SIMPLE DESIGN EQUATIONS FOR BROADBAND CLASS E POWER AMPLIFIERS WITH REACTANCE COMPENSATION

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ABSTRACT

In this paper, a simple analytical design approach to determine the parameters of the loading networks to design broadband class E amplifiers is presented. The design equations are given for each element of single and double resonant loading circuits. The analysis and simulation were performed on the example of high-voltage LDMOSFET power amplifier, which show that in octave-band of 100-200 MHz the power gain of 10 dB with deviation of only ±0.5 dB and the drain efficiency of about 70% and higher can be achieved.

I. INTRODUCTION

One of the key factors of the power amplifier design is to provide its high efficiency operation mode. In order to realize a high efficiency broadband operation with the appropriate current and voltage waveforms it is necessary to use a broadband loading network. For example, a simple network consisting of a series resonant *LC* circuit tuned on the fundamental and a parallel inductance which provides a constant load phase angle of 50 degree in a frequency range of about 50% [1]. For the first time, such a reactance compensation technique using single-resonant circuit was applied to varactor tuned Gunn oscillator and parametric amplifier [2]. Moreover, it is possible to further increase the tuning range of an oscillator by adding some more stages of reactance compensation [3]. Reactance compensation circuit technique can also be applied to the microwave transistor amplifier design because a series or shunt *RLC* circuit generally represents the input and output transistor impedances.

Computer-aided design of wideband microwave transistor amplifiers or oscillators requires, in a common case, considerable time and can be quite tedious. Therefore an analytical approach, which allows defining some common regularities and expressing the ratios between circuit elements in a clear and simple form, can be very useful and desirable to reduce substantially the large amount of the numerical calculation. And in conjunction with the analytical approach computer optimisation technique will give the fastest and most accurate final results.

II. DESIGN EQUATIONS

To describe reactance compensation circuit technique consider the simplified equivalent loading circuit with a series L_1C_1 resonant circuit tuned on the fundamental and a shunt L_pC_p circuit providing a constant load phase angle relatively the device output terminals shown in Figure 1. The reactances of the series and shunt resonant circuits vary with frequency as shown in Figure 2 by curve 1 and curve 2, respectively.

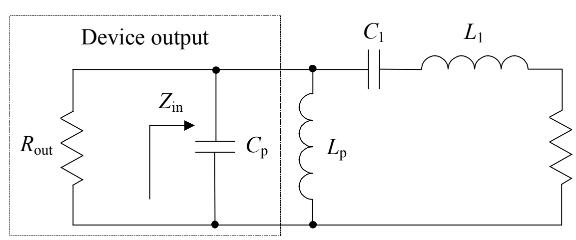

Fig. 1. Single reactance compensation circuit

Near the resonant frequency ω_0 of the series circuit with positive reactance slope (curve 1) the slope of shunt circuit reactance is negative which reduces the overall reactance slope of the loading network (dotted line). By proper choice of the circuit elements a constant load angle can be provided over a very large frequency bandwidth.

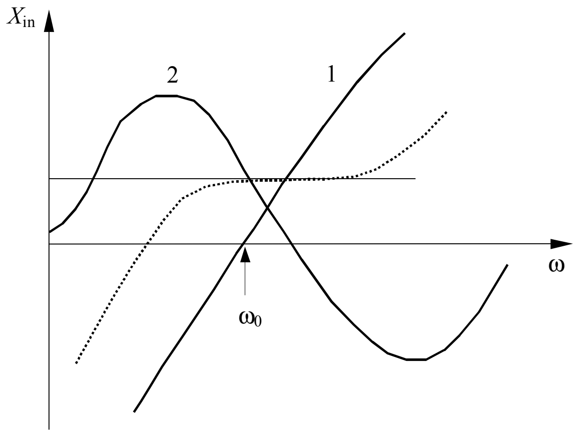

Fig. 2. Reactance compensation principle

First it is necessary to define a value of the constant magnitude and phase angle of the loading circuit input impedance obtained at the device output terminals. So, at resonant frequency for the loading circuit with shunt capacitance C_p and series L_s-L_0C_0-R resonant circuit shown in Figure 3, the magnitude and the phase of the loading network impedance Z_{in} are given by

$$\frac{|Z_{in}|}{R} = \frac{\sqrt{1 + \left(\frac{\omega L_s}{R}\right)^2}}{\sqrt{\left(1 - \frac{\omega L_s}{R}\omega C_p R\right)^2 + \left(\omega C_p R\right)^2}} \qquad (1)$$

$$\phi = \tan^{-1}\left(\frac{\omega L_s}{R}\right) - \tan^{-1}\left(\frac{\omega C_p R}{1 - \frac{\omega L_s}{R}\omega C_p R}\right) \qquad (2)$$

Consequently, taking into account that, for an idealized optimum Class E operation mode,

$$\tan\varphi = \omega L_s / R = 1.1521$$

$$\omega R C_p = 1/5.4466$$

an optimum phase angle of overall loading network is equal to

$$\phi = \tan^{-1}(1.1521) - \tan^{-1}(0.2329) = 35.945^\circ \qquad (4)$$

that defines the input impedance of overall loading network as an inductive.

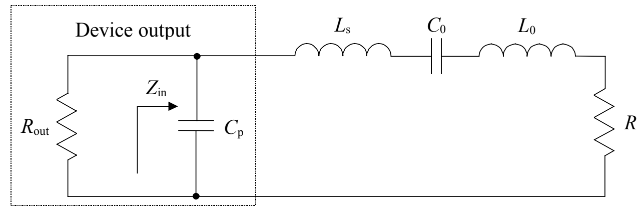

Fig. 3. Class E equivalent circuit

For a broadband-loading network shown in Figure 1, the normalized input impedance can be written by

$$\frac{Z_{in}}{R} = \frac{j\frac{\omega L_p}{R}}{1 + j\frac{\omega L_p}{R}\left(1 + j\omega C_p R\right)} \qquad (3)$$

where, for an optimum Class E operation mode,

$$C_p = \frac{\tan 49.052^\circ - \tan 35.945^\circ}{\omega R} \qquad (4)$$

$$L_p = \frac{R}{\omega \tan 49.052^\circ} \qquad (5)$$

The parameters of the series L_1C_1 resonant circuit must be chosen to provide a constant phase angle of the loading network over a broadband frequency bandwidth. This bandwidth will be maximized if at resonance frequency ω_0

$$\left.\frac{dB(\omega)}{d\omega}\right|_{\omega=\omega_0} = 0 \qquad (6)$$

where $B(\omega) = -\left(1 - \omega^2 L_p C_p\right)/\omega L_p$ is the loading network susceptance. Then, an additional equation can be written by

$$C_p + \frac{1}{\omega^2 L_p} - \frac{2L_1}{R^2} = 0 \qquad (7)$$

As a result, the series capacitance C_1 and inductance L_1 can be calculated by

$$L_1 = \frac{R}{\omega}\left(\tan 49.052^\circ - 0.5\tan 35.945^\circ\right) \qquad (8)$$

$$C_1 = 1/\omega^2 L_1 \qquad (9)$$

The wider bandwidth can be achieved using double-resonant circuit reactance compensation shown in Figure 4. In this case, equating the third derivative to zero gives

$$C_p + \frac{1}{\omega^2 L_p} - 2\frac{C_2 R^2 - L_1}{R^2} = 0 \quad (10)$$

$$\frac{1}{\omega^2 L_p} + \frac{C_2 R^2 - L_1}{R^2} -$$

$$8\omega^2 L_1 \left[C_2^2 + \frac{(C_2 R^2 - L_1)(L_1 - 2C_2 R^2)}{R^4} \right] = 0 \quad (11)$$

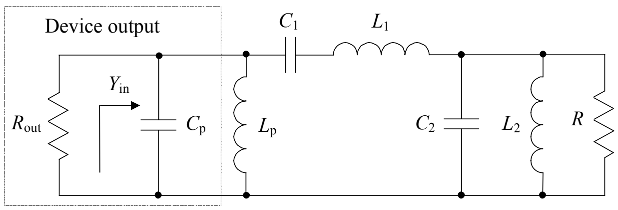

Fig. 4. Double reactance compensation circuit

Solving equations (10)-(11) gives the following simplified equation expressed through the loading network phase angle ϕ in the form of

$$C_2^2 R^4 + (C_2 R^2 - L_1)(L_1 - 2C_2 R^2) -$$

$$\frac{R^3 \tan 35.945^\circ}{16\omega^3 L_1} = 0 \quad (12)$$

It is advisable to rewrite equation (12) through the loaded quality factors $Q_1 = \omega L_1 / R$ and $Q_2 = \omega C_2 R$. Then, for Q_1 and Q_2, which values are close to unity or higher,

$$Q_2 \cong 0.5 Q_1 \left(3 - \sqrt{5}\right) \quad (13)$$

and the parameters of the series and shunt resonant circuits as a starting point for subsequent circuit optimization can be calculated from

$$L_1 = \frac{2R}{\omega} \frac{\tan 49.052^\circ - 0.5 \tan 35.945^\circ}{\sqrt{5} - 1} \quad (14)$$

$$C_1 = \frac{1}{\omega^2 L_1} \quad (15)$$

$$C_2 = \frac{L_1}{R^2} \frac{3 - \sqrt{5}}{2} \quad (16)$$

$$L_2 = \frac{1}{\omega^2 C_2} \quad (17)$$

III. LOADING NETWORK SIMULATION

The circuit simulation for these three loading networks was performed at resonant frequency $f = 150$ MHz for load $R = 50\ \Omega$. The optimum parameters of the loading networks were calculated according to the analytical design criteria for each circuit. So, for the circuit with shunt capacitance they are $C_p = 4$ pF, $L_s + L_0 = 103$ nH, $C_0 = 27$ pF. For single resonant broadband loading network $C_p = 9$ pF, $L_p = 46$ nH, $L_1 = 42$ nH, $C_1 = 27$ pF, whereas for the double resonant broadband one $C_p = 9$ pF, $L_p = 46$ nH, $L_1 = 67$ nH, $C_1 = 17$ pF, $C_2 = 10$ pF, $L_2 = 113$ nH.

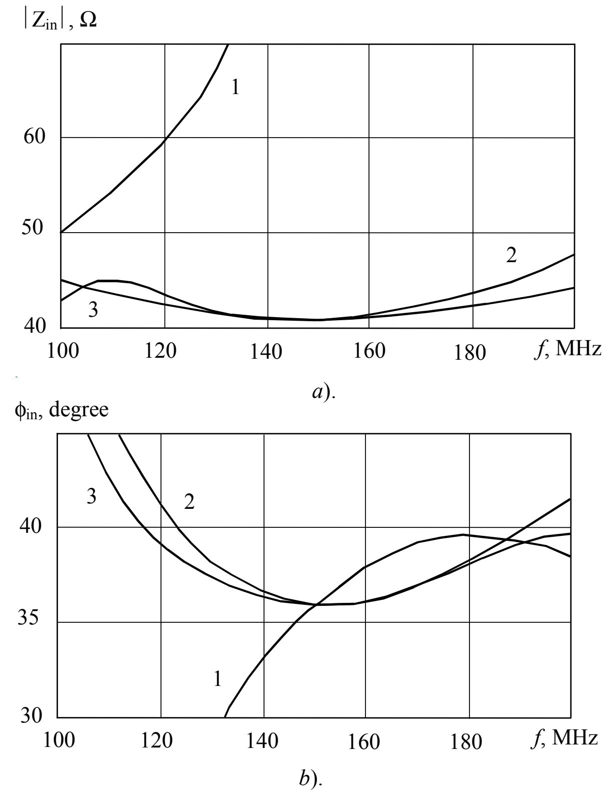

Fig. 5. Loading network broadband performance

In Figure 5, the frequency dependencies of the input impedance magnitudes and phase angles are presented for the loading networks with a shunt capacitance (curve 1), a series resonant circuit (curve 2) and the series and shunt resonant circuits (curve 3). It is clearly seen that reactance compensation technique allows increasing substantially the operating frequency bandwidth. So, using even single resonant loading network gives a significant widening of a frequency bandwidth. Using a double resonant network allows obtaining maximum deviation from the optimum value of about 36° by only 3° in a frequency range of 120-180 MHz.

IV. AMPLIFIER SIMULATION

To achieve high-efficiency broadband operation mode it is very advisable to design power amplifier based on LDMOS transistors. In this case, it is quite easy to provide a very broadband input matching using a dissipative input matching circuit. Besides, it is possible to provide high drain efficiency at high supply voltage. So, using an LDMOSFET device for a Class E power amplifier the drain efficiency of 88% at 144 MHz with an output power level of 14 W was achieved with high-Q inductor in the output series resonant circuit [4].

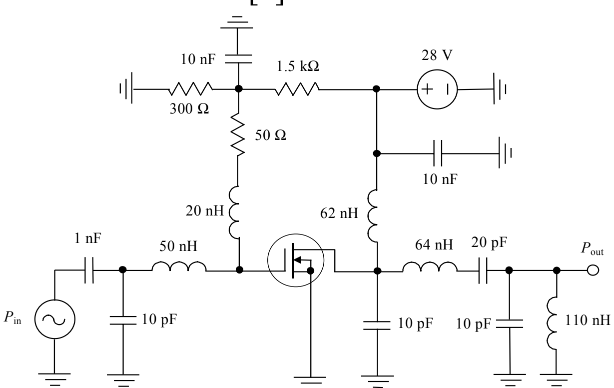

Fig. 6. Simulated broadband Class E amplifier

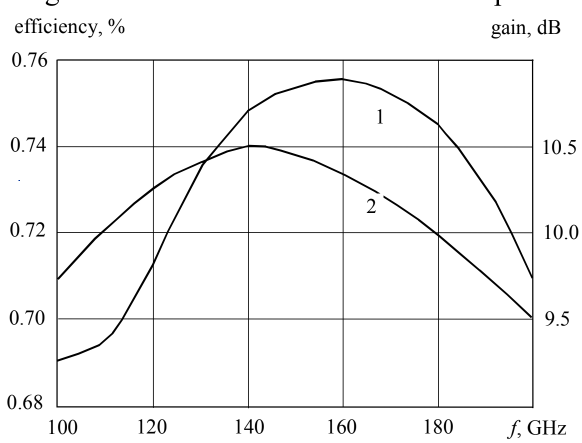

Fig. 7. Broadband Class E amplifier performance

In Figure 6, the schematic of the broadband LDMOSFET power amplifier designed to operate in a frequency bandwidth of 100-200 MHz using a double resonant loading network for input power P_{in} = 1 W is shown. The input matching circuit includes a simple L-transformer as well as a series circuit with the inductance and 50-Ohm resistance connected in parallel which allows to realize the return loss less than -15 dB and, as a result, the input VSWR less than 1.4 in overall octave frequency bandwidth. Figure 7 shows that, for such an octave-band power amplifier, the power gain of 10 dB with deviation of only ±0.5 dB (curve 2) and the drain efficiency of about 70% and higher (curve 1) can be achieved.

V. CONCLUSIONS

A simple analytical design approach to determine the parameters of the loading networks to design broadband class E amplifiers is presented. The design equations are given for each element of single and double resonant loading circuits. The simulation performed on the example of high-voltage LDMOSFET power amplifier verify the possibility of high efficiency Class E operation in a wide frequency range. Such an approach combining the analytical calculations with the computer simulation allows RF and microwave high efficiency power amplifiers to be designed more effectively.

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