CHAPTER 5

BROADBAND CLASS-E AMPLIFIER

5.0 Introduction

Class-E amplifier was first presented by Sokal in 1975. The applications of class-E amplifiers were limited to the VHF band. At this range of frequency, class-E amplifier has shown to exhibit efficiencies as high as 96% [Sokal, 1975]. A few years a go, it was shown that Class E amplifiers can be used at higher frequencies [T. Sowlati, et al, 994]. Several papers have reported class-E amplifiers operating at a frequency above the VHF band [T. Mader and Z. Popovic, 1995; F. Javier, et al, 1998; V. Gudimtla and A. Kain, 1999].

As stated earlier, a class-E is nonlinear amplifier, in the sense that variations in input signal amplitude will not reproduced at the output in any acceptable form. Moreover, class-E configurations prove to have higher efficiency with simpler circuits than conventional reduced conduction angle classes.

New lumped-elements and transmission-line based circuits are presented in this chapter. These circuits show good performance over a wide bandwidth of frequency.

5.1 Class E Operations and Analysis

Figure.5.1 shows an ideal class-E amplifier [R. Zulinsky and J. Steadman, 1987]. It consists of a switch *S*, a bias choke L_b, a capacitance C_s, a tuned circuit *L-C*, and a load R_L. The transistor switch *S* is ON in half of the period, and OFF in the other half. When *S* is ON, the voltage across *S* is zero, and when it is off, the current through *S* is zero. The capacitance C_s includes the parasitic capacitance across the transistor. The *L-C* circuit resonates at the fundamental frequency of the input signal and only passes a sinusoidal current to the load R_L. Figure.5.2 shows ideal class E voltage and current waveforms.

The analysis of the class-E amplifiers has been reported in several papers [Sokal, 1975; M. Kazimierczuk, 1983; F. Raab, 1978]. The analysis is reproduced here. When the switch *S* is off, the voltage V_s, as shown in Fig.5.2, is given by solving the equation

$$C_s \frac{dv_s}{dt} = I_{ds}(1 - a\sin(w_s t + f)) \qquad (5.1)$$

Where, ω_s is the signal frequency, I_{ds} is the dc portion of the drain current, and constants *a* and ϕ are yet to be calculated. V_s can be represented as

$$v_s(t) = \frac{I_{ds}}{w_s C_s}(w_s t + a(\cos(w_s t + f) - \cos f)) \qquad (5.2)$$

Optimum operation of a class-E amplifier requires two condtions [F. Raab, 1989]

$$\frac{dv_s}{dt}(\frac{T_s}{2}) = 0 \qquad (5.3)$$

$$v_s(\frac{T_s}{2}) = 0 \qquad (5.4)$$

These conditons avoid power dissipation due to either shorting the capacitor C_s while V_s has value or nonzero switching time at transition. Using these conditions, constants *a* and ϕ were calculated:

$$a \approx 1.86$$

$$f \approx -32.5^o.$$

The voltage V_s and the capacitor current i_s are known in the whole range:

$$v_s(t) = \frac{I_{ds}}{w_s C_s}((w_s t) + a(\cos((w_s t) + f) - \cos f)) \qquad 0 \le w_s t \le p \qquad (5.5)$$

$$v_s(t) = 0 \qquad p \le w_s t \le 2p \qquad (5.6)$$

$$i_s(t) = 0 \qquad 0 \le w_s t \le p \qquad (5.7)$$

$$i_s(t) = I_{ds}(1 - a\sin(w_s t + f)) \qquad p \le w_s t \le 2p \qquad (5.8)$$

From equations (5.5) and (5.8), the load Z_L at the fundamental frequency is:

$$Z_{net1} = \frac{v_{s1}}{i_{net1}} = \frac{0.28}{w_s C_s} e^{j49^{O}} \qquad (5.9)$$

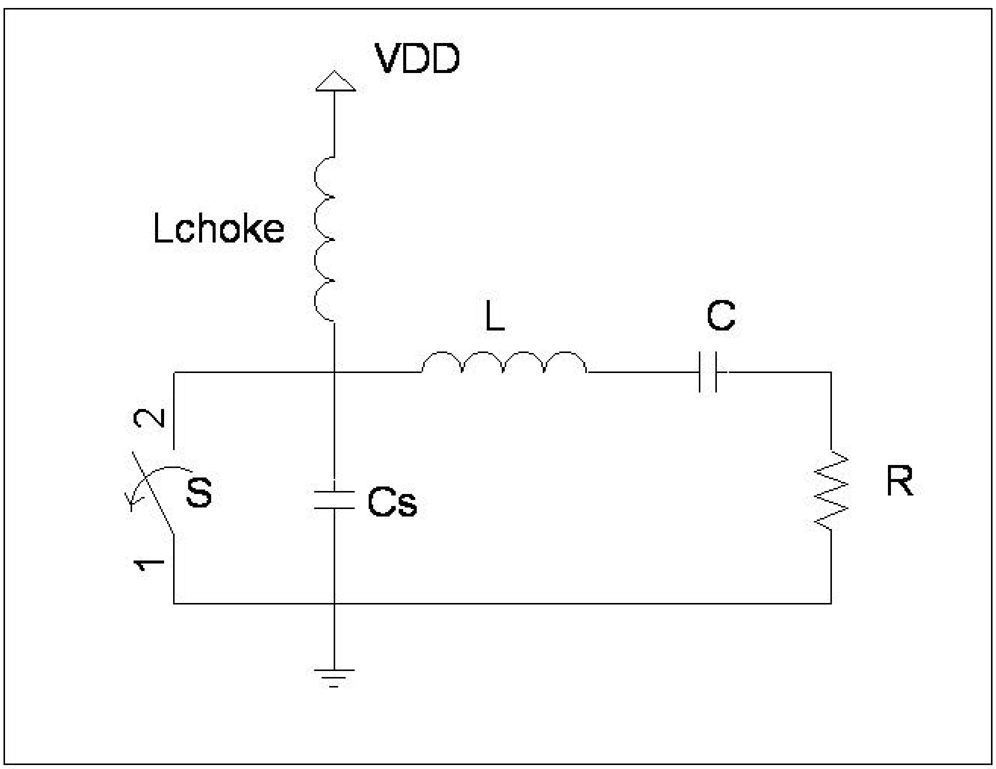


Figure 5.1. Ideal class-E amplifier

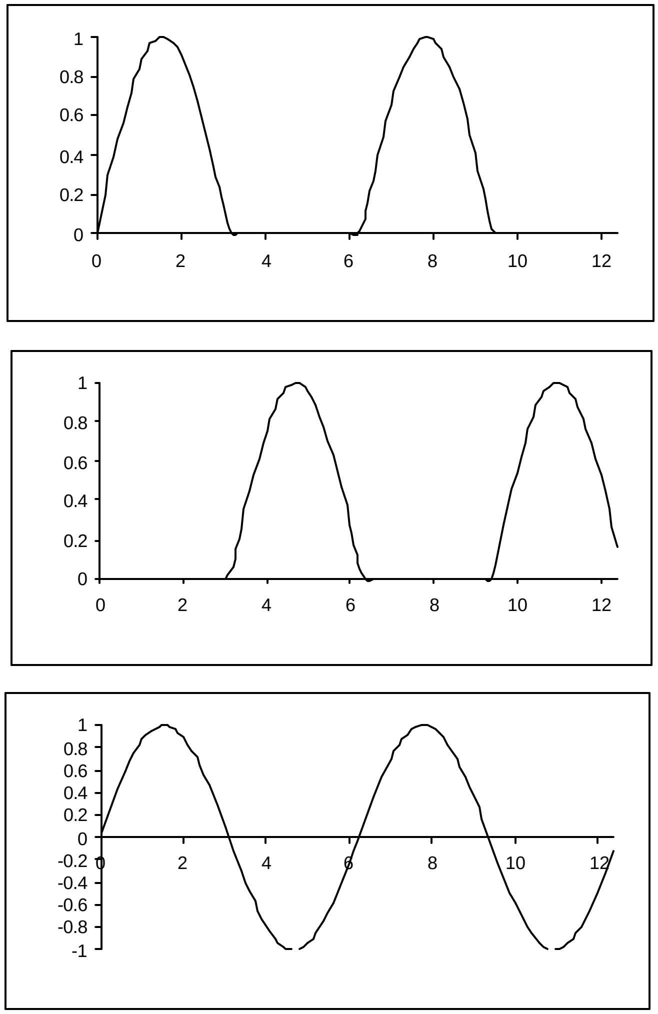


Figure 5.2. Ideal class-E voltage and current waveforms.

Many network configurations can satisfy equation (5.9). To simplify the analysis, the simple load network shown in Fig.5.3 will be used here. The input impedance of the load network is given by

$$Z_{net1} = jw_s L + \frac{1}{jw_s C} + R \qquad (5.10)$$

The load component values are obtained by equating the real and imaginary parts of equations (5.9) and (5.10) [T. Mader, 1995]:

$$C_s = \frac{1}{2pf_s R \cdot (\frac{p^2}{4} + 1)(\frac{p}{2})} \tag{5.11}$$

$$C \approx C_s (\frac{5.447}{Q_L})(1 + \frac{1.153}{Q_L - 1.153}) \tag{5.12}$$

where, Q_L

$$Q_L = \frac{w_s L}{R} \tag{5.13}$$

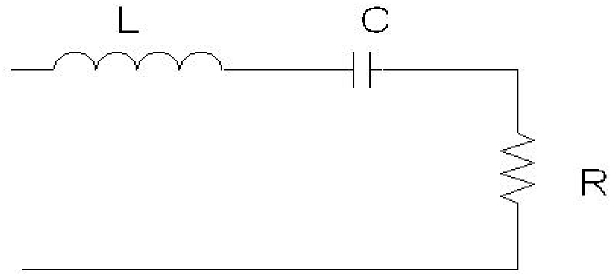


Figure 5.3. Simple RLC load network

5.2 Non-Ideality of Class-E Amplifier

In the ideal situation, 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 result in power dissipation in the switch and reduce the efficiency.

Figures.5.4 (a) and (b) show the transistor's output admittance versus frequency in the ON and OFF states, respectively.

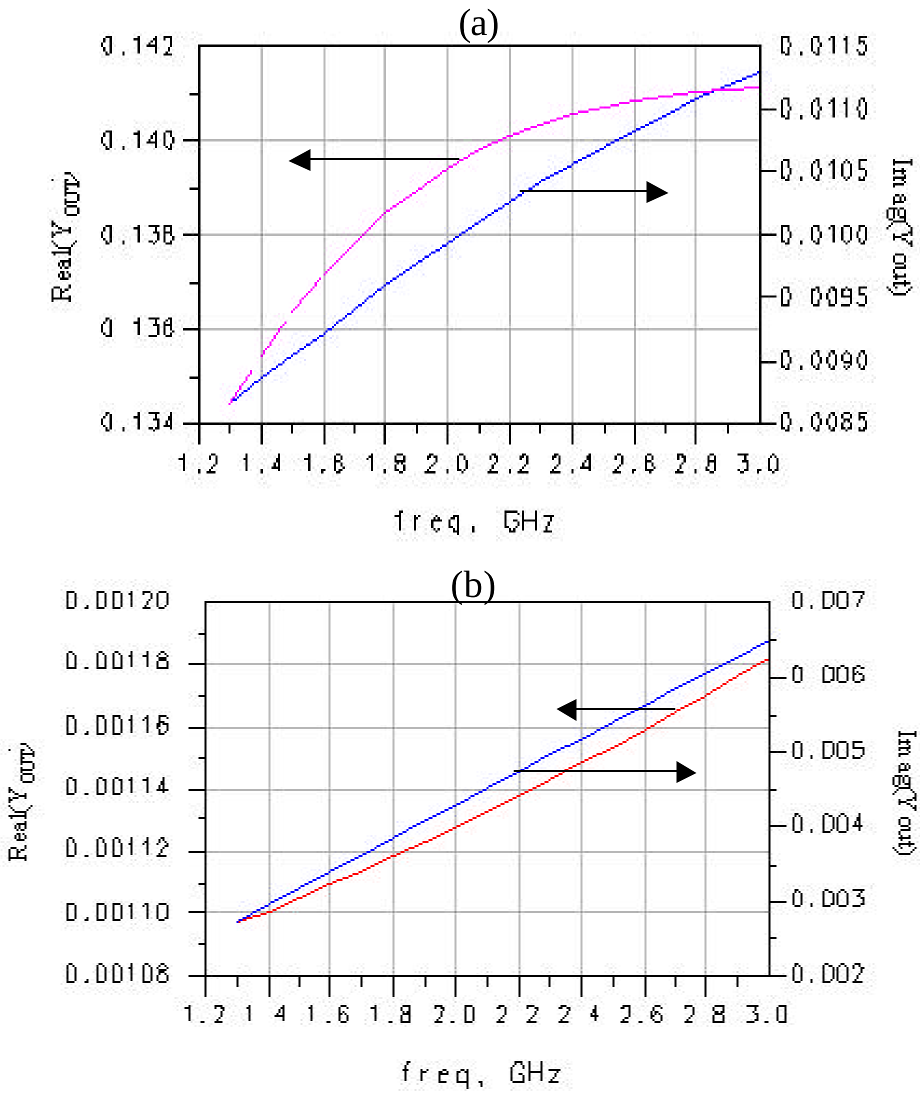


Figure 5.4. Transistor ATF-46100 output impedance:

a) ON state b) OFF state

The transistor can be modeled as a resistor in parallel with a capacitor, as shown in Fig.5.5. During the ON state, the effect of the resistor is the dominant while during the OFF state the capacitor is the dominant one. For the ATF-46100, the ON resistor is around 7 and the OFF capacitor is around 0.9pF.

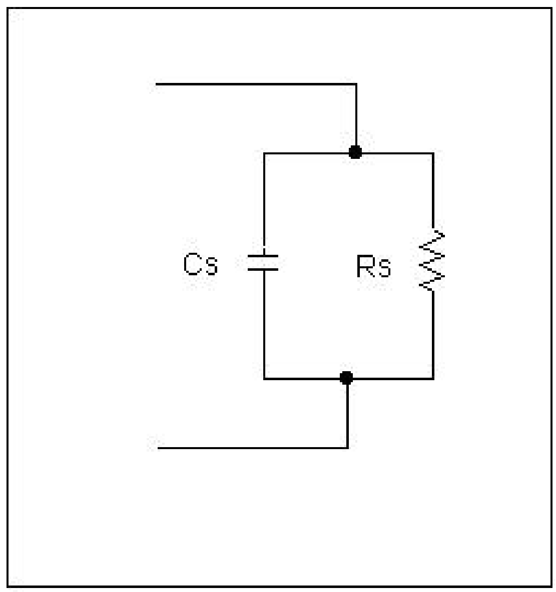


Figure 5.5. Transistor output impedance model

As a result of the switch's non-ideality, the analytical equations described by different authors and reproduced earlier cannot be easily used for the successful optimization of the class-E amplifier. The recent improvement in the modeling of active devices and simulation tools have made it possible to use computer simulation to design such amplifiers with sufficient accuracy. The commercial software used in this work is the ADS (Advanced Design System) by Hewlett Packard.

5.3 L-band Class E Amplifier

As shown in equation (5.9), a transistor requires a specific load to operate in the class-E mode. At the fundamental frequency, the magnitude of the load is frequency dependent and the phase of the load is constant. Also, the load presented at the output terminal of the transistor needs to be high inductive at the harmonics of the input signal. These conditions make the design of a broadband class-E amplifier difficult task.

5.3.1 Lumped Elements Class E Circuit

A class-E amplifier with a 30% bandwidth was proposed in [V. Gudimtla and A. Kain, 1999]. The center frequency was 1GHz and the output power was 23 dB.

The design procedures for a class-E amplifier with a 50% bandwidth are presented in this section. The HP ATF46100 MESFET is used in this design. V_{GS} and V_{DS} are chosen to be –5V and 5.3V, respectively. The optimum load at many frequency points in the band of interest (1.70- 2.7 GHZ) is obtained using the load-pull technique. Another variable that needs to be tuned to obtain better performance in terms of the efficiency and output power is the input power.

Having the optimum load, the next step is to use the E-Syn software to realize the load network. The E-Syn software is a network synthesis program capable of providing a catalog of possible networks with the desired specifications for lumped and distributed components [HP Advance Design System Manual].

A lumped, Butterworth, band-pass response with a passband of 1.7 to 2.7 GHz was selected to realize the load network using the E-Syn. The load network obtained is shown in Fig.5.6.

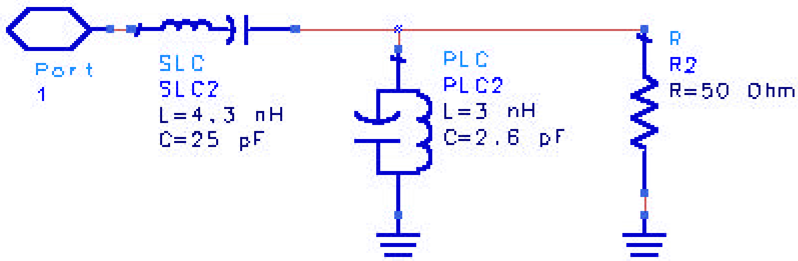


Figure 5.6. Lumped, Butterworth load network

Designing the input-matching network requires knowing the input impedance Z_{in} with the load network connected. Figure.5.7 shows the input impedance versus the frequency. Over the desired band of frequency, Z_{in} can be approximated by a resistor (1.5Ω) in series with a capacitor (1.3 pF) as shown in Fig.5.8 (a). The Chebychef band-pass filter is designed using the E-Syn to realize the input-matching network, as shown in Fig.5.8 (b).

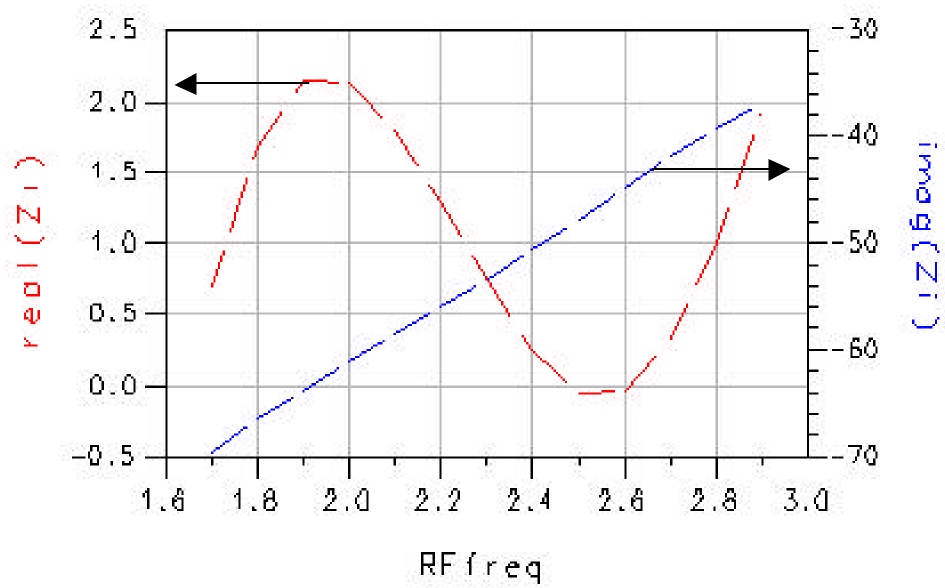
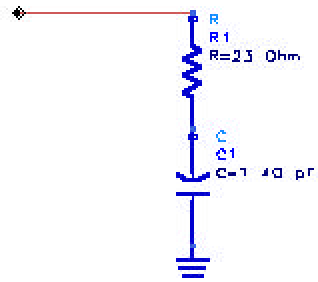
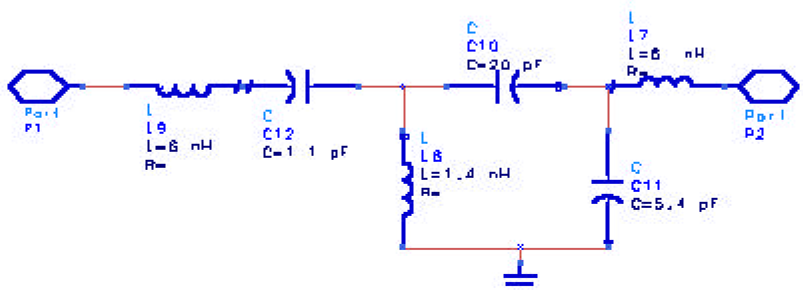


Figure 5.7. The input impedance of the transistor ATF-46100



(a)



(b)

Figure 5.8: (a) Input impedance model of ATF-46100

(b) Input-matching network

As a result of the systematic approach to the design of the amplifier and use of the network synthesis, no further optimization of the complete amplifier (Fig.5.9) was required to give a satisfactory performance. As indicated in Fig.5.10, over most of the range 1.7 to 2.7 GHz, the gain is greater than 16 dB and the output power is almost flat 23 dB. The power added efficiency is greater than 61% over most of the desired band of frequency.

Figure.5.11 shows the output voltage and the drain voltage and current waveforms of the class-E amplifier at 2.2GHz. As 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.

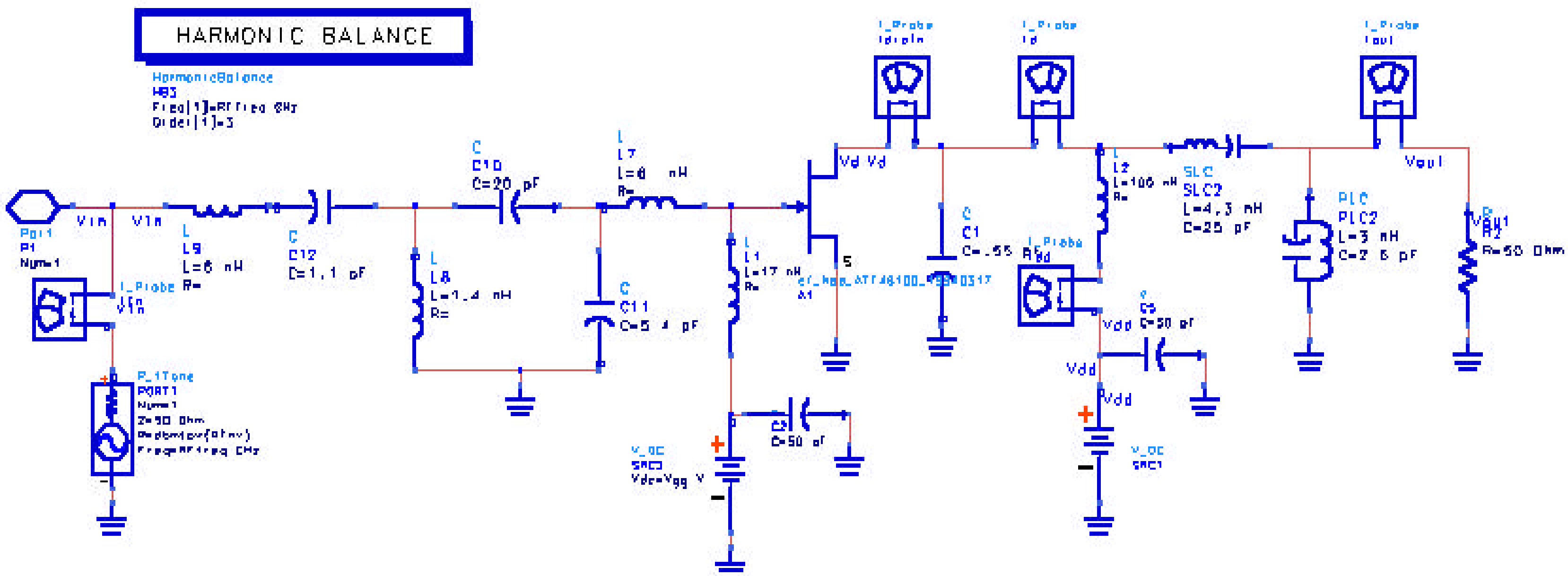


Fig.5.9: **L-band lumped-element class E amplifier**

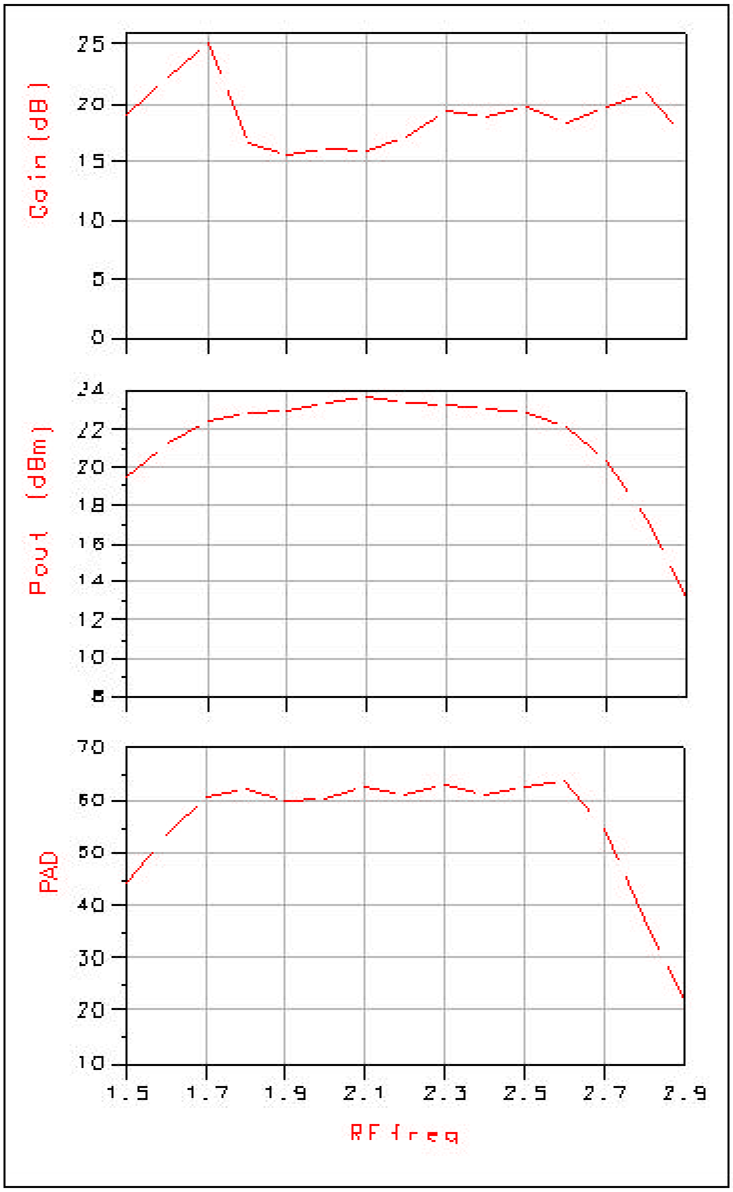


Figure 5.10. Gain (dB), Pout (dBm), and PAD versus Frequency

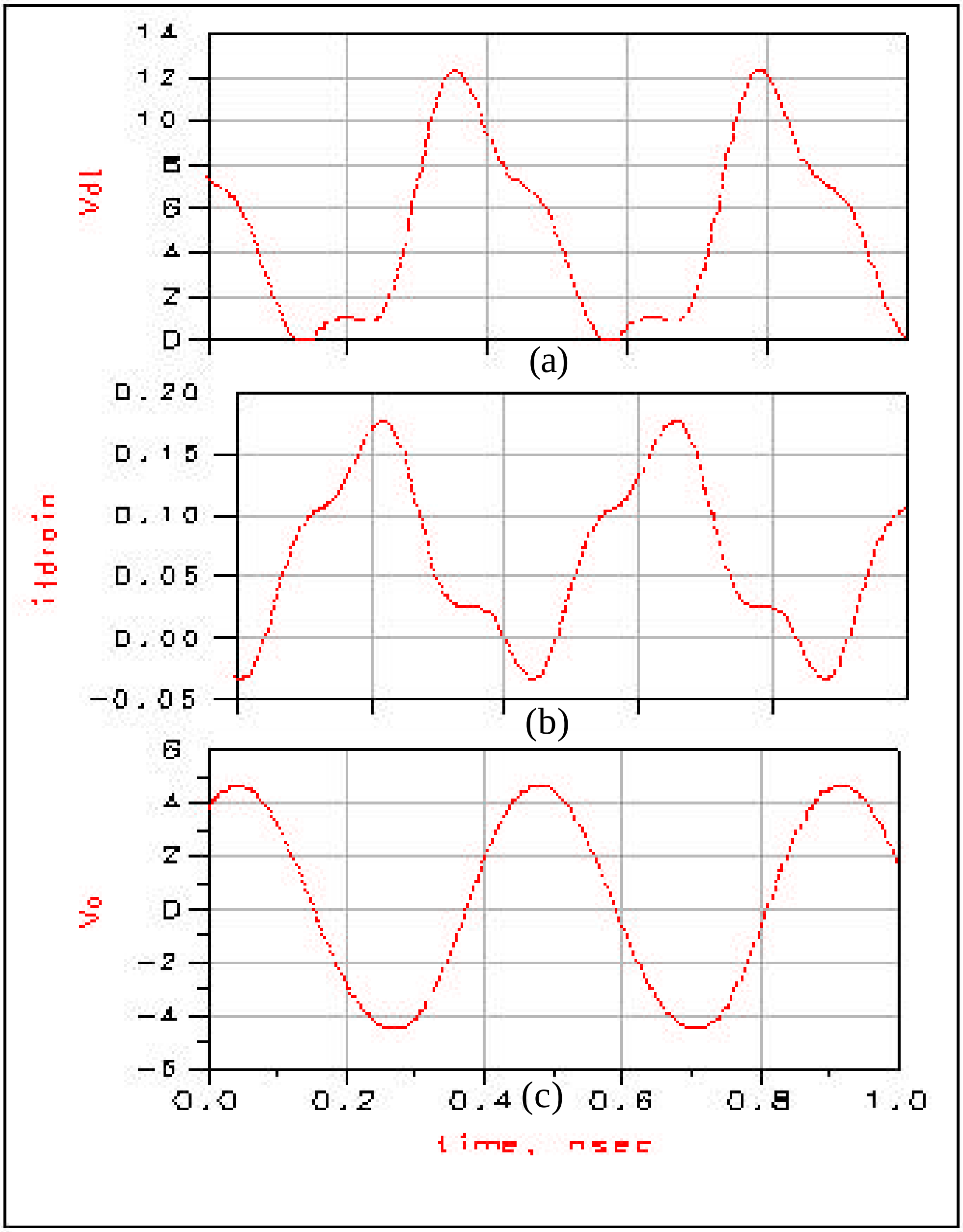


Figure 5.11. Class-E's waveforms:
a) Drain voltage Vdt (V)
b) Drain current (A)
c) Output voltage Vdt (V)

5.3.2 Transmission Line Class-E Circuit

In the previous section, a broadband class-E amplifier based on lumped-elements has been designed. 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.5.12 shows a transmission-line broadband class-E amplifier. This amplifier is designed based on the amplifier shown in Fig.5.9. The lumped elements are replaced with stubs and transmission lines using Richard's transformation method. The lengths and widths of the stubs and transmission lines are adjusted to give better performance. Figure.5.13 shows the performance (the added efficiency and the output power) of the transmission-line broadband class-E amplifier, which is close to the result shown in Fig.5.10.

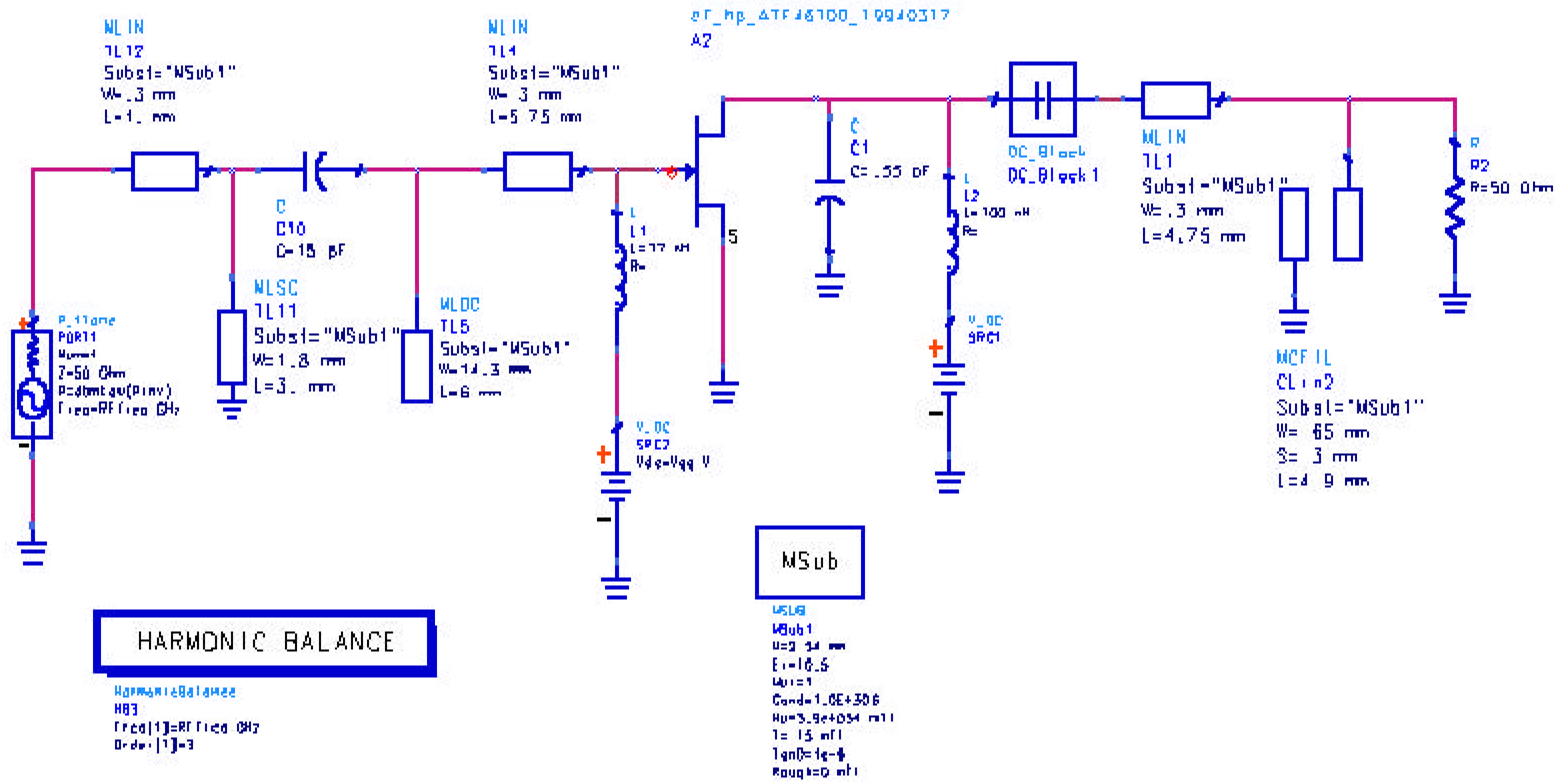


Fig.5.12: **L-band Transmission-line class E amplifier**

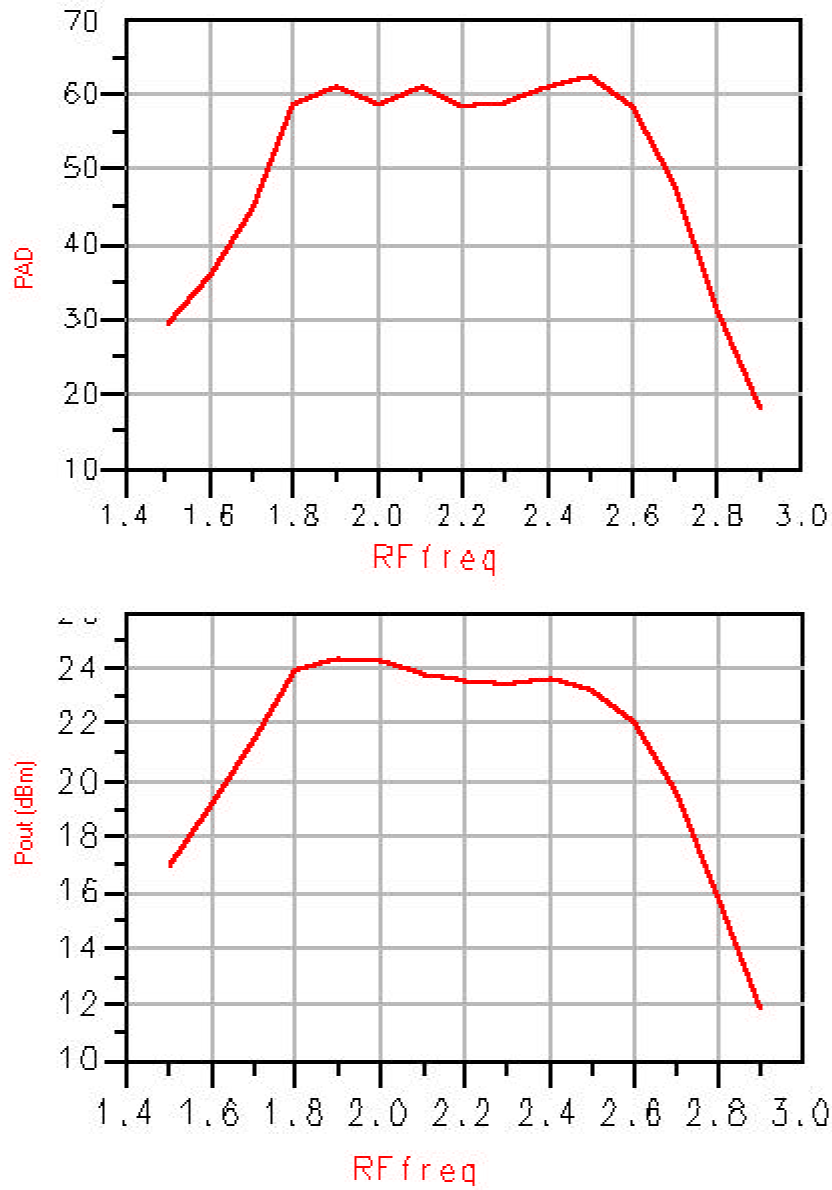


Figure 5.13. Transmission-line broadband class-E amplifier's power added efficiency and output power.

5.4 X-band Class-E Amplifier

Few class-E circuits have been developed at the X-band frequency (8-12GHz). Shijie [L. Shijie, 1998] presented two X-band class-E high efficiency amplifiers. The first one uses the Fujitsu FHX35X HEMT transistor and achieves an output power of 30 mW, and drain efficiency of 80% and PAE (power-added efficiency) of 64% at 11.2 GHz. The second one uses the Fujitsu FLR056XV MESFET transistor and delivers an output power of 186 mW, drain efficiency of 72% and PAE of 56% at 9GHz.

5.4.1 Lumped and Distributed Elements Class-E Circuits

In this section, two X band class-E amplifiers are presented. Both circuits use the Fujitsu FHX35X HEMT transistor [Fujitsu Data Book, 1993]. In contrast to the class-E circuits shown in Figs.5.9 and 5.12, there is no need for the external shunt capacitance C_S. The shunt capacitance C_S consists solely of the output capacitance of the transistor. As shown in Fig.5.14, the first one is based on lumped elements. It achieves drain efficiency above 72%, PAE above 60%, and flat output power 18dBm over wide bandwidth (8.5-13.3 GHz), Fig.5.15. The lumped element class-E amplifier shown in Fig.5.14 is transformed to transmission-lines class-E amplifier shown in Fig.5.16. The transformation process is explained in the previous section. Figure.5.17 shows the transmission-line class E's drain efficiency, PAE, and the output power versus the frequency. Both circuits have a similar performance, which proves the success of the transformation.

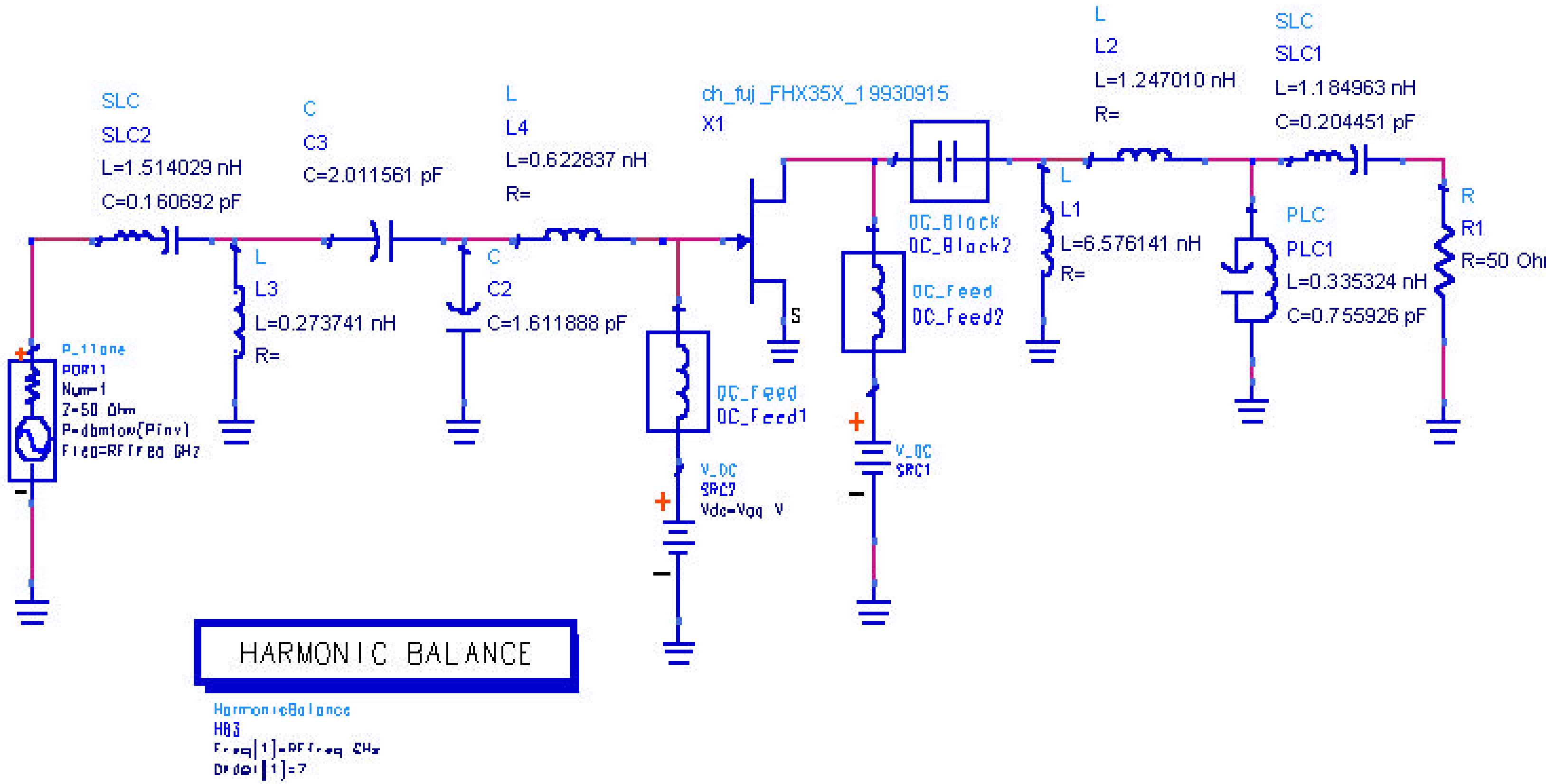


Fig.5.14: **X-band lumped-element class E amplifier**

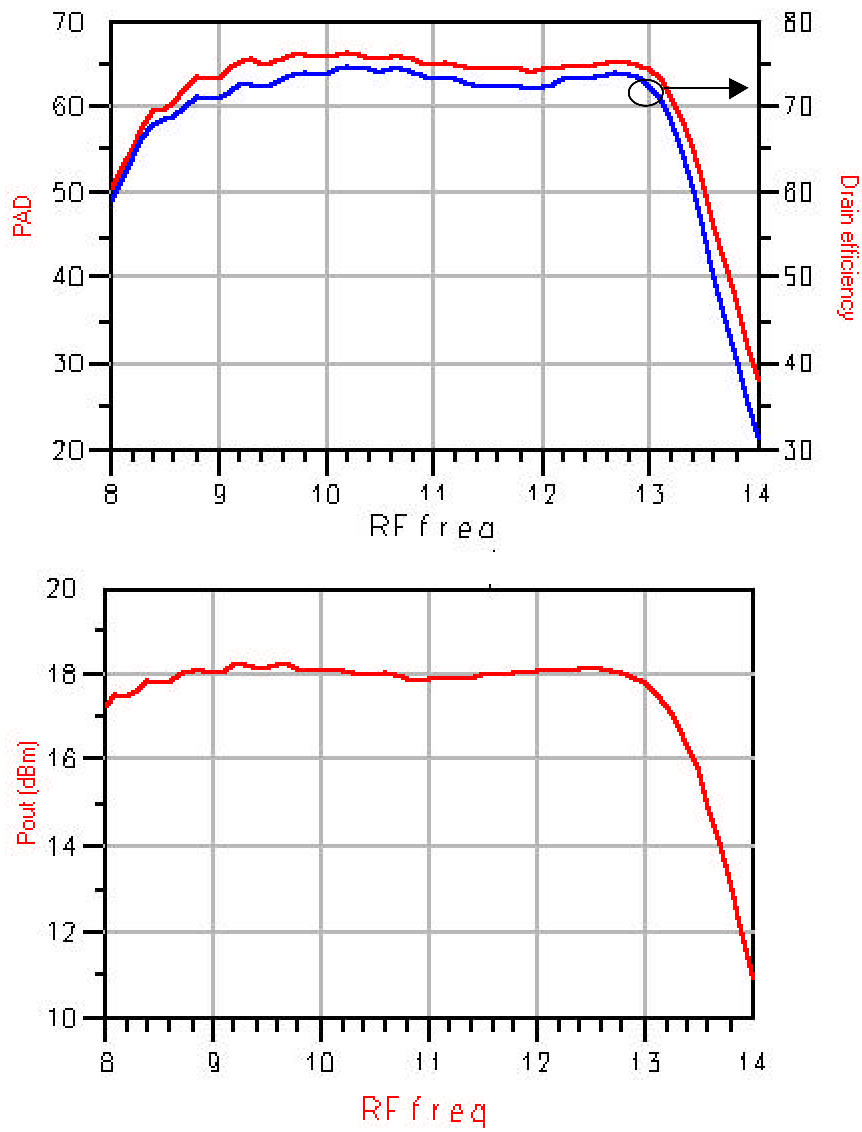


Figure 5.15. The X-band-lumped element class-E's drain efficiency, PAE, and the output power

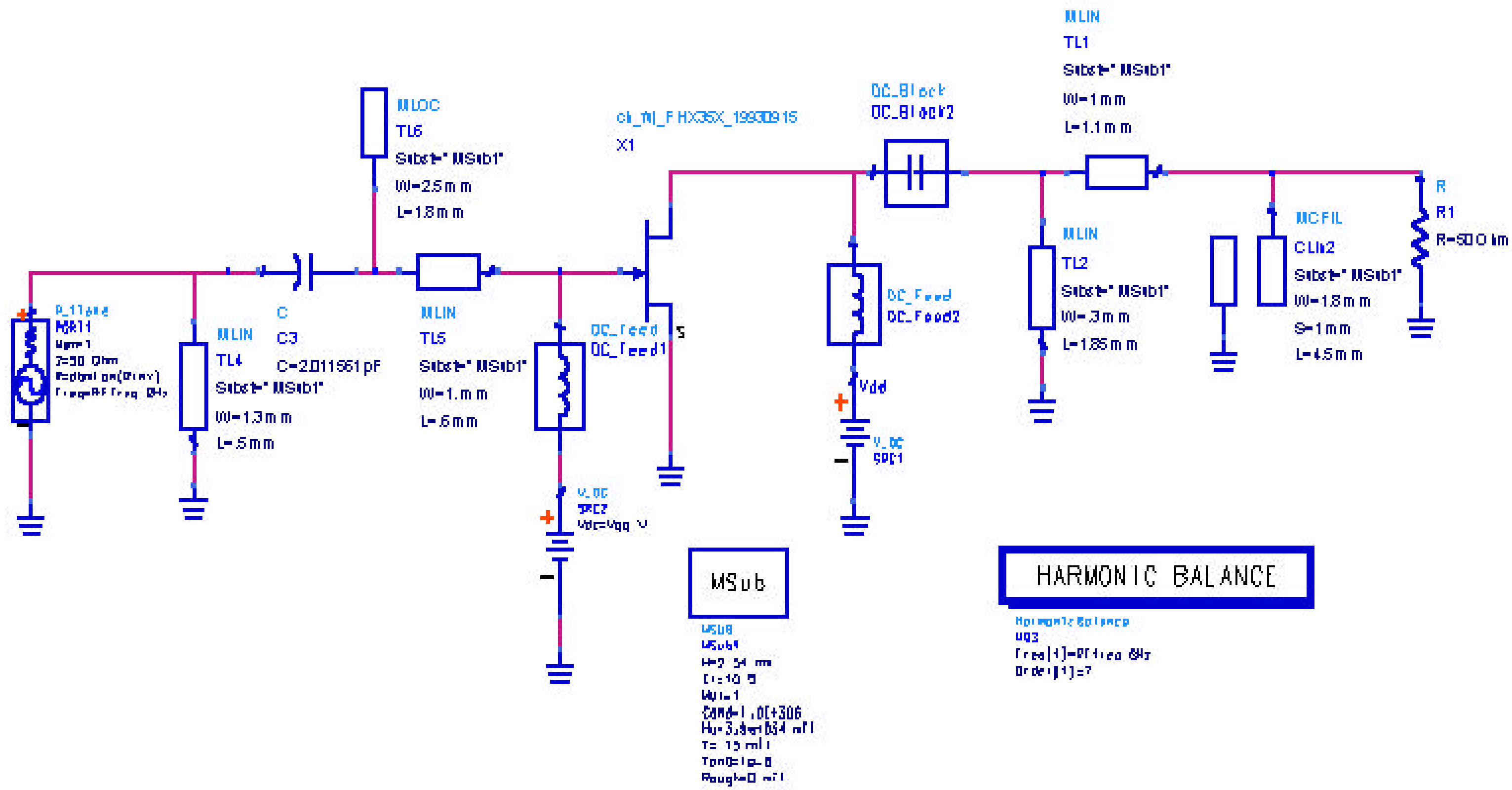


Fig.5.16: **X-band Transmission-line class E amplifier**

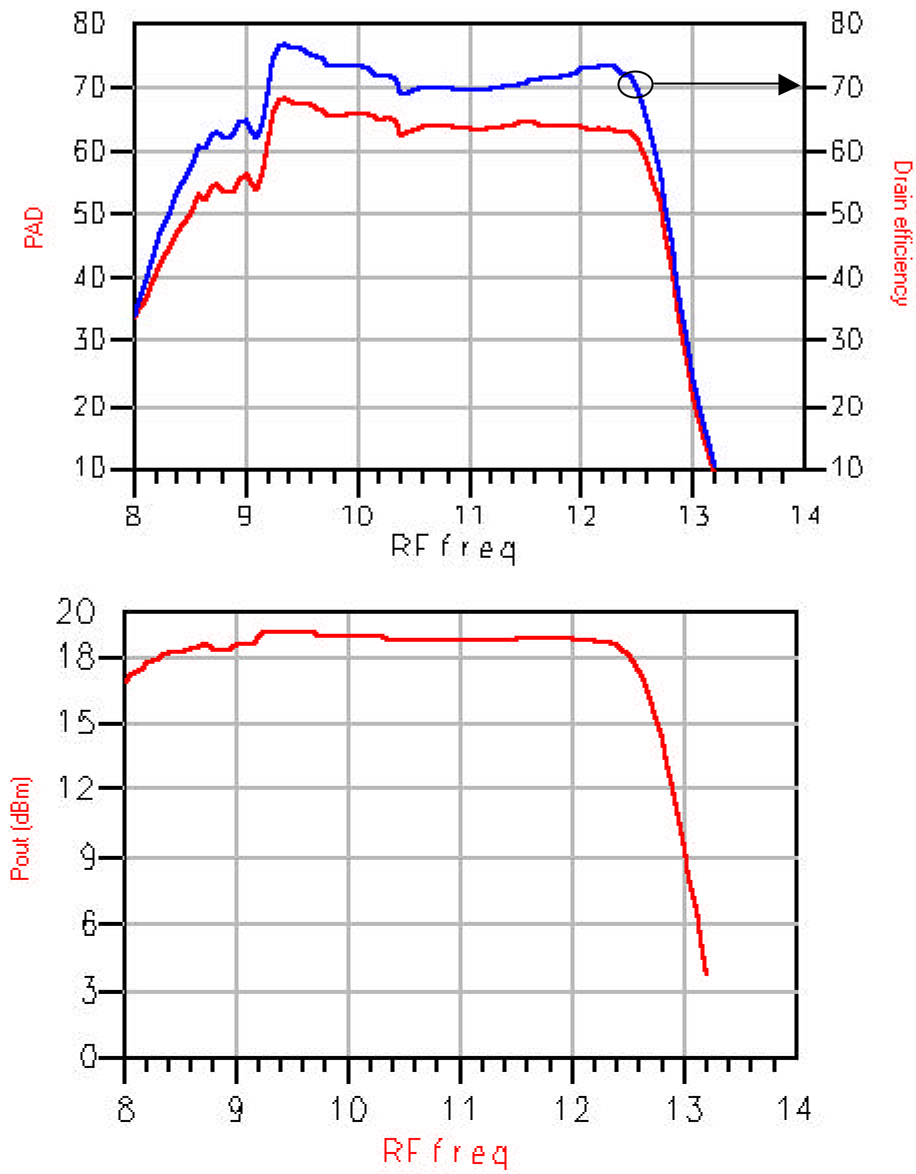


Figure 5.17. The X-band-Transmission-line class-E's drain efficiency, PAE, and the output power

5.5 Technique to Improve Class-E Amplifier's Efficiency

As discussed in section 5.2, the non-ideality of the transistor limits the efficiency of the class-E amplifiers. In this section, a new technique that improves the amplifier efficiency is presented. Two passive networks are added to the class-E circuit. As shown in Fig.5.18, the Z_S network is connected in series with the shunt capacitance and the Z_X network is connected to the transistor source terminal.

To achieve the optimum performance, the characteristic impedance of the passive networks at the harmonics frequencies is obtained using the load-pull technique.

Figure.5.19 shows the drain voltage and current waveforms. The drain efficiency shows an improvement. The drain efficiency is 82 % in contrast to 63% without the Z_S and Z_X networks. The passive networks help in reducing the power dissipation at the output terminal of the transistor by minimizing the overlap of the drain voltage and current waveforms. PAE did not show improvement. The increase in the input power prevents the improvement of PAE.

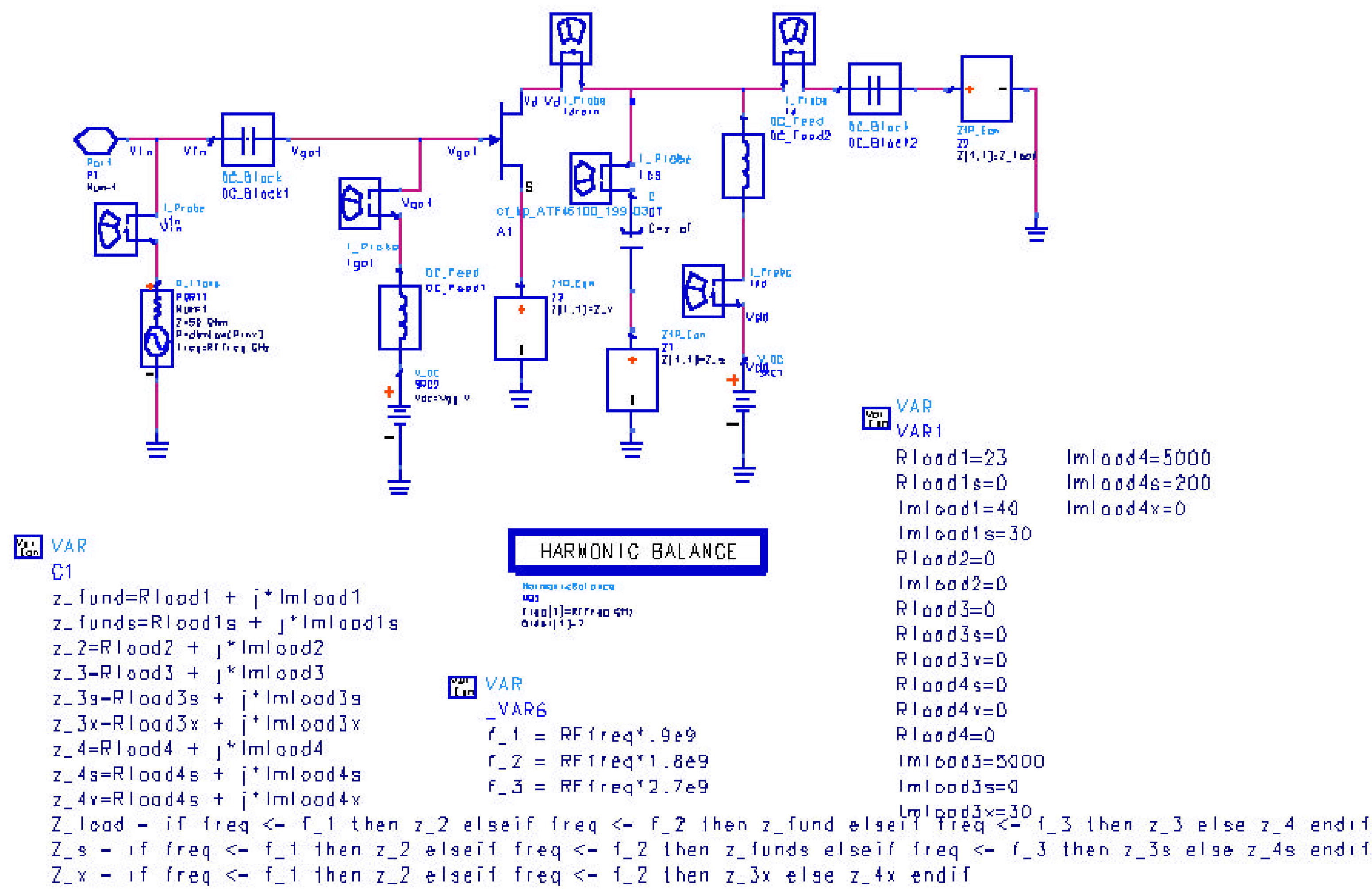


FIG.5.18: CLASS-E AMPLIFIER WITH Z_s AND Z_x NETWORKS.

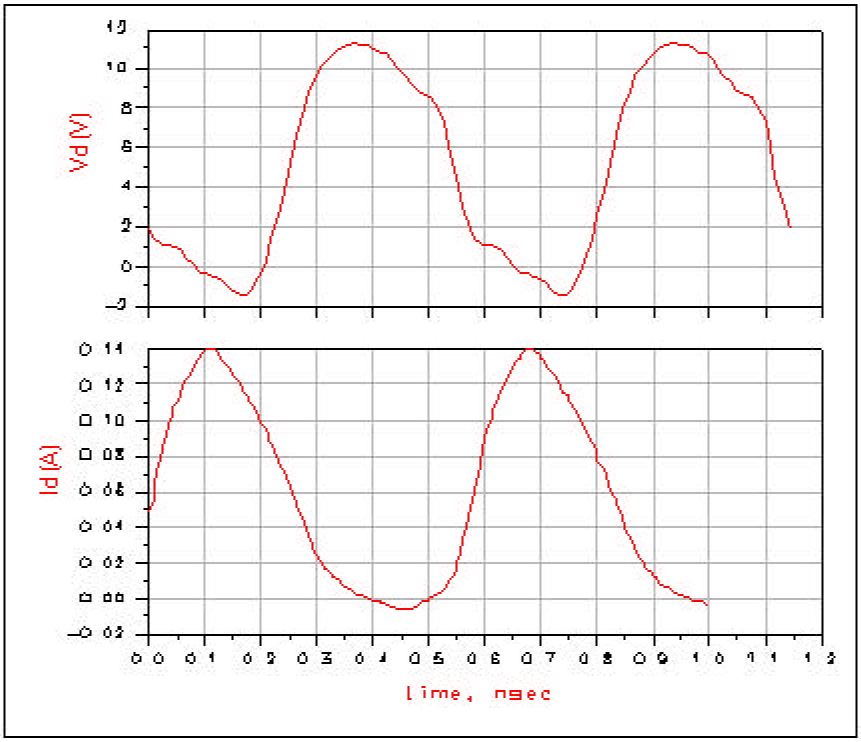


Fig.5.19: Class-E's voltage and current waveforms.

5.6 Class-E versus Class-F amplifiers

This section provides a comparison between the performance of class-E and F amplifiers. As explained in section 2.2.4, the class-F amplifier derives its improved efficiency from the use of resonators to control the harmonic content of the drain (collector) voltage and current.

Various types of transistors (Si bipolar, Si MOSFET, GaAs MESFET, and HEMT) have been used to study the performance of class-E and F amplifiers. Figure.5.20 shows the configuration used to test the performance of class-E and F amplifiers. As indicated in Table 5.1, the two classes have a similar performance (output power, and efficiency).

The drain (collector) peak voltages for class-E and F amplifiers are close to $3V_{ds}$ and $2V_{ds}$, respectively. The breakdown voltage of the transistor puts a limitation on the drain (collector) peak voltage and in sequences puts limitation on the maximum output power. Case 3 in Table 5.1 is an example of the peak voltage limitation, where the maximum bias voltages for class-E and F are 18V and 22V, respectively.

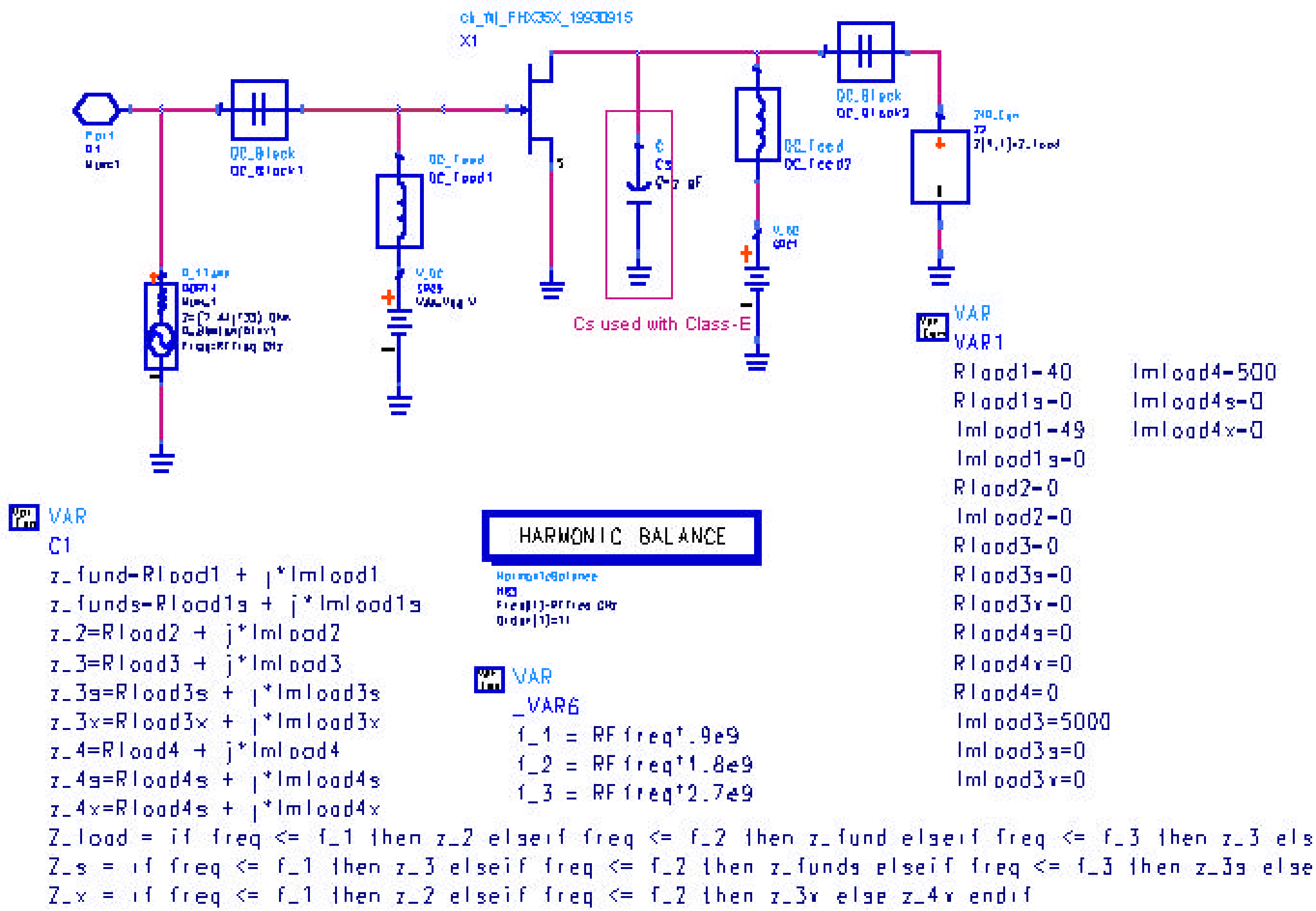


Fig.5.20: Class-E and F amplifiers configuration.

Table 5.1. Summary of results for class-E and F using various transistors.

Transistor	Freq (Ghz)	Amplifier Class	Pout (mW)	PAD %
Si bipolar NEC24600	0.5	E	258	76
		F	446	72
Si MOSFET MRF136	0.1	E	9171	85.5
		F	14060	87.8
GaAs MESFET ATF4600	1.75	E	300	60-
		F	365	62.7
HEMT FHX35X	11	E	79.6	65.9
		F	75	69

Also, the performances of class-E and F amplifiers have been compared at low voltage design. At low voltage design, the class-E amplifier has a better performance than the class-F amplifier, Table 5.2.

Table 5.2. Summary of results for low voltage class-E and F.

Transistor	Freq (Ghz)	Amplifier Class	Pout (mW)	PAD %
GaAs MESFET ATF4600	1.75	E	80	60.7
		F	86	53
Si bipolar NEC24600	0.5	E	85	61
		F	73	59

Another advantage of the class-E amplifier over the class-F amplifier is its simple configuration. As explained earlier, the class-E amplifier requires high load impedance at the harmonics while the class-F amplifier requires high impedance at odd harmonics and low at even harmonics. These requirements make designing a wide-band class-E amplifier much easier than a wide-band class-F amplifier.

The intermodulation test is one of the important tests used to measure the nonlinear behavior of the analog circuits. Commonly the intermodulation test uses a two-tone signal. When the two-tones are applied to a nonlinear system, the output signal exhibits some components that are not the harmonics of the input frequencies. These frequencies are generated from the mixing of the input frequencies and are called intermodulation (IM). Of particular interest are the third-order IM products ($2f_1$-f_2, and $2f_2$-f_1). The importance of these products arises from the small frequency distance between them and the desired signals, which makes filtering off the third-order IM a difficult task. The IM test setup is similar to the one shown in Fig.5.20 with the two-tone input signal applied. Table 5.3 shows comparison between class-A, E, and F interim of the output power, PAD, gain, intermodulation. The intermodulation is expressed in dBc (the difference in dB between the IM power and the fundamental frequency power). Class-A shows better output power, gain, and linearity than the other classes.

As explained earlier, the output voltage waveforms of the ideal class-E and F are half sine and square waveforms, respectively and their Fourier representations are given as:

$$V_{class-E}(t) = A_E + B_E \cdot \sin w + C_E \cdot \sin(2w) + D_E \cdot \sin(4w) + \ldots \qquad (5.14)$$

$$V_{class-F}(t) = A_F + B_F \cdot \sin w + C_F \cdot \sin(3w) + D_F \cdot \sin(5w) + \ldots \qquad (5.15)$$

As indicated in equation (5.15), the output voltage of the ideal class-F amplifier does not contain the second harmonics of the input signal (2ω). The absence of the second harmonics eliminates the third IM ($2f_1$-f_2, and $2f_2$-f_1) at the output. In reality, due to the non-linear behavior of the transistors there will be third IM at the output. This gives class-F the advantage over class-E of having lower third IM.

Table 5.3. Summary of results for class-A, E and F amplifier's linearity.

Amplifier Class	Pout (mW)	PAD %	Gain (dB)	3rd IM dBc	5th IM dBc	7th IM dBc
A	265	20	22	-51	-55	-60
E	195	52	16	-16.5	-34.5	-50
F	220	56	15.5	-26	-34	-51

In general, class-E and F amplifiers show a similar performance. However, in some applications one of them shows a better performance than the other. For example, the class-F amplifier has a better performance in the high-output power amplifier applications and class-E amplifier has better performance in wide-band and low voltage applications.