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A High-Efficiency Transmission-Line GaN HEMT Class E Power Amplifier

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This article describes the design of a Class E wireless power amplifier using transmission lines for output matching, with the circuit implemented using a GaN HEMT transistor

In modern wireless communication systems it is required that the power amplifier could operate with high efficiency, high linearity, and low harmonic output level simultaneously. To increase efficiency of the power amplifier, a switching-mode Class E mode technique can be applied. This kind of a power amplifier requires an operation in saturation mode resulting in a poor linearity, and therefore is not suitable to directly replace linear power amplifiers in conventional WCDMA or CDMA2000 transmitters with non-constant envelope signal. However, to obtain both high efficiency and good linearity, a nonlinear high-efficiency power amplifier operating in a Class E mode can be used in advanced transmitter architectures such as Doherty, LINC (linear amplification using nonlinear components), or ET (envelope tracking) with digital predistortion [1-3]. In this paper, a novel transmission-line load network for a Class-E power amplifier with simple design equations to define its load-network parameters is presented.

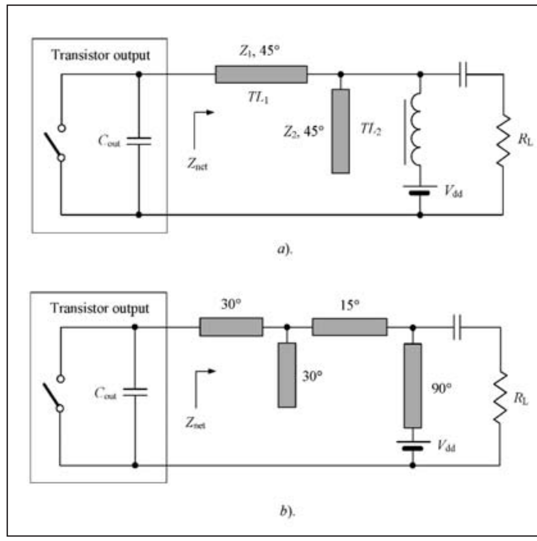

Figure 1 · Transmission-line load networks for Class-E power amplifiers.

Transmission-Line Class-E Load Networks

Generally, the Class E load network can be based on both lumped elements and transmission lines depending on the operating frequency and convenience of practical implementation [4-6]. At higher frequencies, to provide a required inductive impedance at the fundamental and high reactive impedance seen by the shunt capacitance at the second and higher order harmonics, it is preferable to use short-circuit and open-circuit stubs instead of lumped capacitors in the load network for better harmonic suppression and performance predictability [7-10].

Figure 1(a) shows the conventional transmission-line Class E load-network schematic where the series transmission line TL_1 and open-circuited transmission-line stub TL_2, having an electrical length of 45° each, provide high impedance at the second harmonic [5]. This can be considered as a second-harmonic Class E approximation. At the same time, transformation of the optimum Class-E load resistance to the standard 50-ohm load resistance can be realized by proper choice of the transmission-line characteristic impedances of the Z_1 and Z_2. The more complicated Class-E load network which provides the open-circuit conditions simultaneously for

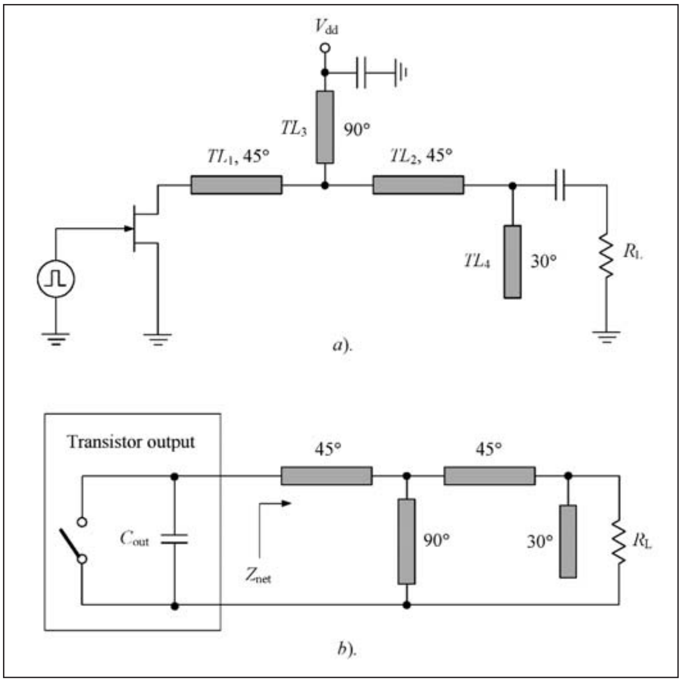

Figure 2 · Modified transmission-line Class E power amplifier.

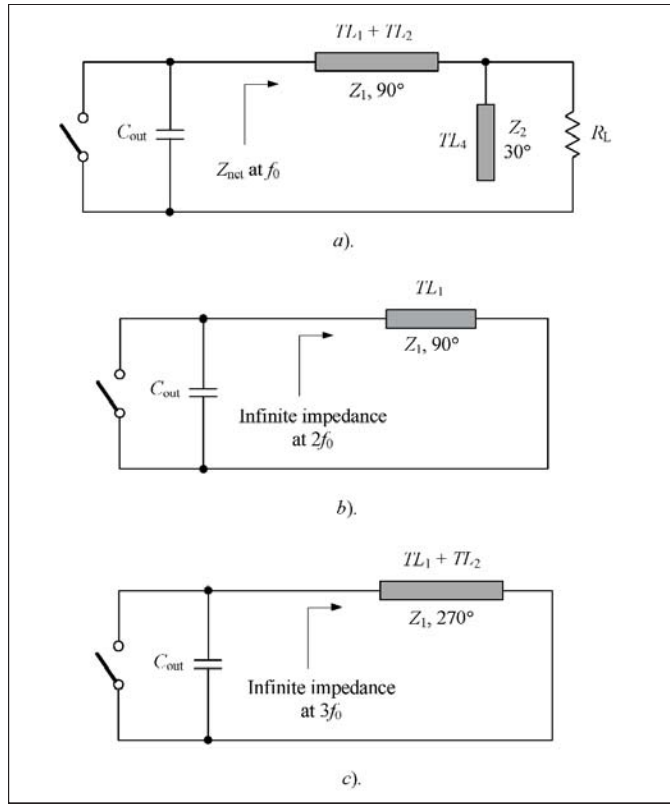

Figure 3 · Load networks seen by the device output at harmonic frequencies.

the second and third harmonics by using a 30-degree open-circuited stub and a short-circuited quarter wave transmission line is shown in Figure 1(b) [11].

In the latter case intended for a conventional Class-E mode, the following harmonic conditions seen by the device output at the fundamental-frequency, second and third harmonic components must be satisfied:

$$Z_{net}(\omega_0) = R\left(1 + j\tan 49.054^{\circ}\right) \tag{1}$$

$$\operatorname{Im} Z_{net}(2\omega_0) = \operatorname{Im} Z_{net}(3\omega_0) = \infty \tag{2}$$

where

$$R = 0.5768\frac{V_{dd}^2}{P_{out}} \tag{3}$$

is the nominal Class E load-network resistance, V_{dd} is the supply voltage, and P_{out} is the fundamental-frequency output power delivered to the load [12]. The device output capacitance C_{out} should be equal to the nominal Class E shunt capacitance C defined by

$$C = \frac{0.1836}{\omega_0 R} \tag{4}$$

and the transformation to the standard load resistance R_L = 50 ohms is provided by the proper choice of the load-network transmission-line characteristic impedances [5, 6]. It should be noted that Eqs. (3) and (4) were obtained for the idealized optimum (or nominal) zero voltage and zero voltage-derivative Class E conditions when device is operated as a lossless switch and a series fundamentally-tuned resonant circuit provides an infinite impedance for the second and higher order harmonics. However, for example, for the practical transmission-line Class-E load networks shown in Figure 1(a), high impedance can be provided at the second harmonic only. In this case, the maximum efficiency can be achieved with nonzero voltage and voltage-derivative conditions, thus providing a second-harmonic Class E approximation when Eqs. (3) and (4) can be considered as an initial guess, with the optimum parameters optimized around these values [4].

Modified Approach: Analysis and Design

The Class-E load network shown in Figure 1(b) can be modified in order to obtain simple analytical equations to explicitly define the transmission line parameters. Such a modified transmission-line Class E load network is shown in Figure 2, where the combined series quarterwave transmission line provides an impedance transformation at the fundamental frequency, and the open-circuited stubs with electrical lengths of 90° and 30° create the open-circuit conditions, seen by the device output at the second and third harmonics, respectively.

Figure 3(a) shows the load network seen by the device output at the fundamental frequency. Here, the combined quarter-wavelength series transmission line $TL_1 + TL_2$, together with an open-circuited capacitive stub TL_4 having an electrical length of 30°, provides simultaneously a required inductive reactance and impedance transformation of the optimum Class E load resistance R to the load resistance R_L by proper choice of the transmission-line characteristic impedances Z_1 and Z_2.

The capacitive load impedance Z_L at the end of a quarterwave line at the fundamental frequency, representing by the load resistance R_L and capacitive stub TL_4, can be written as

$$Z_L = \frac{Z_2 R_L}{Z_2 + jR_L \tan 30^\circ} \tag{5}$$

where Z_2 is the characteristic impedance of a 30-degree open-circuit stub. Generally, the input impedance of the loaded transmission line can be written as

$$Z_{\mathrm{net}} = Z_1 \frac{Z_L + jZ_1 \tan\theta}{Z_1 + jZ_L \tan\theta} \tag{6}$$

where θ is the electrical length of the transmission line. Then, substituting Eq. (5) into Eq. (6) for θ = 90° results in an inductive input impedance

$$Z_{\mathrm{net}} = \frac{Z_1^2}{Z_L} = \frac{Z_1^2}{Z_2 R_L}\left(Z_2 + jR_L \tan 30^\circ\right) \tag{7}$$

when the required optimum Class-E resistance can be provided by proper choice of the characteristic impedance Z_1, while the required optimum Class-E inductive reactance can be achieved with the corresponding value of the characteristic impedance Z_2.

Separating Eq. (7) into real and imaginary parts results in the following system of two equations with two unknown parameters:

$$\mathrm{Re}\, Z_{\mathrm{net}} = \frac{Z_1^2}{R_L} \tag{8}$$

$$\mathrm{Im}\, Z_{\mathrm{net}} = \frac{Z_1^2}{\sqrt{3} Z_2} \tag{9}$$

which allows direct calculation of the characteristic impedances Z_1 and Z_2. As a result, by using Eq. (1),

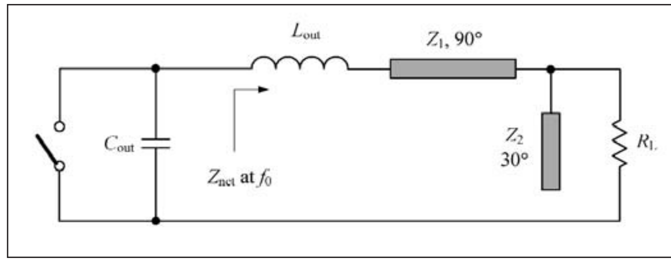

Figure 4 · Load network with series inductance at fundamental.

$$Z_1 = \sqrt{R_L R} \tag{10}$$

$$Z_2 = 0.5 R_L \tag{11}$$

where R_L = 50 ohms and R is calculated from Eq. (3).

The transmission-line Class E load network seen by the device output at the second harmonic is shown in Figure 3(b), taking into account the shorting effect of the quarterwave short-circuited stub TL_3, where the transmission line TL_1 provides an open-circuit condition for the second harmonic. At the third harmonic, the transmission-line Class E load network can similarly be represented, as shown in Figure 3(c), due to the open-circuit effect of the short-circuited quarterwave line TL_3 and short-circuit effect of the open-circuited harmonic stub TL_4 at the third harmonic. In this case, the combined transmission line $TL_1 + TL_2$ provides an open-circuit condition for the third harmonic at the device output being shorted at its right-hand side.

However, in a common case, it is necessary to take into account the transistor output parasitic series bondwire and lead inductance L_{out} shown in Figure 4, which provides an additional inductive reactance at the fundamental and does not affect the open-circuit conditions at the second and third harmonics. The inductive effect at the input of the series quarterwave transmission line should be reduced by proper changing of the characteristic impedance Z_2. In this case, Eq. (9) can be rewritten as

$$\mathrm{Im}\, Z_{\mathrm{net}} = \frac{Z_1^2}{\sqrt{3} Z_2} + \omega_0 L_{\mathrm{out}} \tag{12}$$

Hence, by using Eqs. (1) and (10), the characteristic impedance Z_2 can now be calculated from

$$Z_2 = \frac{R_L}{2} \frac{1}{1 - \dfrac{\omega_0 L_{\mathrm{out}}}{1.1586\mathrm{R}}} \tag{13}$$

resulting in higher characteristic impedance of the open-circuited stub for greater values of series inductance L_{out}.

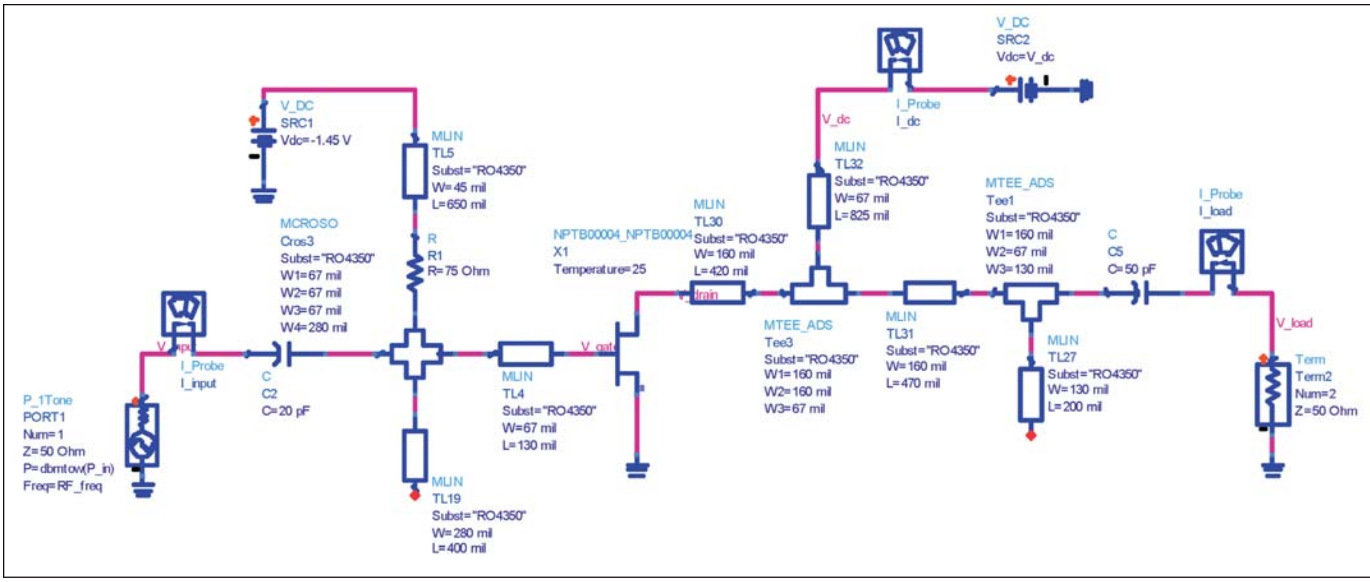

Figure 5 · Circuit schematic of Class E GaN HEMT power amplifier.

Simulation

Figure 5 shows the simulated circuit schematic of a transmission-line Class E power amplifier based on a 28 V 5 W Nitronex NPTB00004 GaN HEMT power transistor. The input matching circuit with an open-circuited stub and a series transmission line provides a complex-conjugate matching with the standard 50-ohm source. The load network represents the modified transmission-

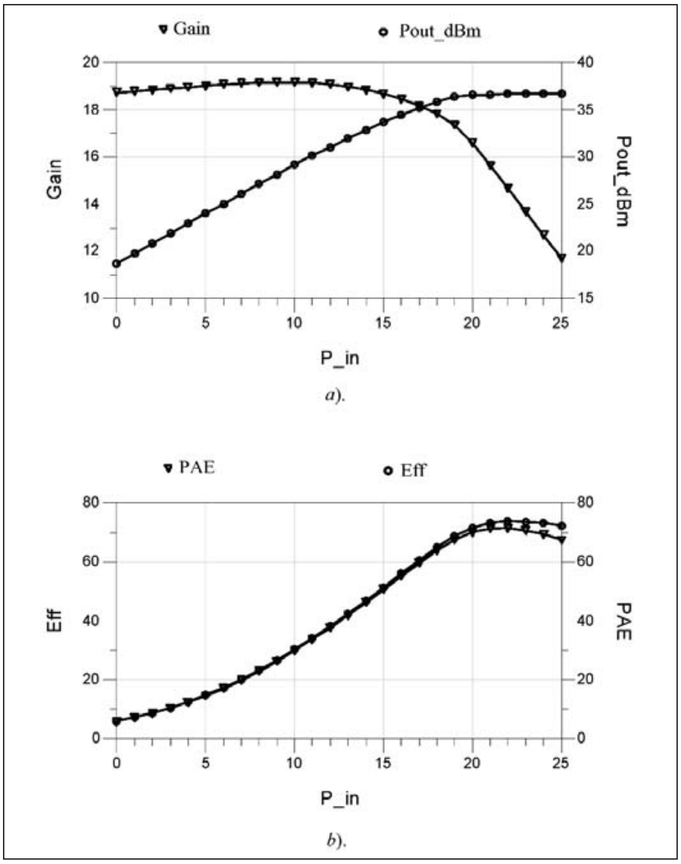

Figure 6 · Simulated results for Class E GaN HEMT power amplifier.

line Class-E load network shown in Figure 2.

Figure 6 shows the simulated results of a transmission-line Class E power amplifier using a RO4350 30-mil substrate. The maximum output power of 37 dBm, drain efficiency of 73% and power-added efficiency (PAE) of 71% at the center bandwidth frequency of 2.14 GHz are achieved with a power gain of 14 dB (linear gain of 19 dB) and a supply voltage of 25 V.

Implementation and Test

The transmission-line Class E power amplifier was fabricated on a RO4350 30-mil substrate. Figure 7 shows the test board of this power amplifier using a 5 W GaN HEMT NPTB00004 device. The input matching circuit, output load network, and gate and drain bias circuits (with bypass capacitors on their ends) are fully based on microstrip lines of different electrical lengths and characteristic impedances, according to the simulation setup shown in Figure 5.

Figure 8 shows the measured results with a maximum output power of 37 dBm, a drain efficiency of 70%, and a PAE of 61.5% with a power gain of 9.5 dB at the operating frequency of 2.14 GHz (gate bias voltage $V_g = -1.4$ V,

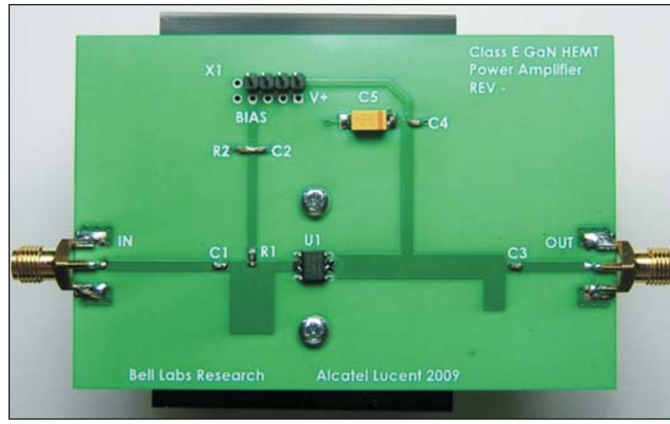

Figure 7 · Test board of Class E GaN HEMT power amplifier.

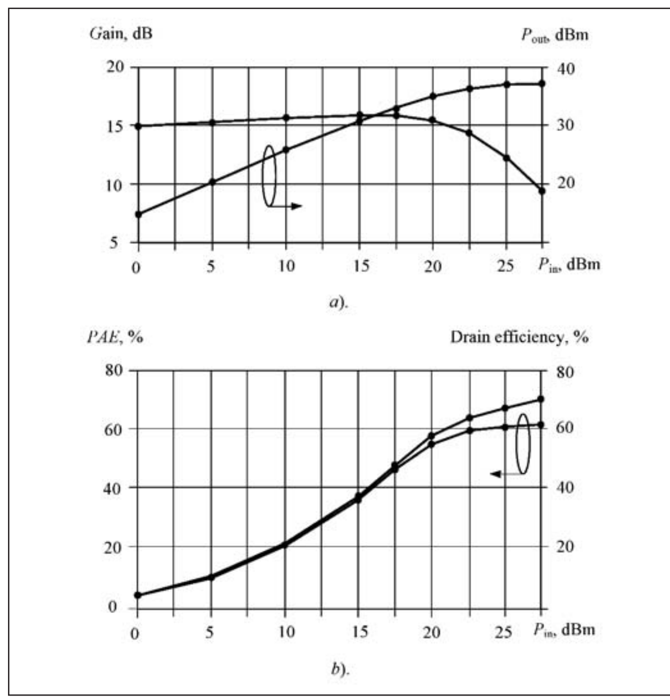

Figure 8 · Measured results for Class E power amplifier at 2.14 GHz.

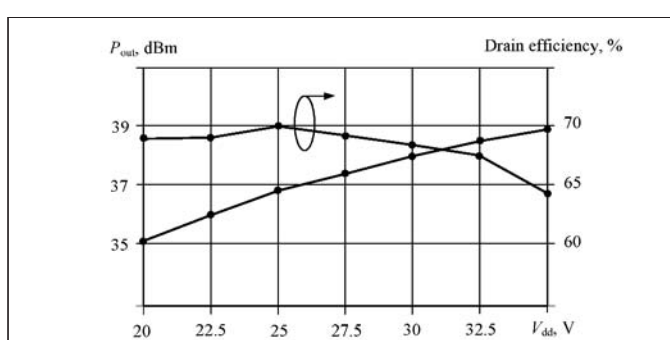

Figure 9 · Measured output power and drain efficiency versus supply voltage.

quiescent current I_q = 20 mA, and drain supply voltage V_{dd} = 25 V), achieved without any tuning of the input matching circuit and load network. In this case, the deeper the saturation mode, the lower DC supply current is measured, resulting in an increasing drain efficiency (70% and higher) with almost constant fundamental output power. The slightly lower power gain is explained by some mismatch at the input due to effect of the lead inductance of the packaged transistor.

Figure 9 shows the measured output power and drain efficiency versus dc supply voltage at the operating frequency of 2.14 GHz when an input power P_{in} was set to 27.5 dBm. The fundamental output power is varied almost linearly from 35 dBm at V_{dd} = 20 V up to almost 39 dBm at V_{dd} = 35 V. In this case, the maximum drain efficiency of 70% is achieved at an optimum DC supply voltage of 25 V.

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