduce it to levels considered to be inaudible by the human ear. The advent of feedback correction by H.S. Black has enabled this to be achieved with relative ease.

When considering radio frequency amplifiers, the resolved audio fidelity of the transmitted signal is still of importance, but is no longer the only consideration. Spectral efficiency, interference and the need to be considerate to other users of the spectrum all become important, along with signal vector error considerations for the signal itself [2].

2.2 The Requirement for Linearity

All radio systems are required to cause the minimum possible interference to other users; they must therefore maintain their transmissions within the bandwidth allocated to them and not radiate significant energy outside of it. Nonlinearities within the system components of the radio equipment cause distortion of the transmitted signal and result in the generation of signals outside of the intended frequency channel or band. These unwanted distortion products are potential interfering sources to other radio users and must be reduced to a level where both systems can operate satisfactory.

In the case of a high power broadcast transmitter, this requirement becomes acute, as the distortion products, although many times smaller than the main output signal, may still be quite large in absolute terms and hence cause interference [2].

2.3 Linear Amplifier Input/Output Characteristics

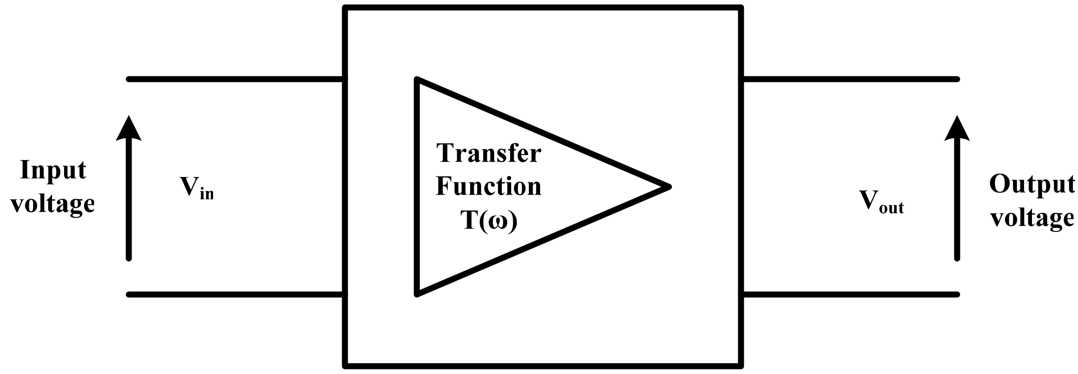

Figure 2.1a: Amplifier two-port representation.

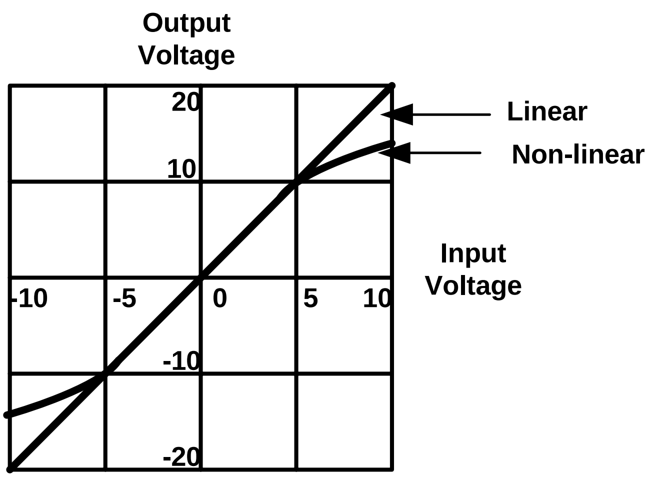

Figure 2.1b: Amplifier input/output characteristic.

Figure 2.1a shows an amplifier represented as a two-port network having an input voltage V_{in}, output voltage V_{out} and transfer function T(ϖ). For a perfectly linear amplifier, the output voltage is simply a constant times the input voltage, that is,

$$V_{out} = G \cdot V_{in} \tag{2.1}$$

Regardless of the signal level, all signals are increased in magnitude by the same factor G and there is a fixed phase shift (equal to the time delay) between input and output for a signal at a given frequency.

In terms of the frequency response T(ϖ), an ideal amplifier has constant characteristics over the bandwidth of the input signal. That is, constant gain, linear phase and hence constant delay. An ideal amplifier is also memoryless; that is, the response of the amplifier at any point in time is determined solely by the value of the input signal at that moment and not by any previous values.

In practice, however, the devices used in amplifiers, such as transistors, have nonlinearities that make the output voltage a function of higher order terms of the input voltage; the input/output characteristics is then said to be nonlinear. Such nonlinear amplifiers also typically have frequency-dependent gain, nonlinear phase and memory [2].

2.3.1 Series Representation of a Nonlinear Amplifier

A nonlinear output voltage can be expressed mathematically as a series such that

$$V_{out} = G_1 \cdot V_{in} + G_2 \cdot V_{in}^2 + G_3 \cdot V_{in}^3 + \cdots G_n \cdot V_{in}^n \qquad (2.2)$$

The amplifier constants $G_{1..n}$ determine the exact shape of the input/output characteristics; for example, Figure 2.1b shows a voltage transfer function having the form given by equation (2.2) (for comparison, the linear response (2.1) is also shown).

As Figure 2.1b illustrates, at high signal levels the output voltage compresses for both positive and negative values. This type of compression (signal clipping) is due to the third-order term G_3 while the-second order term G_2 tends to cause overshoot at one end (gain expansion) and clipping (gain compression) at the other. In practice, both terms are present to a greater or lesser extent resulting in distortion of the output signal regardless of the input signal level [2].

2.3.2 AM-AM and AM-PM Characteristics

Voltages are vector quantities having both amplitude and phase; therefore, an alternative way of looking at the input/output characteristics is to treat amplitude and phase separately. This method is similar to that used for frequency transfer functions, which also have a complex amplitude and phase response – the difference there is that the amplitude and phase responses are functions of frequency and not input level.

For example Figures 2.2(a,b) show the amplifier input/output characteristics in term of the amplitude and phase response for the same case as in Figure 2.1. The amplitude response (Figure 2.2a) is referred to as the AM-AM characteristic and the phase response (Figure 2.2b) is called AM-PM characteristic. The distortion introduced by a nonlinear amplifier is frequently explained in term of AM-AM and AM-PM characteristics and is strongly dependent upon the class of operation in which the amplifier is used [2].

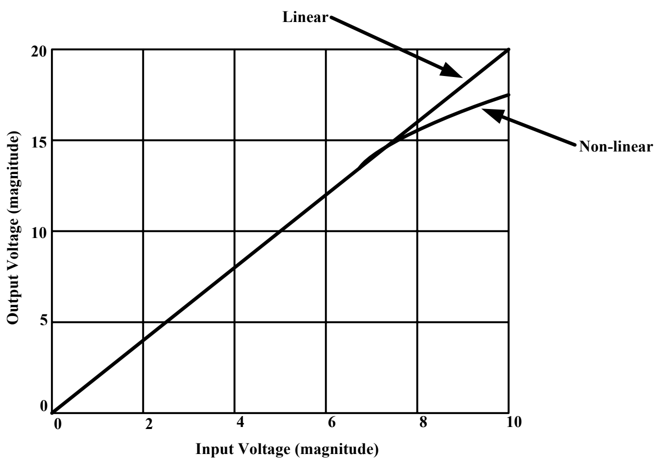

Figure 2.2a: Amplifier AM/AM Characteristic

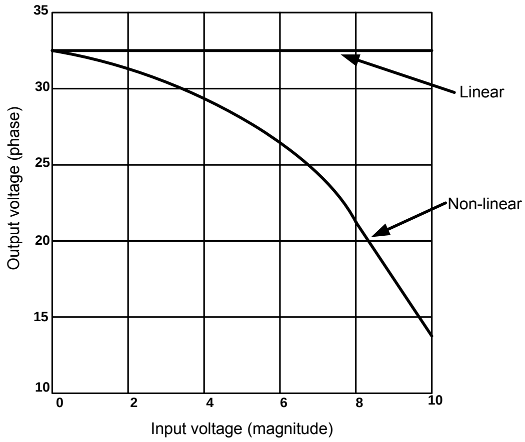

Figure 2.2b: Amplifier AM/PM Characteristic

2.3.3 Single-Carrier Output and Harmonic Distortion

When a single unmodulated CW carrier as the input signal; the input voltage (peak amplitude a, frequency f_1, and arbitrary phase offset ϕ) then has the form

$$V_{in}(t) = a \cdot \cos\left(2 \cdot \pi \cdot f_1 \cdot t + \phi\right) \tag{2.3}$$

The linear output voltage is calculated from (2.1) and the nonlinear output voltage from (2.2). Figures 2.3a and 2.4a show signals in the time domain; as expected, the nonlinear transfer function causes signal clipping (compression) of the output voltage.

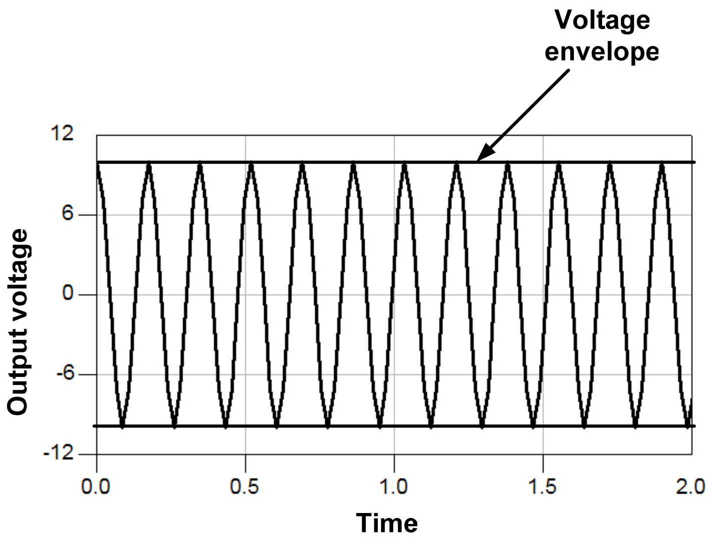

Figure 2.3a: Single carrier linear output – time domain response

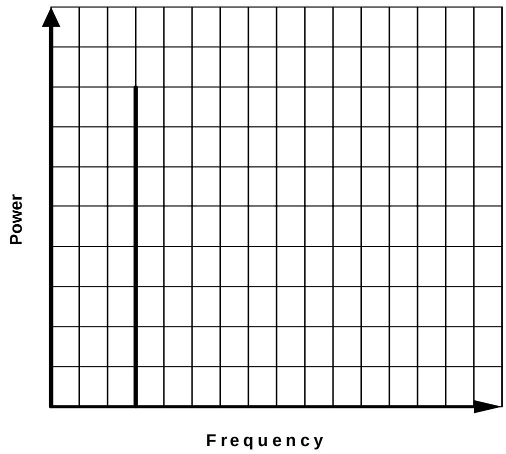

Figure 2.3b: Single carrier linear output – frequency domain response

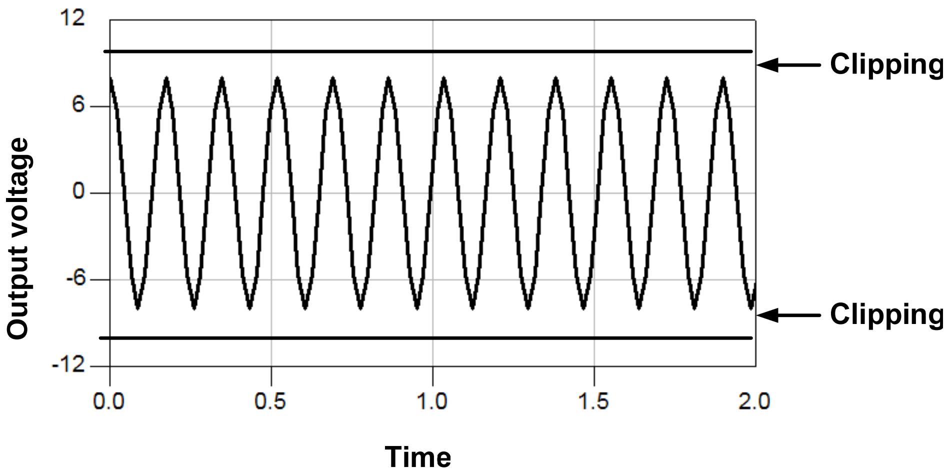

Figure 2.4a: Single carrier nonlinear output – time domain response

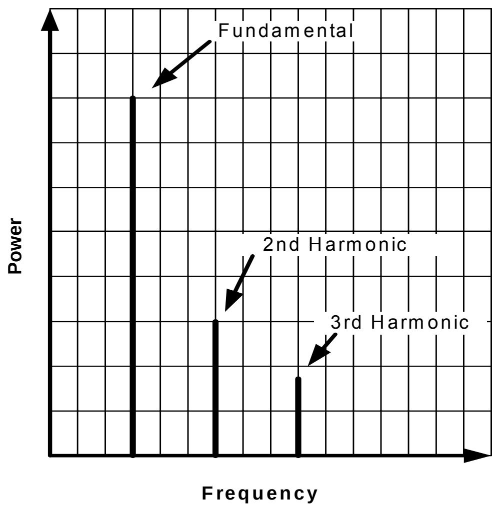

Figure 2.4b: Single carrier nonlinear output – frequency domain response

The frequency-domain response, which is obtained by taking the Fourier transform of the time-domain waveform is usually presented in the form of a power and phase spectrum that is,

$$Power(f) = \overrightarrow{\left(\left|fft\left(v_{out}\right)\right|\right)^2} \qquad (2.4)$$

$$Phase(f) = \overrightarrow{\arg\left(\left|fft\left(v_{out}\right)\right|\right)}$$

Figures 2.3b and 2.4b show the power spectra for linear and nonlinear outputs, respectively. In the linear case only the amplified frequency at f_1 is present, while in the nonlinear example there are additional frequency terms, namely:

- A DC component
- The fundamental tone f_1 - amplitude compressed;
- The second harmonic at 2 f_1;
- The third harmonic at 3 f_1.

The response shown in Figure 2.4b is example of harmonic distortion and occurs even with relatively simple signal such as a single unmodulated carrier. Alternatively as Table 2.1 shows, the same result can be presented in a different from by evaluating in the time domain with $\theta = 2\pi f_1 t$ and collecting terms of similar order.

Table 2.1: Harmonic Distortion

DC term	$\frac{G_2 \cdot a^2}{2}$
Fundamental	$G_1\left(1+\frac{3\cdot G_3 a^2}{4\cdot G_1}\right)\cdot a\cdot\cos\ \theta$
Second Order	$\frac{G_2 \cdot a^2}{2}\cdot\cos\ 2\theta$
Third Order	$\frac{G_3 \cdot a^3}{4}\cos\ 3\theta$

Note that for a 1-dB increase in input signal level, the second-order harmonic terms goes up by 2dB (proportional to a^2) and third-order harmonic by 3dB (proportional to a^3) [2].

2.3.4 Two-Tone Test – Harmonic and Intermodulation Distortion

Now consider two tones of equal amplitude "a" having frequency f_1 and f_2, respectively, that is,

$$V_{in}(t) = a \cdot \cos\left(2 \cdot \pi \cdot f_1 \cdot t\right) + a \cdot \cos\left(2 \cdot \pi \cdot f_2 \cdot t\right) \tag{2.5}$$

Figure 2.5a shows the time-domain waveform for linear output and it is evident that the signal envelope is no longer constant as in the single-carrier case but varies between maximum and minimum values. This particular type of nonconstant envelope behavior makes two-tone test very useful signal for test and measurement purposes since amplifier is driven through the whole range of its transfer characteristics (from zero to the signal envelope maximum). There is also an important practical advantage associated with a two-tone test, the ease of signal generation.

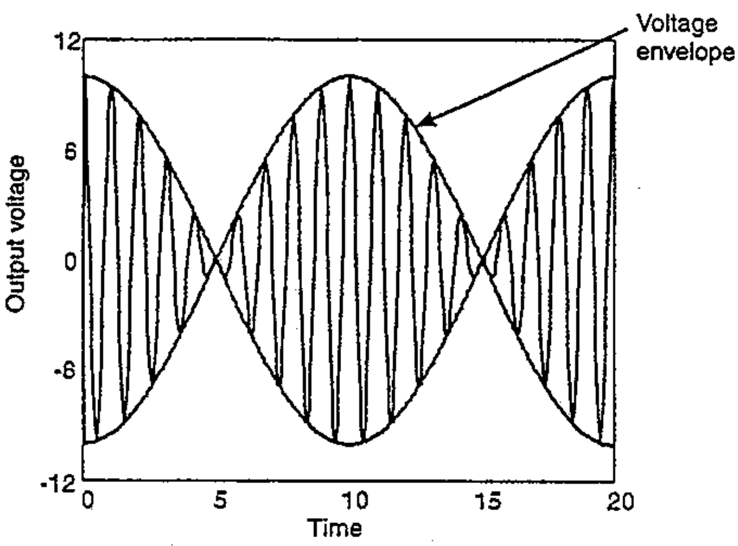

Figure 2.5a: Two-carrier linear output – time domain response

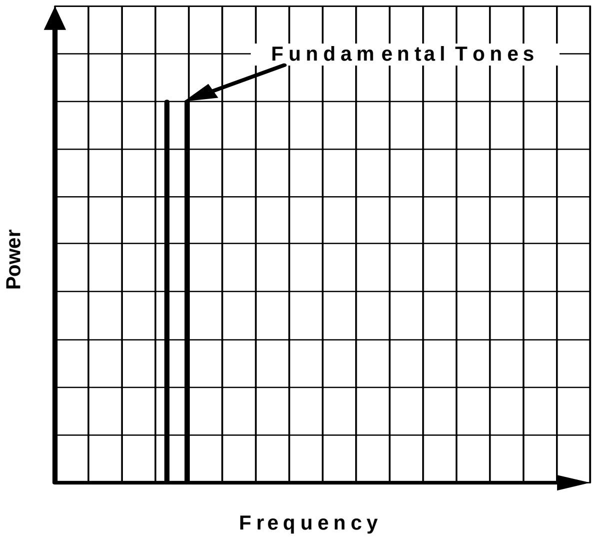

Figure 2.5b: Two-carrier linear output – frequency domain response

For a nonlinear output (Figure 2.6a), the signal no longer follows the true envelope shape and there is asymmetrical signal clipping resulting in distortion. The Fourier transform representation of this distorted time-domain waveform is shown in Figure 2.6b and, in addition to harmonic distortion, other frequency components or intermodulation (IM) products are also present.

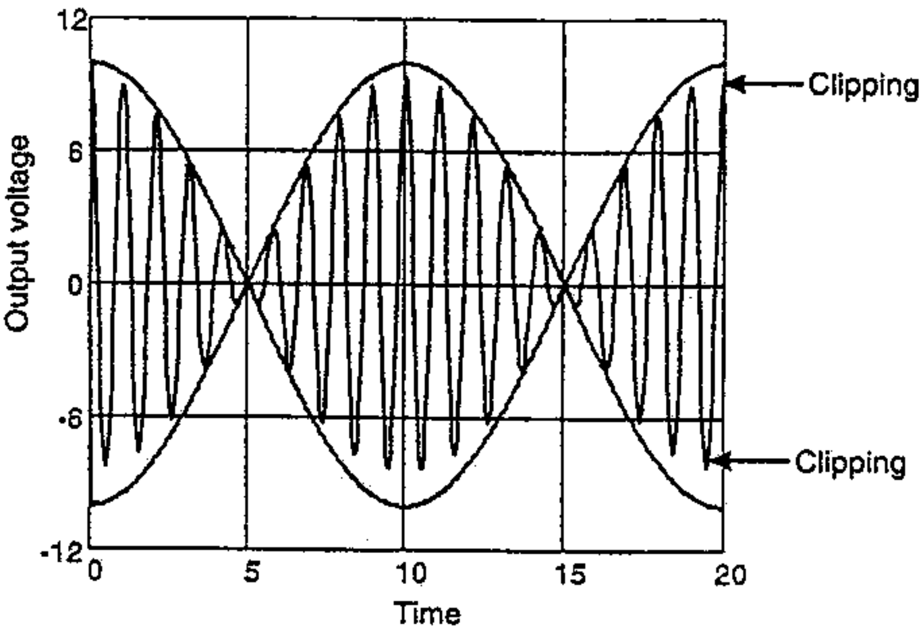

Figure 2.6a: Two-carrier nonlinear output – time domain response

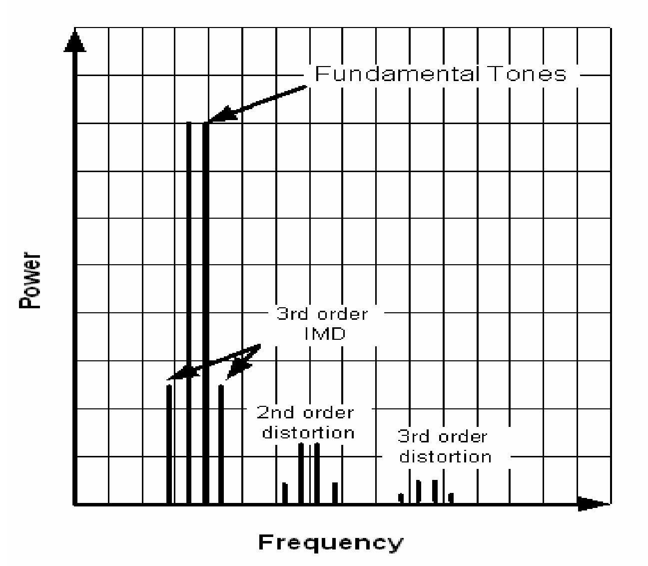

Figure 2.6b: Two-carrier nonlinear output – frequency domain response

Thus, for two unmodulated tones the frequency spectra consists of:

- A DC term;
- Fundamental tones f_1 and f_2 – compressed;
- Harmonics
- Intermodulation products.

Alternatively, Table 2.2 shows the results of evaluating in the time domain with a two-tone signal as the input and collecting terms of similar order.

As before, if the input signal level increased by 1dB, the second-order terms increase by 2dB (proportional to a^2) and the third-order terms increase by 3dB (proportional to a^3) [2].

Table 2.2: Two-Tone Test Frequency Components

DC term	$G_2 \cdot a^2$
Fundamental	$G_1 \cdot a \cdot \left(1 + \frac{9 \cdot G_3 \cdot a^2}{4}\right) \cdot \left(\cos\ \theta_1 + \cos\ \theta_2\right)$
Second Order	$\frac{G_2 \cdot a^2}{2} \cdot \left(\cos\ 2\theta_1 + \cos\ 2\theta_2\right)$ $+ G_2 \cdot a^2 \cdot \left(\cos\left(\theta_1 + \theta_2\right) + \cos\left(\theta_1 - \theta_2\right)\right)$
Third Order	$\frac{G_3 \cdot a^3}{4} \cdot \left(\cos 3\theta_1 + \cos 3\theta_2\right)$ $+ \frac{3 \cdot G_3 \cdot a^3}{4} \cdot \left[\cos\left(2\theta_1 + \theta_2\right) + \cos\left(2\theta_1 - \theta_2\right)\right]$ $+ \frac{3 \cdot G_3 \cdot a^3}{4} \cdot \left[\cos\left(2\theta_2 + \theta_1\right) + \cos\left(2\theta_2 - \theta_1\right)\right]$

2.3.5 Third-Order Intercept Point (IP3)

In order to characterize the third-order distortion of an amplifier, the term $\cos(2\theta_1 - \theta_2)$ and $\cos(2\theta_2 - \theta_1)$ are often used. Since they are proportional to a^3, these intermodulation products increase by 3dB when the fundamental goes up by 1dB. The third-order intercept point is then defined as the theoretical level at which the intermodulation products are equal to the fundamental tone (Figure 2.7) [2].

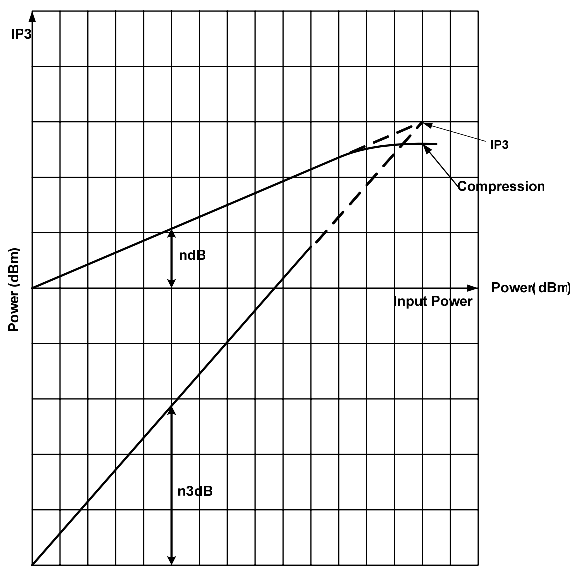

Figure 2.7: Third-order intercept point (IP3)

2.3.6 Distortion of Multicarrier Signals

For a multicarrier signal composed of evenly spaced tones (spacing Δf), the intermodulation products also fall on a Δf grid. IM products thus appear within the same band as the carriers themselves, and hence any thoughts about a filter to remove unwanted intermodulation products must now clearly be abandoned.

As the number of tones is increased, the number of third-order beats (IM products due to third-order distortion) also increases and, theoretically, the highest intermodulation level occurs in the center of the band. In order to measure this "worst case", a gap is often left in the middle of the carriers. For example, Figure 2.8 shows two groups of four tones separated by gap of $2\Delta f$ and the intermodulation product in the center is clearly visible in Figure 2.9. Note that if the number of tones is increased but the total average power is kept constant, the intermodulation performance is degraded [2].

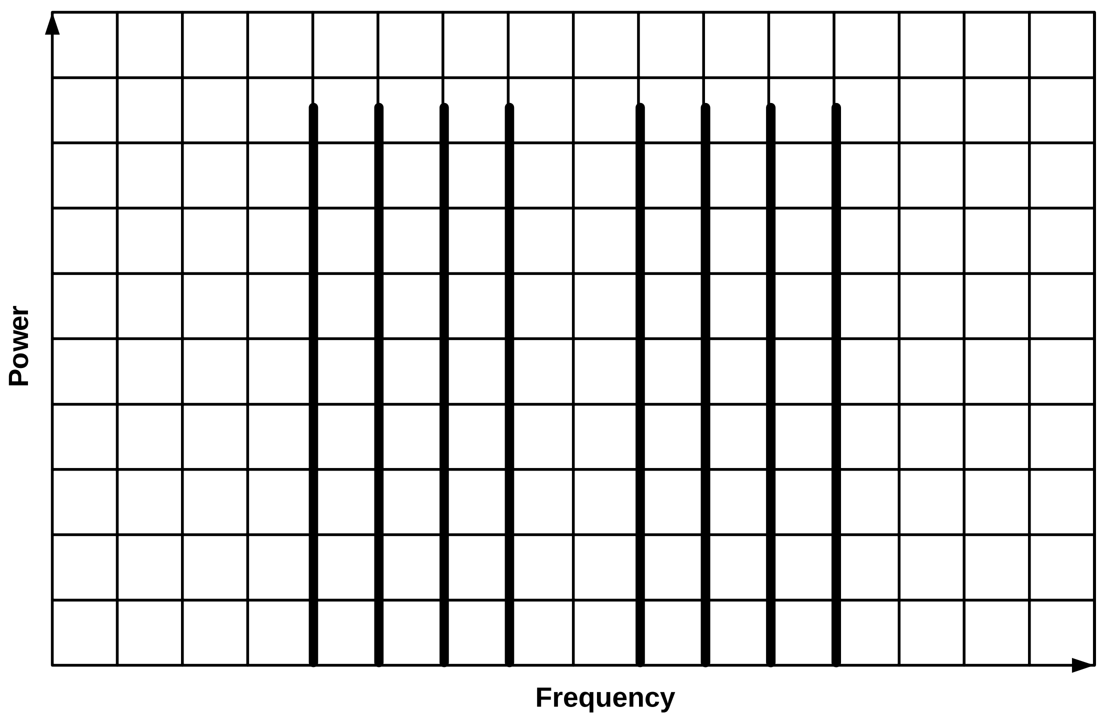

Figure 2.8: Eight-carrier linear output

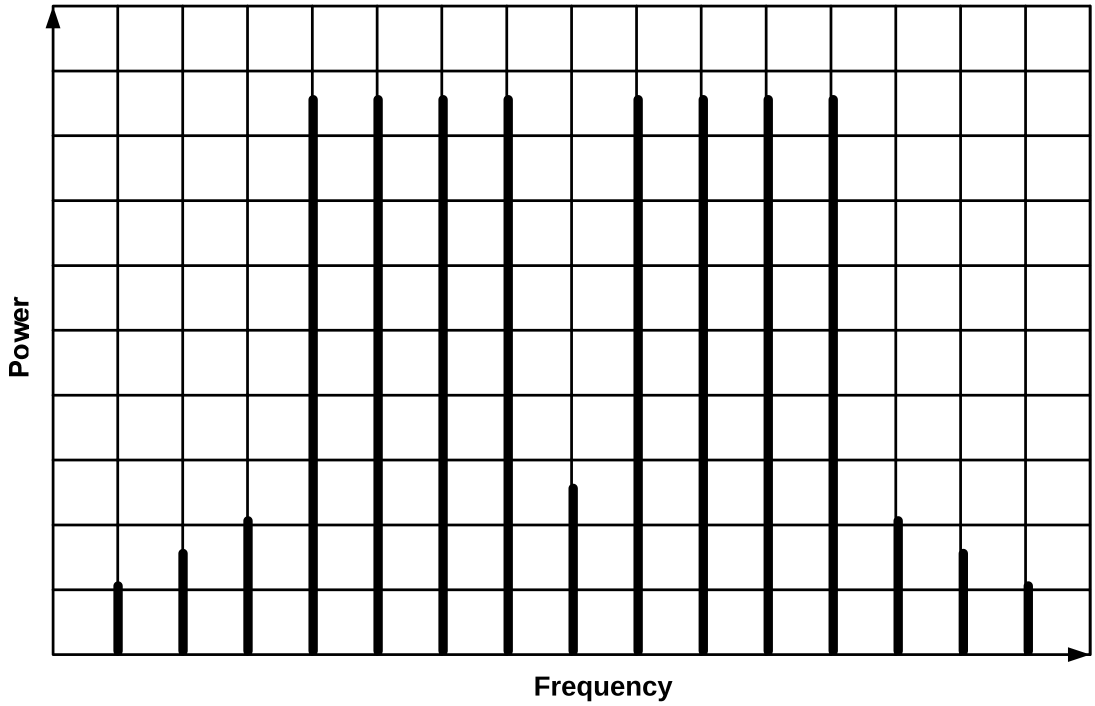

Figure 2.9: Eight-carrier nonlinear output

2.3.7 Intermodulation and Spectral Regrownth

The nonlinearities produced by discrete signals such as unmodulated carriers are known as intermodulation products and appear at discrete frequencies. For more complex (modulated) signals, however, the nonlinearities appear over a continuous band of frequencies and are often referred to as spectral regrownth. For example, the level of adjacent channel power (a measure of the spectral regrownth) is often used as a measure of linearity for complex modulated signals as opposed to the level of discrete intermodulation products for a simple two-tone test [2].

3. POWER AMPLIFIERS AND SYSTEM DESIGN

Radio frequency power amplifiers (PAs) may be broadly defined two categories; those which attempt to preserve the original wave shape of the input signal at the output and those which make no attempt at its preservation. The former category is termed linear amplifiers and the latter constant envelope or nonlinear amplifier.

Within the broad categories outlined above, there are number of sub-divisions or classes of amplifier commonly used. The distinction between the various classes occurs, for example, because of their circuit configurations, operational topologies, linearity and efficiency.

There are three main classes of linear amplifier; A, AB and B, with Class A generally being the most linear and least efficient of the three. Such amplifiers are tradionally most commonly employed in SSB or AM transmitters, where the modulation is transmitted at least partially by means of the amplitude of the RF signal and preservation of the signal envelope is important. In such transmitters, the nonlinearities which are inevitably present, produce distortion in the form of splatter onto adjacent channels as well as distortion within the wanted channel. The linearity performance of these transmitters is therefore an important parameter [1].

The basic building block of a power amplifier is a power transistor. A single transistor can act as an amplifier, but to meet a certain gain or power output requirement, several stages containing one or more transistors are usually cascaded together. For example, Figure 3.1 shows a three-stage power amplifier; the first or input stage has high gain and low output power since the signal level is low, while the final or output stage typically has low gain but high output power. Output stages often use two or more devices in parallel to increase the available output power. The purpose of the second or driver stage is to provide sufficient input power to the output stage; if the driver is not powerful enough, then the potential high-power output will never actually be achieved [2].

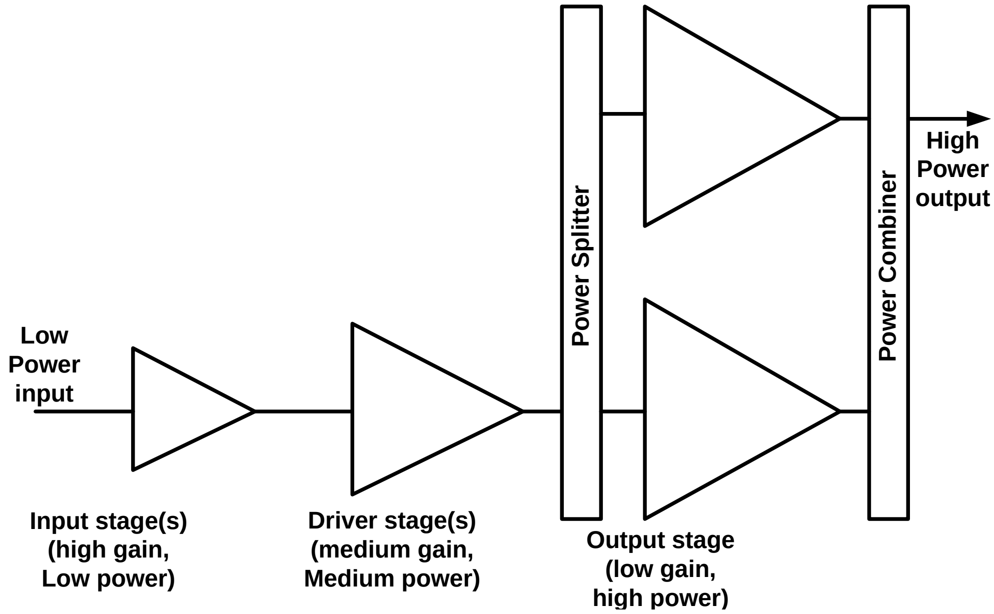

Figure 3.1: Typical power amplifier line-up

3.1 Amplifier Efficiency

One of the many important parameters of an amplifier is its power conversion efficiency (symbol η, units %). Power conversion efficiency is a measure of how effectively an amplifier converts power drawn from the dc supply to useful signal (RF) power delivered to a load, that is,

$$\eta = \frac{P_{load}}{P_{dc}} \tag{3.1}$$

Power that is not converted to useful signal power is dissipated as heat; and for power amplifiers that have a low efficiency; the thermal and mechanical requirements resulting from high levels of heat dissipation are often a limiting factor in a particular design [2].

3.2 Gain-Bandwidth Product

Amplifiers are normally designed for operation over a specific bandwidth, the transmitter band, and ideally have gain that is constant over this bandwidth. Outside of the transmitter band, the gain response tends to drop off at both low and high frequencies (dc amplifiers are a special case). At low frequencies, components such as dc coupling and bypass capacitors have increasing impedances; at high frequencies, a similar effect is cause by internal device capacitances.

The bandwidth over which the gain is within some specified limit, for example 3dB, can be used together with the value of midband gain to define a parameter called the gain-bandwidth product. Another example of the gain-bandwidth product is the transition frequency f_T of a transistor, that is, the theoretical frequency at which the common-emitter current gain is unity.

For given transistor, the gain-bandwidth product is often constant; hence, bandwidth and gain can be traded for each other. For example, the use of negative feedback in an amplifier allows the bandwidth to be increased at the expense of reduced gain. Constant gain over a wide bandwidth is also an important feature of Feedforward amplifiers; thus, gain is often reduced in favor of more broadband operation [2].

3.3 Power Semiconductors

The amplifier linearization techniques are applicable to all forms of amplifier, however, it is anticipated that they will most frequently be used in power amplifier and transmitter design. It is therefore worth examining briefly some of the characteristics of the semiconductor devices which are usually employed in such circuits.

There are two main types of power semiconductor currently employed in high-power RF power amplifiers: bipolar junction transistors (BJTs) and MOSFET devices (in the form of VMOS, TMOS and LDMOS technologies).

The more recently adopted MOS devices have a number of advantages over the more traditional BJT. They are considerably easier to bias, both when used as linear devices in Class A amplifiers and when negatively biased for Class C operation. The bias circuitry does not need to take account of the temperature of the power device since most MOSFET devices have a negative temperature coefficient and are thus thermally stable. Close thermal coupling is required in BJT bias circuits due to their positive temperature coefficient and consequent tendency toward runaway. This close coupling (usually of a biasing diode to the transistor case) is generally unnecessary for MOS devices.

MOSFET devices are also less susceptible to secondary breakdown effects occurring within the device. Such breakdown is caused by excessive instantaneous power levels, due to the peak voltage-current product, which cause hot-spots within the transistor and subsequent device failure. One disadvantage, however, is the variability of the threshold voltage, V_T, between different examples of the same device. This can result in biasing differences between production units and must be taken into account in the amplifier design. In particular, there is much concern over 'V_{GS} drift' in LDMOS devices, with various techniques being suggested to overcome the problem. Essentially the problem amounts to a change in the threshold voltage of 20% or more over the lifetime of the device, with much of this change occurring within the first few hours after first use. Techniques, which have been suggested to overcome the problem, include:

1. A period of 'burn-in' of a given amplifier during post manufacturing 'test', to allow the bulk of the charge to occur prior to final bias setting. This can prove expensive in test time (and hence product cost).
2. Automatic (feedback-based) bias control usually based upon a measure of the average drain current under particular operating conditions.
3. Open-loops bias trimming based on elapsed time from first switch on. The drift profile of a range of samples of a given device may be used to determine a 'typical' drift characteristic and this can be stored in a look-up table, which is accessed based on the elapsed time since the device was first used.
4. Bias control based upon gain monitoring. A change in the bias level will cause associated change in amplifier gain and this can be monitored by comparing the

input and output power levels to the amplifier (or just the output power level, if the input level is known to be fixed and constant).

5. Intelligent monitoring of any linearization scheme surrounding the amplifier. As the bias level changes, the degree of linearization required to meet a given specification will also change and a well-designed linearization scheme will be aware of this. It can therefore re-adjust the bias level to return the degree of correction required to that determined during its design.

Metal semiconductor field effect transistors (MESFETs) are increasingly of interest in the RF power amplifier field. They are widely used in GaAs integrated circuit power amplifiers; particularly for handset applications and silicon carbide (SiC) devices are now being fabricated experimentally for use in higher power systems. SiC devices, in particular, have a number of interesting and useful properties with regard to high power RF and microwave amplification:

1. SiC has a wide bandgap of 3.2eV (compared with 1.1eV for silicon and 1.4eV for Gallium Arsenide), this wide bandgap gives rise to a very high breakdown electric field, some ten times greater than that of either Si or GaAs.
2. The saturated electron velocity is predicted to be around three times that of GaAs at high electric fields.
3. It has a high drain efficiency (largely due to the high breakdown voltage of ~100V).
4. Combining 3 and 4 above yields a large RF power density of 3W/mm.
5. Finally, it has a very thermal conductivity (roughly three times that of silicon and ten times that of GaAs).

The above properties lead to a predicted ultimate power level for a SiC MESFET being al least five times that available from GaAs. This is clearly a significant benefit and hence SiC may well become a key technology in RF amplifier designs in the near future.

Many RF transistors are designed with their application in mind and often unsuitable for other applications seemingly within their ratings. Devices intended for Class C operation, for example, may be destroyed by even modest amounts of standing bias intended to induce Class A operation. Selection of devices for a particular application is therefore not as straightforward as is in the case with audio-frequency power amplifier designs.

A large signal semiconductor may be characterized by three regions of operation: Cut-off, linear or active and saturation.

Cut-off refers to the region of operation in which there is insufficient forward bias on the device for conduction to occur. In the case of a silicon NPN BJT, cut-off occurs when the base-emitter voltage is less than about 0.7V. The equivalent situation for a MOSFET occurs when a gate-source voltage of less than the threshold voltage, V_T, is applied. The value of V_T is around 2V to 3V for typical MOS power FET. In the cut-off region, the device may essentially be considered as an open circuit between all terminals.

As the forward bias on the device is increased and eventually exceeds the relevant threshold ($V_{BE} > 0.7V$ or $V_{GS} > V_T$), the active region is entered. In the case of a BJT, the base-emitter junction becomes a forward bias diode and the collector-emitter junction becomes a current source whose value is equal to the base current times the current gain of the device, β-a linear relationship. To operate in this region the base current must be small enough such that the device does not enter saturation, in other words the collector-emitter voltage must be larger that at saturation, V_{SAT}. A typical value of this saturation voltage is around 0.2V, although it will vary depending upon the device in question.

In the case of a large-signal FET, the active region is represented by a current source between the drain and source terminals of value $i_D = g_m (V_{GS} - V_T)$.

When an FET is driven into saturation, it appears as pure resistance, R_{on} (to a first approximation). It will enter saturation if a drain voltage of less than $i_D R_{on}$ appears due to the load, whilst the device is operating in the active region.

A BJT will enter saturation when the voltage across the load, V_L, is such that:

$$V_{CE} = V_{sat} \qquad (3.2)$$

In this region, the collector-emitter voltage is roughly constant ($=V_{sat}$). The limiting condition for entering this state is that I_B (or V_{BE}) and therefore I_C are increased to a level such that V_C falls to V_B. In other words the base-collector junction is on the verge of becoming forward biased. Further increases of I_B (or V_{BE}) and therefore I_C, lead to V_C falling very close to V_E and limiting at $V_{CE,sat}$ [1].

3.4 Classes of Amplifier Operation

The manner in which transistors are operated or biased is called the class of operation and refers to the output current waveform when input signal is applied. The class of operation, or, more specifically, the transistor conduction angle (the portion of the input cycle for which the transistor conducts and an output current flows), has very important implications for power amplifiers in term of linearity and efficiency. In any given design there is always a tradeoff between linearity and efficiency; as linearity increases, efficiency decreases and vice versa.

There are many different classes of amplifier operation and bias techniques, however, the discussion here is limited primarily to Class A, Class B and Class AB amplifiers. Other classes of amplifier operation are possible. For example; Classes C, D, E and F. However, they are not commonly used for applications requiring high linearity [2].

3.4.1 Class A

In Class A operation (Figure 3.2a, common emitter configuration) the transistor is biased with a dc current I_q greater than the amplitude of the signal current and therefore current (i_c) flows in the collector for the complete duration of the input cycle; the conduction angle is thus 360 degrees. The collector output current waveform for Class A operation is shown in Figure 3.2b.

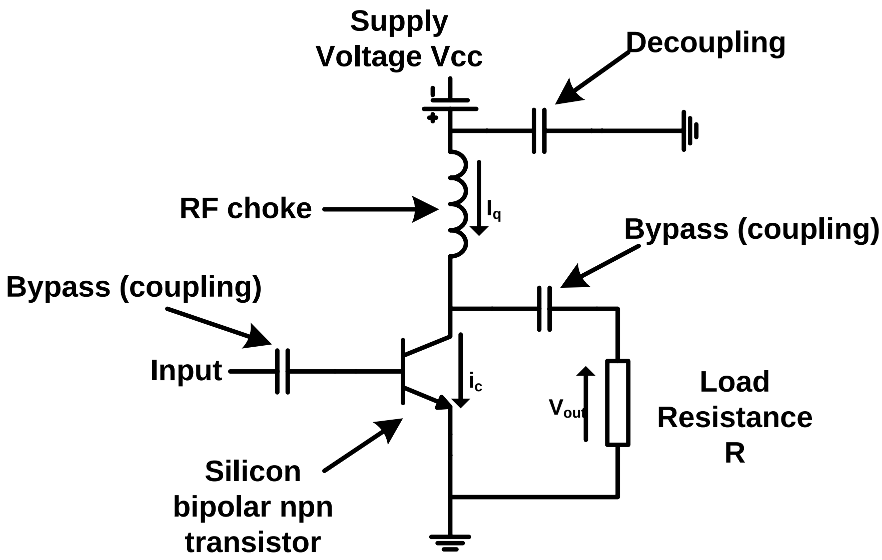

Figure 3.2a: Common emitter Class A amplifier circuit diagram.

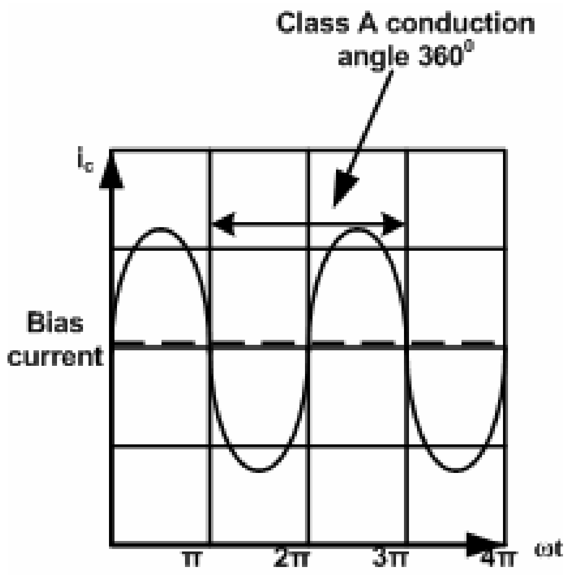

Figure 3.2b: Common emitter Class A amplifier output current.

The power consumption of a Class A amplifier is independent of the output signal amplitude and can be shown to be

$$P_{dc} = \frac{V_{cc}^2}{R} \tag{3.3}$$

The signal power is given by

$$P_{load} = \frac{V^2}{2 \cdot R} \tag{3.4}$$

where V is the maximum ac voltage flowing in the load, that is, $V_L = V\sin(\omega t)$. Therefore, the efficiency is ($V \leq V_{cc}$)

$$\eta_{ClassA} = \frac{V^2}{2 \cdot V_{cc}^2} \tag{3.5}$$

The theoretical maximum efficiency is 50% (Figure 3.3); however, in practice, the efficiency is typically less at around 30%; for signals with a high peak-to-mean ratio the efficiency becomes much lower.

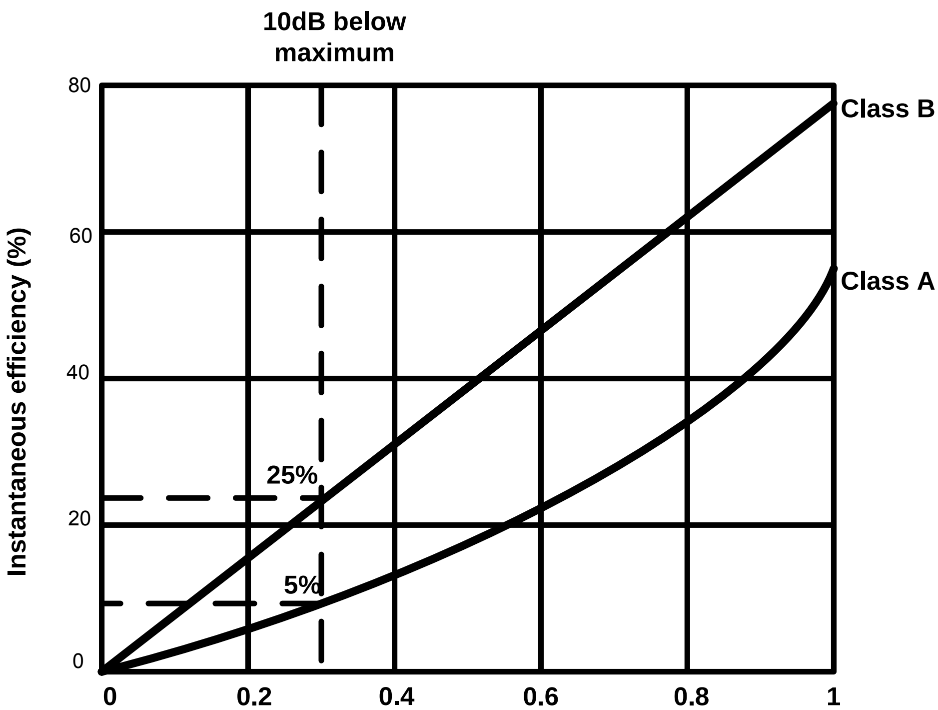

Figure 3.3: Instantaneous Efficiency.

Class A amplifiers are very useful, however, when output levels are low because of their good linearity characteristics. For example, Class A amplifiers are widely used with linearization techniques such as feedforward, which require a second or error amplifier that is very linear but has low output power [2].

3.4.2 Class B

In Class B operation (Figure 3.4a, common emitter configuration), the quiescent dc bias current is set to zero and the conduction angle is 180 degrees (Figure 3.4b). The output waveform is no longer a pure sinusoid.

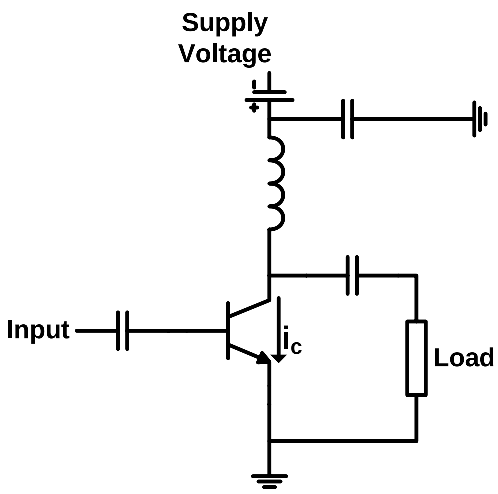

Figure 3.4a: Common emitter Class B circuit diagram

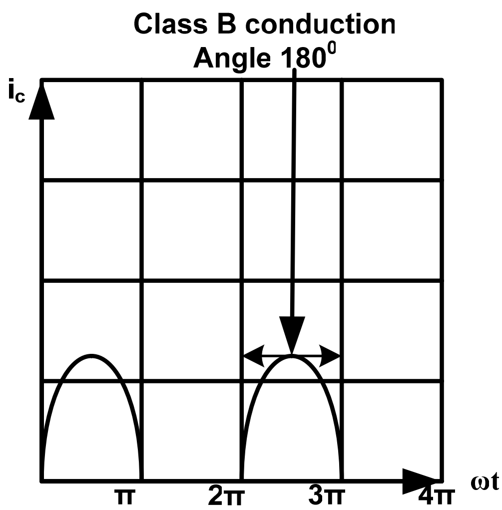

Figure 3.4b: Common emitter Class B output current waveform

A solution that allows full cycle (negative and positive cycles) of operation in Class B is to use a second transistor that conducts for the negative half-cycles of the sinusoidal input (Figure 3.5). The transistors are said to operate in push-pull mode whereby one transistor pushes or sources into the load when the input is positive and the other pulls or sinks current when the input goes negative. Note that when the input is close to zero in Class B push-pull operation, neither transistor is conducting and the output is distorted. This effect is often referred to as crossover or deadband distortion and if the goal is to make the amplifier as linear as possible, then a method must be found to reduce or completely avoid this type of distortion.

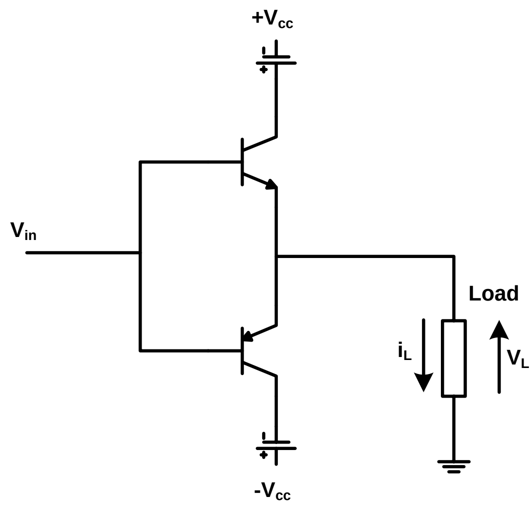

Figure 3.5: Class B push-pull configuration

The dc power consumption of a Class B amplifier is proportional to the signal amplitude V and can be shown to be

$$P_{dc} = \frac{2 \cdot V_{cc} \cdot V}{\pi \cdot R} \tag{3.6}$$

The signal power is the same as for a Class A amplifier, that is,

$$P_{load} = \frac{V^2}{2 \cdot R} \tag{3.7}$$

Therefore, the efficiency is

$$\eta_{ClassB} = \frac{\pi \cdot V}{4 \cdot V_{cc}} \tag{3.8}$$

The efficiency of a Class B amplifier is thus proportional to the output signal level and the maximum theoretical efficiency is 78% (Figure 3.3). Unlike in Class A, where the dc power consumption is constant even if there is no input signal, Class B there is zero power consumption when the input signal is zero. Class B amplifiers are less linear their Class A counterparts, but their much higher efficiency is a real advantage [2].

3.4.3 Class AB

As the same suggests, Class AB is an intermediate class between Class A and Class B. As with Class A, the transistor is biased with a nonzero dc current but the amplitude in Class AB is much less than the peak value of the output sine wave signal. Figure 3.6 shows the output waveform for Class AB operation; the conduction angle is greater than 180 degrees but much less than 360 degrees.

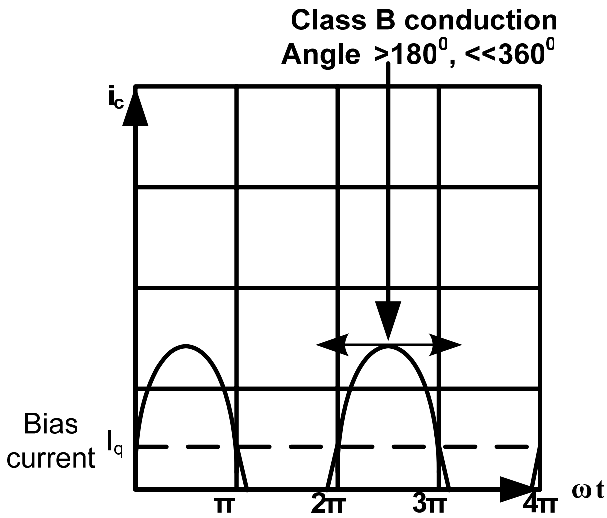

Figure 3.6: Class AB output waveform

As with Class B, Class AB stages are not usually operated at single-ended stages; instead two transistors are used, one conducts for slightly longer than the positive half-cycle of the input signal and the other for slightly longer than the negative half-cycle. When the input signal is close to zero, both transistors conduct and crossover distortion is thus virtually eliminated. The efficiency of Class AB amplifiers is very similar to that Class B (Figure 3.3) except that under quiescent conditions (no input signal), a Class AB amplifier dissipates a small amount of power.

Note that the peak current demands on a device are different for different classes of operation. For example, in Class A, the peak current is determined by the bias current and is independent of the output power. In Class B and Class AB, the peak current is a function of the output power and efficiency (a more efficient amplifier has a lower peak current for a given output power) [2].

3.4.4 Class C

The output current waveform for a Class C amplifier is shown in Figure 3.7; the conduction angle is less than 180 degrees, resulting in good efficiency but poor linearity. Class C amplifiers are good for applications that require high efficiency, but their poor linearity is a significant disadvantage in many cases [2].

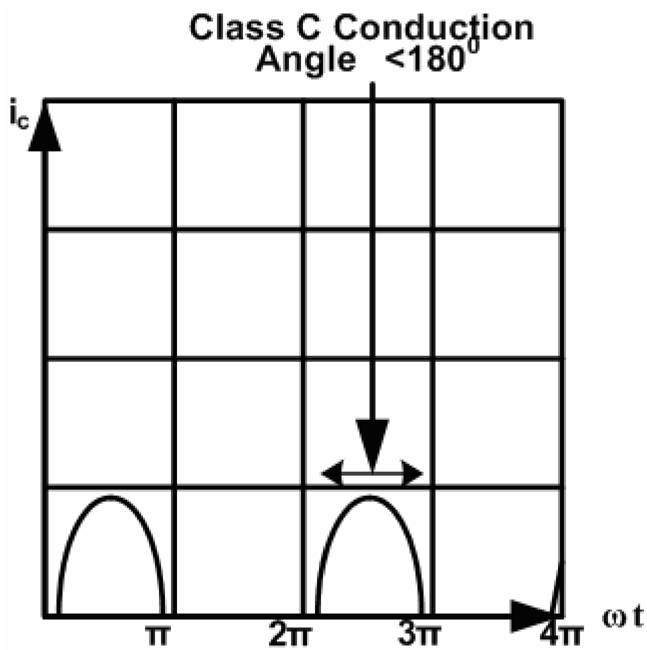

Figure 3.7: Class C output waveform

3.5 Efficiency and Peak-to-Mean Ratio

The instantaneous efficiency of an amplifier is most easily calculated assuming a CW signal. A narrowband signal can be thought of as a CW signal with a nonconstant envelope; for example, a multicarrier signal has an envelope that follows the Rayleigh distribution. If the peak voltage is unity and the peak-to-mean ratio is ψ, the amplitude (probability) function $f(v)$ is given by

$$f(v) = 2 \cdot \psi \cdot v \cdot e^{-\psi \cdot v^2} \qquad (3.9)$$

For each envelope amplitude, v $(0 \le v \le 1)$, the instantaneous efficiency can thus be calculated in the same way as the CW case.

3.5.1 Class A

For a signal having an envelope that follows the Rayleigh distribution, the average (load) power of a Class A amplifier is calculated from

$$P_{av} = \frac{1}{2 \cdot R} \int_0^1 v^2 \cdot f(v) dv \qquad (3.10)$$

A close form of the integral can be found if the upper limit of the integral is taken as infinity rather than unity. As long as the probability of any values exceeding unity very low, this method is valid; alternatively, numerical integration can be used to evaluate the equation as shown in (3.10).

Substituting for $f(v)$ from (3.9) and evaluating for $(0 \le v \le \infty)$ gives

$$P_{av} = \frac{\psi}{R} \cdot \int_0^\infty v^3 \cdot e^{-\psi \cdot v^2} dv = \frac{1}{2 \cdot \psi \cdot R} \qquad (3.11)$$

The dc power consumption of a Class A amplifier with V_{cc} normalized to unity is given by

$$P_{dc} = \frac{V_{cc}^2}{R} = \frac{1}{R} \tag{3.12}$$

Therefore, the efficiency η is

$$\eta_{Class\_A} = \frac{1}{2 \cdot \psi} \tag{3.13}$$

Figure 3.8 shows the Class A efficiency as function of the peak-to-mean ratio ψ. A signal with a 10dB peak-to-mean ratio has an efficiency η = 5% and if, for example, the average output power is 30W, then the power drawn from the dc supply is 600W (570W continuously dissipated as heat).

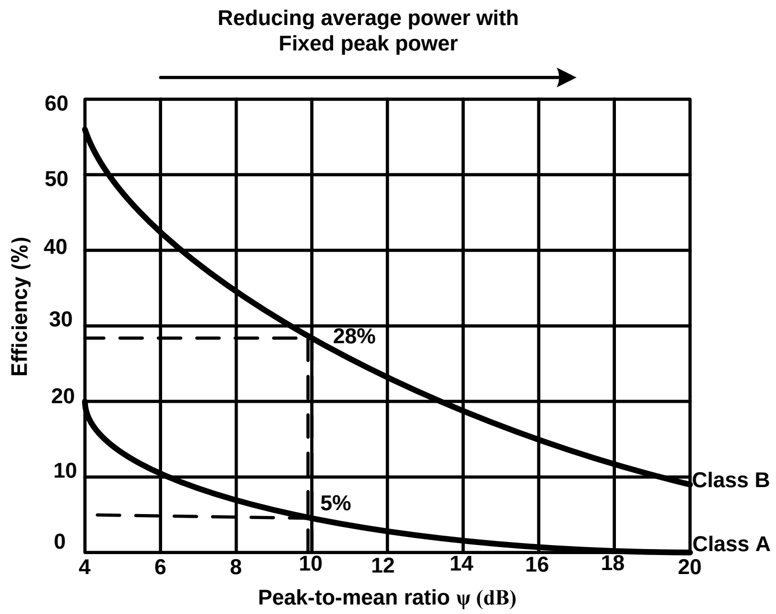

Figure 3.8: Efficiency and peak-to-mean ratio

3.5.2 Class AB

For Class AB amplifier the average signal power is the same as for Class A (3.11), however the dc power consumption in Class AB is a function of the signal amplitude. That is, with V_{cc} normalized to unity,

$$P_{dc} = \frac{2}{\pi \cdot R} \cdot \int_0^1 v \cdot f(v) dv \tag{3.14}$$

Evaluating (3.14) for ($0 \le v \le \infty$) to find a close-form solution gives

$$P_{dc} = \frac{4 \cdot \psi}{\pi \cdot R} \cdot \int_0^\infty v^2 \cdot e^{-\psi \cdot v^2} dv = \frac{1}{\pi \cdot R} \cdot \sqrt{\frac{\pi}{\psi}} \tag{3.15}$$

The efficiency is then equal to

$$\eta_{Class\_AB} = \sqrt{\frac{\pi}{4 \cdot \psi}} \tag{3.16}$$

Figure 3.8 shows for a Class AB amplifier and peak-to-mean ratio of 10dB the efficiency is 28%. Using the same example of 30W average output power, Class AB power consumption is ≈110W with 80W dissipated as heat. This is considerably less than if the amplifier was operated in Class A, but the penalty is reduced linearity.

In practice, the efficiency for both Class A and Class AB is somewhat less than the theoretical values since the output voltage saturates before the supply voltage is reached. For example, with bipolar transistors, (3.13) and (3.16) become

$$\eta_{Class\_A} = \frac{1}{2 \cdot \psi} \cdot \left(\frac{V_{cc} - V_{sat}}{V_{cc}} \right) \tag{3.17}$$

$$\eta_{Class\_AB} = \sqrt{\frac{\pi}{4 \cdot \psi}} \cdot \left(\frac{V_{cc} - V_{sat}}{V_{cc}} \right) \tag{3.18}$$

Practical values of efficiency for Class A amplifiers at maximum power are around 30% rather than the theoretical 50%, and at 10dB "back-off", the efficiency is around 3% rather than 5%. Class AB amplifiers at 10dB below maximum power have a practical efficiency of ≈15%.

The exact operating point of a transistor in terms of class of operation is a compromise between many factors, most notably linearity and efficiency. Once the mechanical design has been fixed (e.g., size, material and profile of the heatsink, maximum airflow, maximum safe operating temperature and maximum ambient temperature), the linearity is largely fixed. Increasing the bias current for better linearity (closer to Class A operation) results in lower efficiency and more heat dissipation, or alternatively lowering the bias current to improve the efficiency (closer to Class B) results in reduced linearity.

Note that in addition to the class of operation, the overall efficiency of an amplifier is affected by factors such as dielectric and conductor losses. For example, components such as power combiners and splitters have loss, printed circuit board materials have loss, and even signal tracks are lossy. A common part of the design procedure for power amplifiers is to first quantify any loss in the circuit, then attempt to minimize it, and finally ensure that the mechanical and thermal design is adequate under all conditions [2].

3.6 Compression Point and Peak Envelope Power

The out of a nonlinear amplifier compresses at high signal levels; that is, the gain drops. The output power level at which the gain has dropped by 1dB compared to the linear value is called the 1dB compression point, P_{1dB}. For Class A amplifier, which has a constant gain at lower output power, this definition is straightforward (Figure 3.9a) but Class AB the gain varies and hence the concept of compression point is more difficult to use (Figure 3.9b). A reference level, such as the gain 10 dB below maximum power, could be defined or alternatively the rated PEP could be specified.

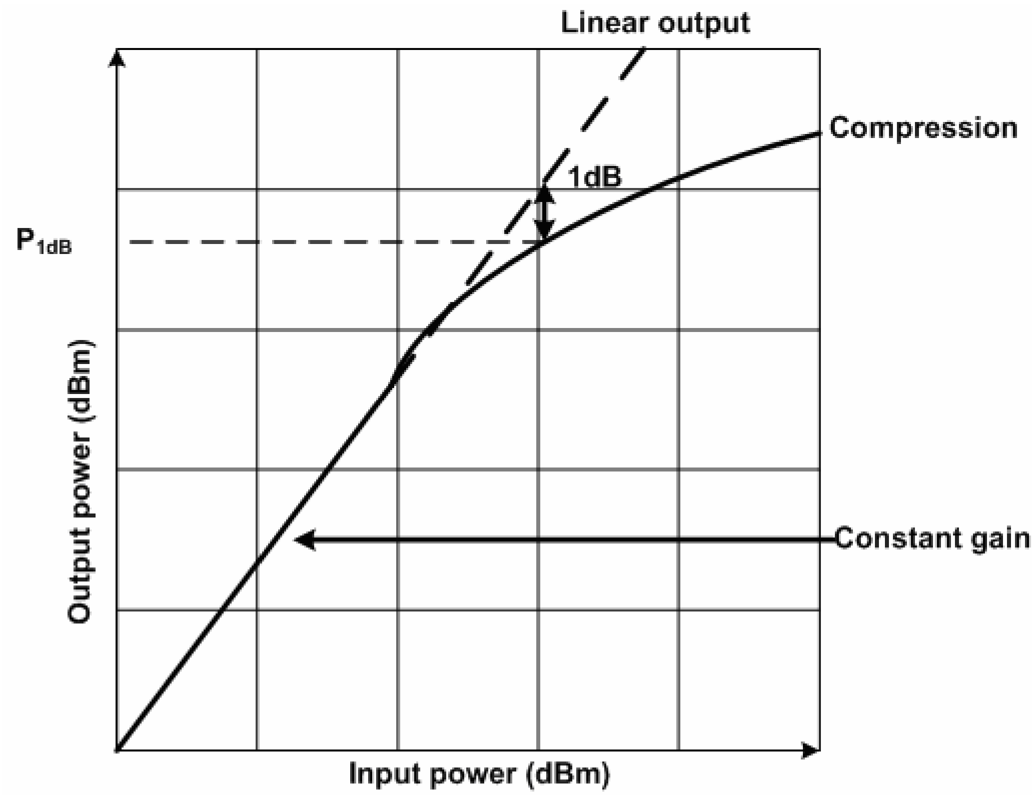

Figure 3.9a: Bipolar 1dB compression point example – Class A

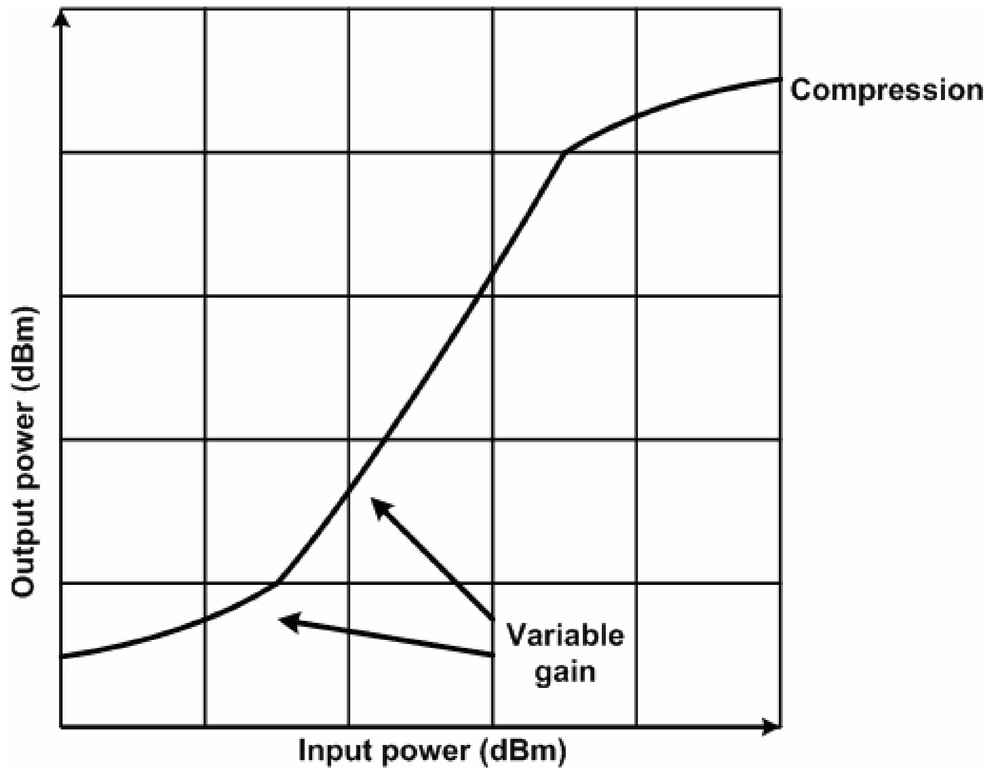

Figure 3.9b: Bipolar 1dB compression point example – Class AB

The rated PEP of an amplifier is typically defined in terms of the maximum envelope power for a given linearity; it is common practice to specify the linearity level as –30dBc for a two-tone signal. According to this definition, the rated PEP of an amplifier is the maximum envelope power of a two-tone signal for which the amplifier intermodulation level is –30dBc.

In practice, the 1dB compression point and rated PEP of an amplifier are approximately equal, and hence either PEP or P_{1dB} can be used as a figure of merit. Note however, that in general, a transistor used in Class B or Class AB has higher compression point than the same transistor used in Class A. For example, in silicon devices the compression point in Class A is related to the maximum current flowing through the device, which in turn is related to the physical size of the semiconductor material. A silicon device, for example, may have a PEP of 30W in Class AB, but in Class A the corresponding figure may be only 10W to 15W.

3.7 Factors Affecting Choice of Transistor

Currently, LDMOS appears to be the best choice for Class AB amplifiers while Gallium Arsenide seems preferable for Class A; however, the final choice of transistor for a particular application will depend upon many factors. First, there are performance criteria, primarily:

- Frequency of operation,
- Average power output,
- Efficiency,
- Linearity (intermodulation performance),
- Peak-power requirement (signal peak-to-mean ratio).

There is a whole host of other factors to consider when choosing a transistor, for example:

- Cost (component cost for prototype and volume production),
- Availability (number of suppliers, lead times, quantity),

- DC power requirements (availability of power modules, current capacity),
- Previous experience,
- Process reliability (consistency from one device to another – RF and dc performance),
- Device reliability,
- Production time,
- Ruggedness (e.g., handling in production, maximum VSWR, overdrive capability),
- Maximum operating temperature and cooling considerations,
- Physical size and packaging.

In practice, several different kinds of transistors are used in the same amplifier since input driver and output stages all have varying requirements in terms of power-handling capability [2].

4. LINEARIZATION TECHNIQUES

A number of linearization techniques exists; feedback, predistortion and feedforward can be used (separately, or sometimes in combination) to linearize an inherently nonlinear amplifier. It is also possible to generate a linear signal using the synthesis of other nonlinear signals, for example using techniques such as RF synthesis and envelope elimination and restoration.

An important property of all techniques is their linearization bandwidth, which often described as being narrowband or wideband. The actual definition of wideband and narrowband is open to interpretation and depends upon the specific application. For example, a single carrier with a 5MHz channel bandwidth generally regarded as a "wideband" signal, whereas a single sideband signal (5kHz channel bandwidth) is regarded as narrowband. In general, however, whether one particular technique or signal is narrowband or wideband is a question of definition; for example RF feedback, which can have a linearization bandwidth of several megahertz, may be regarded as either wideband or narrowband [2].

4.1 Feedback

In the late 1920s, an electronics engineer named Harold Black proposed using feedback as a useful circuit function and the feedback amplifier, which has since become a fundamental building block in modern electronic circuits, came into existence. A few earlier Harold Black also proposed a technique called Feedforward and received a patent relating to it in 1928. Feedforward, however, was largely ignored until alternative linearization methods were required for amplifiers with high delay where stability considerations precluded the use of feedback.

4.1.1 Principle of Operation

Two form of feedback exists: positive feedback and negative feedback. Positive feedback in amplifier is undesirable because the amplifier response can become oscillatory rather than stable; therefore, when referring to amplifiers, it is generally assumed that feedback is negative. The terms feedback amplifier and negative feedback amplifier thus become interchangeable. In essence, negative feedback allows a designer to trade gain for some other desirable property, for example, reduced nonlinear distortion or increased bandwidth. Negative feedback can also be used to control input and output impedances, reduce the effects of noise and make the gain of an amplifier less sensitive to variations in circuit components, for example, due to temperature effects. A limitation of negative feedback is that under certain conditions the feedback can become positive and be of sufficient magnitude to cause oscillations; there is thus stability criteria associated with a (negative) feedback amplifier.

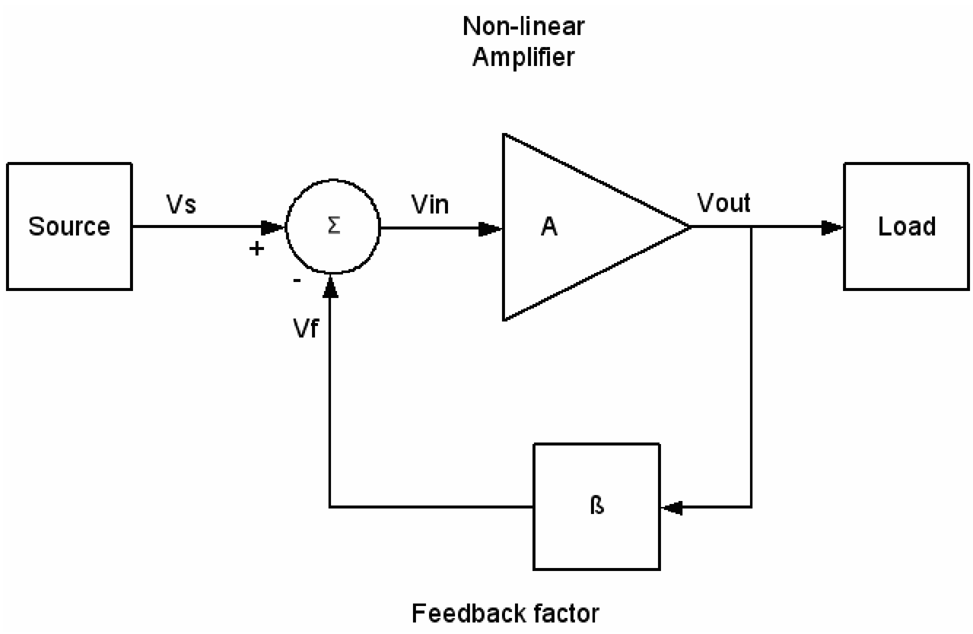

Figure 4.1: Feedback components

Figure 4.1 shows principle of a feedback amplifier; the source and load are assumed to be perfect and do not affect the open loop gain, ‘A’, of the amplifier in any way. With no feedback applied, the open-loop transfer function (or the simply the amplifier gain) is given by

$$A = \frac{V_{out}}{V_{in}} \tag{4.1}$$

In feedback, that is, closed-loop configuration, the output signal V_{out} is reintroduced (fed back) at the input after scaling by a factor β, the feedback factor, that is

$$V_f = \beta \cdot V_{out} \tag{4.2}$$

The feedback signal V_f is subtracted from the source signal V_s, generating a difference signal V_{in} that becomes the input signal to the basic amplifier.

$$V_{in} = V_s - V_f \tag{4.3}$$

Substituting for V_f from (4.2) gives

$$V_{in} = V_s - \beta \cdot V_{out} \tag{4.4}$$

The transfer function of the amplifier with feedback, that is, the gain V_{out}/V_s, is obtained by combining (4.1) to (4.3) to give

$$A_f = \frac{V_{out}}{V_s} = \frac{A}{1 + A\beta} \tag{4.5}$$

Note that it is the subtraction of the signals (the negative sign in (4.3)) that makes the feedback negative; negative feedback thus always acts to reduce the signal at the input to the amplifier. The feedback remains negative as long as the feedback gain, Aβ, is a positive quantity; that is, V_f and V_s have the same sign. If loop gain becomes negative for some reason, then feedback becomes positive and oscillation may occur.

In general, Aβ >> 1 and (4.5) becomes

$$A_f = \frac{1}{\beta} \tag{4.6}$$

That is, the gain of a feedback amplifier almost independent of the open loop gain and depends only on the feedback network, which can be chosen with a high degree of accuracy and implemented with linear passive elements. The penalty for better linearity is that gain of the feedback amplifier is reduced by the feedback factor.

4.2 RF Synthesis

Linear amplification with nonlinear components (LINC) and combined analog locked loop universal modulator (CALLUM) are example of narrow band linearization schemes that RF synthesis techniques.

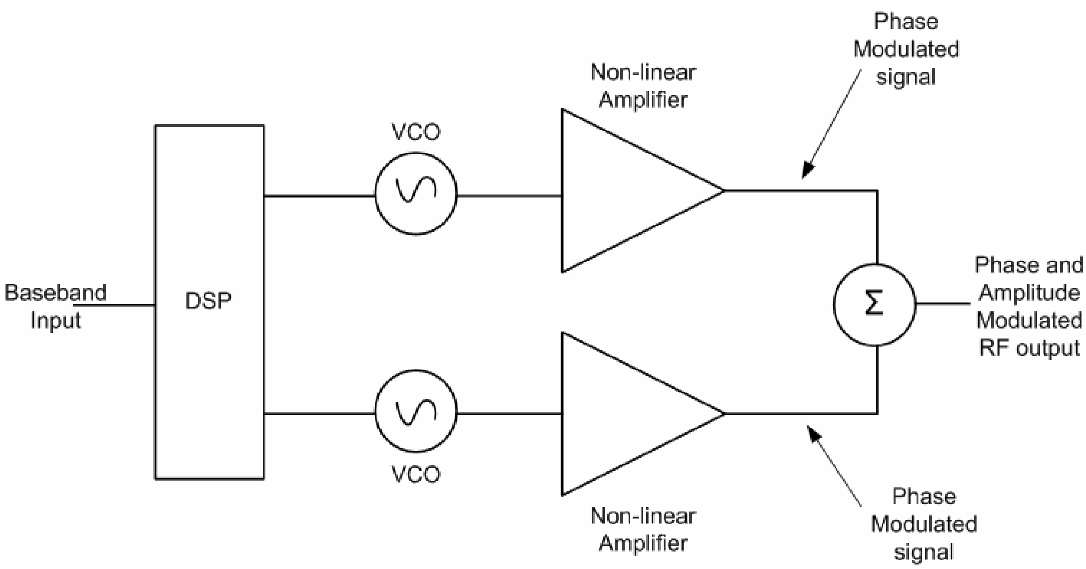

Figure 4.2: Linear amplification using nonlinear components (LINC).

Figure 4.2 shows principle of a LINC transmitter whereby voltage controlled oscillators (VCOs) are used to produce two phase modulated signals. The output signal amplitude (after combining the amplifier outputs) is a function of the phase in the two signal paths and can vary from zero (opposite phase) to maximum (phase alignment). Disadvantages with the LINC approach include the fact that it is an open-loop system and that the two signal paths must be very accurately matched. The bandwidth of the phase modulated signals can also be large and the vector can also be problem. CALLUM solves some of these problems by adding feedback, however, neither LINC or CALLUM have yet been commercially proven.

4.3 Envelope Elimination and Restoration

Another narrowband synthesis method is envelope elimination and restoration, also referred to as the Kahn method. Figure 4.3 shows the principle of operation whereby the input signal is first split into two paths, one going to a limiter, the other to an envelope detector. The limiter removes the amplitude modulation component of the input signal, leaving a constant envelope phase modulated signal. The output of the envelope detector is an amplitude modulated signal, that is, a nonconstant envelope signal. An efficient but nonlinear RF amplifier is used to amplify the phase modulated signal, while the amplitude modulated signal is amplified using a low-frequency amplifier. The amplified, amplitude modulated signal is the used to remodulate the amplified phase modulated signal. The envelope wave shape of the high-power output signal after the modulator stage is thus the same as that of the input signal, as is the phase modulation component. Note that to ensure linear output, the time relationships between amplitude and phase modulation signals must also be properly maintained.

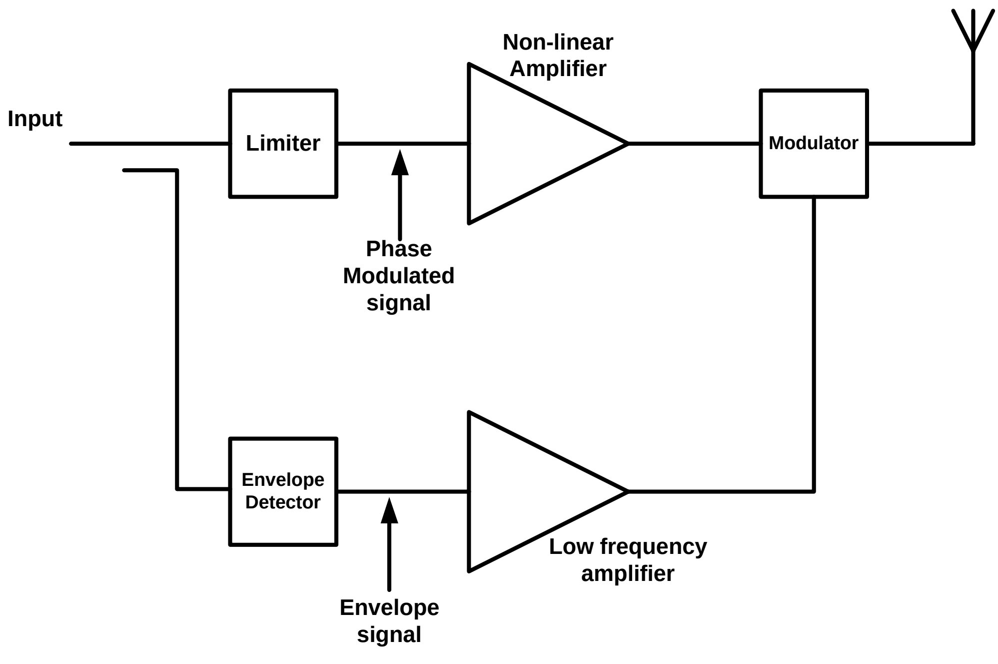

Figure 4.3: Envelope elimination and restoration.

4.4 Predistortion

The nonlinear output of an amplifier can be represented by a polynomial, that is,

$$V_{out} = G_1 \cdot V_{in} + G_2 V_{in}^2 + \cdots + G_n \cdot V_{in}^n \qquad (4.12)$$

The amount of AM/AM and AM/PM distortion introduced by the amplifier is a function of the signal level and relative contributions of the amplifier coefficients $G_1 \ldots G_n$. If the coefficients are known (e.g., from measurement and/or simulation), then the amplifier distortion can be compensated by introducing a nonlinearity, which when cascaded with the amplifier provides linear gain; this is the principle of predistortion. Figure 4.4 shows a typical implementation whereby a predistortion circuit operating at low power introduces nonlinearity before amplifier; the optimal nonlinearity is the inverse of the amplifier transfer characteristic.

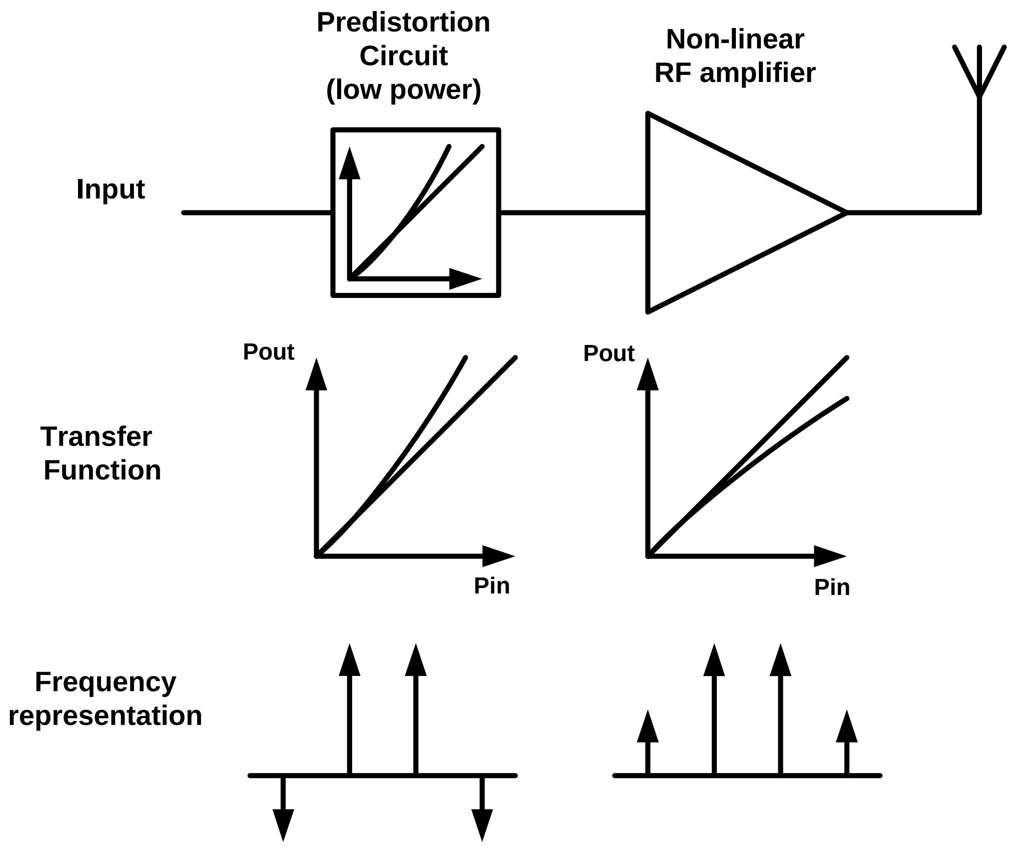

Figure 4.4: Predistortion.

The predistortion function can be implemented at baseband (e.g. adaptive baseband predistortion) or at IF/RF. Predistortion can correct for both AM/AM and AM/PM distortion and is not restricted in bandwidth since there is no inherent feedback path (note, however, that baseband and DSP predistortion are bandwidth limited). Fixed or adaptive predistortion schemes can be used, with the latter being able to compensate for changes in the amplifier characteristics over time, for example, due to temperature.

Disadvantages of predistortion include the fact that predistorters are normally optimized for a specific power level and that can typically only provide limited reduction in distortion, normally only third-order distortion products. Unlike many other linearization techniques, however, predistortion does not significantly reduce the efficiency of an amplifier. For example, when used in combination with feedforward linearization, even modest improvements in third-order distortion levels can lead to improved overall system efficiency.

4.5 Feedforward

Feedback compares the output of a nonlinear amplifier with its input and uses the same amplifier to amplify the difference signal; in contrast, feedforward uses two amplifiers and there is a continuous forward signal flow, that is, no feedback path. The lack of an inherent feedback path means that feedforward is unconditionally stable and allows operation over theoretically unrestricted bandwidth. Feedforward is thus classed as a wideband linearization technique unlike feedback, which inherently more narrowband.

4.5.1 Principle of Operation

Figure 4.5 shows a simplified representation of a feedforward amplifier. The input is first split into two paths by a power splitter (usually a directional coupler) with one path going to the nonlinear main amplifier and the other going to a delay element. A portion of the distorted main amplifier output is separated from the main amplifier path using a second coupler as a power divider and after appropriate scaling subtracted from the delayed feedforward input. The resulting error signal, ideally containing only distortion components, is then amplified by an error amplifier before being subtracted from a delayed version of the main amplifier output, thus canceling the distortion components in the main path. For example, Figure 4.6b illustrates how the distortion that is added to a two-tone signal by a nonlinear amplifier is canceled using feedforward; the distortion is isolated in the first feedforward loop (Loop1) and canceled in the second loop (Loop 2). Figures 4.6 (a, b) show the frequency spectrum before and after cancellation.

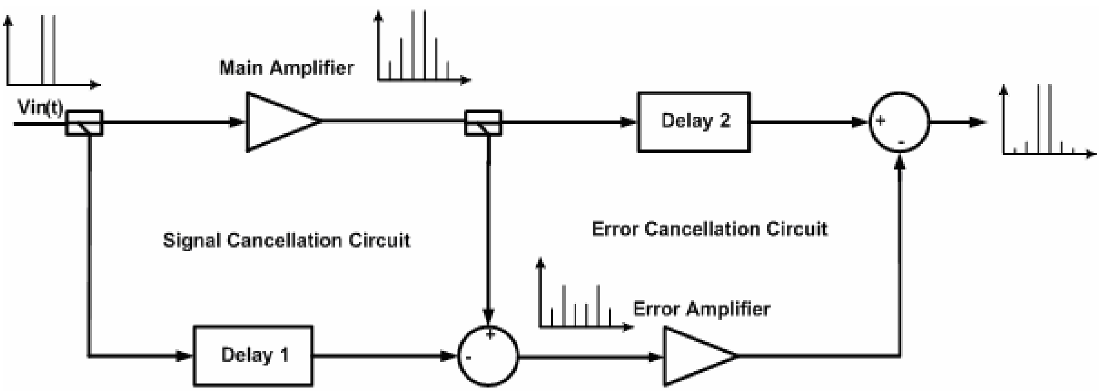

Figure 4.5: Feedforward components

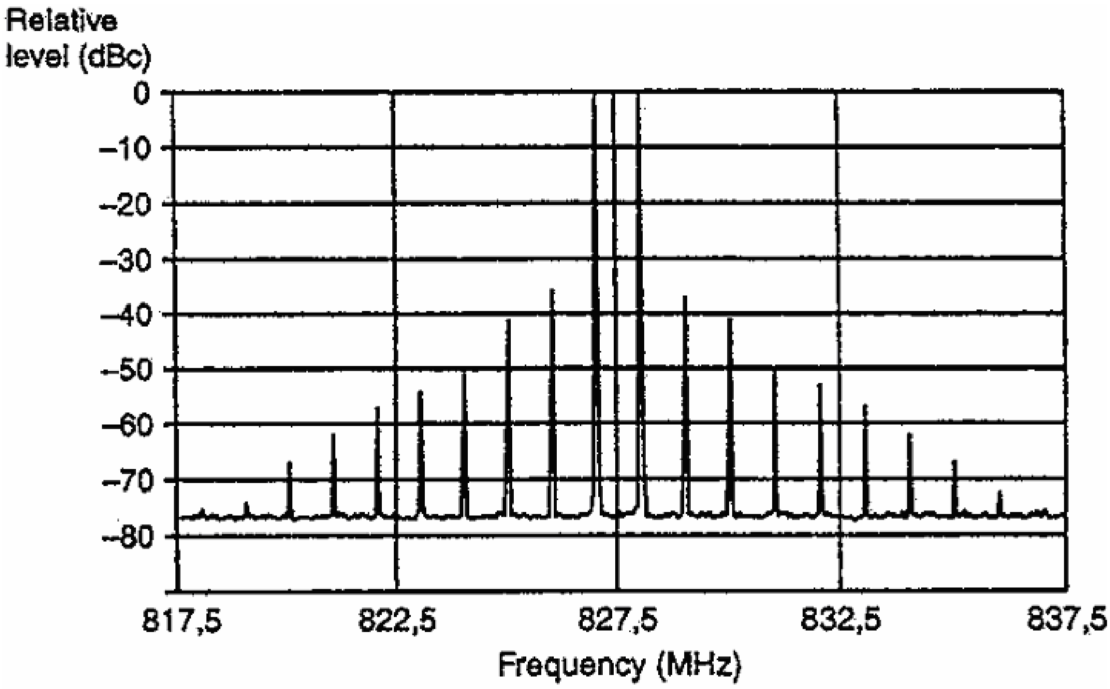

Figure 4.6a: Two-tone distortion before feedforward correction

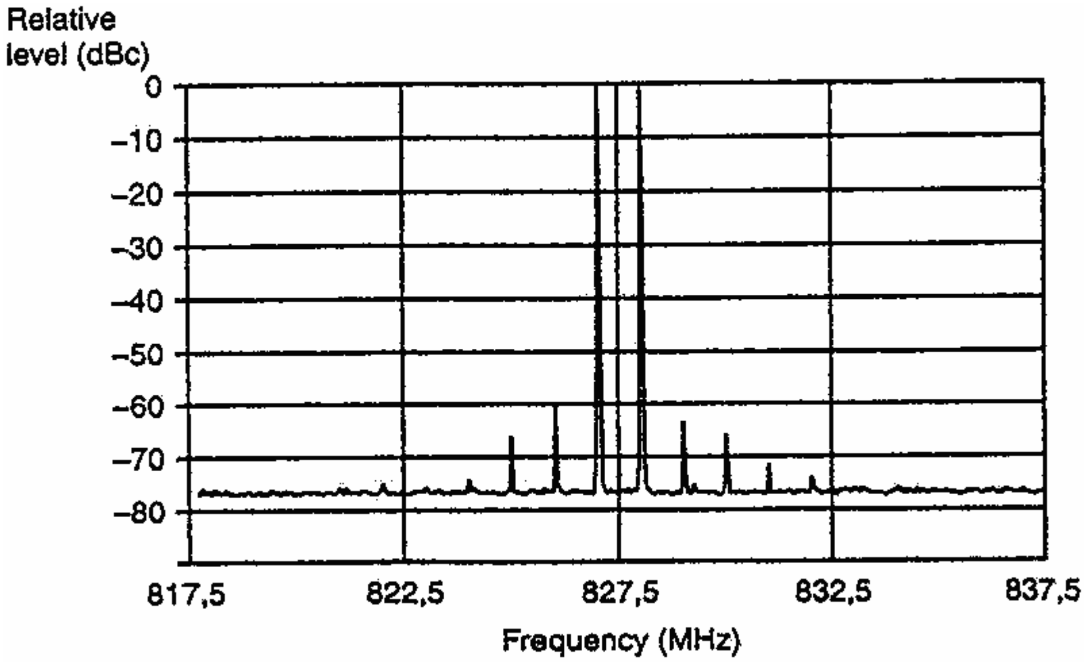

Figure 4.6b: Two-tone distortion after feedforward correction

4.5.2 Input-Output Signal Linearity

In practice, any feedforward input signal (transceiver output signal) has a finite level of distortion and a feedforward amplifier cannot reduce this level, but only maintain or increase it. The output of a feedforward amplifier is therefore only completely distortion-free when the input signal contains no distortion (an ideal case and not possible in practice). For example, if a particular input signal has a distortion level of –60dBc, then the output signal has a linearity not better than –60dBc.

An ideal feedforward amplifier does not add any distortion to an input signal, it is a true linear amplifier. In practice, however, feedforward amplifiers inevitably introduce some distortion and strictly speaking, the output is no longer linear. Terms such as “very-linear” and “ultra-linear” are therefore sometimes used to emphasize that the practical linearity of a particular amplifier is close to but not equal to ideal.

4.5.3 Multicarrier Input and Noise Performance

As previously noted, for an input signal that consists of many carriers with uniform carrier spacing Δf, the intermodulation products also appear on a Δf grid and cannot be removed by filtering. Removing distortion components that appear at center frequencies does not, however, present a problem in feedforward since the distortion is isolated in the first loop and canceled at the output of the second. The carriers at the output are unaffected as long as Loop 1 cancellation is maintained.

Another very useful property of feedforward is that in addition to canceling intermodulation distortion, the second loop also cancels distortion in the form of noise added in the main amplifier path. Feedforward amplifiers with their attendant low-noise figure are therefore particularly suited to applications where signal levels are low, such as cable systems. Indeed one of the first large-scale commercial applications for feedforward was cable TV systems.

4.5.4 Signal Cancellation

Feedforward depends upon the successful isolation of an error signal and the removal of distortion components, both of which involve signal cancellation over a band of frequencies. Mathematically, signal cancellation at a single frequency is represented by the subtraction of two signals with equal amplitude; the resultant has a magnitude equal to zero or $-\infty$ dB. In practice, cancellation is achieved by the vector addition of signals, or more specifically voltages, with equal amplitude but opposite phase.

Perfect broadband cancellation, that is, vector cancellation over a band of frequencies, occurs only when signals have:

- Equal amplitude;
- 180 degree phase difference;
- Equal delay

Equal amplitude and opposite phase are sufficient for cancellation at a single frequency, however, the requirement of equal delay is necessary for broadband signal cancellation.

4.5.5 Gain and Phase Adjustment

The gain and phase through the various feedforward signal paths (Figure 4.5) is, in general, a function of many parameters, for example, signal level and temperature; therefore, some form of gain phase adjustment is required to ensure that the signals are always matched.

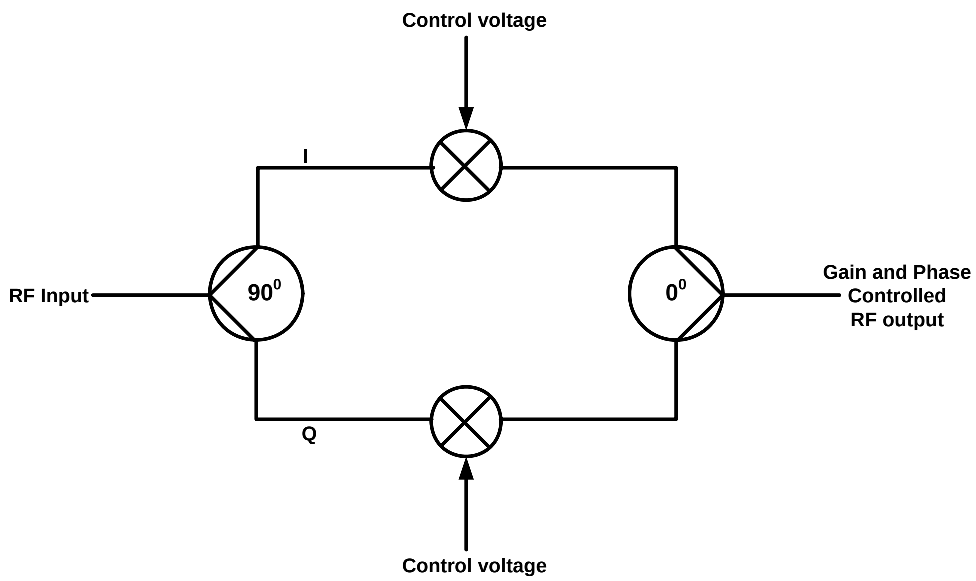

Figure 4.7: Gain/phase control network – IQ modulator example.

Gain and phase adjustment of an RF signal can be done in several ways, including using separate amplitude and phase control networks or IQ networks; the circuits themselves can be implemented as discrete components. Figure 4.7, for example, shows an IQ control circuit that uses two dc voltages to control the gain and phase of the RF signal. Table 4.1 shows mathematical representation of how IQ control circuit adjusts gain and phase. '|a|' changes amplitude of RF input signal and 'θ' changes phase of RF input signal. This circuit is only for simulation purposes and all components in this circuit are assumed ideal. Important parameters of such gain and phase adjustment networks include:

- The dynamic range over which the amplitude can be varied;
- The dynamic range over which the phase can be varied;
- Linearity and power-handling capability;

- Noise figure;
- Frequency response (amplitude and phase ripple);
- Physical size and cost.

Table 4.1: Mathematical Representation of IQ Control Circuit

a_I	: In phase control voltage (DC)
a_Q	: Quad phase control voltage (DC)
$\lvert a \rvert = \left\lvert \sqrt{a_I^2 + a_Q^2} \right\rvert$	: Amplitude control component
$\theta = \tan^{-1}\left(-\dfrac{a_Q}{a_I} \right)$	: Phase control component

For example, a certain gain phase adjustment network may have a dynamic range of 20dB with 360-degree phase control, a 1dB compression point power of 10dBm, and a noise figure of 10dB. Another network may only have 90-degree phase control, a compression point power of 0dBm, and a noise figure of 20dB. An example study in [11] is presented for phase and amplitude control.

The position of gain and phase control networks in feedforward amplifier is also important.

4.5.6 General Properties and Advantages of Feedforward

Some of the advantages of feedforward as an amplifier linearization technique are detailed below.

1. Feedforward correction does not (ideally) reduce amplifier gain. This is in contrast to feedback systems in which linearity achieved at the expense of gain.
2. Gain-bandwidth is conserved within the band of interest. This is again in contrast to feedback systems, which often require very wide feedback bandwidths in order to provide the required levels of correction.
3. Correction is independent of the magnitude of the amplifier delays within the system. A high-gain RF amplifier will often have a significant group delay and

this is potentially disastrous for any form of feedback system due to the large potential for instability.

4. Correction is not attempted based on past events (unlike feedback). The correction process is based on what is currently happening rather than what has happened in the recent past.
5. The basic feedforward configuration is unconditionally stable. This is one of the most important advantages and follows from the points raised above.
6. Cost is the main limiting factor to the number of stages (or loops) and hence the level of correction which may be achieved, although size and efficiency may also be important in some applications. In other words, an arbitrarily high level of correction is possible, as there is no theoretical limitation on the number of times which feedforward correction may be applied. In an ideal system shown in 4.5, however, in reality the error amplifier itself will distort the error signal and this will appear directly at the output. Gain and phase matching throughout the system also affect the performance.
7. The error amplifier, ideally, need only process the main amplifier distortion information and hence can be of a much lower power than the main amplifier. Thus it is likely that a more linear and lower noise error amplifier can be constructed. This is in turn will result in a lower overall system noise figure.

Fault tolerance: In a single loop feedforward system, the failure of either amplifier will result in a degradation of performance and possibly a lowering of the final output power; however, the system will not fail together. In the case of a feedback system there is only one forward-path amplifier and if this fails then the whole system has failed [1].

5. ADAPTIVE FEEDFORWARD SYSTEMS

5.1 Need for Adaptation

The idea of using additional circuitry to allow the feedforward system to monitor, and correct for its own performance has received considerable interest. There are a number of configurations to fulfill this function ([3…10]), as it is arguably the most elegant solution to the problem.

The idea of allowing the system to monitor its own performance and then perform the necessary correction implies some form of feedback system around the feedforward loop. This feedback system is required to control the gain and phase matching of the two halves of the feedforward loop, termed the error loop and the correction loop for convenience. Thus, it is evident that two separate feedback systems are required: one to correct for the gain and phase mismatches in the error loop in order to minimize the amount of input signal component of the error signal, and the second to correct for the gain and phase mismatches in the correction loop in order to minimize the amount of distortion present in the final output signal. These two loops are shown conceptually in Figure 5.1.

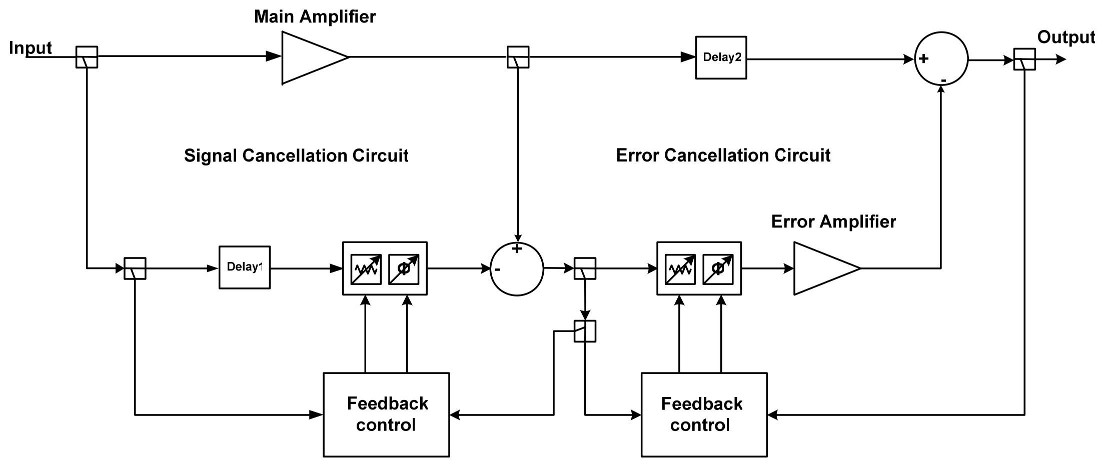

Figure 5.1: Feedback control applied to a feedforward amplifier

The gain and phase adjustment components (compensation circuits) shown in Figure 5.1 could appear in a number of different locations around the two halves of the feedforward loop. For example, these components in the error loop could be placed in the lower path (along with a number of other positions as shown in Figure 5.2). Ultimately, the decision regarding the placement of these components rests with the system designer; however, a number of practical points must be considered.

1. Power handling. Since feedforward linearization is most often applied to amplifiers having a significant power output, the power handling capabilities of the gain and phase compensation components must be considered. A common method of construction of these components involves the use of 3dB quadrature hybrid couplers together with either PIN of varactor diodes respectively, however, both of these devices will only operate al low signal levels. It is thus necessary to place them in small-signal parts of the system, that is, at the inputs to rather than the outputs from either of the amplifiers.
2. Linearity. The linearity of the adjustment components is critical to the overall system performance since any distortion generated by either will translate to the output. In the case of the correction loop, distortion from these components will be amplified by the error amplifier (assuming that they are placed before it) and appear directly in the output signal. In the case of error loop, if they are placed in the time delay path then their distortion will appear in the error signal and be amplified by the error amplifier as above. If, however, they are placed before the main amplifier, then their distortion will appear as part of that from the main amplifier and it will be corrected as such at the final output. For this reason, a position prior to the main amplifier is usually considered to be optimum for these components.

It is also possible to locate the adjustment components for error loop between the sampling coupler and the subtracter (with suitable fixed attenuation as necessary). The signal levels at this point will in general be small and will be amplified and translated to the output signal. A summary of the possible locations of the adjustment components is shown in Figure 5.2.

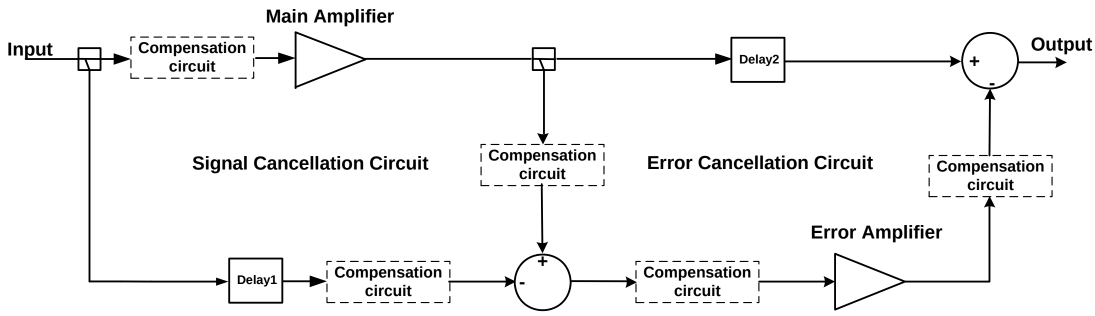

Figure 5.2: Summary of the available locations for the gain and phase adjustment components in the error compensation loops.

5.2 Adaptation Techniques

5.2.1 Adaptation with Pilot Signal (Carrier Injection)

One of the main compensation techniques involves the injection of an additional carrier (or carriers) into the error loop immediately prior to the main amplifier (although anywhere in that path will produce similar results). The level of this signal is usually significantly less than the level of the main input signal(s); 10dB or 20dB being typical.

The injected carrier will be amplified by the amplifier and will pass to the output coupler via the time-delay element in the usual manner. A sample of the carrier will also be present in the error signal and will be amplified by the error amplifier before being fed to the output coupler in anti-phase to its main-path counterpart. If the compensation loop is adjusted correctly, then this injected signal will be cancelled at the final output, and hence the compensation circuitry must aim to minimize its level in the output signal.

This may be achieved by the use of a narrow-band receiver tuned to the injected carrier frequency and hence receiving the residual carrier level. Some form of intelligent controller can then monitor the output of the narrowband receiver and adjust the gain and phase components to minimize the residual carrier level. The configuration of this control strategy is shown in Figure 5.3.

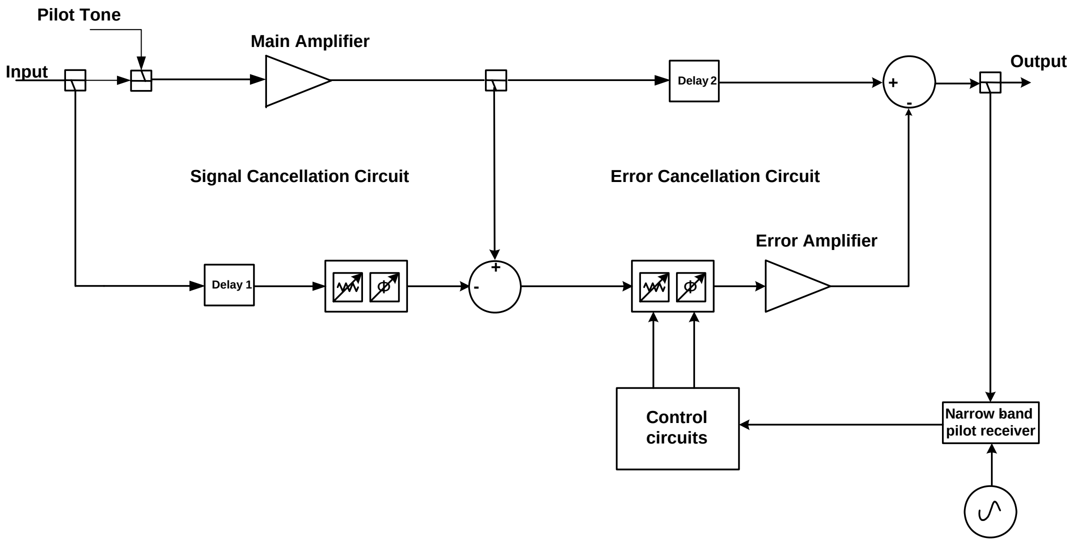

Figure 5.3: Compensation of a feedforward amplifier using pilot-injection technique

There are a number of disadvantages with this technique. It does not help in any way with the elimination of the wanted signals from the error signal. This must still be done manually with the attendant problem of drift over time. The error amplifier must therefore be made more powerful than is strictly-necessary for the required for the required performance (or a different form of control must be used for this loop).

The injected carrier will intermodulate with the wanted signals in the main amplifier create additional (unnecessary) intermodulation products. These should be eliminated from the final output by the action of the feedforward loop, but must still be amplified by the error amplifier, again adding to its power rating (usually negligible).

Incomplete removal of the pilot is also a problem in some systems and an out of band pilot plus additional output filtering is sometimes employed. This is clearly an expensive and inefficient technique and also potentially erroneous as it relies on the correlation between an out-of-band pilot and amplifier performance within the wanted band. This correlation will, at best, be limited [1].

5.2.2 Power Minimization Method

The most straightforward method of assessing the level of the tones within the error signal is simply to detect the overall energy of the complete signal. In general, the level of the distortion product is small relative to the wanted signals, and thus the wanted signal energy will dominate. Thus, a system can be envisaged in which the overall energy of the error signal is minimized by automatic adjustment of the gain and phase components using voltage control.

The use energy minimization in both parts of a feedforward correction loop is illustrated in Figure 5.4. The detector could be any form of broadband energy detector; a simple example being an envelope detector.

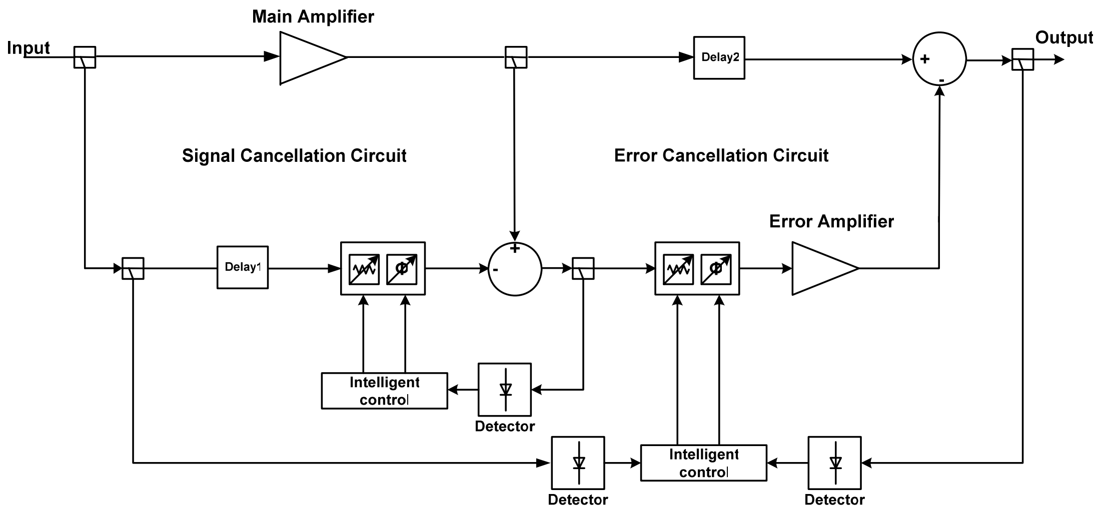

Figure 5.4: Compensation of a feedforward amplifier using energy minimization technique

This somewhat crude approach may well be all that is required for the error signal in some systems, as large amounts of rejection of the remaining input signal energy below the level of the intermodulation products is not usually required. Once this level is low enough such that error amplifier power is dominated by the intermodulation products this is generally acceptable. If the error amplifier specification can be relaxed still a little, such that it can be more powerful than is strictly necessary, then the input signal rejection specification can be relaxed still further. The ultimate consideration now becomes the cancellation of the wanted signal energy in the final output. It may be acceptable to sacrifice a few tenths of a dB of output power to the cause of reduced system complexity.

In the case of the correction loop, two detectors are required. An output detector is required to ascertain the level of the signal-plus-distortion present at the output and input detector necessary to provide an accurate indication of the input power level, and more importantly, any changes in that level (e.g., due to power control variations). Whilst it would be possible to utilize an output detector alone, the extremely long time-constant required in that detector (due to the extremely small changes in output level resulting from distortion level changes) mean that it could easily be fooled by even small changes in input (and hence output) level.

Utilizing this technique for the correction loop is usually unsatisfactory for the reason outlined above. The wanted signal energy present at the output will be very large (hopefully) with respect to the remaining distortion and any changes in this distortion level will thus have an almost negligible effect on the output signal energy. As a result the problem of detecting these small changes in energy becomes almost impossible and an alternative solution is required.

One possibility is that of generating a second ‘error signal’ from the output signal. The main signal energy should be substantially cancelled in this case and hence energy detection could yield a more realistic result [1].

5.2.3 Gradient Adaptation (Coherent Detection)

The obvious solution to the problems inherent in broadband energy detection systems is to employ some form of coherent or correlation process instead. Thus in the case of the error loop, the error signal could be correlated with original input signals to generate a suitable feedback error signal to control the gain and phase components. This is the basic approach, which many of the patented schemes employ.

The basic configuration of this approach is shown in Figure 5.5. The correction loop correlation process utilizes the final output signal and the error signal and this process assumes that the rejection of the original signal components from the error signal is very good. It is necessary to ensure good rejection of the original signal components in order to subsequently ensure that the result of the correlation process is influenced only by the distortion components and not by the wanted signals.

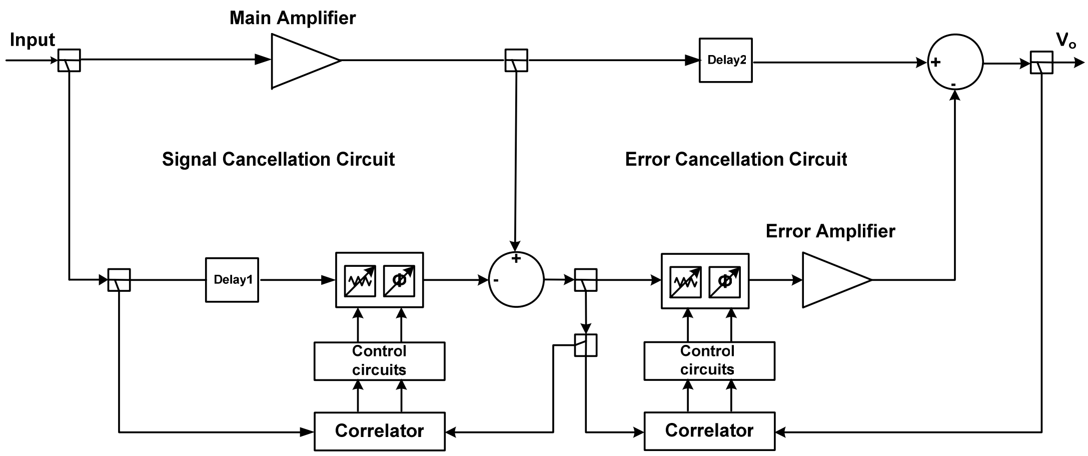

Figure 5.5: Compensation of a feedforward amplifier using correlation technique.

A number of other systems work on a similar principle but utilize different positions for their gain and phase adjusting circuits, along with techniques for reducing the tight constraints on the level of wanted signal components present in the error signal.

5.3 Mathematical Representation of Adaptation for Coherent Detection

5.3.1 Mathematical Representation of Feedforward Technique

Figure 5.6 shows a feedforward model for mathematical representation of feedforward technique.

$$V_{in} = ce^{j(\omega t+\phi)} \tag{5.1}$$

$$V_a = G_1 V_{in} + v_{IMD} \tag{5.2}$$

$G_1 = |G_1| e^{j\alpha}$: The linear gain of PA.

$G_1 v_{in}$: Amplified version of input signal plus a certain phase shift α.

v_{IMD} : Intermodulation products.

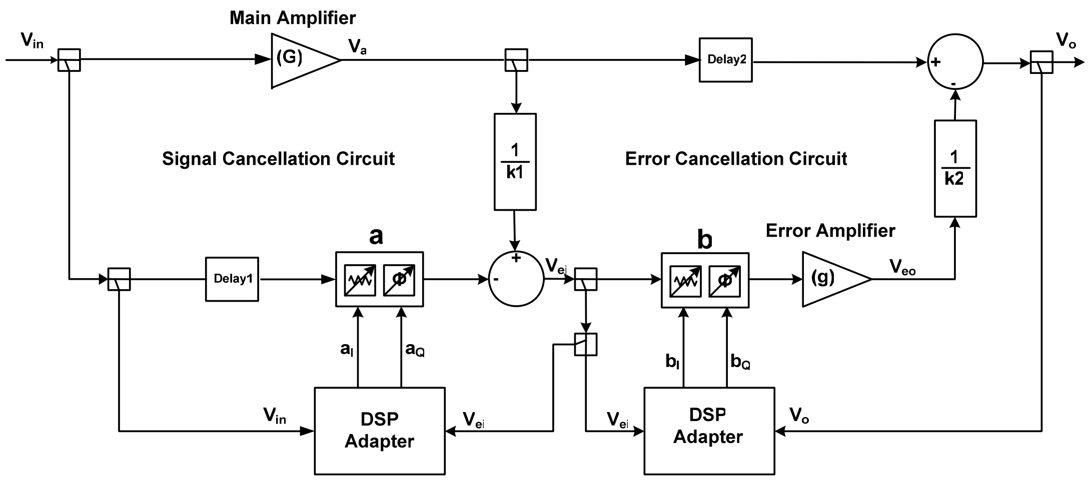

Figure 5.6: A feedforward model for mathematical representation of feedforward technique.

In the signal cancellation circuit a fraction of the PA output signal v_a/k_1, with k_1 (coupling coefficient of first loop) real, is compared with a sample of the input av_{in}, with 'a' (adaptation coefficient of first loop) complex, resulting in an error signal v_{ei} given by

$$v_{ei} = \frac{G_1}{k_1} v_{in} + \frac{1}{k_1} v_{IMD} - a v_{in} \tag{5.3}$$

If 'a' is adjusted in such a way that

$$a = \frac{G_1}{k_1} \tag{5.4}$$

The error signal v_{ei} will contain only the intermodulation products.

$$v_{ei} = \frac{1}{k_1} v_{IMD} \tag{5.5}$$

In order to adjust these gain and phase imbalance accurately, the complex parameter 'a' will be altered by means of an adaptive procedure. In a similar way, in the error canceling circuit, the error signal is amplified in a second amplifier, the error amplifier, to obtain

$$v_{eo} = g v_{ei} \tag{5.6}$$

$g = |g| e^{j\beta}$ is the gain of the second amplifier. In a way similar to the previous case, the gain and phase of the signal v_{eo} will be adjusted by means of the complex parameter 'b', before comparing it to the signal v_o free of distortions

$$v_o = V_a - \frac{v_{eo}}{k_2} \tag{5.7}$$

where k_2 (coupling coefficient of second loop) is real.

$$v_o = G_1 v_{in} + v_{IMD} - b \frac{g}{k_1 k_2} v_{IMD} \tag{5.8}$$

If 'b' is adjusted such that

$$b = \frac{k_1 k_2}{g} \tag{5.9}$$

The equation for v_o becomes

$$v_o = G v_{in} \tag{5.10}$$

The complex parameter 'b' will be adjusted by means of an adaptive algorithm similar to the one used with the parameter 'a'.

5.3.2 Mathematical Representation of Adaptive Solution

An effective and recent solution for an adaptive estimate of the complex parameters 'a' and 'b' includes the use of digital signal processing (DSP) in both cancellation circuits.

The complex parameters a and b will be adjusted according to the LMS (Least Mean Square) algorithm implemented in the numerical domain (Help of Agilent ADS simulator). Only the equations corresponding to the adaptation of parameter ‘a’ are developed with the assumption that the same ones are completely applicable to the parameter ‘b’.

According to the LMS algorithm, the equations for a are given by

$$a(n) = a(n-1) + \Delta \bullet v_{in}(n) \bullet v_{ei}^{*}(n) \tag{5.11}$$

$$a(n) - a(n-1) = K \bullet \Delta\tau \bullet v_{in}(n) \bullet v_{ei}^{*}(n) \tag{5.12}$$

where,

K: constant of adaptation that fixes the algorithm adaptation speed.
V_{in}: PA input signal.
v_{ei}^{*}: complex conjugate of the error signal described by (5.3).

Equation (5.12) can take the equivalent form

$$\frac{a(n) - a(n-1)}{\Delta\tau} = K \bullet v_{in}(n) \bullet v_{ei}^{*}(n) \tag{5.13}$$

Taking the limit for Δτ approaching zero and assuming a(τ) to be analytic,

$$\frac{da}{d\tau} = K \bullet v_{in}(\tau) \bullet v_{ei}^{*}(\tau) \tag{5.14}$$

and the actual expression for a is obtained as

$$a(t) = K \int_{0}^{t} v_{in}(\tau) \bullet v_{ei}^{*}(\tau) d\tau \tag{5.15}$$

This algorithm can be simulated using DSP based Agilent Ptolemy simulator of Agilent's ADS. Figure 5.7 shows circuit diagram of complex correlator. Control voltages of vector modulators can be obtained by this control circuit and output of this circuit (two circuits for each loop) directly feeds the vector modulators in both loops. Components of this circuit is assumed ideal only for simulation purposes.

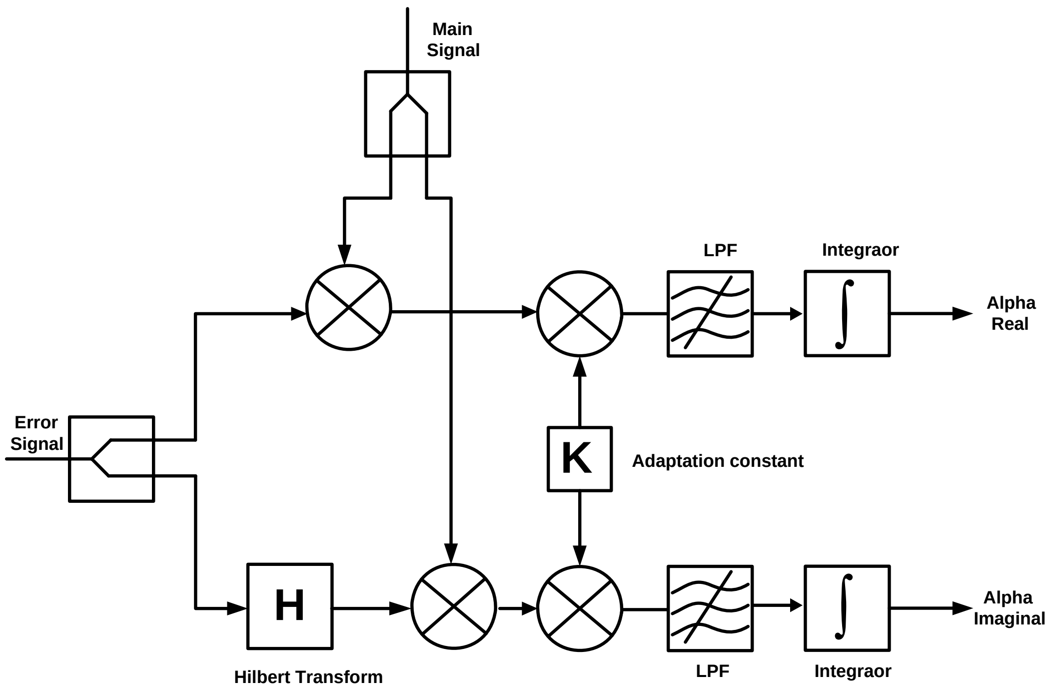

Figure 5.7: Complex Correlator

6. SIMULATION RESULTS

Toshiba Microwave Semiconductor's TMD0507-2A Microwave MMIC Amplifier is used for the simulation. Amplifier supplies typically 25 dB gain. The amplifier offers +33dBm P1dB and 22.0dB G1dB. In Figure 6.1, delay characteristics of TMD0507-2A is shown. Delay elements are used in both loops to compensate this delay.

A complex correlator is used to simulate the LMS algorithm in DSP controller unit. Basic correlator unit has two inputs and outputs. For the first loop, inputs are main and error signals to be correlated. Moreover, for the second loop, inputs are error and feedforward output signals to be correlated. The outputs of correlators are adaptation coefficients, 'a' and 'b'. These adaptation coefficients applied directly to complex gain adjusters (vector modulators) of both loop which is used for adjusting gain and phase to match both upper and lower branches. In Figure 6.2 and 6.3, adaptation coefficients 'a' and 'b' are shown with respect to time.

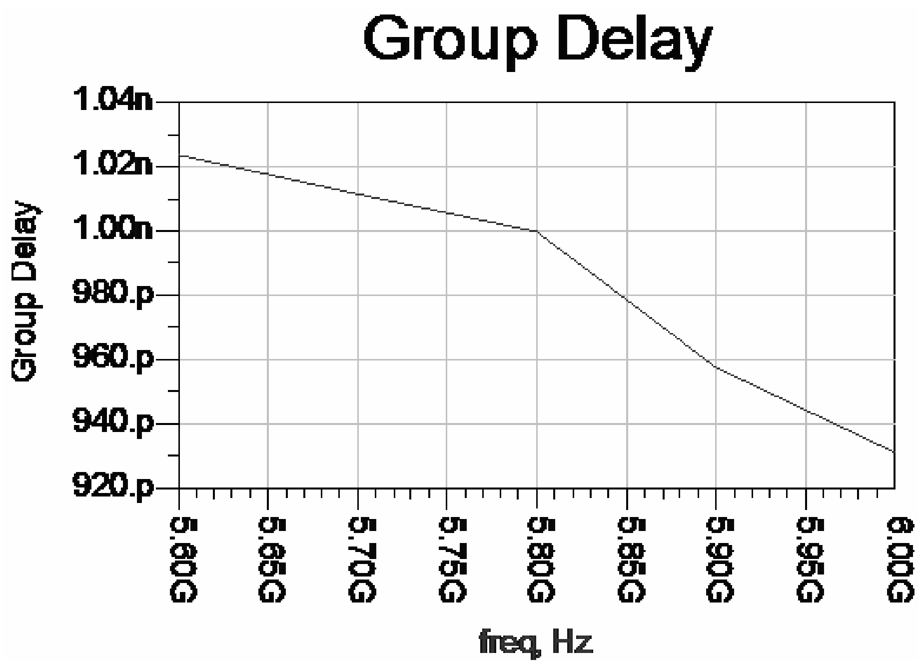

Figure 6.1: Delay Characteristic of TMD0507-2A

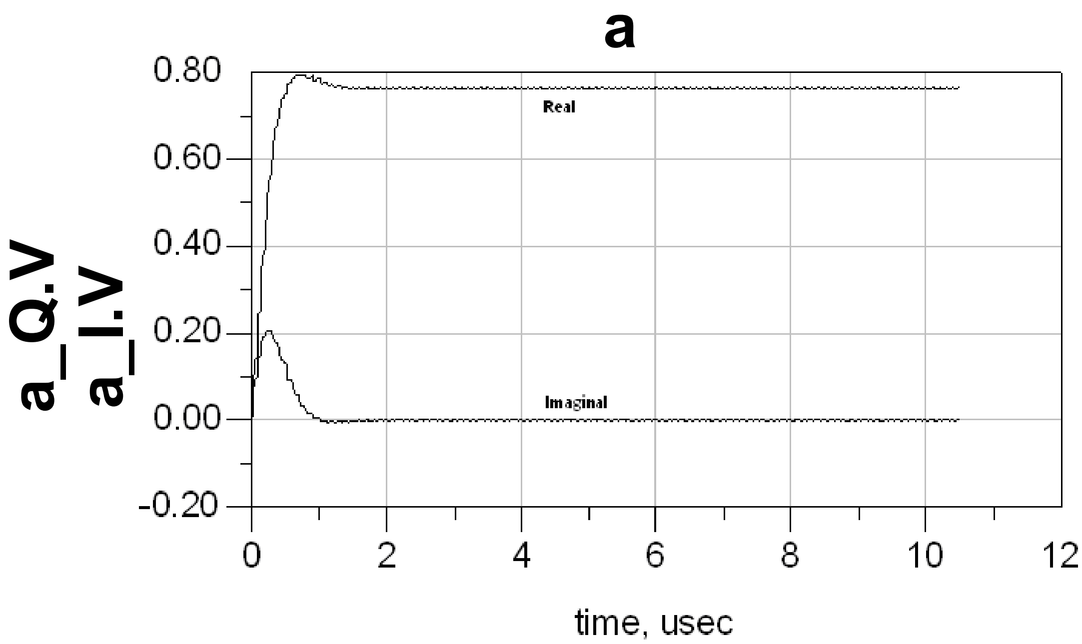

Figure 6.2: Adaptation Coefficient 'a'

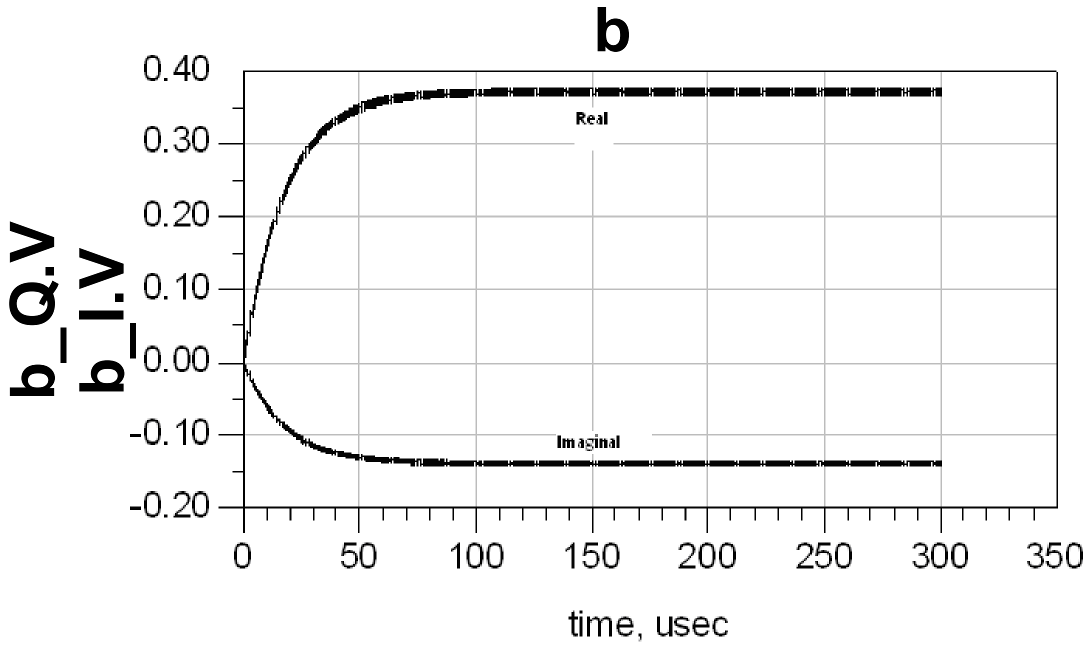

Figure 6.3: Adaptation Coefficient 'b'

When two-tone signal is applied to input of the amplifier, the output of the amplifier is in Figure 6.4. In figure, 3^{rd} order IMD is about -30dBc.

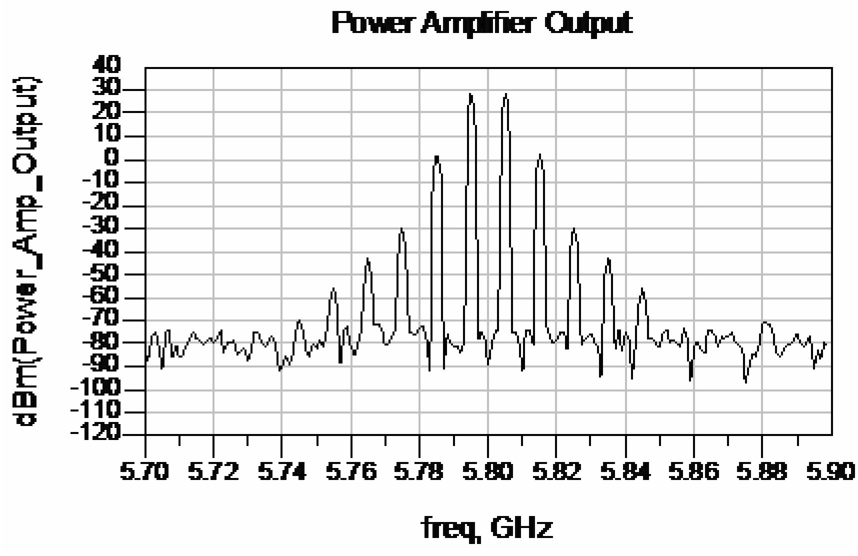

Figure 6.4: Power Amplifier Output

In Figure 6.5 and 6.6, signal cancellation output (error signal) and feedforward output are shown, respectively. It is clearly seen that 3rd order IMD is about -55dBc and 25dB improvement is achieved in IMD performance of amplifier.

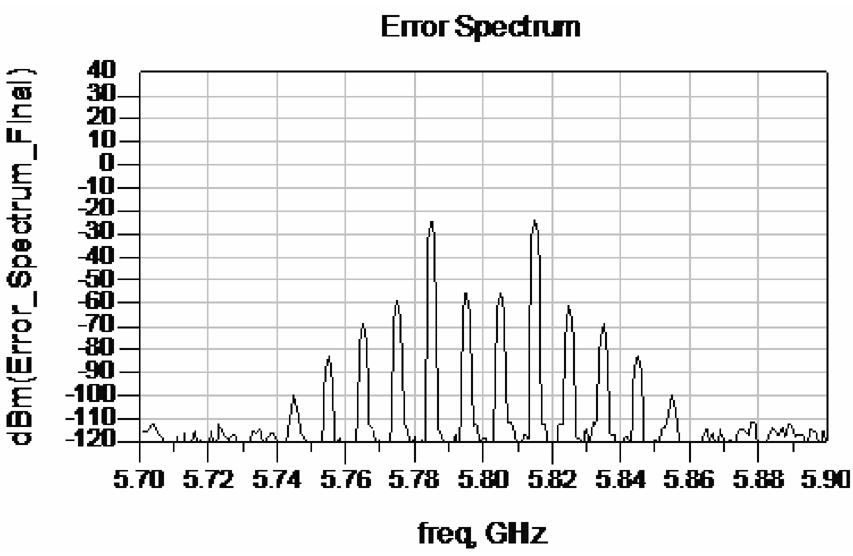

Figure 6.5: Error Signal (Signal Cancellation Output)

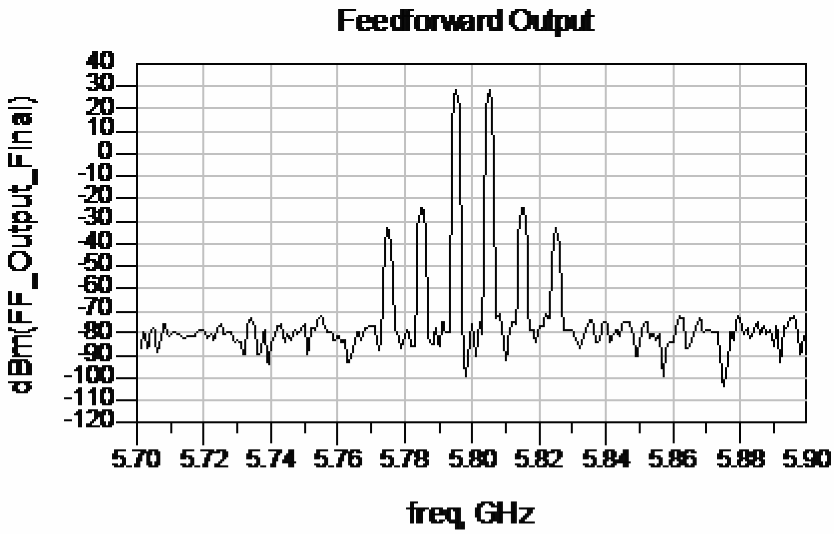

Figure 6.6: Feedforward Output

7. CONCLUSION

In this thesis, an adaptive feedforward amplifier is designed and simulated with the help of Agilent ADS simulation tool.

By using vector modulators, gain and phase adjustment is made to match the upper and lower branches in both loops. Vector modulator is a good gain and phase adjustment circuit in terms of speed and performance. Because of gain and phase adjustment has a very important role in a circuit like feedforward vector modulators is used in it. In practice, a vector modulator gives a 360^0 phase shift and provides about 20dB dynamic amplitude variations.

Adaptive circuit structure automatically controls the gain and phase adjustment in both loops in feedforward amplifier by using a DSP based complex correlater (LMS algorithm used). In consequence of this control, obtained adaptation coefficients at the output of both control circuit are dc voltages and vector modulators are driven by these control voltages to adjust gain and phase.

Consequently, intermodulation performance of a 5.8GHz power amplifier is improved 25dB by adaptive feedforward scheme using vector modulator and DSP based controller.

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AUTOBIOGRAPHY

Engin Kurt was born in İstanbul in 1979; he concluded his high school education in Şişli Yunus Emre High School. He had graduated from the Electrical and Electronics Engineering Department of İstanbul University at year 2001. He started his education for master degree at the Electronics Engineering Department of İstanbul Technical University. He is currently working as a researcher at Tübitak Marmara Research Center, Information Technologies Institute.