İSTANBUL TECHNICAL UNIVERSITY ★ INSTITUTE OF SCIENCE AND TECHNOLOGY

OPTIMIZATION OF 5.7 GHz CLASS E POWER AMPLIFIER FOR THE APPLICATION OF ENVELOPE ELIMINATION AND RESTORATION

M.Sc. Thesis by
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Department : Electronics & Communication Engineering

Programme: Electronics Engineering

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CONTENTS

ABBREVIATIONS

PA	: Power Amplifier
DE	: Drain Efficiency
PAE	: Power Added Efficiency
RF	: Radio Frequency
IF	: Intermediate Frequency
RFC	: Radio Frequency Choke
THD	: Total Harmonic Distortion
ADS	: Advanced Design System
LINC	: Linear Amplification using Non-Linear Components
EER	: Envelope Elimination and Restoration
PWM	: Pulse Width Modulation
IMD	: Intermodilation Distortion
IP	: Intercept Point
IIP	: Input Intercept Point
OIP	: Output Intercept Point
TWG	: Triangle Wave Generator
VGA	: Variable Gain Amplifier
ELGF	: Envelope Limiter Gain Function
AM	: Amplitude Modulation
PM	: Phase Modulation
QAM	: Quadrature Amplitude Modulation
DSP	: Digital Signal Processor
SSB	: Single Sideband
LO	: Local Osilator

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η	: Efficiency
θ	: Angular Time
φ	: Phase Shift Respect to the Input
ω	: Operating Angular Frequency
A	: Constant Amplifier Gain
V_o	: Output Signal
V_i	: Input Signal
P_{out}	: Output Power
P_o	: The Overall Efficiency
P_{dc}	: Total Power Taken from the DC Supply
P_{in}	: Input Power
V_{DD}	: Supply Voltage
R_L	: Output Resistance
L_0	: Series Inductor in the Output Resonator
C_0	: Series Capacitor in the Output Resonator
jX	: Series Reactance
L_1	: Inductance of the RFC
C_1	: Shunt Capacitor
R_{dc}	: Equivalent Resistance of the Amplifier to the Supply
Q_L	: Quality factor of the Output Resonator

5.7GHz E-SINIFI GÜÇ KUVVETLENDİRİCİNİN ZARF YOKETME VE YENİDEN OLUŞTURMA TEKNİĞİNİN UYGULANIMI AMACIYLA OPTİMİZE EDİLMESİ

ÖZET

Rekabetin yoğun olduğu günümüzde tasarımcılar hafif, boyutları daha küçük ve düşük güçle çalışan yüksek performanslı ürün geliştirmenin yollarını aramaktadırlar. RF alıcı uygulamalarında güç kuvvetlendiricileri en fazla güç sarfiyatının olduğu bölümdür. Kablosuz iletişim sistemlerinde güç küvvetlendiricisi verimi maliyeti direkt olarak etkilemektedir.

Teorik olarak %100 verim elde edilebilen E-sınıfı güç kuvvetlendiricileri transistorların açık/kapalı durum geçişlerinde güç sarfiyatını minimize edebilmektedir. Ayrıca çıkış gerilimi kaynak gerilimi ile doğrusal değişmektedir.

Bu çalışmada E sınıfı güç kuvvetlendiricilerinin tasarım metodları ele alınmıştır. 5.7 GHz de çalışan birinde toplu devre elemanları, diğerinde transmisyon hattı elemanları kullanımış E sınıfı güç kuvvetlendiricileri tasarlanmıştır. Her iki devrede de %50 güç ekli verim (GEV) ve 500mW çıkış gücü elde edilmiştir. Sinyaldeki bozulmayı azaltmak için başvurulan doğrusallaştırma yöntemi Zarf Yoketme ve Tekrar Oluşturma metodudur. E sınıfı kuvvetlendiricinin Zarf Yoketme ve Tekrar Oluşturma yöntemi kullanılarak doğrusallaştırılmasıyla IMD bileşenlerinde 7.5 dB azalmış olup seviyesi gerçek işaretin 20dB altındadır.

OPTIMIZATION OF 5.7 GHz CLASS E POWER AMPLIFIER FOR THE APPLICATION OF ENVELOPE ELIMINATION AND RESTORATION

SUMMARY

In today's competitive, manufactures and product developers are seeking ways to build high performance devices that are lighter in weight, smaller in size and operating at lower power. In transceiver applications one module is responsible for a large portion of the power consumption - the power amplifier. The efficiency of the power amplifier has a direct impact on the cost of the wireless communication system.

The class-E amplifier has a maximum theoretical efficiency of 100%. Class E power amplifiers have the ability to minimize power loss during on/off transitions of the transistor. Also, the output voltage varies linearly with the supply voltage.

This thesis describes the design and the linearization methodology of the Class E amplifiers. Two class-E amplifiers operating at 5.7 GHz are presented. One of them is a lumped elements based circuit and the other is a transmission lines based circuit. Both circuit show good performance with 50% PAE and have 500mW output power. Envelope elimination and restoration is the linearization method chosen to achieve reduction of signal distortion. Linearization Class E PA using EER system provides an additional 7.5 dB reduction in intermodulation distortion products, achieving a minimum distortion level of 20 dB below the fundamental signals.

1. INTRODUCTION

In today's competitive, cost-driven, rapidly growing wireless personal communication market, one apparent strategy for a company to stay competitive and profitable is to lower product development cost. Manufactures and product developers are seeking ways to build high performance devices that are lighter in weight, smaller in size and operating at lower power. In both handheld and base-station applications one module is responsible for a large portion of the power consumption - the power amplifier. In general, the higher the output power of the transmitter, the higher the power amplifier consumption as a percentage of the total.

The efficiency of the power amplifier has a direct impact on the cost of the wireless communication system. Increasing the efficiency of the power amplifier in a handheld transmitter results in reduced DC power drain on the batteries, reduced handset weight resulting from reduced heat sinking requirements, and increased reliability due to reduced junction operating temperatures.

In a base-station application, increased efficiency results in benefits similar to those accrued to handset using efficient power amplifiers. In addition, the reduction in input power significantly reduces the cost of amplification. As a result, efficient power amplification is highly desirable regardless of the final application.

Because of its superior performance over the MOS transistors, Gallium Arsenide (GaAs) transistors have been used extensively to build the RF power amplifiers. GaAs based power amplifiers have several drawbacks: costly to implement, require high level power supply, and have large size. CMOS power amplifier's performance is limited because of its low breakdown voltage, low current drive, and lossy substrate.

There are two general types of power amplifiers (PA)- linear PAs (Class A, Class B, Class AB, Class C) and switch mode PAs (Class D, Class E, and Class F). The Class A amplifier topology while exhibiting excellent linearity suffers from the poorest efficiency. Classes B, and AB have slightly better efficiency: however, they require complementary devices or transformers to drive the inputs of the devices out of phase and for generating the complete output waveform. Class D amplifiers have a similar requirement for transformers - a difficult device to operate over wide bandwidths. The Class C amplifier exhibits high efficiency and is relatively simple to construct. The Class F and E amplifiers also exhibit high efficiency and consist of relatively few (and simple) components.

Among switch mode amplifiers, Class E power amplifiers have the ability to minimize power loss during on/off transitions of the transistor. Secondly, the output voltage varies linearly with the supply voltage. In addition, given the trend towards low-voltage, low power operation, Class E power amplifiers have clear advantages than other types of power amplifiers.

Using nonlinear amplifiers requires a linearization technique to reduce the distortion of the output signal to an acceptable level. The linearization technique chosen for implementation in this work is called Envelope Elimination and Restoration. This method makes use of input signals which possess both amplitude (envelope) and phase information. The envelope component of the input signal is amplified in a high efficiency modulator. The phase component is recombined with the envelope component in a second high efficiency amplifier. The result is an amplified replica of the input signal with very low distortion level. Not only is this technique suitable for use with cellular modulation schemes, it can also be used to efficiently amplify combined signals, such as exist in a cellular base station, with reduced power consumption and space requirements.

The organization of this study is explained below.

Chapter 1 is the introduction part.

Chapter 2 surveys and discusses the definitions and circuit operations of the power amplifier classes.

Chapter 3 focuses on the Class E power amplifiers. The operation principle and design methodology of the Class E amplifiers are presented. Two class-E amplifiers operating at 5.7 GHz are presented in section 3.4. One of them is a lumped elements based circuit and the other is a transmission lines based circuit.

Chapter 4 describes prior linearization methods including feedback, feedforward, predistortion and EER methods.

Chapter 5 details the analysis of the EER system and the PA designed in chapter 3 is linearized using this method.

Finally Chapter 6 summarizes the simulation results and provides conclusions.

2. POWER AMPLIFIERS

2.1 Introduction

A power amplifier is the final stage of amplification in the transmission path for all wireless communication systems. The main purpose of the PA is to amplify the transmit signal so that it can reach the receiver at a specified distance. Depending on the distance between transmitter and receiver, the output power of PA is determined, and depending on the method of modulation, the type of power amplifier is decided. If the signal contains both amplitude and phase modulation, a linear amplification is required. Since the PA dissipates a lot of the power, high efficiency is necessary to improve the battery life for mobile wireless systems. So, the power efficiency and linearity of the PA are very important.

2.1.1 Linearity

An amplifier is said to be linear if it preserves the details of the signal waveform, that is to say,

$$y(t) = Kx(t) \qquad \textbf{(2.1)}$$

where, x(t) and y(t) are the input and output signals respectively, and K is a constant gain representing the amplifier gain. Fig. 2.1 illustrates the linear behavior of an amplifier in time and frequency domain. In Fig. 2.1(a), the input signal x(t) on the left side is equal to a signal of the form $A\sin(\omega_0 t)$ where A is the maximum amplitude and equal to one and ω_0 is the frequency in radians per second. If K=10, then the output signal y(t), depicted on the right side in Fig. 2.1(a) equals 10 $A\sin(\omega_0 t)$. Fig. 2.1(b) and (c) depict the frequency spectra of input and output signals, respectively. The input and output spectra only have one frequency, the same frequency at input and output. The amplitude of the output signal is ten times greater than the amplitude of the input signal.

A non-linear amplifier may cause distortion in the output signal by modification of the amplitude and or time delay information. These modifications result in output power compression, phase distortion and excess frequency terms known as intermodulation and harmonic distortion products.

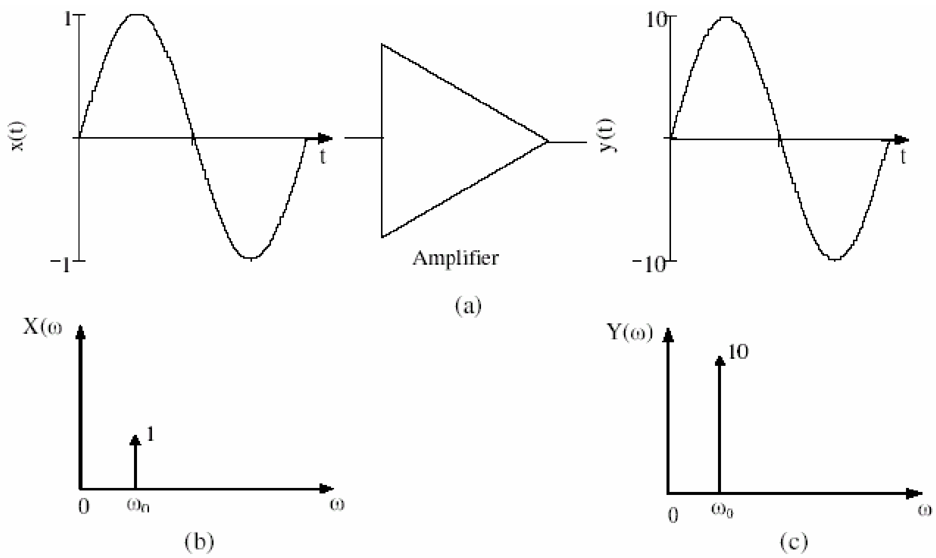


Figure 2.1: Input and Output Signals of Linear Amplifier: The Left Side In (a) Shows The Input Signal x(t) In Time Domain With Amplitude 1. The right Side In (a) Depicts The Output Signal y(t) In The Time Domain. (b) Displays The Frequency Spectrum of The Input Signal X(ω) (c) Shows The Spectrum of The Output Signal Y(ω)

2.1.1.1 Harmonic distortion

The simplest form of amplitude non-linearity may be illustrated by the addition of a second term to the transfer characteristic.

$$y(t) = k_1 x(t) + k_2 x^2(t) \tag{2.2}$$

This form of transfer characteristic is referred to as second-order due to the power of two. This non-linear amplifier will introduce an extra frequency component, which will appear at two times the original frequency.

Examination of the amplitude of the second harmonic component indicates that it will increase in proportion to the square of the input signal (and also in proportion to the constant, k_2). The amplitude of the fundamental frequency component, however, will only increase in proportion to the voltage gain, k_1. As a result, the amplitude of the second harmonic will increase at a greater rate than that of the fundamental component and at some input level both the fundamental and the second order harmonic will have the same output.

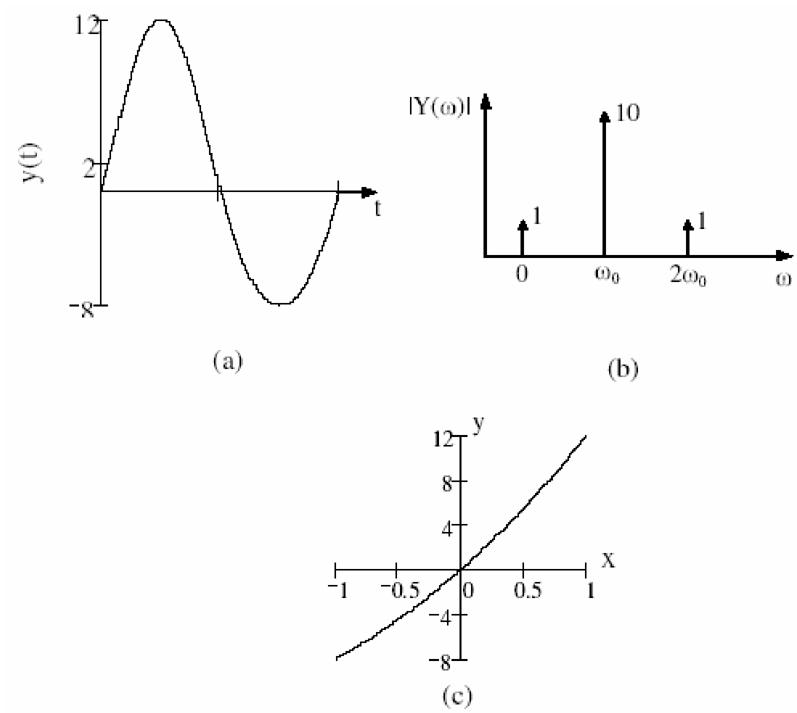


Figure 2.2: Second Order Nonlinearity In Amplifier: (a) Output Signal y(t) In Time Domain, Signal Exhibits Positive DC Offset; (b) Output Signal $|Y(\omega)|$ In Frequency Domain, DC Component at $\omega=0$ and Second Harmonic at $\omega=2\omega_0$ With Fundamental at ω_0; (c) Input-Output Transfer Characteristic Has Nonlinear Behavior; $k_1=10$, $k_2=2$.

2.1.1.2 Intermodulation Distortion

Intermodulation Distortion is a phenomenon of generation of undesirable mixing products, which distort the fundamental tones and gives rise to intermodulation products. The third order intermodulation products have the maximum effect on the signal, as they are the closest to the fundamental tone. The unwanted spectral components, such as the harmonics, can be filtered out. But the filtering does not work with the third order intermodulation products, as they are too close to the fundamental tone. Figure 2.3 shows the frequency domain representation of the intermodulation distortion caused due to a two-tone signal. Assuming that the output voltage is an instantaneous function of the input voltage and the non-linearity is weak, then the output voltage y(t) can be related to the input voltage x(t) by a simple power series:

$$y(t) = k_1 x(t) + k_2 x(t)^2 + k_3 x(t)^3 \qquad \textbf{(2.3)}$$

Applying the input signal $x(t) = A\cos\omega_1 t + A\cos\omega_2 t$ to the device transfer function defined by equation 2.3 will yield an output voltage as follows:

$$\begin{aligned} y(t) &= k_2 A^2 + k_2 A^2 \cos(\omega_1 - \omega_2)t + \left(k_1 A + \frac{9}{4}k_3 A^3\right)\cos\omega_1 t \\ &+ \left(k_1 A + \frac{9}{4}k_3 A^3\right)\cos\omega_2 t + \frac{3}{4}k_3 A^2 \cos(2\omega_1 - \omega_2)t \\ &+ \frac{3}{4}k_3 A^3 \cos(2\omega_2 - \omega_1)t + k_2 A^2 \cos(\omega_1 + \omega_2)t + \frac{1}{2}k_2 A^2 \cos 2\omega_1 t \\ &+ \frac{1}{2}k_2 A^2 \cos 2\omega_2 t + \frac{3}{4}k_3 A^3 \cos(2\omega_1 + \omega_2)t + \frac{3}{4}k_3 A^3 \cos(2\omega_2 + \omega_1)t \\ &+ \frac{1}{4}k_3 A^3 \cos 3\omega_1 t + \frac{1}{4}k_3 A^3 \cos 3\omega_2 t \end{aligned} \qquad \textbf{(2.4)}$$

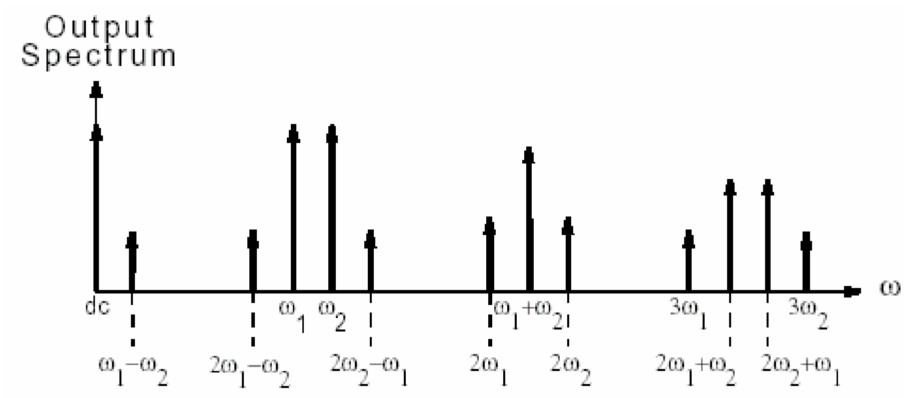


Figure 2.3: Output Spectrum of A Device With Two Input Sinusoids and Third Order Distortion

The output signal consists of six different types of components. The first component occurs at dc followed by the fundamental frequency components ω_1 and ω_2. The second and third harmonic components are located at frequencies $2\omega_1$, $2\omega_2$ and $3\omega_1$, $3\omega_2$. The second order intermodulation product components occur at frequencies of $\omega_1 \pm \omega_2$. And finally the third order intermodulation product components occur at $2\omega_1 \pm \omega_2$and $2\omega_2 \pm \omega_1$.

Practical systems are not concerned with the second and third harmonic components, and the second order intermodulation components. These components can be removed through appropriate filtering. However, the third order components fall within the pass band of the system and cannot be removed by filtering. This means that the system 3rd order intermodulation products (IMD3s) must be reduced to as much as possible to reduce distortion in the output signal.

2.1.1.3 Desensitization

In radio receivers, one effect of nonlinearity is when weak signals are being blocked by strong signals at different frequencies. This situation can occur when a radio receiver has to process a weak signal radiated from a remote radio sender in the presence of a strong signal emitted from a close radio transmitter. When this strong signal is processed along with the weak desired signal, a radio system can exhibit an effect called desensitization. This effect can be explained with the fundamental term of the full expansion of equation (2.3), which is described with (2.5).

$$\left(k_1 A_1 + \frac{3}{4}k_3 A_1^3 + \frac{3}{2}k_3 A_1 A_2^2\right)\cos(\omega_1 t) \qquad \textbf{(2.5)}$$

where (2.5) represents the linear component of the output signal. Coefficients k_1 and k_3 represent the linear component and third order nonlinearity of the nonlinear amplifier, respectively. If A_1 is the amplitude of the weak desired signal and A_2 the amplitude of a strong interferer and $A_1 << A_2$, (2.5) can be approximated as shown in (2.6).

$$\left(k_1 + \frac{3}{2}k_3 A_2^2\right)A_1 \cos(\omega_1 t) \qquad \textbf{(2.6)}$$

For $k_3<0$, the gain as given by $k_1 + \frac{3}{2}k_3 A_2^2$ a decreasing function of A_2. This decrease in gain decreases the desired signal strength at the output of the amplifier.

2.1.1.4 1dB Compression Point

All amplifiers have some maximum output-power capacity, referred to as saturated power or simply saturation. Driving an amplifier with a greater input signal will not produce an output above this level. As an amplifier is driven closer to saturation, its deviation from a straight-line response will increase. The output level will increase by a smaller amount for a fixed increase in input signal and then reaching saturation.

A good way of measure distortion is to measure the input or output power at which an amplifier's gain reduces by 1dB from the strait line. This point is referred to as the 1dB compression point, illustrated in Figure 2.4.

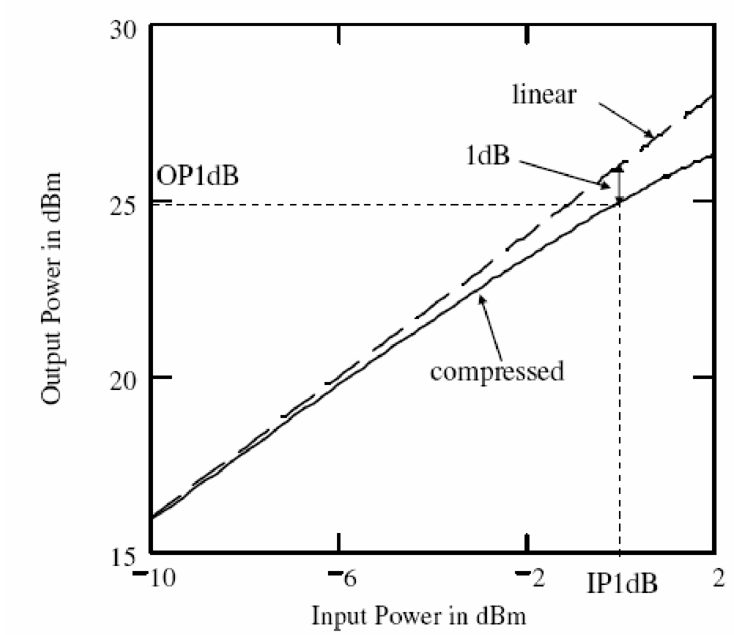


Figure 2.4: 1dB-Compression Point: Linear Transfer (dashed line) and Compressed Transfer Characteristic (solid line); Input Referred 1dB Compression Point, IP1dB is The Input Power Level Where The Gain is Compressed by 1 dB, The Compressed Output Power Level at This Input Power Level is Called Output Referred 1dB Compression Point OP1dB.

2.1.1.5 Intercept Point

Intermodulation products, particularly third order distortion products, are of major concern in communication systems. As discussed above, the frequencies of these products appear close to the desired band and are difficult to filter. The level of intermodulation products is dependent on input power level. The intercept point (IP) is a single parameter that characterizes the behavior of intermodulation products independent of input power level. The intercept point can be input or output referred. Of most interest are second order intercept point (IP2) and third order intercept point (IP3).

The intercept point is defined as the point where the fundamental linear component and intermodulation products have equal amplitude at the output of a nonlinear circuit. In most practical circuits, intermodulation products will never be equal to the fundamental linear term because both amplitudes will compress before reaching this point. Nevertheless, the intercept point is useful to characterize circuits.

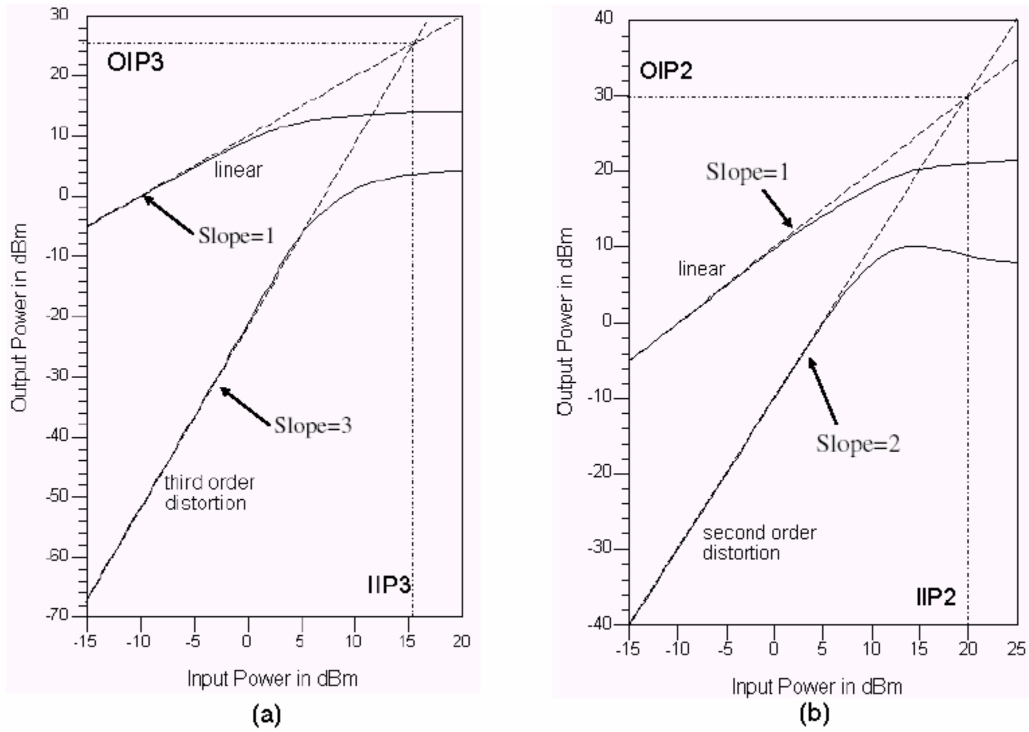


Figure 2.5: Intercept Point a) Third Order Intercept Point IP3 b) Second Order Intercept Point IP2

Figure 2.5(a) illustrates the circumstances of the third order intercept point graphically. The plot displays the output power versus the input power in dBm. The upper solid line describes the linear component and the lower solid line describes the third order distortion product dependent on input power levels. Both solid lines compress and do not increase further due to compression and the limit in supply voltage. The dashed lines in Figure 2.5(a) indicate extrapolation of the solid lines. The point where the dashed lines intersect is called third order intercept point. The horizontal coordinate of the intercept point is usually referred to as third order input intercept point IIP3 and the vertical coordinate is the third order output intercept point OIP3. As stated before, as the input power increases by 1 dB, the linear output power also increases by 1 dB, but the third order distortion increases by 3 dB due to a slope of three of the third order distortion component on a logarithmic scale. In Figure 2.5(a), the input IP3 equals 15 dBm and the output IP3 is 26 dBm.

A similar intercept point can be defined for the second order intermodulation. Similar to Figure 2.5(a), the solid lines in Figure 2.5(b) are linear and second order distortion components. Both solid curves are extrapolated as shown with the dashed lines. The point where the dashed lines intersect is called second order intercept point. As demonstrated before, the fundamental component increases linearly until clipping occurs. The second order intermodulation product level rises with a slope of two due to the second order term. IP2 can be referred to input (IIP2) or output (OIP2).

2.1.1.6 Phase Distortion

A phase distortion called AM to PM conversion occurs when the magnitude of the input signal can affect changes in the phase of the output signal. The AM to PM conversion is defined as the change in output phase for a 1 dB increment of output power. Phase distortion also generates IMD products that are similar in nature to amplitude distortion. Thus the reduction of these products will also ensure reduction of phase distortion.

2.1.2 Efficiency

Efficiency is another important parameter when comparing the performance of different PA's. It is the ratio of power delivered to the load to the power consumed from the source. Ideally the best PA is the one which all the power consumed from the supply is delivered to the source, i.e. 100% efficiency. In this case, no power would be consumed by the PA circuit. However, in reality, PA efficiencies never reach 100%. Power is dissipated by the circuits in order to amplifying the circuits or through losses mechanisms. In practice, three types of efficiency measures are used to characterize PAs. The first one is drain efficiency,

The DE is defined as:

$$DE = \frac{P_{out}}{P_{DC}} \tag{2.7}$$

where is the power delivered to the load at the desired RF frequency, and is the total power taken from the DC supply.

The PAE, the most commonly used metric in literature, is defined to be:

$$PAE = \frac{P_{out} - P_{in}}{P_{DC}} \quad \textbf{(2.8)}$$

where is the power needed to drive the input at the RF frequency of interest.

The overall efficiency is defined as:

$$P_o = \frac{P_{out}}{P_{DC} + P_{in}} \quad \textbf{(2.9)}$$

Both the PAE and the overall efficiency can be seen as better gauges of the true performance of a PA, since they include the power needed to drive the PA.

Power amplifiers are classified according to their mode of operation. Amplifier operating modes are determined from active device bias conditions and the output network topology. The bias conditions determine whether the active device is operated as a current source or as a switch.

2.2 Current Source Power Amplifiers

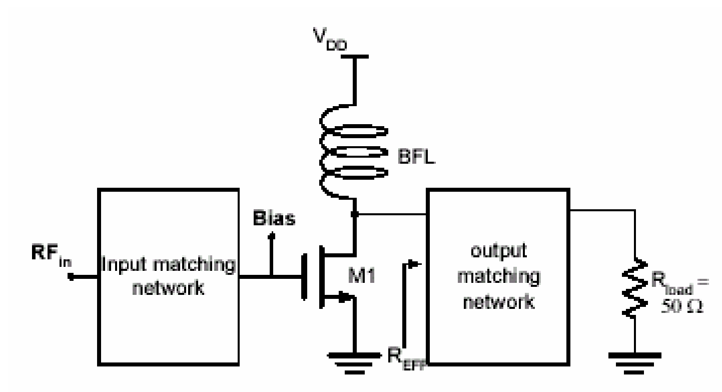


Figure 2.6: Basic Circuit of A Linear PA

Figure 2.6 shows a basic circuit of a linear PA where in the transistor is used as a current source. The transistor changes the RF input voltage into an output drain current. The RF choke inductor should be large enough to block passage of the RF signal to V_{DD}. Class-A, class-B, class-AB or class-C operation is determined by the input DC bias voltage, which sets the conduction angle. The conduction angle represents the fraction of time that a transistor is on during one complete RF cycle. If the transistor is on during the entire RF cycle, the conduction angle is 360 degrees, and if it is on during half the RF cycle, the conduction angle is 180 degrees. Drain efficiency, total harmonic distortion (THD), output power, and power gain are determined by the conduction angle.

There are two key matching networks in Figure 2.6; one is the input matching network and the other is output matching network. The input-matching network is usually designed based on the maximum power transfer theorem, and the output-matching network is designed to achieve the specified output power.

2.2.1 Class A Power Amplifier

A Class-A PA is the simplest and most basic form of power amplifier. In the Class-A operation, the transistor is in its active region for the entire input cycle, and thus is always conducting current. The class-A PA approximately maintains the same voltage gain throughout the entire working cycle since it is linear in that region in case of a MOS device. Figure 2.7 shows the schematic and the corresponding waveforms for a Class-A PA.

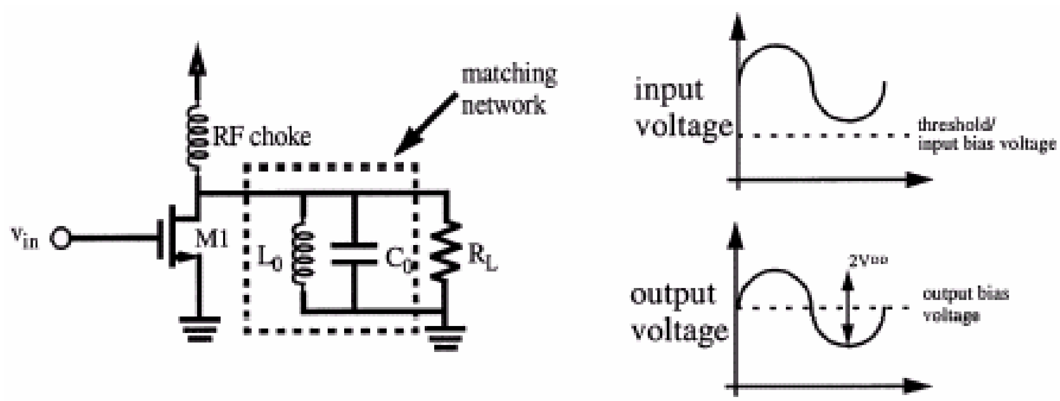


Figure 2.7: Class-A PA Circuit and Voltage Waveforms

The main limitation of a class A PA is its poor efficiency. Since the device is conducting current at all times, the PA consumes power continuously, whether the power is used to drive a load or not. In the ideal case, the drain efficiency is given by

$$\eta = \frac{P_o}{P_{dc}} = \frac{\frac{V_{DD}^2}{2R_L}}{\frac{V_{DD}^2}{R_L}} = \frac{1}{2} \qquad \textbf{(2.10)}$$

In an ideal class-A PA, the output is biased at the supply voltage, and the output swing has amplitude of V_{DD}, thus the efficiency is limited to 50%. However, if the load is resistive, the max efficiency cannot even be higher than %25 because the output voltage will be limited to less than V_{DD}.

2.2.2 Class B Power Amplifier

A class B PA provides higher efficiency than a class A circuit, by using two transistors which are connected in parallel and biased at the edge of conducting current. This class is also known as a "push-pull" stage. In a Class B PA there is no standing continuous current flowing in the PA at all times. M2 pushes current to the load during the positive half cycle with M1 being nearly off. Similarly during the negative half cycle M1 pulls current from the load and M2 is off.

The output power for the single ended class-B amplifier is

$$P_{out} = \frac{1}{2} I_{om} V_o \tag{2.11}$$

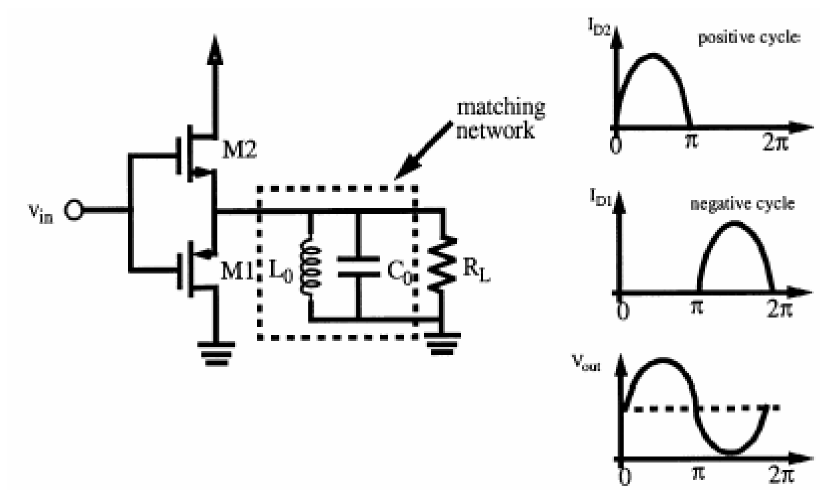


Figure 2.8: Class-B PA Topology and Voltage and Current Waveforms

The dc drain current is

$$I_{dc} = 2\frac{I_{om}}{\pi} \tag{2.12}$$

The dc input power is

$$P_{dc} = 2\frac{I_{om}V_{DD}}{\pi} \quad \textbf{(2.13)}$$

and the maximum efficiency when $V_{om} = V_{DD}$ is

$$\eta = \frac{P_{out}}{P_{dc}}100 = \frac{\pi}{4}\frac{V_{om}}{V_{DD}}100 \leq 78.53\% \quad \textbf{(2.14)}$$

The class-B PA allows for higher efficiencies in contrast to the class A. In practice, the efficiency of a Class B implementation may reach as high as 60% in GaAs implementations. However, it suffers from the crossover distortion problem. Crossover distortion can occur when the two transistors change their phases. This reduces the overall linearity of the configuration.

2.2.3 Class AB Power Amplifier

Class AB amplifiers take both advantages of higher linearity of class A, higher efficiency of class B topologies. It utilizes the push-pull concept to improve the efficiency, and use slightly higher bias voltage to alleviate the crossover distortion that can cause problems in a class-B structure. As its name implies, a class AB is a cross between a class-A and a class-B structure. Circuit implementations of a Class AB PA are similar to those of the class-B architecture; the only difference is the biasing and output waveforms. Examples of these are given in below.

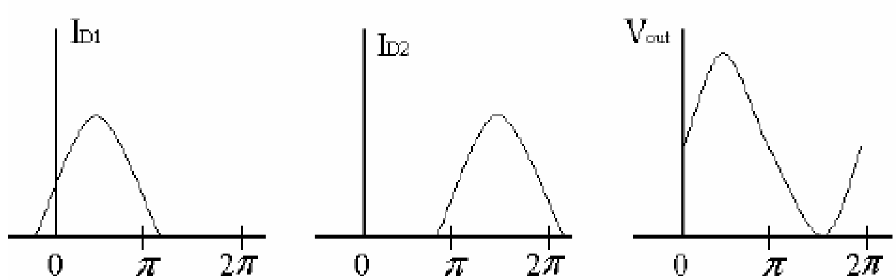


Figure 2.9: Class-AB PA Voltage and Current Waveforms

In a class-AB PA, since M1 and M2 are allowed to conduct current for a short period with higher bias voltage, the output voltage waveform during the crossover period can be smoothed out and thus can reduce the distortion at the output. Moreover, to achieve higher efficiency, the bias point can be chosen to be close to the threshold (the Class B bias point), therefore the efficiency and linearity would approach the Class B levels.

Alternatively, to achieve better linearity, the bias can be chosen such that the device remains on for longer of the input cycle (closer to the Class A bias point), therefore the efficiency and linearity would start to approach the Class A levels. Several Class AB PA's have been reported in the literature, with efficiencies ranging between 30% and 60%.

2.2.4 Class C Power Amplifier

The Class C PA has the same structure as a class A PA, which biasing the transistors at the edge of the conducting current. In this class, the transistor is only on for less than a half cycle, which increases the efficiency. Figure 2.10 shows the circuit and the waveforms of a Class C PA.

The biasing voltage, which is negative, is applied to the input gate. Transistor turns on every time the input voltage is larger than the sum of the absolute value of the biasing voltage and threshold voltage. The matching network not only matches the output impedance, but also filters out the harmonics other than the fundamental. The power dissipation is proportional to the conduction angle, and the efficiency is dependent on θ and is given by

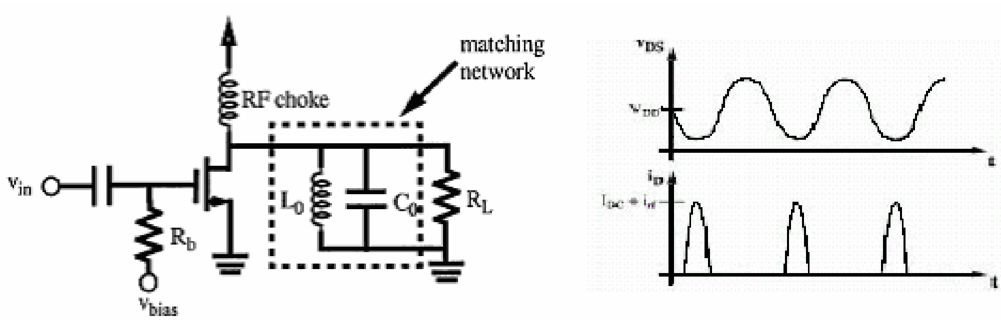


Figure 2.10: Class-C PA Topology and Waveforms

By varying the conduction angle, the efficiency can be changed according to the requirement. Conduction angle is between 180º and 0º for class C, corresponding to efficiencies ranging from 79% to 100%.

$$\eta = \frac{\theta - \sin\theta}{4\left(\sin\left(\frac{\theta}{2}\right) - \frac{\theta}{2}\cos\left(\frac{\theta}{2}\right)\right)} \tag{2.15}$$

2.3 Switch-Mode Power Amplifiers

Switching mode power amplifiers (class D, E, and F) have higher power efficiency than the current mode amplifiers because the transistors are acting as low loss switches. Ideal lossless switches have no overlap between nonzero switch voltage and nonzero switch current; thus efficiency could reach 100%. Practical switch mode power amplifiers have significant power loss at high frequency due to parasitic and non-ideal switching transistors.

2.3.1 Class D Power Amplifier

The class-D PAs are often built on a push-pull pair. In the input stage, two switch transistors M1 and M2 which are driven by a differential input feed the primary stage of a center-taped transformer, with the center winding connected to the power supply directly. M1 and M2 are switched on and off periodically, and thus the drain voltage demonstrates a square wave.

In the output stage, a series LC filter resonated at the fundamental frequency is employed to reject the high-order harmonics delivered to the load, resulting in a sine wave current at the fundamental frequency in the output. This sine wave current in turn generates a sine wave current across the primary stage of the transformer. Schematics and the waveforms of the class-D PA are shown in Figure 2.11.

Ideally there is no overlap between the drain voltage and the drain current, and thus the ideal drain efficiency can be 100%. However, the applications of the voltage-mode class-D PA in GHz range are limited. Due to the parasitic capacitance across the drain and the source of the switch transistors, the drain voltage will be charged or discharged periodically, and thus the voltage waveform cannot have a perfect square shape and some transient current spikes occur when the transistors are turned on. Therefore the overlapping of the nonzero voltage and the nonzero current induces extra loss, which becomes the dominant loss starting at hundreds of MHz.

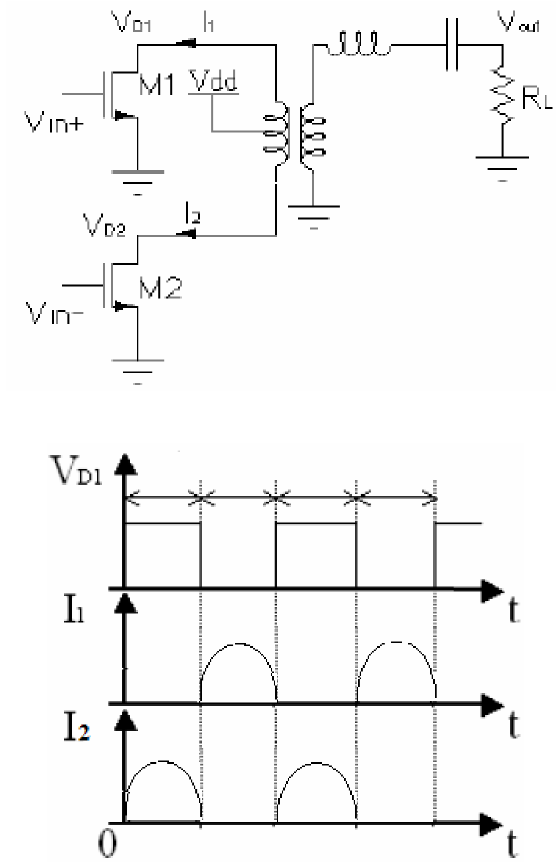


Figure 2.11: Schematics and The Waveforms of The Class-D PA

2.3.2 Class E Power Amplifiers

The transistor in a class E amplifier acts as a switch, rather than as an amplifier. When the transistor switch is closed, the drain voltage ideally is zero, and a large drain current can exist. When the switch is open, no current flows, but a large drain voltage can exist. Thus, simultaneous nonzero voltage and current is avoided, eliminating transistor power losses in the fully-open and fully-closed states. The capacitor C_p acts to hold the drain voltage V_{ds} at zero volts during the on-to-off switch transition, to avoid switching losses. The C_S, L_S, L_X, R_L network is designed such that the drain voltage falls back to zero just before the off-to-on transition, again to avoid switching losses. So all the power drawn from the DC source is driven into the output node; that is, the efficiency is theoretically 100%. General circuit diagram and typical drain voltage and current waveforms for an optimally tuned class E amplifier are shown in Figure 2.12 and Figure 2.13 respectively.

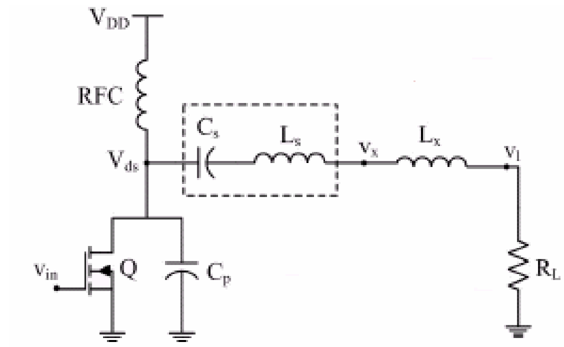


Figure 2.12: Schematic of Class E PA

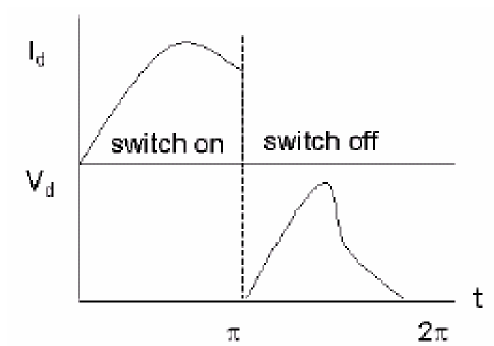


Figure 2.13: Voltage and Current Waveforms of A Class E PA

2.3.3 Class F Power Amplifiers

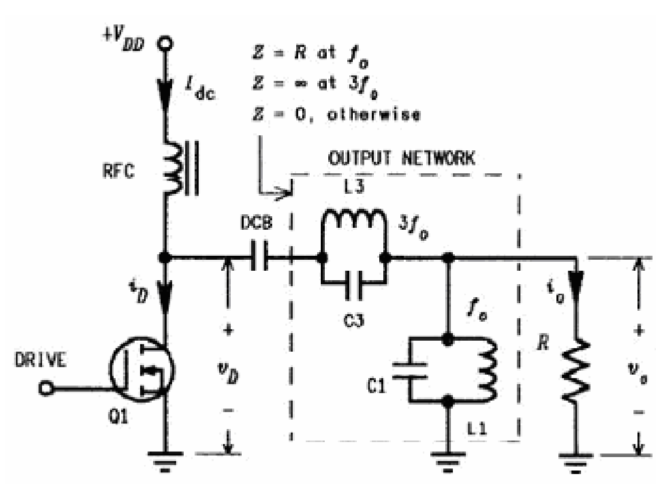


Figure 2.14: Schematic of Class F PA

The class-F amplifier is one of the highest efficiency amplifiers. It uses harmonic resonators to achieve high efficiency, which resulted from a low dc voltage current product. In other words, the drain voltage and current are shaped to minimize their overlap region. The inductor L_3 and capacitor C_3 are used to implement a third harmonic resonator that makes it possible to have a third harmonic component in the drain voltage. The output resonator is used to filter out the harmonic, keeping only the fundamental frequency at the output. The magnitude and the phase of the third harmonic control the flatness of the drain voltage and the power of amplifier. The efficiency of a class F PA can be as high as 88%.

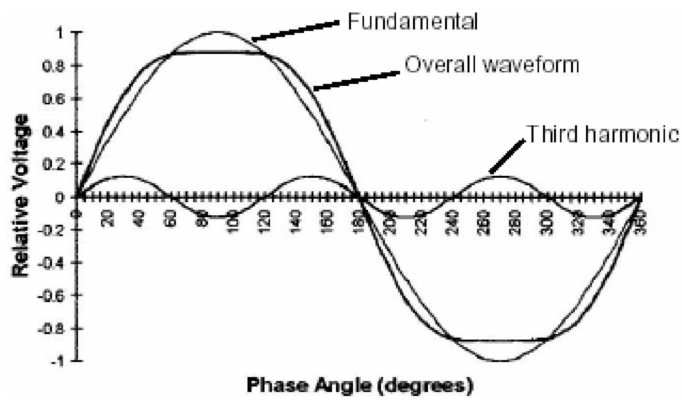


Figure 2.15: Phase Angle vs. Relative Voltage of The Class F PA

2.3.4 Class S

The Class S technique was invented in 1932 by B.D. Bedford [11]. The modern design incorporates a switching transistor and tuned output filter as in the Class D and E systems. Unlike these other systems, the input signal is modulated using pulse width modulation techniques.

The transistor acts as a single pole switch. The switch replicates the incoming PWM pulses, alternately connecting and disconnecting the lowpass output filter, formed by Lo and Co, to the power supply. The lowpass filter acts as an averaging stage, smoothing out the transitions and provides a slowly varying current to the load. Different pulse widths produce different average output values. During the times when the switch is “off”, the inductor of the low pass filter will continue to draw current up through the diode. To minimize current drooping during the off periods a multiple stage filter can be used, however, it is necessary to make the first component an inductor to ensure that high impedance is presented to the switch. The result is an efficiently amplified replica of the input envelope.

The output voltage of the modulator can have any value between 0 and Vcc. Consequently the output power is:

$$P_o(\theta) \leq \frac{V_{cc}^2}{R} \tag{2.16}$$

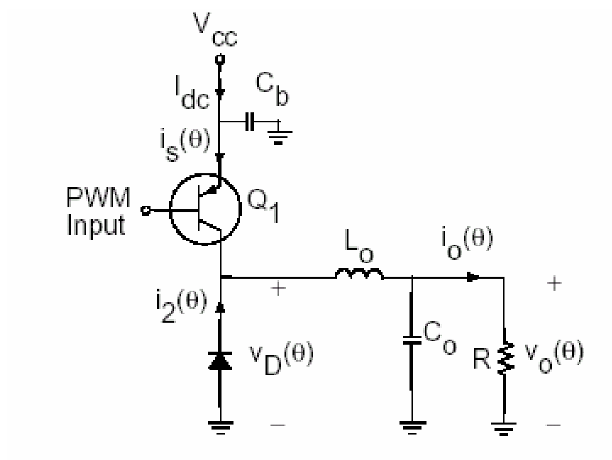


Figure 2.16: Schematic of Class S PA

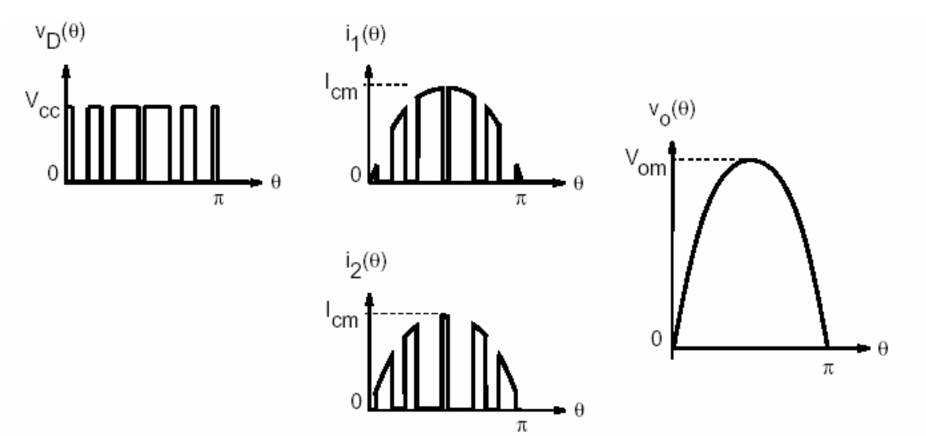


Figure 2.17: Voltage and Current Waveforms of A Class S PA

The active device never experiences non-zero voltage and non-zero current at the same time. As a result, the Class S modulator has a theoretical efficiency of 100 %.

3. CLASS E POWER AMPLIFIERS

3.1 Introduction

The class E power amplifier was first presented by Sokal et al in 1975 [7]. Sokal noted the majority of the power dissipation in conventional power amplifiers occurred in the active device. To minimize this loss in the Class E configuration, the transistor is used as a switch that is turned on and off at the carrier frequency. Under ideal conditions, the transistor would dissipate no power because the voltage and current waveforms would never overlap. The circuit operation is controlled by the load network when switch is on, and by the transient response of the network when the switch is off. The load network shapes the current and voltage waveforms such that minimum power is dissipated in the transistor itself.

In the following sections, the basic operation and mathematical analysis of the class E power amplifier is presented. Then, a design example based on the developed algorithm is described and simulation results are discussed.

3.2 Class E Theory of Operation

The ideal topology of the Class E amplifier is shown in Figure 3.1. The circuit includes a transistor operated as a switch, a shunt capacitor, C_1, an RF choke, L_1, a tuned circuit L_0-C_0, and the load resistor, R_L. The capacitance C_1 includes the parasitic capacitance across the transistor. The L_0-C_0 circuit resonates at the fundamental frequency of the input signal and only passes a sinusoidal current to the load R_L. L_0 and C_0 are modeled as ideal components. Nonidealities associated with implementing the series resonator are lumped into jX, which is termed the excessive reactance. jX primarily serves to adjust the phasing in the L-C harmonic resonator. The transistor switch S is ON in half of the period, and OFF in the other half. During the time interval t_1, the switch is open and the current through it is zero. During the time interval t_2, the switch is closed and the voltage across it is zero. Since the voltage and current waveforms of the switch do not overlap, the power dissipation in the switch is ideally zero. When the switch is off, the current through the RF choke splits between the two branches containing C_1 and R_L. The capacitance C_1 starts charging and produces the voltage across the switch. When the switch turns on, any charge stored on the capacitor C_1 will be discharged to the ground resulting in a power loss. In order to avoid this power loss, the circuit must be designed such that the voltage across the switch reaches zero at the turn on time of the switch as shown in Figure 3.2. In the ideal case the efficiency of a class E amplifier is 100%. However, in practice the switch has a finite on-resistance and the transition times from the off-state to the on-state and vice versa are not negligible. Both of these factors results in power dissipation in the switch and reduce the efficiency.

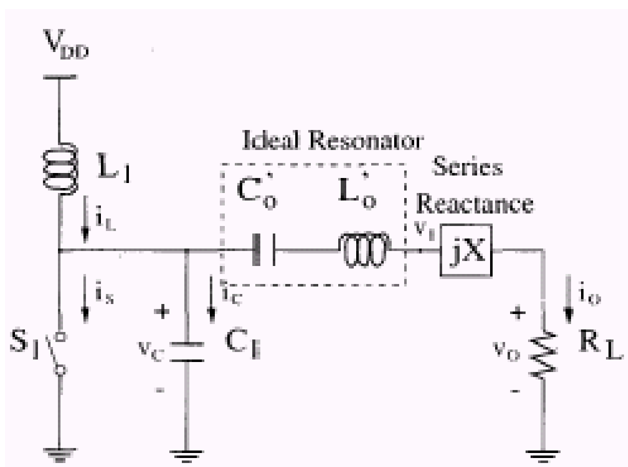


Figure 3.1: Ideal Class-E Amplifier

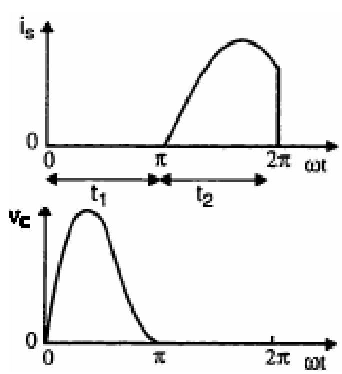


Figure 3.2: Ideal Class-E Voltage and Current Waveform

3.3 Derivation of the Ideal Class E Design Equations

The assumptions made in this analysis are [5];

1. The active device acts as an ideal switch.
2. RF choke allows only DC (I_{dc}) current and has no resistance
3. Loaded quality factor Q_L of series fundamentally tuned resonant L_0C_0 -circuit is infinite to provide pure sinusoidal current flowing into load.
4. Total shunt capacitance is assumed to be linear
5. All components are ideal.
6. The switching duty cycle is 50%.

The amplifier is driven by a large, periodic square-wave voltage signal to obtain the switching performance. Consequently, the steady state output current and voltage is also periodic and can be approximated as a sinusoidal waveform due to the high Q of the output-matching network. The output current and voltage can be written as

$$v_o(\theta) = c\sin(\theta + \varphi) \tag{3.1}$$

$$i_o(\theta) = \frac{c}{R_L}\sin(\theta + \varphi) \tag{3.2}$$

c is the amplitude of the output voltage, $\theta=\omega t$ is the angular time and φ is the phase shift respect to the input. The voltage at the output of the ideal resonator, v_1 (θ), is also a sinusoidal waveform, but is not in phase with v_o (θ) because of reactance jX.

$$v_1(\theta) = c_1\sin(\theta + \varphi_1) \tag{3.3}$$

where

$$c_1 = c\sqrt{1 + \left(\frac{X}{R_L}\right)^2} \tag{3.4}$$

and

$$\varphi_1 = \varphi + \arctan\left(\frac{X}{R_L}\right) \tag{3.5}$$

When the switch turns OFF, the sinusoidal current continues to flow, but now it flows through the shunt capacitor. During the OFF interval,

$$i_c(\theta) = I_{dc} - \frac{c}{R_L}\sin(\theta + \varphi) \qquad \textbf{(3.6)}$$

$$v_c = \frac{1}{\omega C_1}\int_0^{\theta} i_c(\phi)\,d\phi \qquad \textbf{(3.7)}$$

$$= \frac{1}{B}\left\{ I_{dc}\theta + \frac{c}{R_L}\left[\cos(\theta + \varphi) - \cos(\varphi)\right]\right\} \qquad \textbf{(3.8)}$$

where B=ωC1.

For optimal Class E operation, both the drain voltage and the rate of change of the drain voltage should be zero at switch turn on [5].

$$\frac{dv_c}{dt}\left(\frac{T}{2}\right) = 0 \qquad \textbf{(3.9)}$$

$$v_c\left(\frac{T}{2}\right) = 0 \qquad \textbf{(3.10)}$$

The first condition avoids shorting the capacitor C_1 when there is voltage across it during switching, and the second condition ensures a soft turn on condition for the switching device. Using these conditions,

$$v_c(\pi) = 0 \qquad \textbf{(3.11)}$$

$$\frac{1}{B}\left\{ I_{dc}\theta + \frac{c}{R_L}\left[\cos(\theta + \varphi) - \cos(\varphi)\right]\right\} = 0 \qquad \textbf{(3.12)}$$

$$\cos(\varphi) = \frac{\pi I_{dc} R_L}{2c} \qquad \textbf{(3.13)}$$

and

$$I_{dc} - \frac{c}{R_L}\sin(\theta + \varphi) = 0 \qquad \textbf{(3.14)}$$

$$\sin(\varphi) = -\frac{I_{dc}R_L}{c} \tag{3.15}$$

From equations (3.13) and (3.15), the phase shift, φ,

$$\varphi = \tan^{-1}(-2/\pi) = -32.48^{\circ} \tag{3.16}$$

From equations (3.15) and (3.16),

$$\frac{c}{I_{dc}R_L} = \sqrt{1+\frac{\pi^2}{4}} \tag{3.17}$$

Since there is no DC voltage drop across the RF choke, the average of the drain voltage of the transistor must be equal to the supply voltage V_{DD},

$$V_{DD} = \frac{1}{2\pi}\int_0^{2\pi} v_C(\theta)d\theta \tag{3.18}$$

$$= \frac{I_{dc}}{2\pi B}\left[\frac{\pi^2}{2} - \frac{c\pi}{I_{dc}R_L}\cos(\varphi) - \frac{2c}{I_{dc}R_L}\sin(\varphi)\right] \tag{3.19}$$

$$= I_{dc}R_{dc} \tag{3.20}$$

where R_{dc} is the equivalent resistance the amplifier shows to the DC power supply

$$R_{dc} = \frac{1}{2\pi B}\left[\frac{\pi^2}{2} - \frac{c\pi}{I_{dc}R_L}\cos(\varphi) - \frac{2c}{I_{dc}R_L}\sin(\varphi)\right] \tag{3.21}$$

Substitute equations (3.13) and (3.15) into equation (3.21)

$$R_{dc} = \frac{1}{\pi B} = \frac{1}{\pi\omega C_1} \tag{3.22}$$

The output power (dissipated in the load resistor, R_L) is obtained by

$$P_{out} = \frac{c^2}{2R_L} \tag{3.23}$$

The DC input power is given by

$$P_{dc} = \frac{V_{DD}^2}{R_{dc}} = I_{dc}^2 R_{dc} \tag{3.24}$$

Therefore the drain efficiency is obtained as

$$\eta = \frac{P_{out}}{P_{dc}} = \left(\frac{c}{I_{dc} R_L}\right)^2 \frac{R_L}{2R_{dc}} \tag{3.25}$$

For 100% efficiency R_{dc} can be written as

$$R_{dc} = \left(\frac{c}{I_{dc} R_L}\right)^2 \frac{R_L}{2} \tag{3.26}$$

Substitute equation (3.17) into equation (3.26),

$$R_L = \frac{2R_{dc}}{\left(1 + \frac{\pi^2}{4}\right)} \tag{3.27}$$

Using equations (3.22) and (3.27)

$$C_1 = \frac{1}{\pi \omega R_{dc}} = \frac{1}{5.4466 \omega R_L} \tag{3.28}$$

The values of the series inductor L_0 and capacitor C_0 in the output resonator can be calculated using the formula for RLC networks,

$$L_o = \frac{Q R_L}{\omega} \tag{3.29}$$

$$C_o = \frac{1}{\omega Q R_L} \tag{3.30}$$

At the switching frequency, *f*, there is no voltage drop across the ideal series-tuned circuit in Figure 3.1. Therefore, the voltage at the output of the ideal resonator, $v_1(\theta)$, which is a sinewave of phase φ_1, is the fundamental of the drain voltage, $v_c(\theta)$, and its amplitude is given by

$$c_1 = \frac{1}{\pi}\int_0^{2\pi} v_c(\theta)\sin(\theta+\varphi_1)d\theta \tag{3.31}$$

The fundamental component of $v_c(\theta)$ has no cosine (quadrature) component with respect to phase φ_1,

$$\frac{1}{\pi}\int_0^{2\pi} v_c(\theta)\cos(\theta+\varphi_1)d\theta = 0 \tag{3.32}$$

$$\frac{2c\cos(\varphi)\sin(\varphi_1)}{\pi B R_L} - \frac{2I_{dc}\cos\varphi_1)}{\pi B} - \frac{I_{dc}\sin(\varphi_1)}{B} + \frac{c\cos(\varphi_1-\varphi)}{2BR_L} = 0 \tag{3.33}$$

$$\frac{c}{I_{dc}R_L} = \frac{2+\pi\tan(\varphi_1)}{2\cos(\varphi)\tan(\varphi_1)+\frac{\pi}{2}\cos(\varphi)+\frac{\pi}{2}\sin(\varphi)\tan(\varphi_1)} \tag{3.34}$$

Substitute equations (3.13) and (3.15)

$$\tan(\varphi_1) = \frac{\pi}{2} - \frac{4}{\pi} \tag{3.35}$$

From equations (3.5) and (3.16)

$$\frac{X}{R_L} = \tan(\varphi_1 - \varphi) \tag{3.36}$$

$$= \frac{\pi}{8}\left(\frac{\pi^2}{2} - 2\right) \tag{3.37}$$

Finally the series reactance can be obtained as

$$X = R_L\frac{\pi}{8}\left(\frac{\pi^2}{2}2\right) = 1.1525R_L \tag{3.38}$$

3.4 Design Example

Using the procedure mentioned in Section 3.3, a Class E power amplifier was designed. The design goal was to achieve 500mW at a supply voltage of 5V operating at 5.7 GHz.

Table 3.1: The Parameters of The Class E PA

Parameter	Value	Equation
V_{DD}	5 V	Assumed
P_{out}	500mW	Assumed
Q	5	Assumed
R_{dc}	50 Ω	Equation 3.24
R_L	28.84 Ω	Equation 3.27
C_1	0.18 pF	Equation 3.28
C_0	0.19 pF	Equation 3.30
L_0	4 nH	Equation 3.29
X	33.3 Ω	Equation 3.38

Simulation of the circuit performed with the commercial software ADS (Advanced Design System) by Hewlett Packard. Ideal models are used for all the passive components and FLK052 MESFET model is used for the transistor. Table 3.1 summarizes the parameters of the class E, their values and the equation used to calculate them.

The real Class E circuit used in this paper is slightly different compared to the circuit in figure 1. First the parallel capacitor C_p is completely absorbed by the output capacitance of the transistor. Secondly the R_F load, which is about 28.84 Ω, is transformed to 50 Ω by a 2-stage transformation network. The input match of the amplifier was implemented with a series inductor and a parallel capacitor. Finally the Q_L-value of the series resonant network was chosen to be low, of the order 5, in order to have low sensitivity of the circuit to the series resonator values. The complete Class E circuit can be seen in Figure 3.

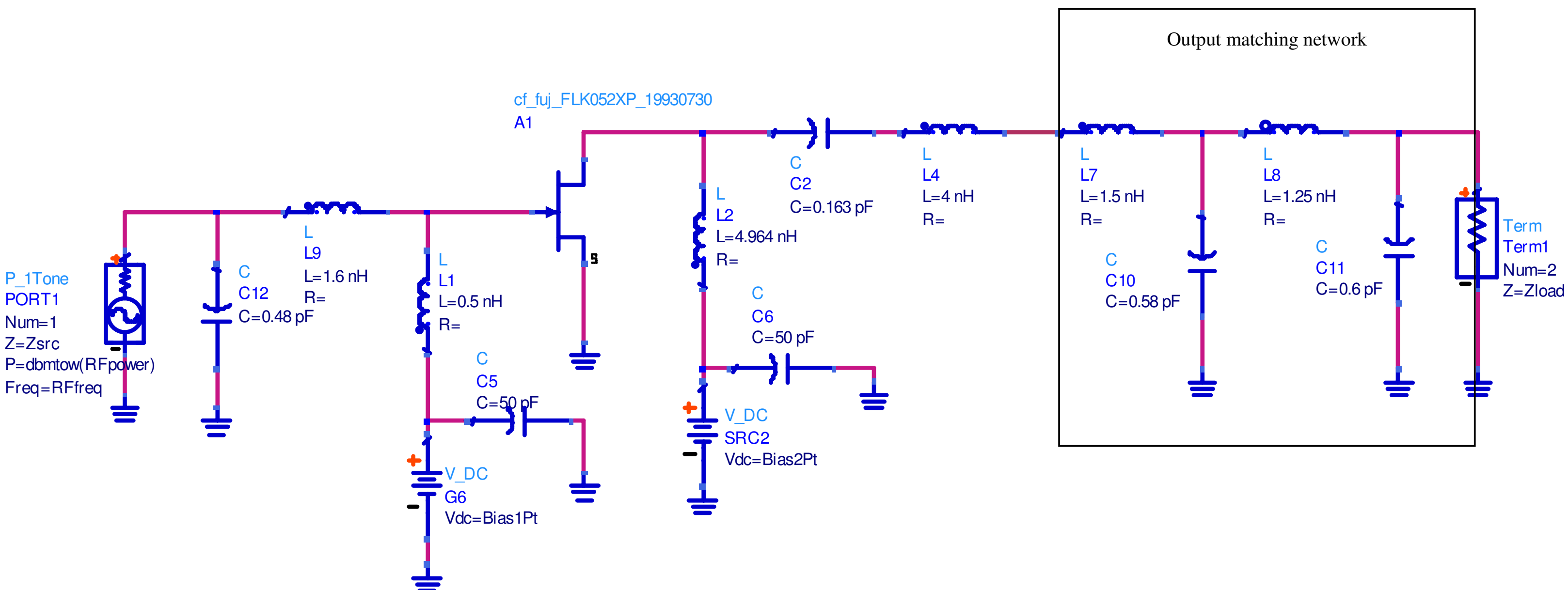


Figure 3.3: Designed Circuit Diagram of Class E Amplifier at 5.7 GHz

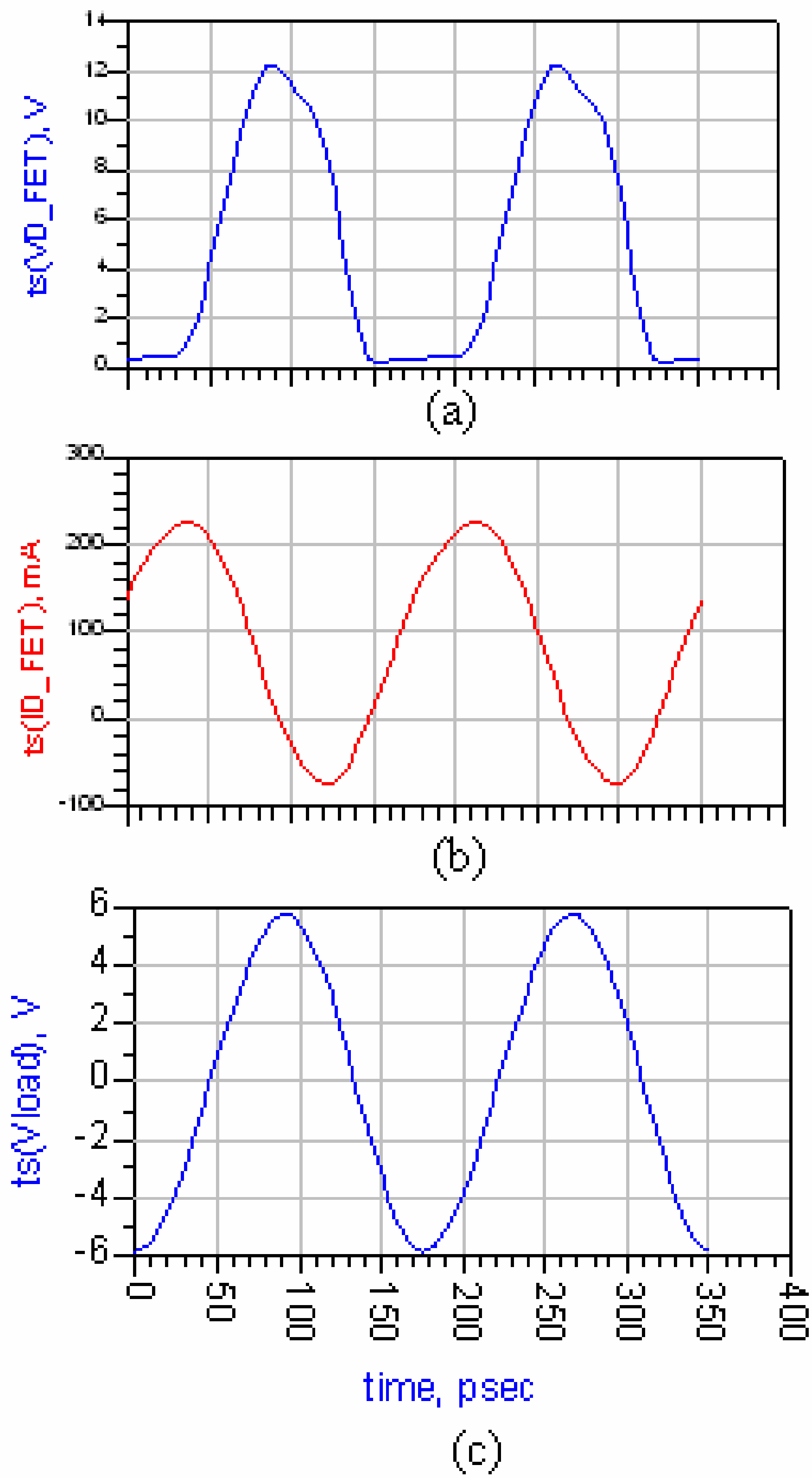


Figure 3.4: Voltage and Current Waveforms of Lumped Elements Based PA a) Drain Voltage VD_FET (V) b) Drain Current ID_FET (mA) c) Output Voltage Vload (V)

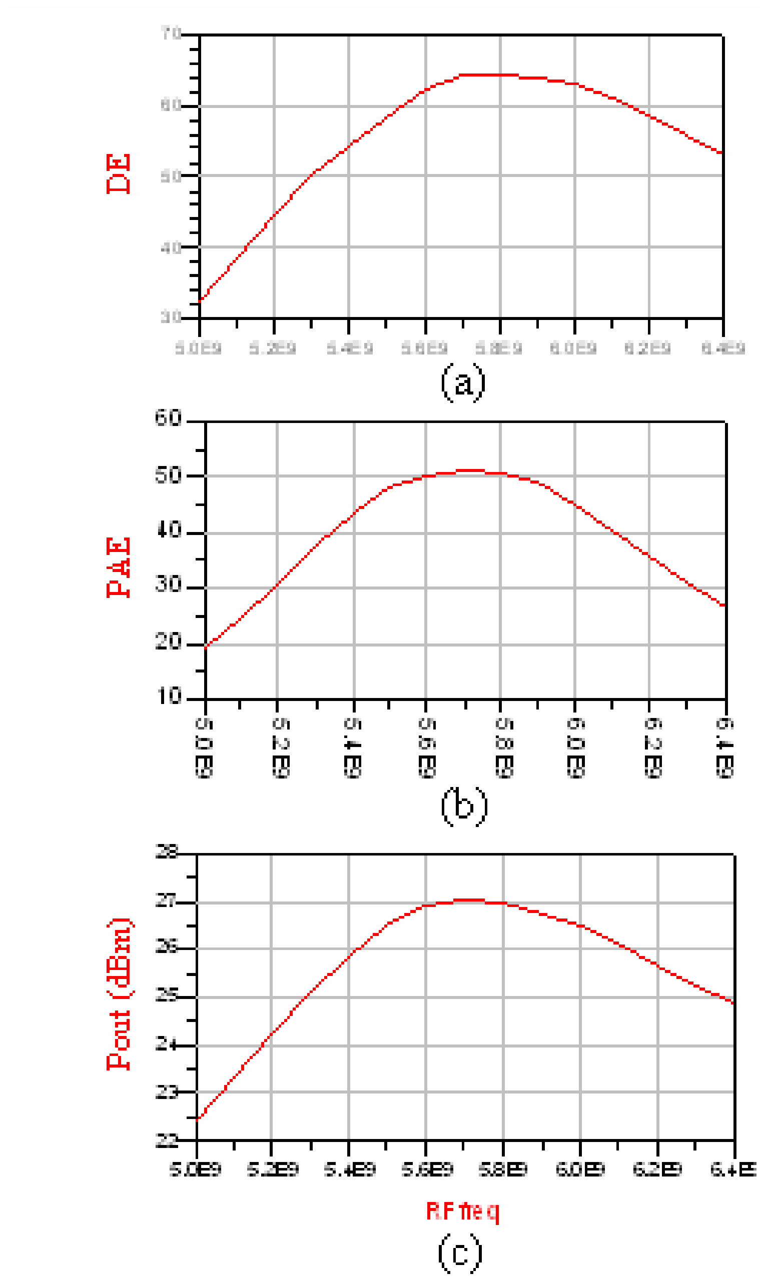


Figure 3.5: P_{out} and Efficiency Waveforms of The Lumped Elements Based PA: a)Drain Efficiency (DE) b) Power Added Efficiency (PAE) c) Output Power (Pout(dBm))

The lumped-element design generally works well at low frequencies, but two problems arise at microwave and millimeter-wave frequencies. First, lumped elements such as inductors and capacitors are generally available only for a limited range of values and are difficult to fabricate at the microwave and millimeter frequencies. In addition, at the microwave and upper frequencies the distances between a circuit's components are not negligible. For these reasons, transmission lines are often preferred over lumped elements at the microwave and upper frequencies.

Based on Richard's transformation, a shunt inductor can be replaced with a short-circuited stub, while a shunt capacitor can be replaced with an open-circuited stub. Moreover, a series element (inductor or capacitor) can be transformed to a shunt one using a transmission line.

Figure 3.6 shows a transmission- line class-E amplifier. This amplifier is designed based on the amplifier shown in Figure 3.3. The inductors are replaced with microstrip lines using Richard's transformation method and series capacitor is replaced with a microstrip interdigital capacitor. The lengths and widths are adjusted to give better performance. Figure 3.7 and 3.8 shows the performance of the class-E amplifier.

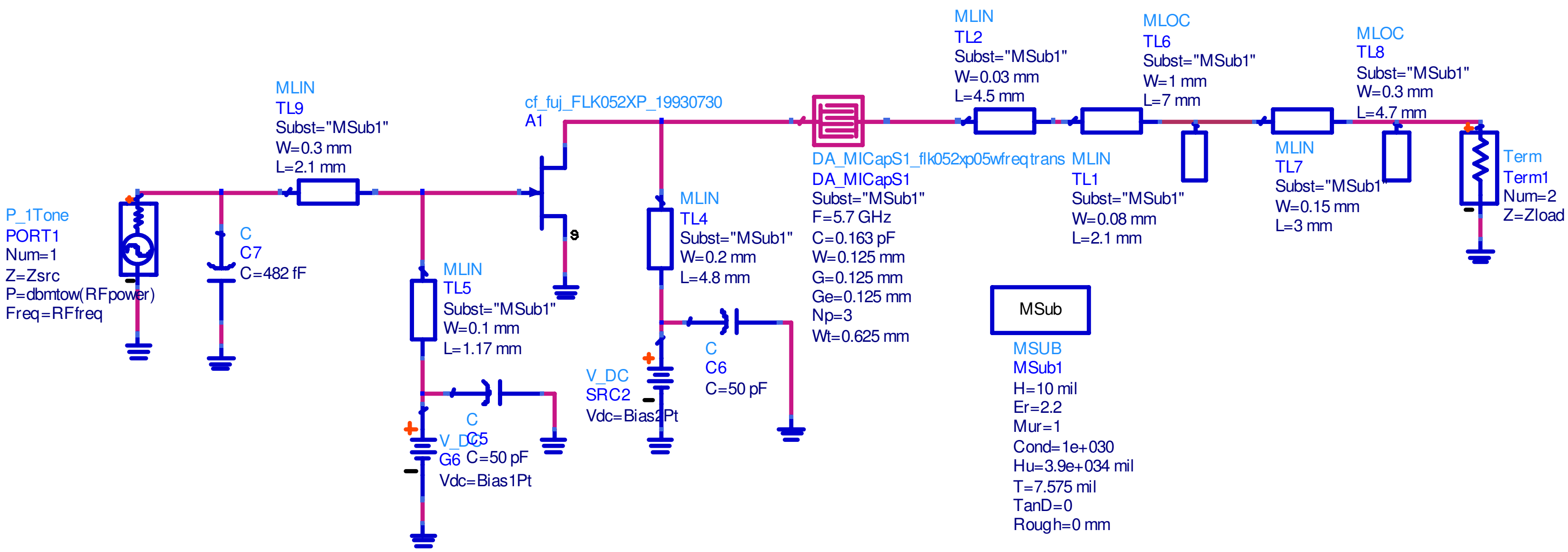


Figure 3.6: Transmission-Line Class E Amplifier at 5.7 GHz

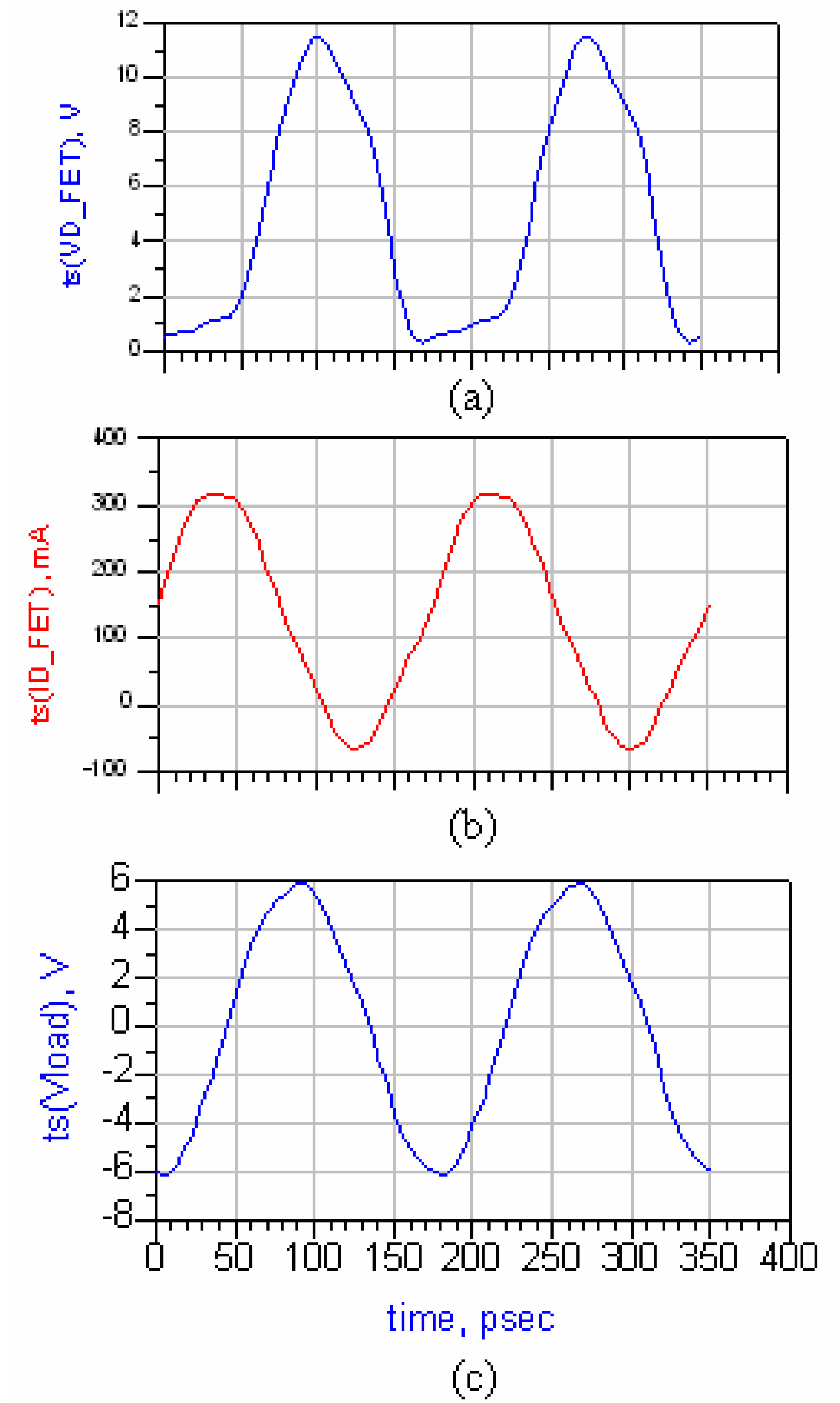


Figure 3.7: Voltage and Current Waveforms of Transmission-Line Based PA a) Drain Voltage VD_FET (V) b) Drain Current ID_FET (mA) c) Output Voltage Vload (V)

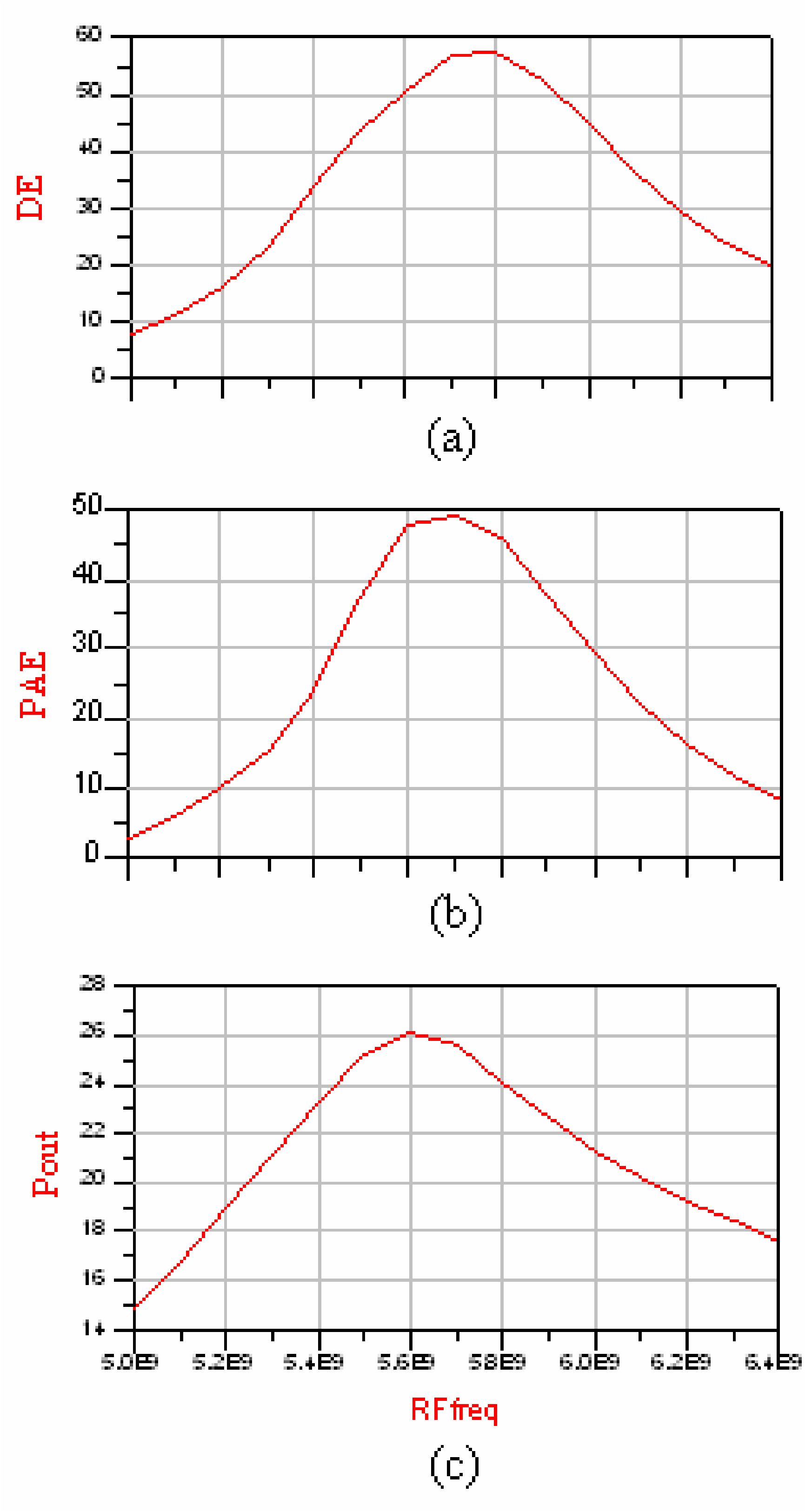


Figure 3.8: P_{out} and Efficiency Waveforms of Transmission-Line Based PA a) Drain Efficiency (DE) b) Power Added Efficiency (PAE) c) Output Power (Pout (dBm))

Figure 3.4 and Figure 3.7 show the output voltage and the drain voltage and current waveforms of the class-E amplifier at 5.7 GHz. As a result of the non- ideality of the transistor switch, an overlap occurs between the drain voltage and current. This overlap causes a power dissipation that degrades the efficiency.

Figure 3.5 and Figure 3.8 show the performance of both lumped element and transmission-line class E amplifiers. The PAEs are around 50%. The output powers are 27 dBm and 26 dBm respectively which are matched with our assumption.

4. LINEARIZATION TECHNIQUES

4.1 Introduction

Amplifiers which possess theoretical power efficiencies of 100% do exist. These amplifiers are used in a “switched” mode to perform the operation of single pole and double pole switches. The currents and voltages through the devices are controlled by the drive signals and tuned output filters to ensure that a finite current and a finite voltage on the device output cannot occur simultaneously. Thus no power is dissipated in the active device. Unfortunately this highly desirable characteristic comes bound with the undesirable characteristic of a highly non-linear transfer function. When amplifying signals with varying envelopes, effciency is less of a concern than signal distortion, which renders high efficiency PAs useless due to their inherent nonlinear behavior.

Using these classes of amplifiers requires a linearization technique to reduce the distortion of the output signal to an acceptable level. This chapter provides an overview of the types of linearization techniques available.

4.2 RF Feedback

The most commonly used linearization technique is the traditional direct negative feedback. Such a system can represented as shown in Figure 4.1. The RF signal is input to a subtractor on the left side in Fig. 4.1. The output signal of the amplifier, on top of Fig. 4.1, is fed back to the subtractor and subtracted from the RF input signal. The feedback network, at the bottom of Fig. 4.1 can either be passive or active. An amplifier can be used for an active feedback network or resistors or transformers can be deployed as passive feedback networks. The feedback network can reduce distortion appearing at the output of the nonlinear amplifier.

Voltage-controlled current feedback and current-controlled voltage feedback are commonly used for this method because they are simple and their distortion performance is predictable. However, due to the time delays in the feedback network, loop stability is a problem in this design. As a result, RF feedback is limited to narrowband systems. A loss of gain is also an issue in this technique, particularly in transmitters, where high output power is desired [4].

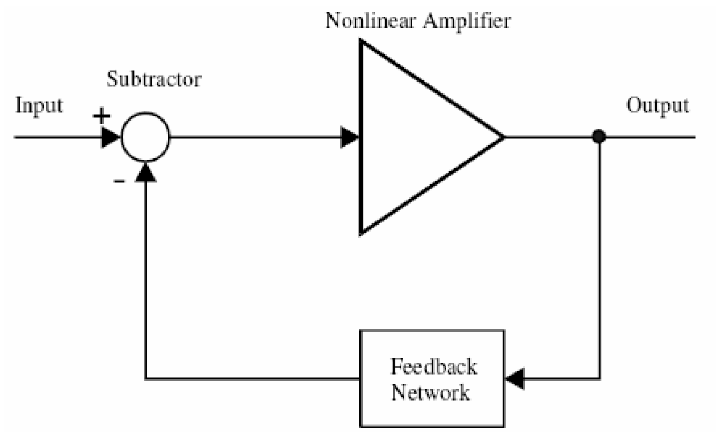


Figure 4.1: RF Feedback Linearization Method

4.3 Cartesian Loop Feedback

Cartesian Loop is a form of feedback that involves linearization of the complete transmitter (Figure 4.2). Baseband I and Q signals are upconverted to the carrier frequency and then amplified to the desired power level. This signal is then sampled and downconverted back into quadrature components. The resulting I and Q signals are fed back to the transmitter input, where they are compared to the original baseband inputs with error amplifiers.

With this technique, any nonlinearity in the transmitter is effectively cancelled out. The entire process of upconversion and any intermediate stages of amplification are included in the linearization process. One of the main drawbacks of this technique is a limited bandwidth due to delay around the loop. Thus, a compromise must be made between the bandwidth of the feedback loop and linearity improvement. Due to the addition of the feedback demodulators and error amplifiers, the PAE of this system is generally not improved unless the additional components can be implemented in an IC with low power dissipation and a high efficiency power amplifier (e.g. Class C) is used.

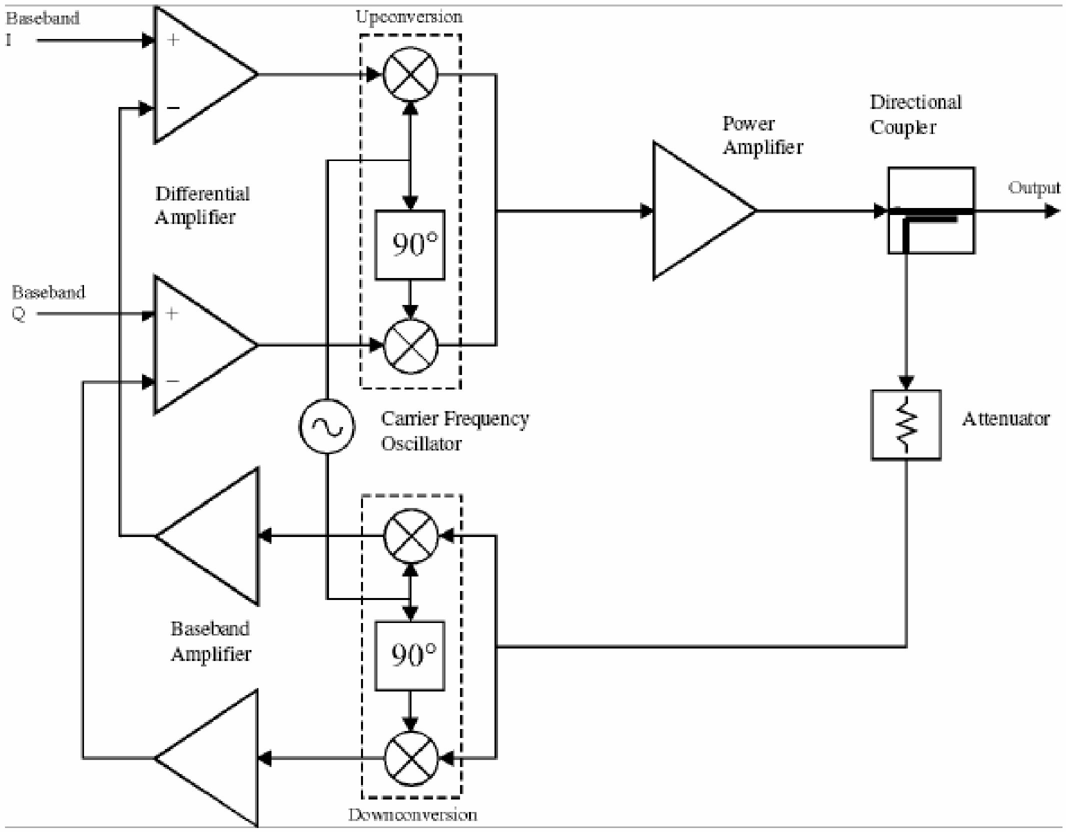


Figure 4.2: Cartesian Loop Feedback Linearization

4.4 Polar Loop Feedback

Another form of feedback that may be used for linearization is polar loop. This technique is similar to the Cartesian loop except that amplitude and phase are fed back rather than I and Q. A sample of the output signal is downconverted to a convenient IF by the local oscillator (LO) at the right bottom corner of Fig. 4.3. The IF signal is then separated into phase and amplitude by a limiter and demodulator, respectively. On the left side in Fig. 4.3, a single sideband (SSB) modulated signal, as input signal, is split and separated into phase and amplitude by a limiter and demodulator, respectively, similar to the sampled output signal. In the center of Fig. 4.3, both output and input amplitudes are compared with an error amplifier and the resulting error signal controls a modulation amplifier. The phase signals of input and output are multiplied utilizing a mixer. The resulting signal controls a voltage-controlled oscillator (VCO) after passing a loop filter and being amplified. The new-formed phase and amplitude error signals are combined with a modulating amplifier. Finally, the combined signals are amplified with a power amplifier [4].

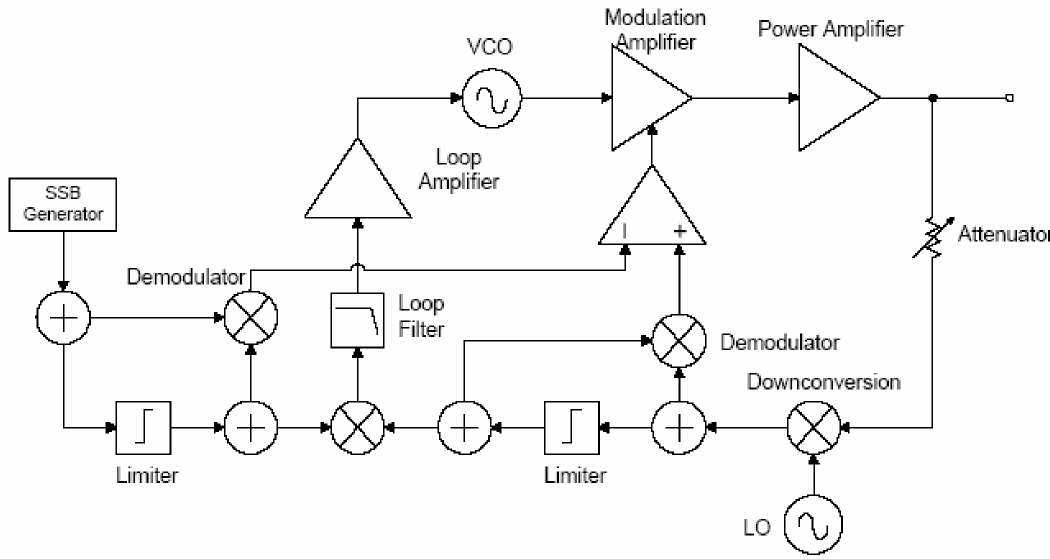


Figure 4.3: Polar Loop Feedback Linearization

A resulting problem with this method is that the required feedback bandwidths for the amplitude and phase components are different from each other for most modulation formats. This limits the available loop gain to either the amplitude or phase path since one path will require a feedback bandwidth that reduces the available loop gain, while the other path may need a larger loop gain. This effectively limits the overall linearity improvement. Essentially, the operation of the phase-feedback path relies on a phase-locked loop; the loop can experience locking problems at low amplitude levels and also have problems tracking abrupt changes in phase.

4.5 Feedforward

The linearized PA with feedforward correction consists of a main and error amplifier, directional couplers, and delay lines. The incoming signal is split into two paths with one path going to the main amplifier, and the other path going to a delay element, shown in Figure 4.4 on the right side. The signal at the output of the nonlinear amplifier contains the desired and distortion components. The delayed input signal is subtracted from a sample of the output signal of the nonlinear amplifier. The sample of the output signal of the nonlinear amplifier is taken by means of a coupler at the top of Fig. 4.4. Ideally, the carrier is canceled and the signal at the output of the subtractor contains only the distortion component. The error signal is amplified by an amplifier depicted at the bottom of Figure 4.4 and combined with a delayed version of the output signal of the main amplifier. This second combination ideally cancels the distortion components of the main amplifier and leaves the desired signal unchanged. The first loop cancels the carrier and the second loop reduces the distortion component.

The feedforward technique doesn't ideally reduce the amplifier gain. It is also independent from amplifier delays unlike the feedback technique. The correction is based on the current events, not on past events (feedback). The feedforward loop is unconditionally stable, which is not the case with feedback systems.

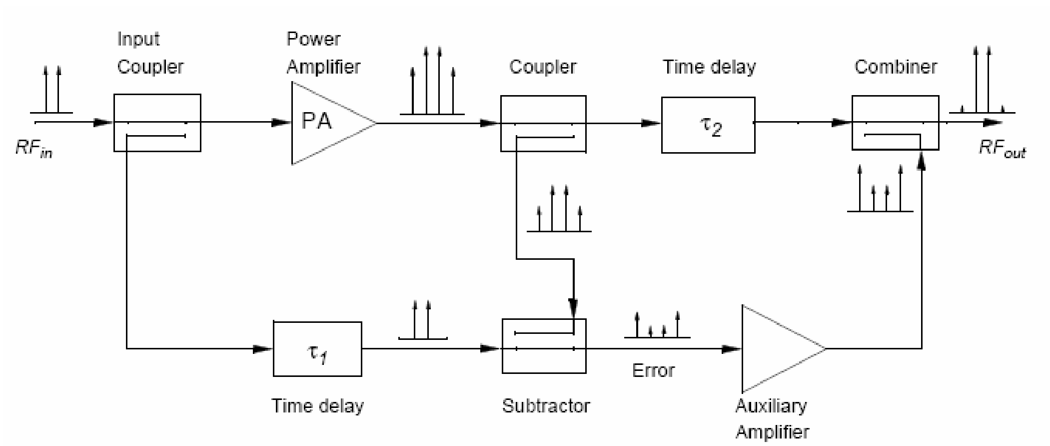


Figure 4.4: Feedforward Linearization

Feedforward can be used over a wide bandwidth of about 10-100 MHz. However, inaccurate matching of devices in amplitude and phase can impair the performance of the feedback system. Since this system is feedforward in nature, alterations over aging and temperature degrade the correction of linearity. A way to reduce these effects is to use multiple feedforward loops [4].

4.6 Analog Predistortion

The analog predistortion technique connects a predistorting amplifier in front of the main amplifier. Compared to the compressive main amplifier, this additional amplifier has the opposite output distortion characteristic, i.e. its nonlinearity is expansive, not compressive. These two nonlinear distortions cancel each other when summed, resulting in a linear and distortion-free output from the main RF amplifier. Predistortion can be accomplished at either RF, IF or baseband and has the ability to linearize the entire bandwidth of an amplifier or system. There is also the possibility to cascade the distortion element at the output of the PA. This case is called postdistortion [4].

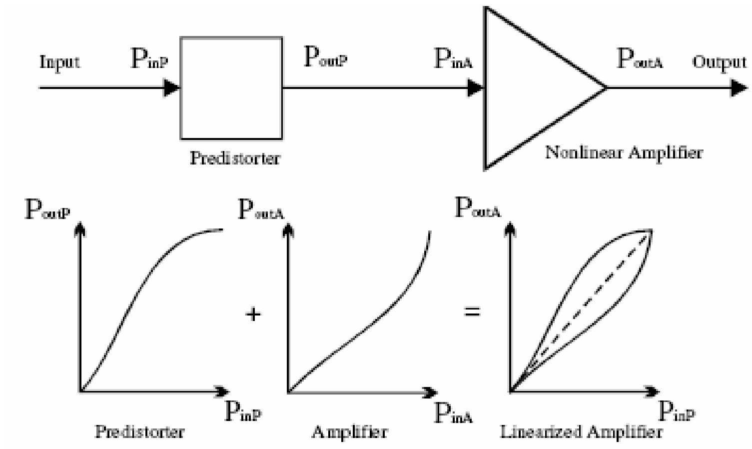


Figure 4.5: Analog Predistortion Linearization

This conceptually simple technique can in linearize a PA system quite well, if the predistorter and main amplifier are sufficiently well matched. Because of its open-loop nature there are no stability issues and the usable bandwidth is much greater than that of closed-loop feedback systems. Because of its simplicity, the technique can be applied to microwave circuits without too great difficulties.

The predistorter and main amplifier have to automatically match up perfectly, because there is no inherent adaptivity in this technique. There is also AM-PM distortion in the RF amplifier, so in addition to cancelling the AM distortion the predistorter should cancel the PM distortion as well. Therefore this technique is not very robust from the manufacturing point of view.

4.7 Adaptive Baseband Predistortion

A second type of predistortion is called adaptive baseband predistortion (also called digital predistortion). Digital predistortion is basically a Cartesian feedback with digital signal processing (DSP) added. This technique linearizes complete transmitters. Fig. 4.6 displays an adaptive baseband predistortion scheme. Predistortion occurs at the baseband level and manipulates usually I and Q components. On the right side of Fig. 4.6, the output signal is sampled, downconverted by a QAM down-converter (into components I and Q), and input to a digital signal processing (DSP) unit. Similarly, input I and Q components are directly fed into the DSP block. In order to create predistortion, weighting coefficients are stored in lookup tables in the DSP section. These coefficients can be updated by new coefficients derived from the fed back in-phase and quadrature components. The output of the DSP is the predistorted signal. This predistorted signal is then up-converted by a QAM and amplified with a PA illustrated at the top of Fig. 4.6.

The primary disadvantages of digital predistortion are its relative complexity and bandwidth limitations tied to the accuracy and computational rate of the specific DSP [25]. Furthermore, power consumption is increased due to the digital signal processor.

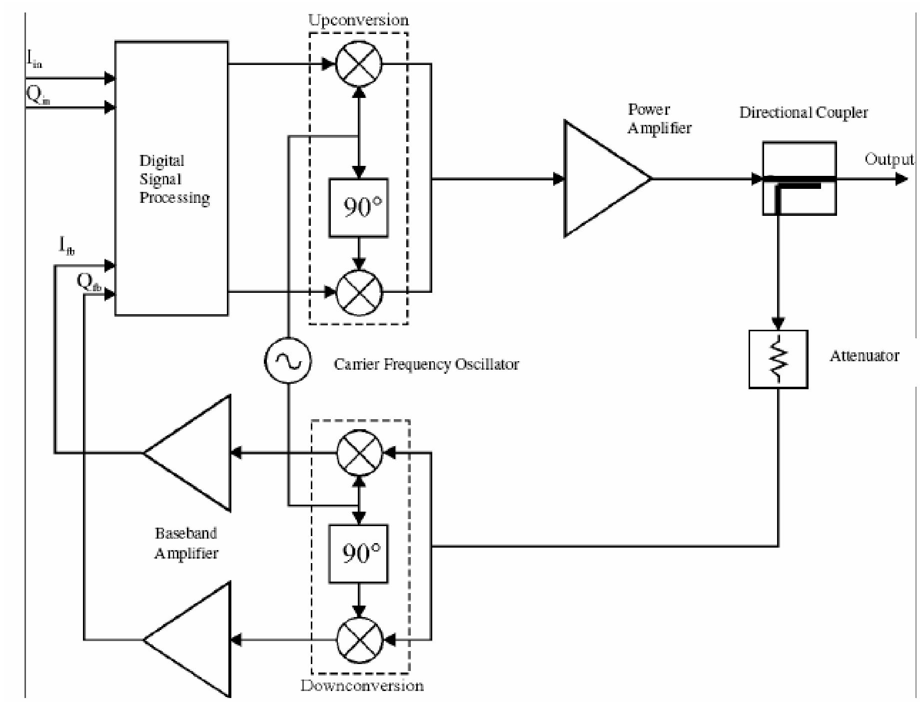


Figure 4.6: Adaptive Baseband Predistortion Linearization

4.8 Envelope Elimination and Restoration

Figure 4.7 shows the block diagram of the EER linearization scheme as first proposed by Khan. The EER technique consists of a limiter and envelope detector to extract the phase and envelope information, respectively. The limiter, shown at the

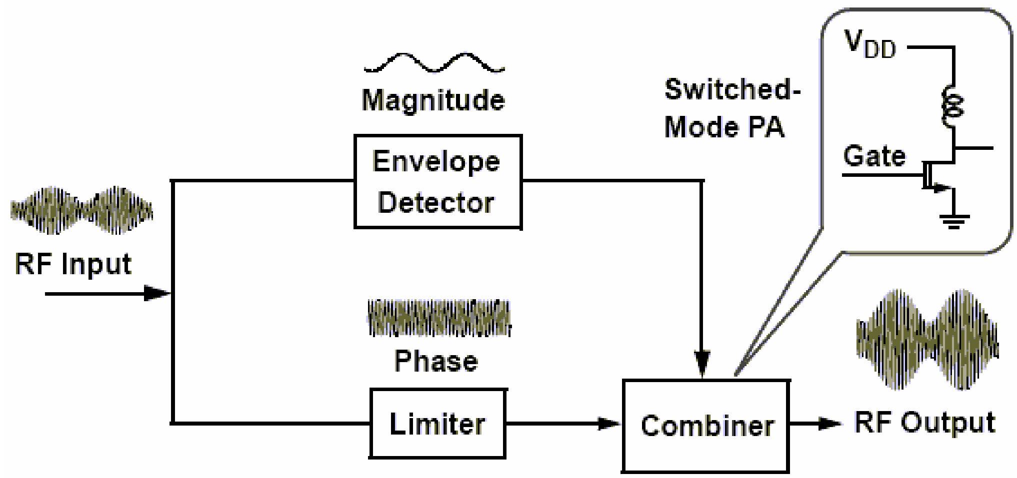


Figure 4.7: Envelope Elimination and Restoration linearization

bottom left of Fig. 4.7, eliminates the envelope and thus makes it possible for a high-efficient nonlinear PA to amplify the constant-envelope signal. Finally, the envelope amplifier (class-S modulator), depicted on the top of Fig. 4.7, modulates the final RF power amplifier and creates an amplified replica of the input signal at the output. A detailed analysis of the EER system is reserved for Chapter Five.

4.9 Linear Amplification using Non-Linear Components

The LINC system (Linear Amplification using Non-Linear Components) is fundamentally different from the previously mentioned linearization methods in that there is no feedback used and the amplifier itself can be highly non-linear. The LINC technique was originally called "Outphasing". These systems were first developed during the 1930's by H. Chireix [34] to improve the efficiency and linearity of AM-broadcast transmitters.

The basic principle for the LINC system is to separate the baseband input signal, which contains either or both amplitude and phase information, into two constant amplitude signals. These two constant amplitude signals can be separately amplified by a pair of highly non-linear amplifiers. The amplified component signals are passively re-combined to produce an amplified replica of the input signal. A block diagram of a simple LINC system is shown in Figure 4.8.

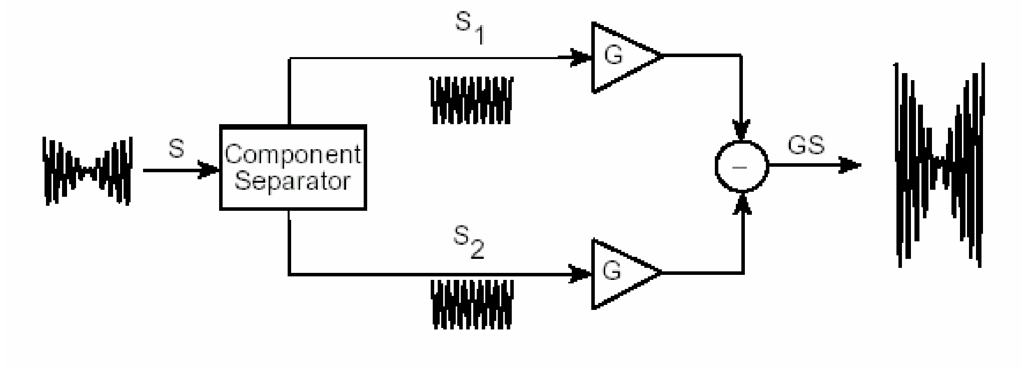


Figure 4.8: LINC Linearization

To gain a basic understanding of the system operation, consider first the input signal.

$$S(t) = E(t)\cos(\omega_0 t) \tag{4.1}$$

where: $E(t) = E_m \sin\phi(t)$ and is defined as the envelope

E_m is the maximum value of the envelope

The component separator produces the two constant amplitude signals S1 and S2 which are related to the input signal as follows:

$$S_1(t) = \frac{E_m}{2}\sin[\omega_0 t + \phi(t)] \tag{4.2}$$

$$S_2(t) = \frac{E_m}{2}\sin[\omega_0 t - \phi(t)] \tag{4.3}$$

The output is generated after amplification of the two constant amplitude signals and recombination.

$$GS_1(t) - GS_2(t) = GE_m \sin\phi(t)\cos\omega_0 t = GS(t) \tag{4.4}$$

The LINC technique is susceptible to amplitude and phase differences in each of the paths. Differences in these paths can severely degrade system performance and some form of feedback is usually necessary in order to compensate for variations in the amplifiers.

5. ENVELOPE ELIMINATION AND RESTORATION

5.1 Introduction

The EER technique was originally designed by L. R. Kahn during the 1952's and has been initially used in base stations to increase linearity [24]. The first practical system using EER was a 100 kW AM transmitter which, at the time, was the most powerful single-sideband transmitter in operation.

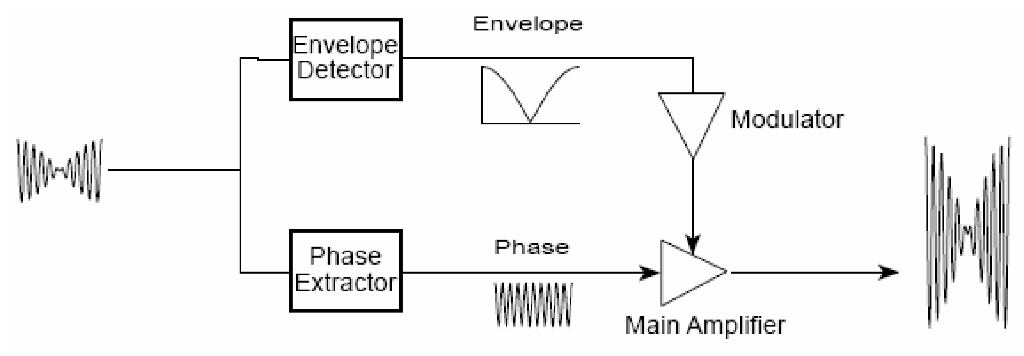


Figure 5.1: EER System Block Diagram

EER is a technique which tries to increase linearity and power efficiency simultaneously. This technique combines a highly efficient but nonlinear RF power amplifier (PA) with a highly efficient envelope amplifier to implement a high-efficiency linear RF power amplifier. As the name "envelope elimination and restoration" implies, the envelope of the RF input is first eliminated by a limiter to generate a constant-amplitude phase signal. At the same time, the magnitude information is extracted by an envelope detector. The magnitude and phase information are amplified separately and then recombined to restore the desired RF output. A way to recombine the magnitude and phase components is to use an efficient switched-mode RF PA. The envelope of the RF output of a switched-mode PA is directly proportional to its supply voltage. The envelope and phase components can therefore be recombined if the phase signal (RF) is applied at the gate of the transistor and the magnitude signal (low frequency) directly modulates the supply. A block diagram of a simple EER system is shown in Figure 5.1.

Consider the RF input signal $a(t)e^{j\varphi(t)}$ is applied to the system. Here a(t) is the amplitude and φ(t) is the phase information of the signal. The envelope detector extract the magnitude information a(t) of the input signal and φ(t) is equal to zero. The envelope signal a(t) is then amplified by the modulator so that A(t) is used to drive the high efficiency main amplifier. At the same time, the limiter at the bottom side of the system eliminates the envelope information. Only the constant-amplitude phase information $e^{j\varphi(t)}$ is obtained at the output of the limiter. A(t), the amplifier envelope signal and $e^{j\varphi(t)}$, constant-amplitude phase signal are recombined using an efficient switched-mode RF PA. The phase signal is used to initiate the switching action of the highly non-linear main power amplifier. The amplified envelope component is used to control the power output from the main power amplifier. The active recombination of the phase and envelope signals in the main power amplifier generates $A(t)e^{j\varphi(t)}$ amplified replica of the input signal.

The key advantage of this EER approach is that the RF PA always operates as an efficient switched-mode amplifier. That is, the EER system can linearize the switched-mode RF PA without compromising its efficiency.

One major concern in the EER system is that the mismatch between the total phase shift and gain of the two paths should be minimized, which is hard to achieve because the delay in RF path is smaller than the low-frequency amplitude path. It can be shown that the intermodulation term due to the delay mismatch is given by [23]:

$$IMD \approx 2\pi B_{RF}^2 \Delta\tau^2 \qquad \textbf{(5.1)}$$

Second, the dc controller that generates control current and voltage for the PA may not operate at 100% effciency and has limited bandwidth which may not be adequate for multichannel signal. Other problems are the limiter circuit and nonlinear capacitance of the PA device that, when operated at large-signal conditions, may introduce undesirable AM-PM distortions.

In the following sections, the components of the EER system will be discussed and a linearization example will be presented using the Class E Power Amplifier which is designed in chapter 3. A transient simulation is used to simulate this system between

200 ns and 400 ns. A 5.7 GHz AM signal with a bandwidth of 10 MHz is applied to the input of the EER system.

5.2 Envelope Detector

The purpose of the envelope detector is to extract the amplitude information embedded in the combined input signals. Figure 5.2 shows the envelope detector circuit. The envelope detector consists of a transistor M1 operating as a diode, capacitor C1, and current source I1. Unfortunately, the gate-to-source voltage of transistor M1 is rather large and introduces distortion. A first-order cancellation of the dc voltage and distortion of transistor M1 is provided by a pseudoreplica circuit consisting of transistor M2 and current source I2. Amplifier Amp keeps the voltages at nodes and equal by driving the gate of transistor M2 to be equal to the envelope of the RF input signal at the gate of M1. A key feature of this envelope detector is that the pseudoreplica circuit only needs to operate at the envelope frequency, not at RF[23].

A 5.7 GHz AM signal with a bandwidth of 10 MHz is applied to the input of the circuit. The performance of this envelope detector in Fig. 5.3b shows that the 10MHz envelope of the RF signal can be extracted from an AM modulated RF input at 5.7GHz. The 10 MHz envelope signal obtained here can be used as an input signal of the modulator.

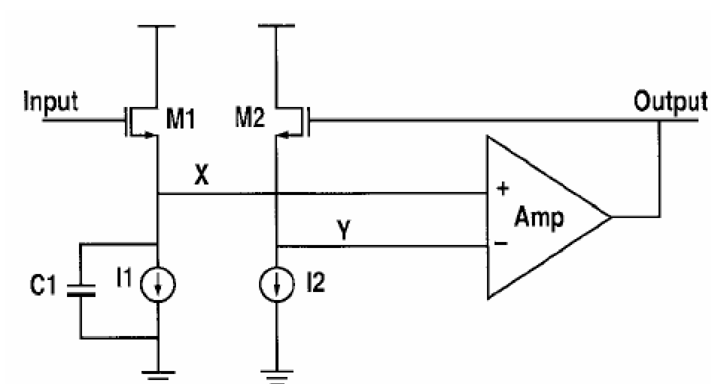
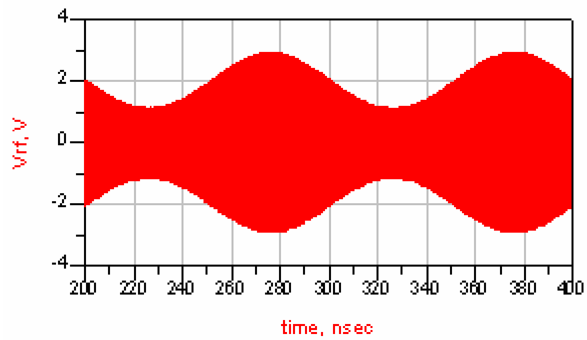
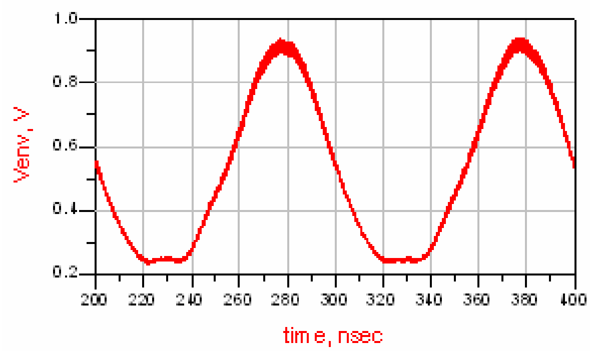


Figure 5.2: Basic Envelope Detector Circuit



(a)



(b)

Figure 5.3: Input and Output Voltages of The Envelope Detector Circuit: (a) AM Modulated RF Input Signal (b) Envelope Signal

5.3 Limiter

The purpose of the limiter is to remove the amplitude variation effects embedded in the input signal. In order to produce an output signal of constant amplitude, the limiter must apply a varying gain to the input signal; large gain when a low level signal is present and small gain when a high level signal is present. An ideal limiter can therefore be thought of as a variable gain device. The variable gain of the limiter is denoted the "Envelope Limiter Gain Function", ELGF, and is equal to the inverse of the envelope of the input signal [17].

$$\mathrm{elgf(t)} = \frac{1}{env(t)} \qquad \textbf{(5.2)}$$

Modulation of the input signal with the envelope limiter gain function removes the amplitude modulation and produces the phase only information shown in Figure 5.4(c).

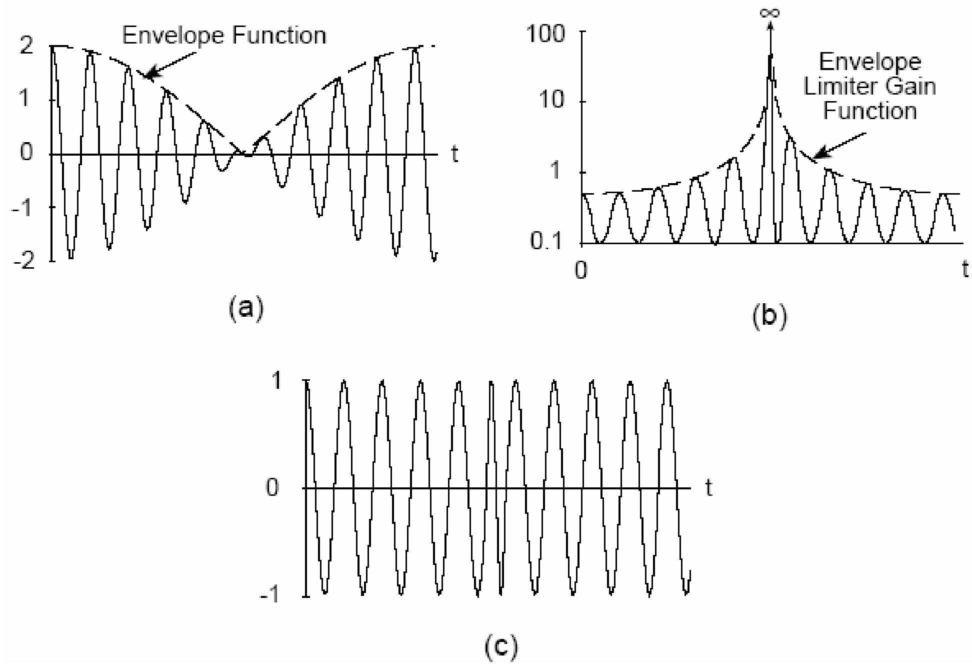


Figure 5.4: Ideal Limiter Waveforms (a) Input signal (b) Envelope Limiter Gain Function (c) Phase Information at Limiter Output

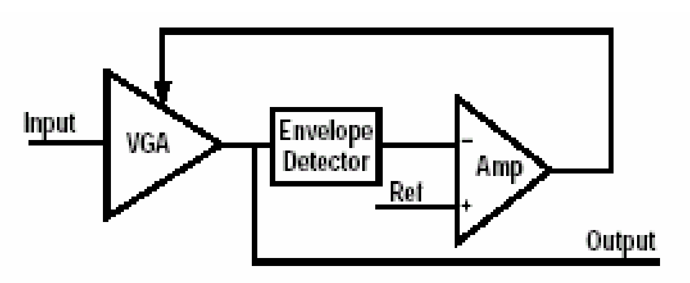
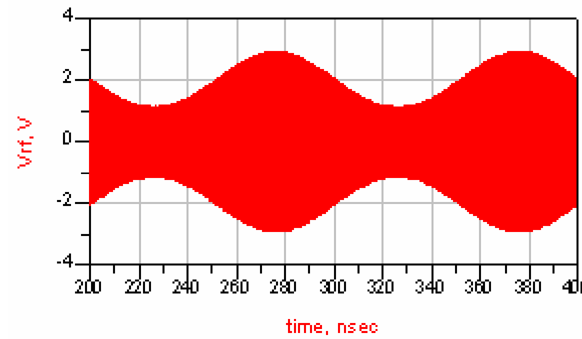


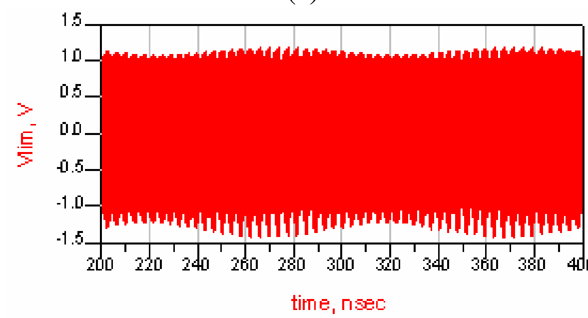
Figure 5.5: Limiter Block Diagram

The Block diagram of a limiter is shown in Fig 5.5. The circuit consists of a VGA, an envelope detector, and an error amplifier. The envelope of the VGA output is detected and compared with a dc reference signal. The error amplifier adjusts the gain of the VGA to provide a constant envelope RF output.

A 5.7 GHz AM signal with a bandwidth of 10 MHz is applied to the input of the circuit. The limiter is removed the envelope variation from an AM modulated RF input at 5.7GHz. It is seen from Figure 5.6 that the envelope variation of the limiter output is almost zero. Therefore, the phase signal obtained here can be applied to the input of the main amplifier.



(a)



(b)

Figure 5.6: Input and Output Voltages of Limiter Circuit (a) AM Modulated RF Input Signal (b) Phase Signal

5.4 Modulator

The purpose of the modulator is to provide sufficient gain to the relatively weak incoming envelope signal so that it may be used to drive the high efficiency main amplifier. The modulator must provide linear amplification of the entering envelope signal while also maintaining high efficiency. The modulator must also possess a bandwidth that exceeds the envelope signal bandwidth. A different type of switching amplifier, which readily provides these characteristics, is known as a Class S amplifier and is shown in Figure 5.7.

The transistor replicates the Pulse Width Modulation (PWM) signal, which has first been provided by comparing the envelope signal with TWG output signal. The lowpass filter averages the signal to provide an output proportional to the input. The active device never experiences non-zero voltage and current at the same time, as a result, the Class S modulator has a theoretical efficiency of 100%.

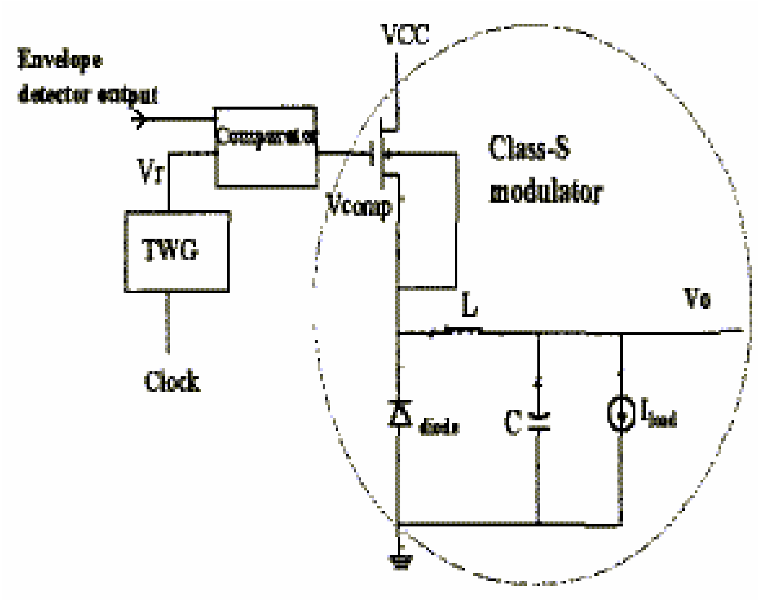
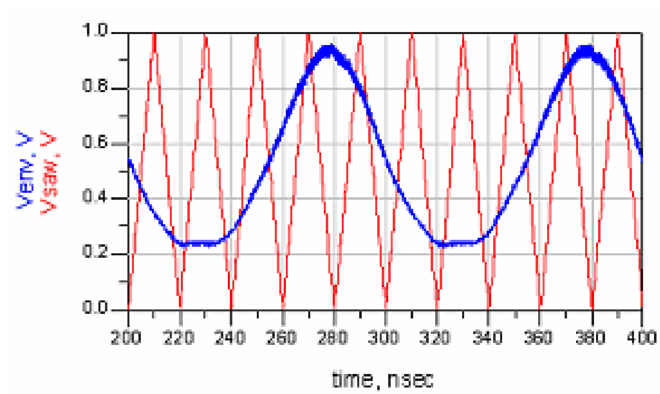
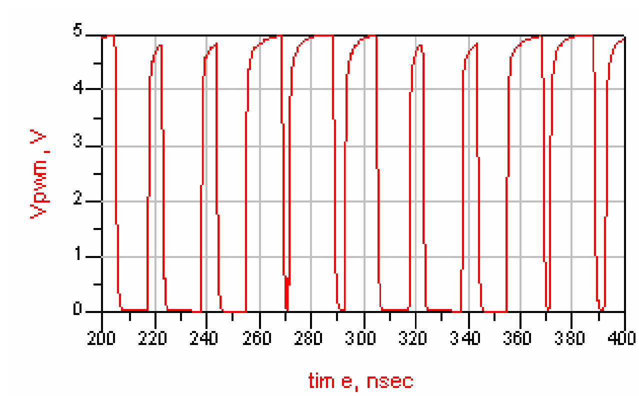


Figure 5.7: Modulator



(a)



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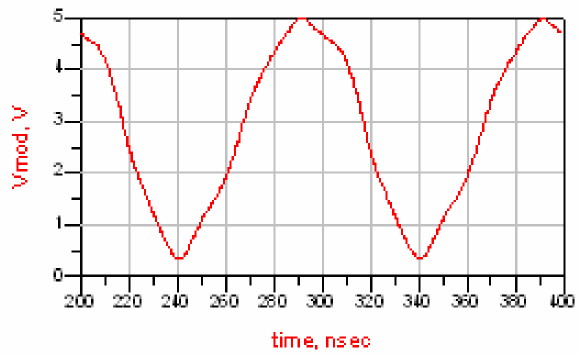

(c)

Figure 5.8: Modulator Waveforms (a) Sampling The Envelope Signal Using Triangular Waveform (b) PWM Signal (c) Output Signal of The Modulator

The Figure 5.8 (a) shows the reference triangular waveform. This triangular wave is obtained integrating a square wave with a simple RC integrator. Since this reference voltage is used for sampling the envelope signal, its frequency is chosen as 100MHz. The PWM signal, which is shown in Figure 5.8 (b), is then amplified using the switching-mode Class S PA. The output voltage of the modulator is shown in Figure 5.8 (c).

When operating from a 5V supply, the switching power supply can provide an output voltage that ranges from 0.3V to 4.98V. Output voltage range of the modulator can be adjusted by changing the dimensions of the transistor and the values of the passive elements. Now this signal can be used as the supply voltage of the Class E power amplifier. The output power of a switched-mode PA is directly proportional to $(V_{DD})^2$. That is, the EER system can linearize the Class E PA without compromising its efficiency.

Recombination of the envelope signal and phase signal at the main power amplifier allows for the possibility that a timing error may occur. The dominant source of delay in the envelope path is the lowpass filter of the modulator. To minimize the distortion a delay block is used to add a time delay to the phase signal. This delay cancels out the time delay produced by the modulator on the envelope signal. This block is mandatory to synchronize the modulated phase signal with the modulated amplitude signal before the power amplifier. Figure 5.9 shows the time delay (13 ns) between input and output of the modulator.

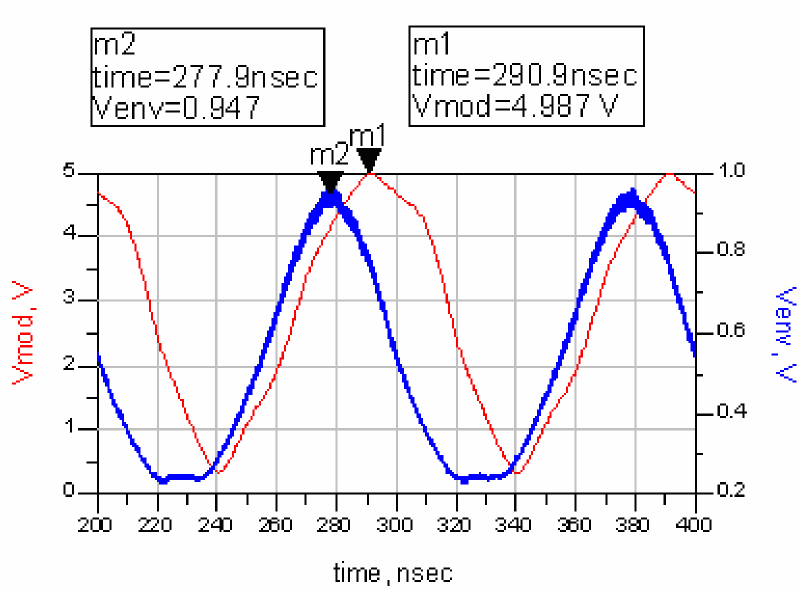

Figure 5.9: Input and Output Signals of The Modulator

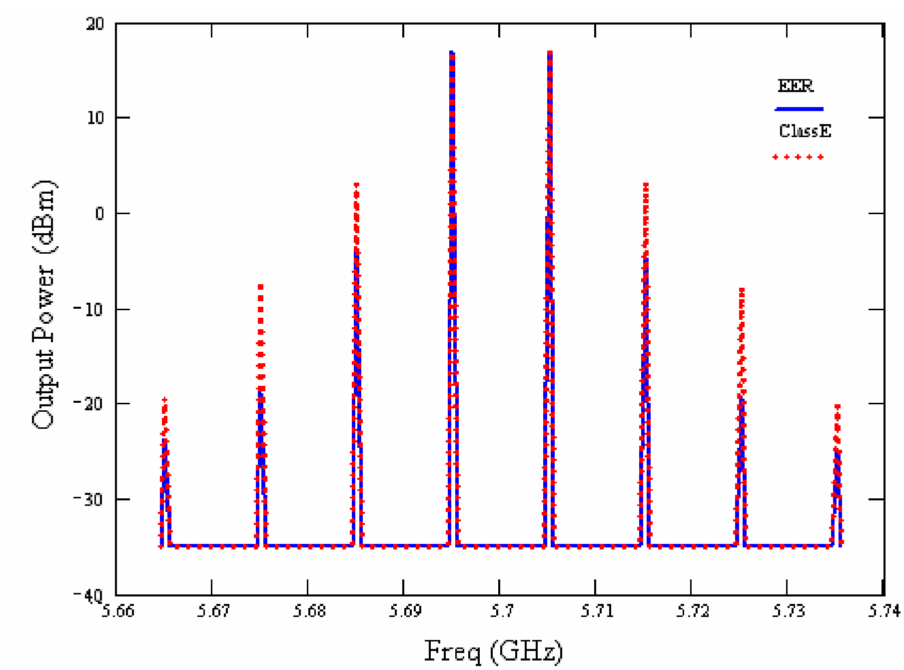

Figure 5.10: Performance Measurement with and without Distortion Reduction System Enabled

The final performance characteristic of the system is the amount of reduction of intermodulation distortion products with and without the modulator and phase extractor active. The Class E amplifier's performance is first measured with the phase extractor and modulator removed from the system. The Class E amplifier's main power is provided from a fixed dc power supply. The combined input signals are used to drive the Class E amplifier directly. In the second case the Class E amplifier performance is measured with the phase extractor and modulator active in the system. Figure 5.10 shows the system performance improvement when the phase extractor and modulator are active in the system.

The results of this final experiment indicate that the distortion reduction system provides an additional 7.5 dB reduction in intermodulation distortion products, achieving a minimum distortion level of 20 dB below the fundamental signals.

6. CONCLUSION

One of the challenging blocks in the designing the handset for wireless communication transceiver is the power amplifier. This thesis investigated the design and linearization of a 5.7 GHz Class E power amplifier using Envelope Elimination and Restoration.

Chapter 2 provided the basic concepts of the power amplifiers. Different power amplifiers were described and trade-offs and efficiencies were presented.

The efficiency of the power amplifier has a direct impact on the cost of the wireless communication system. Increasing the efficiency of the power amplifier in a handheld transmitter results in reduced DC power drain on the batteries, reduced handset weight resulting from reduced heat sinking requirements, and increased reliability due to reduced junction operating temperatures.

The focus of Chapter 3 was Class E power amplifiers. The Class E power amplifier is one of the switch-mode power amplifiers in which the transistor acts as a switch. Under ideal conditions, the transistor would dissipate no power because the voltage and current waveforms would never overlap. Therefore Class E amplifiers can achieve 100% ideal efficiency. The circuit operation is controlled by the load network when switch is on, and by the transient response of the network when the switch is off. The load network shapes the current and voltage waveforms such that minimum power is dissipated in the transistor itself. In the chapter 3, the basic operation and mathematical analysis of the class E power amplifier was presented. A design example based on the developed algorithm is described. Two class-E amplifiers operating at 5.7 GHz are presented. One of them is a lumped elements based circuit and the other is a transmission lines based circuit. Both circuit show good performance with 50% PAE and have 500mW output power.

A non-linear amplifier may cause distortion in the output signal by modification of the amplitude and or time delay information. These modifications result in output power compression, phase distortion and excess frequency terms known as intermodulation and harmonic distortion products.

As it was described in chapter 2, there is a trade off between the linearity and the efficiency. The linear amplifiers have poor efficiency on the other hand the switching amplifiers have poor linearity but high efficiency. Therefore if high efficiency and high linearity is necessary for a system, an additional linearization method for the switching amplifier should be used.

In Chapter 4 prior linearization methods were described. There have been many methods proposed for the purpose of amplifier linearization. These techniques can be grouped according to the following characteristics: 1) feedforward, 2) feedback, 3) predistortion and 4) signal separation and recombination. Feed forward techniques traditionally require two power amplifiers and high power combining devices. Feedback techniques offer a reduction of the distortion proportional to the loop gain but at the cost of reduced output gain. Predistortion techniques require complex modulation circuitry and or Digital Signal Processing (DSP) to achieve a reduction in the distortion level. For the strongly non-linear amplifiers only signal separation and recombination linearization techniques can be used to produce low distortion systems. These techniques offer reduced distortion with a simple open loop architecture.

The basic principle for the EER system is seperation of the baseband input signal, which contains both amplitude and phase information, into a phase only component and envelope only component, then recombination of these components in the main power amplifier to generate an amplified replica of the input signal. In chapter 5 the analysis of the EER system was presented and the PA designed in chapter 3 was linearized using this method. . Linearization Class E PA using EER system provides an additional 7.5 dB reduction in intermodulation distortion products, achieving a minimum distortion level of 20 dB below the fundamental signals.

The key advantage of this EER approach is that the RF PA always operates as an efficient switched-mode amplifier. That is, the EER system can linearize the switched-mode RF PA without compromising its efficiency. One major concern in the EER system is that the mismatch between the total phase shift and gain of the two paths should be minimized, which is hard to achieve because the delay in RF path is smaller than the low-frequency amplitude path.

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