High-Efficiency Class-E Power Amplifier With Shunt Capacitance and Shunt Filter

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***Abstract*—An analysis of a novel single-ended Class-E mode with shunt capacitance and shunt filter with explicit derivation of the idealized optimum voltage and current waveforms and load-network parameters with their verification by frequency domain simulations with 50% duty ratio is presented. The ideal collector voltage and current waveforms demonstrate a possibility of 100% efficiency. The circuit design with transmission lines at 2.14 GHz is discussed and analyzed. In order to reduce the voltage peak factor, the load network parameters can be rearranged to correspond to Class-E/F$_3$ mode by providing a short-circuit condition at the third harmonic when the second-harmonic tank is connected in series to the shunt filter. Broadband capability of a Class-E mode with shunt filter using reactance compensation technique has been demonstrated by two examples, one with lumped elements and the other with transmission-line elements. The test board of a transmission-line broadband Class-E GaN HEMT power amplifier with shunt filter was measured and high-performance results with the output power of around 41 dBm, average drain efficiency of 68%, and power gain of about 9 dB were achieved across the frequency band from 1.4 to 2.7 GHz.**

***Index Terms*—Circuit design, Class E, efficiency, GaN HEMT, harmonic balance technique, resonant circuit, RF power amplifier, time domain analysis, transmission line.**

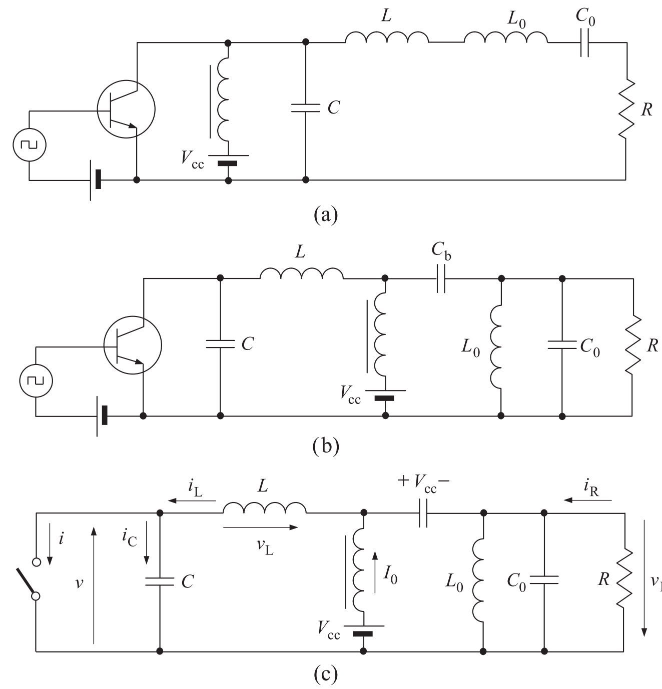

Fig. 1. Basic circuits of Class-E power amplifier with shunt filter.

I. Introduction

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, it is possible to apply a switchmode Class-E, inverse Class-F, or mixed Class-E/F mode technique [1], [2]. Such kinds of the power amplifiers require an operation in saturation mode resulting in a poor linearity, and therefore are not suitable to directly replace linear power amplifiers in WCDMA/LTE transmitters with non-constant envelope signal. However, to obtain both high efficiency and good linearity, the nonlinear high-efficiency power amplifier operating in an inverse Class-F or Class-E mode can be used in advanced transmitter architectures such as Doherty, LINC (linear amplification using nonlinear components), or EER (envelope elimination and restoration) with digital predistortion [3], [4].

The switchmode Class-E power amplifiers with shunt capacitance have found widespread application due to their design simplicity and high-efficiency operation [2], [5]. Their load-network configuration shown in Fig. 1(a) consists of a shunt capacitor, a series inductor, and a series fundamentally tuned L_0C_0 filter to provide high level of harmonic suppression. In this case, the transistor operates as an on-to-off switch and the shapes of the current and voltage waveforms provide a condition when the high current and high voltage do not overlap simultaneously that minimize the power dissipation and maximize the power amplifier efficiency. Such an operation mode can be realized for the tuned power amplifier by an appropriate choice of the values of the reactive elements in its load network. However, the circuit schematic with a shunt capacitor and series filter that can provide ideally 100% collector efficiency is not a unique.

This paper presents a novel alternative to Class E with shunt capacitance and series filter where a shunt filter is used instead of a series filter, thus resulting in a new load-network configuration, which can provide not only Class-E switching conditions but also broadband capability with high efficiency across the wide frequency band. The theoretical assumptions, simulation and experimental results for Class-E mode with shunt capacitance and shunt filter by providing a detailed analytical insight into the circuit analysis, explicit derivation of optimum voltage

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and current waveforms and load-network parameters, and load-network transmission-line design techniques are explained and discussed.

II. Circuit Analysis

The optimum parameters of a single-ended Class-E power amplifier with shunt capacitance and shunt filter can be determined based on an analytical derivation of its steady-state collector voltage and current waveforms. Fig. 1(b) shows the basic circuit configuration of a Class-E power amplifier with shunt capacitor and shunt filter, where the load network consists of a shunt capacitor C, a series inductor L, a blocking capacitor C_b, a shunt fundamentally tuned L_0C_0 circuit, and a load resistor R. In this case, the shunt L_0C_0 circuit operates as a harmonic filter creating zero impedance at the second- and higher-order harmonics instead of the open-circuit harmonic conditions corresponding to the conventional Class-E power amplifier with shunt capacitance and series filter. In a common case, a shunt capacitance C can represent the intrinsic device output capacitance and external circuit capacitance added by the load network. The dc power supply is connected by an RF choke with infinite reactance at the fundamental and any higher-order harmonic component. The active device is considered an ideal switch that is driven at the operating frequency to provide instantaneous switching between its on-state and off-state operation conditions.

In order to simplify the analysis of a single-ended Class-E power amplifier with shunt filter, whose equivalent circuit is shown in Fig. 1(c), the following several assumptions are introduced:

— The transistor has zero saturation voltage, zero saturation resistance, infinite off-resistance, and its switching is instantaneous and lossless.
— The shunt capacitance is assumed to be constant.
— The shunt L_0C_0 filter has zero impedance at the second- and higher-order harmonics.
— There is no loss in the circuit except the load R.
— For simplicity, a 50% duty ratio is used.

The idealized optimum (or nominal) Class-E switching conditions for the transistor switch can be written at $\omega t = 2\pi$ as

$$v(\omega t)|_{\omega t=2\pi} = 0 \quad (1)$$

$$\left.\frac{dv(\omega t)}{d\omega t}\right|_{\omega t=2\pi} = 0 \quad (2)$$

where v is the voltage across the switch. Note that these nominal Class-E switching conditions do not correspond to minimum dissipated power losses for the non-ideal transistor switch with finite value of its saturation resistance [2].

Let the output current i_R flowing through the load be sinusoidal

$$i_R(\omega t) = I_R \sin(\omega t + \varphi) \quad (3)$$

where I_R is the fundamental-frequency current amplitude and φ is the initial phase shift.

When the switch is turned on for of $0 \le \omega t \le \pi$, the voltage on the switch is

$$v(\omega t) = V_{cc} - v_L(\omega t) - v_R(\omega t) = 0 \quad (4)$$

where

$$v_L(\omega t) = \omega L \frac{di_L(\omega t)}{d\omega t} \quad (5)$$

$$v_R(\omega t) = V_R \sin(\omega t + \varphi) \quad (6)$$

where $V_R = I_R R$ is the voltage amplitude across the load.

Since the current flowing through the shunt capacitance C is

$$i_C(\omega t) = \omega C \frac{dv(\omega t)}{d\omega t} = 0 \quad (7)$$

consequently

$$i(\omega t) = i_L(\omega t) \quad (8)$$

where

$$i_L(\omega t) = \frac{1}{\omega L}\int_0^{\omega t} V_{cc}\, d\omega t + \frac{V_R}{\omega L}\cos(\omega t + \varphi) + i_L(0). \quad (9)$$

Since $i(0) = 0$, the initial value for the current $i_L(\omega t)$ flowing through the series inductor L at time instant $\omega t = 0$ can be defined using (8) and (9) as

$$i_L(0) = -\frac{V_R}{\omega L}\cos\varphi. \quad (10)$$

As a result, from (8) through (10), it follows that

$$i(\omega t) = \frac{V_{cc}}{\omega L}\omega t + \frac{V_R}{\omega L}[\cos(\omega t + \varphi) - \cos\varphi]. \quad (11)$$

When the switch is turned off for $\pi \le \omega t < 2\pi$, there is no current flowing through the switch, i.e., $i(\omega t) = 0$, and the current flowing through the capacitor C can be written as $i_C(\omega t) = i_L(\omega t)$, or

$$\begin{aligned} i_C(\omega t) &= \omega C \frac{dv}{d\omega t} \\ &= \frac{1}{\omega L}\int_\pi^{\omega t} [V_{cc} - v(\omega t) - v_R(\omega t)]\, d\omega t + i_L(\pi) \end{aligned} \quad (12)$$

under the initial off-state conditions $v(\pi) = 0$ and

$$i_C(\pi) = \frac{\pi V_{cc} - 2V_R\cos\varphi}{\omega L}. \quad (13)$$

Equation (12) can be represented in the form of the linear nonhomogeneous second-order differential equation by

$$\omega^2 LC \frac{d^2 v(\omega t)}{d(\omega t)^2} + v(\omega t) - V_{cc} + V_R \sin(\omega t + \varphi) = 0 \quad (14)$$

whose general solution can be given in the normalized form as

$$\frac{v(\omega t)}{V_{cc}} = C_1\cos q\omega t + C_2 \sin q\omega t + 1 + \frac{q^2}{1-q^2}\frac{V_R}{V_{cc}}\sin(\omega t + \varphi) \quad (15)$$

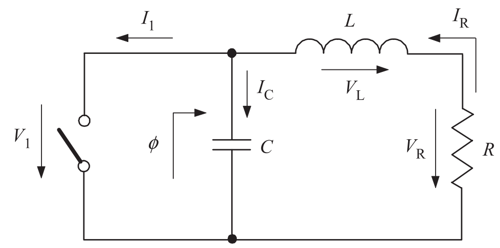


Fig. 2. Equivalent Class-E load network at fundamental frequency.

where

$$q = \frac{1}{\omega\sqrt{LC}} \tag{16}$$

and the coefficients C_1 and C_2 are determined from the initial off-state conditions by

$$C_1 = -\cos q\pi - q\pi \sin q\pi + \frac{q}{1-q^2}\frac{V_R}{V_{cc}} \times \left[q\cos q\pi \sin\varphi + (1-2q^2)\sin q\pi\cos\varphi\right] \tag{17}$$

$$C_2 = -\sin q\pi + q\pi\cos q\pi + \frac{q}{1-q^2}\frac{V_R}{V_{cc}} \times \left[q\sin q\pi\sin\varphi - (1-2q^2)\cos q\pi\cos\varphi\right]. \tag{18}$$

In this case, when compared to Class E with shunt capacitance and series filter where the circuit electrical behavior is described by the first-order differential equation, the general solution is derived from the second-order differential equation, similarly to that for Class E with parallel circuit [2], [6]. The fundamental-frequency voltage $v_1(\omega t)$ across the switch consists of the two quadrature components, as shown in Fig. 2, with an amplitude of the real component defined by Fourier formula as

$$V_R = -\frac{1}{\pi}\int_0^{2\pi} v(\omega t)\sin(\omega t + \varphi)d\omega t. \tag{19}$$

As a result, by solving a system of three equations—two of them defined by the Class-E switching conditions given by (1) and (2) and the third one for V_R given by (19)—the three unknown parameters can be calculated as

$$\varphi = -41.614^\circ \tag{20}$$

$$q = 1.607 \tag{21}$$

$$\frac{V_R}{V_{cc}} = 0.9253. \tag{22}$$

For an idealized collector efficiency of 100%, the dc power $P_0 = I_0 V_{cc}$ and the fundamental-frequency output power $P_{out} = V_R^2/(2R)$ delivered to the load are equal, i.e.,

$$I_0 V_{cc} = \frac{V_R^2}{2R}. \tag{23}$$

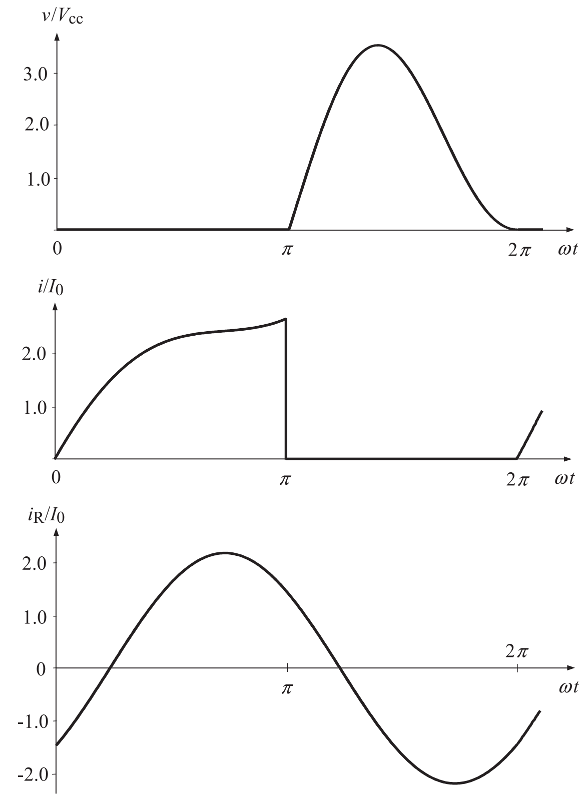


Fig. 3. Ideal waveforms for nominal Class-E mode with shunt filter.

The dc current I_0 can be determined by applying Fourier-series expansion to (11) as

$$I_0 = \frac{1}{2\pi}\int_0^{2\pi} i(\omega t)d\omega t = \frac{V_{cc}}{2\pi\omega L}\left[\frac{\pi^2}{2} - \frac{V_R}{V_{cc}}(2\sin\varphi + \pi\cos\varphi)\right]. \tag{24}$$

By using (23) and (24), equation for the optimum normalized series inductance L can be derived as

$$\frac{\omega L}{R} = \frac{\frac{1}{\pi}\left[\frac{\pi^2}{2} - \frac{V_R}{V_{cc}}(2\sin\varphi + \pi\cos\varphi)\right]}{\left(\frac{V_R}{V_{cc}}\right)^2}. \tag{25}$$

Hence, the optimum normalized series inductance L and shunt capacitance C can be calculated using (16), (20)–(22), and (25) as

$$\frac{\omega L}{R} = 1.4836 \tag{26}$$

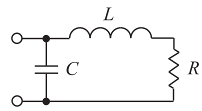
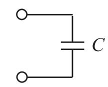
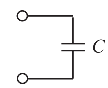
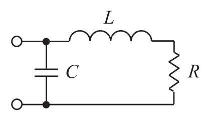
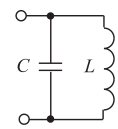
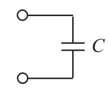
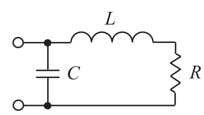
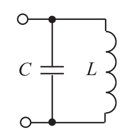
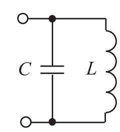
$$\omega CR = 0.261. \tag{27}$$

The optimum load resistance R can be obtained from (22) and (23) for the dc supply voltage V_{cc} and fundamental-frequency output power P_{out} delivered to the load as

$$R = \frac{1}{2}\frac{V_R^2}{P_{out}} = \frac{1}{2}\left(\frac{V_R}{V_{cc}}\right)^2\frac{V_{cc}^2}{P_{out}} = 0.4281\frac{V_{cc}^2}{P_{out}}. \tag{28}$$

Fig. 3 shows the normalized collector voltage and current waveforms and sinusoidal current waveform for idealized

TABLE I
OPTIMUM IMPEDANCES AT FUNDAMENTAL AND HARMONICS

Class-E load network	f_0 (fundamental)	$2nf_0$ (even harmonics)	$(2n+1)f_0$ (odd harmonics)
Class E with shunt capacitance and series filter			
Class E with quarterwave line and series filter [2]			
Class E with shunt capacitance and shunt filter			

optimum Class-E mode with shunt capacitance and shunt filter during the entire interval $0 \leq \omega t \leq 2\pi$. From the collector voltage and current waveforms, it follows that, when the transistor is turned on, there is no voltage across the switch and the current from the inductor flows through the switch. However, when the transistor is turned off, this current flows through the capacitor C. In this case, there is no nonzero voltage and current simultaneously, which means a lack of the power losses that gives an idealized collector efficiency of 100% written by (23). Note that the collector voltage and current waveforms of Class E with shunt capacitance and shunt filter are very close to those corresponding to Class E with quarterwave transmission line [2].

The phase angle ϕ of the load network at fundamental seen by the switch shown in Fig. 2 and required for an idealized optimum (or nominal) Class-E mode with shunt capacitance and shunt filter can be determined through the load-network parameters using (26) and (27) as

$$\phi = \tan^{-1}\left(\frac{\omega L}{R}\right) - \tan^{-1}\left(\frac{\omega CR}{1 - \frac{\omega L}{R}\omega CR}\right) = 32.945^\circ. \tag{29}$$

The peak collector voltage $V_{\max}$ and current $I_{\max}$ can be determined from (11) and (15) using (24) as

$$\frac{V_{\max}}{V_{\rm cc}} = 3.677 \tag{30}$$

$$\frac{I_{\max}}{I_0} = 2.768 \tag{31}$$

that shows that the voltage peak factor is as high as in a conventional Class E with shunt capacitance and series filter.

The maximum operating frequency $f_{\max}$ in a nominal Class-E mode with shunt capacitance and shunt filter is limited by the device output (collector or drain-source) capacitance $C_{\rm out}$. As a result, from (27) and (28) it follows that

$$f_{\max} = 0.097\frac{P_{\rm out}}{C_{\rm out}V_{\rm cc}^2} \tag{32}$$

TABLE II
LOAD-NETWORK PARAMETERS FOR DIFFERENT CLASS-E MODES

Normalized load-network parameter	Class E with shunt capacitance and series filter	Class E with quarterwave line and series filter [2]	Class E with shunt capacitance and shunt filter
$\frac{\omega L}{R}$	1.1525	1.349	1.4836
ωCR	0.1836	0.2725	0.261
$\frac{P_{\rm out}R}{V_{\rm cc}^2}$	0.5768	0.465	0.4281
$\frac{f_{\max}C_{\rm out}V_{\rm cc}^2}{P_{\rm out}}$	0.0507	0.093	0.097

which is 1.91 times higher than that for Class E with shunt capacitance and series filter [2].

In Table I, the optimum impedances seen by the device collector at the fundamental-frequency and higher-order odd and even harmonic components are illustrated by the appropriate circuit configurations. It can be seen that Class-E mode with shunt capacitance and shunt filter shows different impedance properties from other alternative Class-E load networks. At even harmonics, its optimum impedances can be established by the parallel LC circuit, similarly to Class E with quarterwave line and series filter. However, at odd harmonics, the optimum impedances for Class E with shunt capacitance and shunt filter differ from Class E with series filter and Class E with quarterwave line where impedances are defined by the shunt capacitances, because they are also provided by the parallel LC circuits.

The optimized load-network parameters of the different Class-E modes including Class E with series filter, Class E with quarterwave line, and Class E with shunt filter are shown in Table II in a normalized form. As can be seen, Class E with shunt filter offers the larger value of the shunt capacitance C for the same load R and much higher value of the maximum operating frequency $f_{\max}$ for the same dc supply voltage $V_{\rm cc}$,

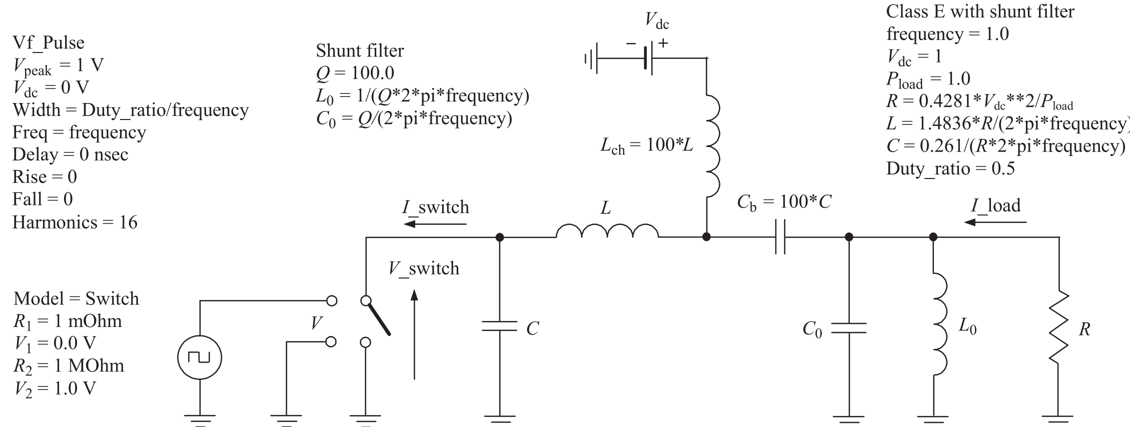


Fig. 4. Simulation setup for idealized Class-E mode with series inductance and shunt filter in frequency domain.

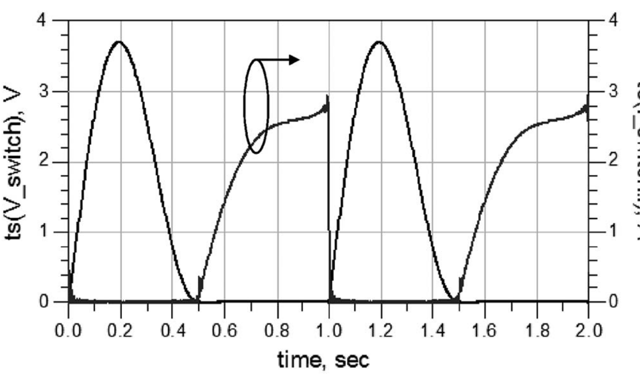


Fig. 5. Simulated switch voltage and current waveforms in frequency domain.

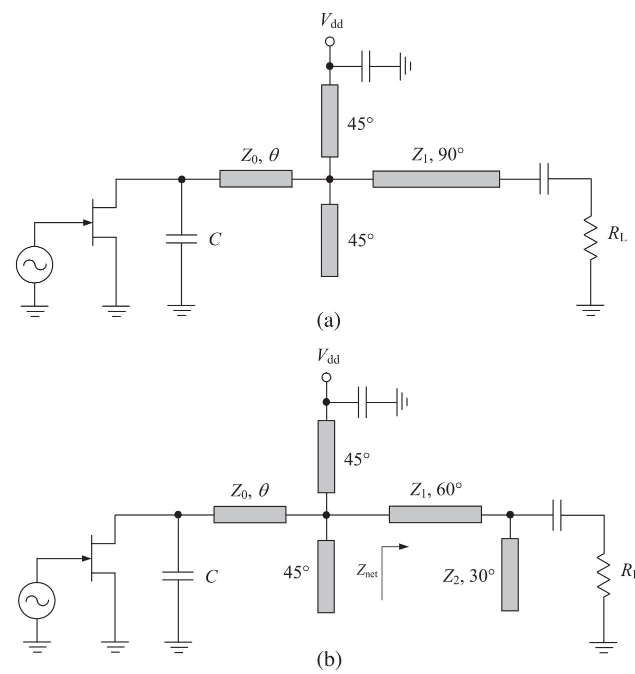


Fig. 6. Schematics of transmission-line Class-E power amplifier with shunt filter.

device output capacitance $C_{\rm out}$, and output power $P_{\rm out}$, compared to Class E with shunt capacitance and series filter. At the same time, difference between Class E with quarterwave line and Class E with shunt filter is not so significant because the quarterwave line being grounded at its end through bypass capacitor operates for even harmonics as a shunt filter.

III. Idealized Simulation

The ADS simulation setup shown in Fig. 4 is used to obtain the nominal Class-E operation mode with infinite RF choke in frequency domain, where the input source Vf_Pulse represents a voltage source with Fourier series expansion of periodic rectangular wave with different pulse width characterized by a duty ratio used in a harmonic balance simulator. Generally, using frequency domain enables the overall simulation procedure to be much faster than that in time domain and can take a few seconds [2]. However, because the number of harmonic components is not infinite, the simulation waveforms and numerical results for the optimum load-network parameters are not so accurate. The harmonic order is chosen to 100.

In this case, the optimization procedure can be applied with respect to efficiency as an optimization parameter. Since the simulation time is very short, a number of iterations can significantly be increased for more accuracy. Fig. 5 shows the normalized switch voltage and current waveforms obtained for the idealized optimum (or nominal) parameters of the Class-E load network with shunt capacitor and shunt filter given by (26) through (28) with a 50% duty ratio.

Unlike the simulations in time domain, the frequency-domain simulations are characterized by smoother transitions between the positions when the switch is turned on and the switch is turned off and vice versa. Nevertheless, for $r_{\rm sat} = 1\ {\rm m}\Omega$ and lossless circuit elements except load resistance R, the efficiency is equal to 98.8%, and the simulated normalized switch voltage and current waveforms shown in Fig. 5 are similar to the same theoretical waveforms shown in Fig. 3.

IV. Design With Transmission Lines

The circuit schematics of a high-efficiency transmission-line Class-E power amplifier with shunt capacitor and shunt filter are shown in Fig. 6, where the optimum load resistance R is

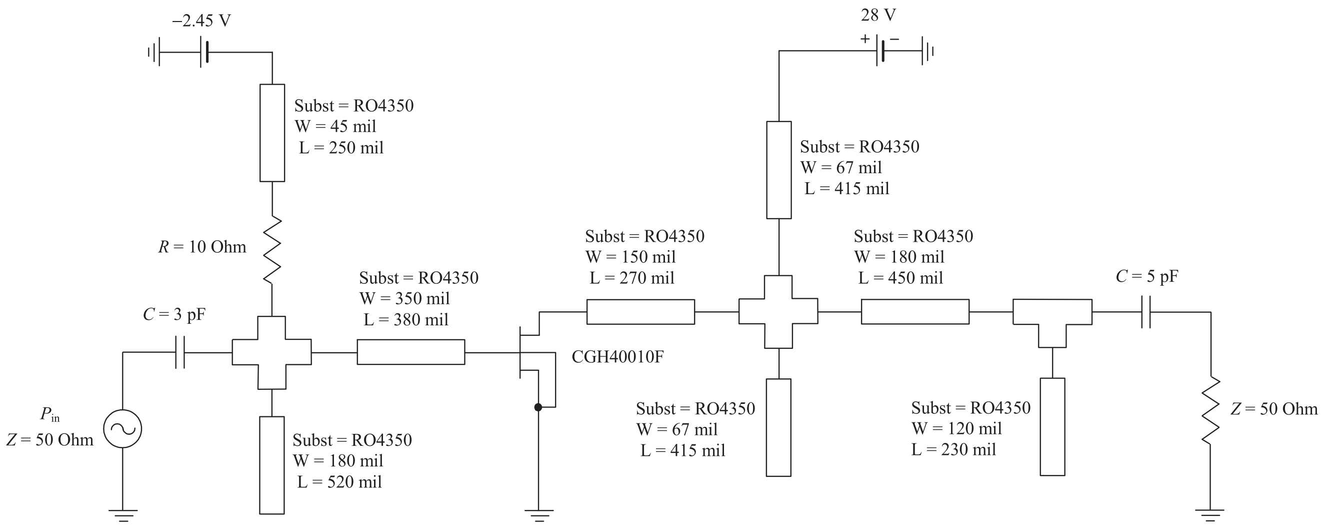


Fig. 7. Circuit schematic of transmission-line Class-E GaN HEMT power amplifier with shunt filter.

matched to the standard load impedance $R_L = 50\ \Omega$ at the fundamental frequency using a quarterwave transmission line, as shown in Fig. 6(a), or using a transmission-line L-type impedance transformer, as shown in Fig. 6(b). Here, the shunt harmonic filter is composed of 45° short-circuit and open-circuit stubs to create short-circuit conditions at even harmonics. To create a short-circuit condition at the third harmonic at the right-hand side of the series inductor represented by a short-length series transmission line with the characteristic impedance Z_0 and electrical length θ, the series transmission line with electrical length of 60° and an open-circuit stub with electrical length of 30° are used.

For electrical length of a sufficiently short series transmission line with the characteristic impedance Z_0 and electrical length θ of less than 45°, the required optimum value of θ for Class-E mode with shunt capacitance and shunt filter using (26) can be approximated by

$$\theta = \tan^{-1}\left(1.4836\frac{R}{Z_0}\right). \tag{33}$$

The output matching circuit is necessary to match to the required optimum Class-E resistance R calculated in accordance with (28) to the standard load resistance of 50 Ω. In addition, it is required to provide a short-circuit condition at the third harmonic component. This can be easily done using the output matching topology in the form of an L-type transformer with the series transmission line and open-circuit stub. Its load-network impedance $Z_{\rm net}$ at the fundamental can be written as

$$Z_{\rm net} = Z_1\frac{R_L(Z_2 - Z_1\tan 30^\circ\tan 60^\circ) + jZ_1Z_2\tan 60^\circ}{Z_1Z_2 + j(Z_1\tan 30^\circ + Z_2\tan 60^\circ)R_L} \tag{34}$$

where Z_1 and $\theta_1 = 60^\circ$ are the characteristic impedance and electrical length of the series transmission line, and Z_2 and $\theta_2 = 30^\circ$ are the characteristic impedance and electrical length of the open-circuit stub.

Hence, the complex-conjugate matching with the load can be provided by a proper choice of the characteristic impedances Z_1 and Z_2. Separating (34) into real and imaginary parts and taking into account that $\mathrm{Re}Z_L = R$ and $\mathrm{Im}Z_L = 0$, the system of two equations with two unknown parameters can be written as

$$(Z_1 + 3Z_2)^2R_L^2R - 3Z_1^2Z_2^2(4R_L - 3R) = 0 \tag{35}$$

$$3Z_1^2Z_2^2 + R_L^2(Z_1 + 3Z_2)(Z_2 - Z_1) = 0 \tag{36}$$

which enables the characteristic impedances Z_1 and Z_2 to be properly calculated.

This system of two equations can be explicitly solved as a function of the parameter $r = R_L/R$, resulting in

$$\frac{Z_1}{R_L} = \frac{\sqrt{4r-1}}{\sqrt{3}r} \tag{37}$$

$$\frac{Z_1}{Z_2} = \frac{r-1}{r}. \tag{38}$$

Consequently, for specified value of the parameter r with the required optimum load resistance R, corresponding to Class E with shunt capacitance and shunt filter, and standard load $R_L = 50\ \Omega$, first the characteristic impedance Z_1 is calculated from (37) and then the characteristic impedance Z_2 is calculated from (38). For example, if the required optimum load resistance is equal to $R = 20\ \Omega$, resulting in $r = 2.5$, the characteristic impedance of the series transmission line is equal to $Z_1 = 35\ \Omega$ and the characteristic impedance of the open-circuit stub is equal to $Z_2 = 58\ \Omega$.

Fig. 7 shows the simulated circuit schematic, which approximates the transmission-line Class-E power amplifier with shunt filter, based on a 28-V 10-W Cree GaN HEMT power transistor CGH40010F and transmission-line load network shown in Fig. 6(b). The input matching circuit provides a complex-conjugate matching with the standard 50-Ω source. The load network was slightly modified by optimizing the parameters of the series and shunt transmission lines because the device output capacitance $C_{\rm out}$ and series inductance $L_{\rm out}$ formed by drain bondwires and package lead do not match the required exact values of C and L for a nominal Class-E mode.

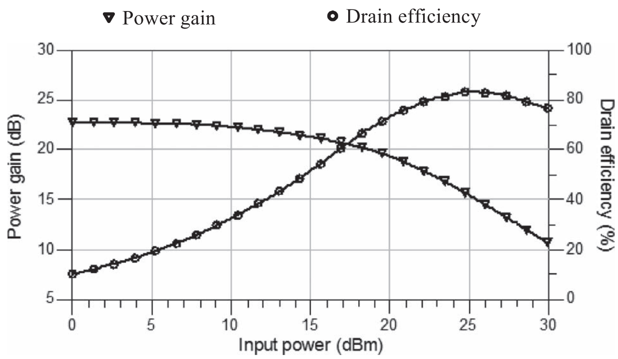


Fig. 8. Simulated results of transmission-line Class-E GaN HEMT power amplifier with shunt filter.

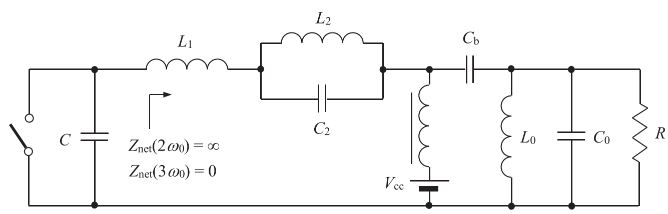


Fig. 9. Load network of Class E/F$_3$ with series tank circuit and shunt filter.

For the transmission-line GaN HEMT Class-E power amplifier with shunt filter using a 30-mil RO4350 substrate, the simulated drain efficiency of 83%, a power-added efficiency (PAE) of 80.4%, and a power gain of 15 dB at an output power of 40.3 dBm with a quiescent current of 30 mA are achieved at an operating frequency of 2.14 GHz with a supply voltage of 28 V, as shown in Fig. 8. The second and third harmonics were suppressed by greater than 50 dB.

V. Class E/F$_3$ With Series Tank Circuit and Shunt Filter

In order to reduce the collector voltage peak factor, a series resonant circuit tuned to the third harmonic can be used in the Class-E load network with series filter, when the voltage peak factor reduces by 13.4% compared to a conventional Class-E mode [7], [8]. Alternatively, a parallel LC resonator tuned to the second harmonic can be added in series to the load network, thus reducing the voltage peak factor by 20% [9]. However, because of a sufficiently low optimum loaded Q_L-factor of the second-harmonic resonator required to provide flat-top voltage waveform, efficiency drops by almost 20% due to power dissipation at harmonics. In this case, the proposed load network with shunt filter can be rearranged to correspond to Class-E/F$_3$ mode without power loss at harmonics by providing a short-circuit condition at the third harmonic when the L_2C_2 tank circuit connected in series is tuned to the second harmonic and provides a zero reactance at the third harmonic together with the series inductance L_1, as shown in Fig. 9.

In this case, the network reactance $\mathrm{Im}Z_{\mathrm{net}}$ at the fundamental frequency at the input of this lumped network can be written as

$$\mathrm{Im}Z_{\mathrm{net}}(\omega_0) = \omega_0 L_1 + \frac{\omega_0 L_2}{1 - \omega_0^2 L_2 C_2} = \omega_0 L \tag{39}$$

where L_1 is the series inductance and the series L_2C_2 tank circuit is tuned to the second harmonic

$$1 - 4\omega_0^2 L_2 C_2 = 0. \tag{40}$$

To provide a short-circuit condition at the third harmonic ($\omega = 3\omega_0$), (39) can be rewritten as

$$\mathrm{Im}Z_{\mathrm{net}}(3\omega_0) = 3\omega_0 L_1 + \frac{3\omega_0 L_2}{1 - 9\omega_0^2 L_2 C_2} = 0. \tag{41}$$

As a result, the ratios between the load-network parameters can be obtained by

$$L_1 = \frac{3}{8} L \tag{42}$$

$$L_2 = \frac{15}{32} L \tag{43}$$

$$C_2 = \frac{8}{15\omega_0^2 L} \tag{44}$$

where L is an optimum inductance defined by (26).

The lumped Class-E/F$_3$ power amplifier with shunt filter based on a 28-V 10-W Cree GaN HEMT power transistor CGH40010F was simulated at an operating frequency of 100 MHz, according to the simulation setup shown in Fig. 10. In this case, the required optimum load resistance given by (28) is close to 30 Ω, therefore it is enough to use a simple low-pass L-type matching section to match to the standard load of 50 Ω. An entire shunt capacitance defined by (27) includes both external capacitance and device output capacitance. The values of the series inductance and third-harmonic tank circuit are obtained by (42) through (44) using (26). The input matching is provided by a simple RL series circuit connected to the device input in parallel and followed by a series resistance.

As a result, for an infinite inductor quality factor in the load network, an output power of 40 dBm, a drain efficiency of 89.2%, and a PAE of 87% with a power gain of 16 dB were simulated for an input power of 24 dBm at a supply dc voltage of 28 V with a quiescent current of 30 mA, as shown in Fig. 11(a). Due to the series second-harmonic tank and shunt filter, the second- and higher-order harmonic components are suppressed by greater than 50 dB, as shown in Fig. 11(b). Fig. 11(c) shows the simulated drain waveform demonstrating the effect of an inverse Class-F$_3$ mode, resulting in a voltage peak factor of $80/28 = 2.86$, which is significantly less (by 1.29 times) than that for Class E with shunt filter defined by (30).

VI. Broadband Capability

The conventional design of a high-efficiency Class-E power amplifier requires a high loaded Q_L-factor of the series filter to satisfy the necessary harmonic impedance conditions at the device output and to provide sinusoidal current flowing to the load. However, if a sufficiently small value of the loaded quality factor Q_L is chosen, a high-efficiency broadband operation of the Class-E power amplifier can be realized by applying the reactance compensation technique. For example, a simple

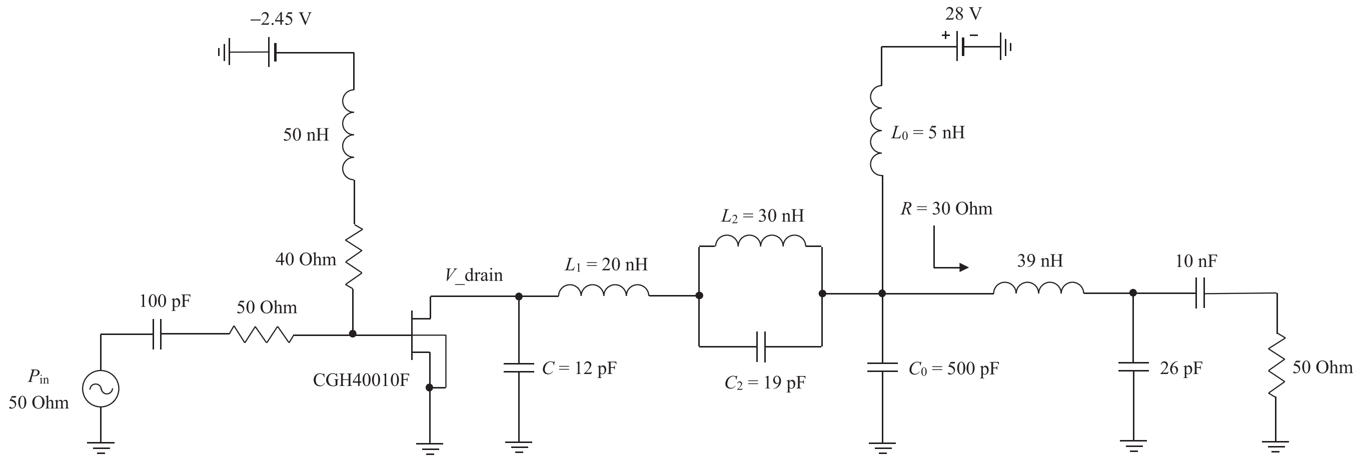


Fig. 10. Simulation setup for lumped GaN HEMT Class-E/F_3 power amplifier with shunt capacitance, second-harmonic tank, and shunt filter.

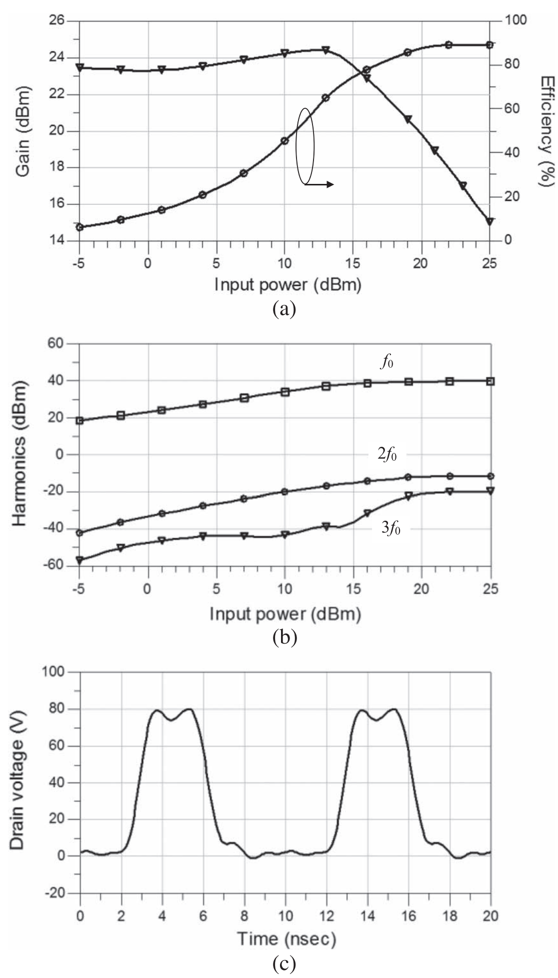


Fig. 11. Simulated results of lumped Class-E/F_3 GaN HEMT power amplifier.

network consisting of a series resonant LC circuit tuned to the fundamental frequency, a parallel inductor, and a load resistor provides a constant load phase angle of 50° in a frequency bandwidth of about 50% around the center bandwidth frequency [10].

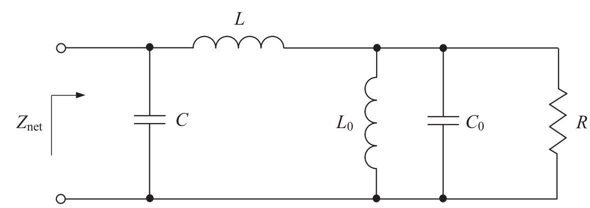


Fig. 12. Equivalent circuit of Class E with shunt capacitor and shunt filter.

Usually, the bandwidth limitation in power amplifiers comes from the device low transition frequency f_T and large output capacitance $C_{\rm out}$. Therefore, silicon LDMOSFET technology has been the preferred choice up to 2 GHz. As an alternative, GaN HEMT technology enables high efficiency, large breakdown voltage, high power density, and significantly higher broadband performance due to higher transition frequency and smaller periphery, resulting in smaller input and output capacitances and less parasitics.

It is very difficult to maintain efficiency at a high level over very wide frequency bandwidth. For a Class-E load network with shunt capacitance, a PAE above 50% was achieved within the frequency range from 1.9 to 2.4 GHz [11]. To increase high-efficiency frequency bandwidth, the broadband Class-E technique based on a reactance compensation principle with a combination of the series and shunt resonant circuits can be used [12]. In this case, a PAE over 53% was observed in a frequency bandwidth of 2.1 to 2.7 GHz with output power variations from 9.3 to 12.7 W at a supply voltage of 40 V [13].

To describe reactance compensation technique, let us consider the simplified equivalent load network with a shunt capacitor C, a series inductor L, and a shunt L_0C_0 circuit tuned to the fundamental, as shown in Fig. 12. In order to maintain the load phase angle constant in a wide frequency range, the positive slope of the reactance provided by the L-type circuit should be cancelled by the negative slope of the reactance provided by the shunt L_0C_0 circuit. In this case, the load-network admittance $Y_{\rm net} = 1/Z_{\rm net}$ can be written as

$$Y_{\rm net} = j\omega C + \frac{1 + j\omega' C_0 R}{R(1 - j\omega\omega' L C_0) + j\omega L} \tag{45}$$

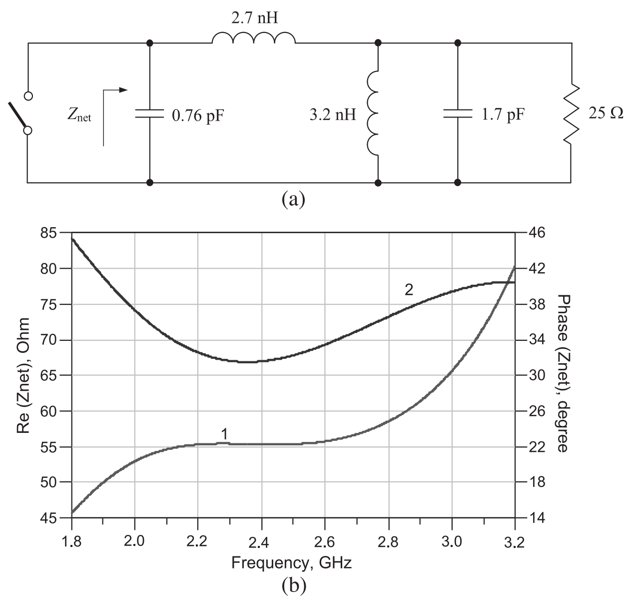


Fig. 13. Reactance-compensation Class-E circuit with lumped elements and its performance.

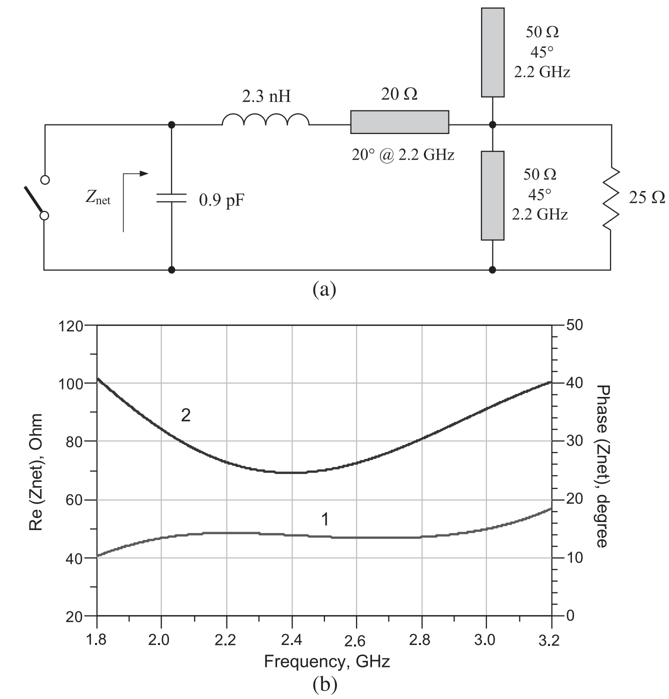


Fig. 14. Reactance-compensation Class-E circuit with lumped elements and transmission lines and its performance.

where

$$\omega' = \omega\left(1 - \frac{\omega_0^2}{\omega^2}\right) \tag{46}$$

$$\omega_0 = \frac{1}{\sqrt{L_0 C_0}} \tag{47}$$

is the radian resonant frequency. For a nominal Class-E mode with shunt capacitance and shunt filter, the normalized inductance L and capacitance C are determined by (26) and (27), respectively.

The frequency bandwidth with a constant load phase angle will be maximized if, at midband frequency ω_0

$$\left.\frac{dB_{\rm net}(\omega)}{d\omega}\right|_{\omega=\omega_0} = 0 \tag{48}$$

where the load-network susceptance $B_{\rm net} = {\rm Im}Y_{\rm net}$ is defined by

$$B_{\rm net} = \omega C + \frac{\omega' C_0 R^2(1 - j\omega\omega' L C_0) - \omega L}{R^2(1 - j\omega\omega' L C_0)^2 + (\omega L)^2}. \tag{49}$$

As a result, an additional equation can be derived as

$$2\omega_0 C_0 R - \frac{\omega_0 L}{R} - \omega_0 C R \frac{\left(\frac{\omega_0 L}{R}\right)^2 + 1}{\left(\frac{\omega_0 L}{R}\right)^2 - 1} = 0. \tag{50}$$

Finally, the shunt capacitance C_0 and inductance L_0 can be obtained by

$$C_0 = \frac{1.0896}{\omega_0 R} \tag{51}$$

$$L_0 = \frac{1}{\omega_0^2 C_0}. \tag{52}$$

Fig. 13(a) shows the example of a reactance-compensation load network for Class-E power amplifier with shunt filter using lumped elements. In this case, the reactance of the circuit with shunt capacitor and series inductor varies similar to that of the series resonant circuit with positive slope, whereas the required negative slope is provided by the parallel resonant circuit. Selection of the proper circuit parameters according to (26) through (28) for Class-E mode together with (51) and (52) for shunt filter elements enables the magnitude of the two slopes to be made identical, so as to achieve a constant total reactance and phase of the load-network impedance $Z_{\rm net}$ over a wide frequency range.

The simulation results at the fundamental frequency show that the resistance ${\rm Re}Z_{\rm net}$ varies from 50 Ω at 1.9 GHz to 65 Ω at 3.0 GHz, as shown in Fig. 13(b) by curve 1, whereas the load-network phase varies between 31.5° and 41° from 1.9 to 3.2 GHz (curve 2).

Fig. 14(a) shows the example of a reactance-compensation load network for a Class-E power amplifier with shunt filter using both lumped elements and transmission lines. In this case, the shunt filter is represented by the open- and short-circuit stubs replacing lumped capacitor and inductor, respectively, each having a characteristic impedance of 50 Ω and electrical length of 45° at 2.2 GHz. The series inductive reactance is created by lumped inductor that can be represented by the device bondwire and package lead and short transmission line. The simulation results at the fundamental frequency show that the load-network resistance ${\rm Re}Z_{\rm net}$ and phase vary between 40 Ω and 57 Ω (curve 1) and between 25° and 41° (curve 2) from 1.8 to 3.2 GHz, as shown in Fig. 14(b).

Generally, very broadband power amplifier design employs an input lossy LCR matching circuit to minimize the input

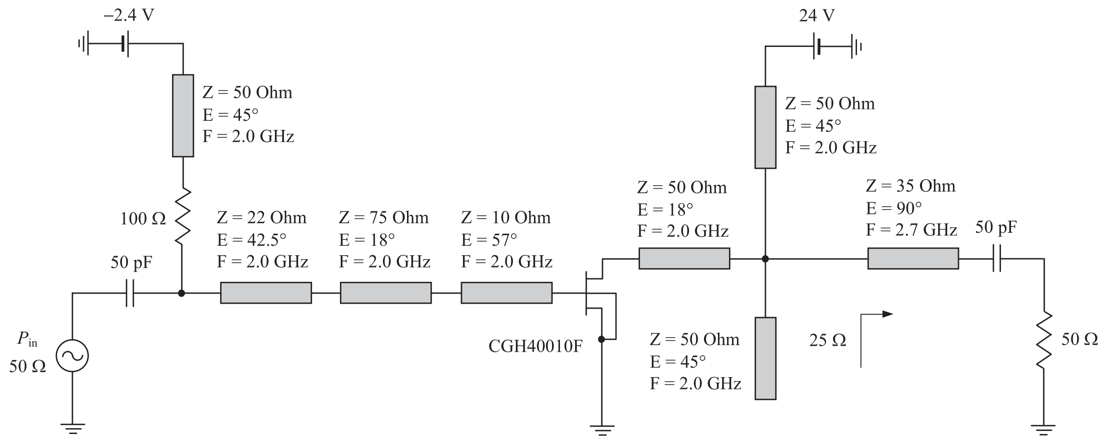


Fig. 15. Idealized circuit schematic of broadband Class-E GaN HEMT power amplifier with shunt filter.

return loss and output power variations over very wide frequency bandwidths with an LC output network to compensate for the device output reactance [14], [15]. As an alternative, to provide an input broadband matching over an octave bandwidth, it is possible to use a multisection matching transformer consisting of stepped transmission-line sections with different characteristic impedances and electrical lengths [12]. Such an input matching structure is convenient in practical implementation because there is no need to use any tuning capacitors.

VII. Simulation and Implementation

Fig. 15 shows the idealized simulation setup of a 10-W broadband Class-E power amplifier circuit with shunt filter designed to operate across the frequency band from 1.7 to 2.7 GHz and based on a GaN HEMT Cree CGH40010 device, where both the input matching circuit and load network are composed of ideal microstrip transmission lines. The nominal Class-E load resistance calculated for $P_{\rm out}$=10 W and $V_{\rm dd}$ = 24 V from (28) is equal to about 25 Ω. In this case, the transmission-line broadband Class-E load network with shunt filter having a 25-Ω load is represented by the open- and short-circuit stubs replacing the lumped capacitor and inductor, respectively, each having a characteristic impedance of 50 Ω and electrical length of 45° at 2.0 GHz. An additional series transmission line with 35-Ω characteristic impedance and of a quarter wavelength at the high bandwidth frequency of 2.7 GHz is used to match an idealized 25-Ω load with a standard 50-Ω load.

As a result, an output power of more than 41 dBm with a power gain of around 10 dB was simulated for an input power of 31 dBm, as shown in Fig. 16(a). In this case, a drain efficiency over 73% and a PAE over 67% were achieved across the required frequency range from 1.7 to 2.7 GHz, as shown in Fig. 16(b). Previously, the drain efficiency greater than 60% was achieved between 1.8 GHz and 2.3 GHz with a 45-W GaN HEMT Cree CGH40045F device using a distributed second-harmonic termination with short-circuit stubs [16].

Fig. 17(a) shows the test PCB board of the implemented broadband Class-E GaN HEMT power amplifier using a Cree CGH40010P device in a pill-type package [17]. The input

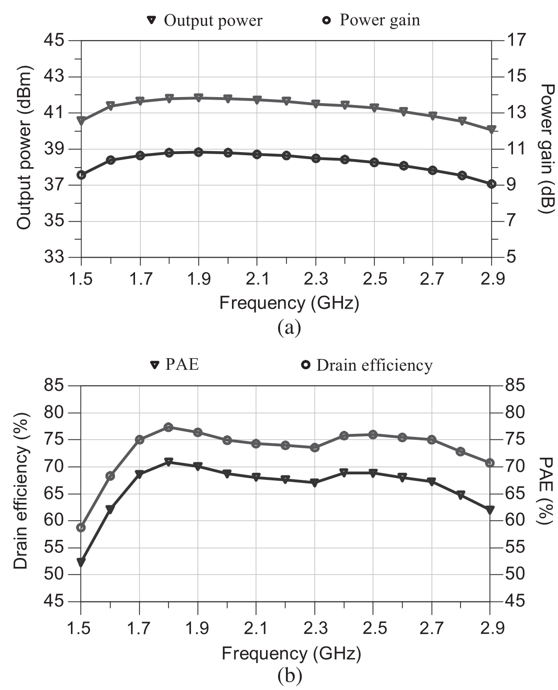


Fig. 16. Simulated results of broadband Class-E GaN HEMT power amplifier with hunt filter.

matching circuit, output load network, and gate and drain bias circuits (having bypass capacitors on their ends) are fully based on microstrip lines fabricated on a 20-mil RO4360 substrate according to their simulation parameters shown in Fig. 15. Only the width of a series microstrip line connected to the drain terminal was slightly increased to make it longer.

Fig. 17(b) shows the measured results of a transmission-line broadband Class-E GaN HEMT power amplifier with shunt filter having an output power of 40.6 ± 0.9 dBm and an average drain efficiency of about 68% (from 63% to 73%) across the frequency bandwidth from 1.4 to 2.7 GHz. In this case, a power gain of around 9.5 dB at a drain supply voltage of 24 V was achieved without any tuning of the input matching circuit and load network.

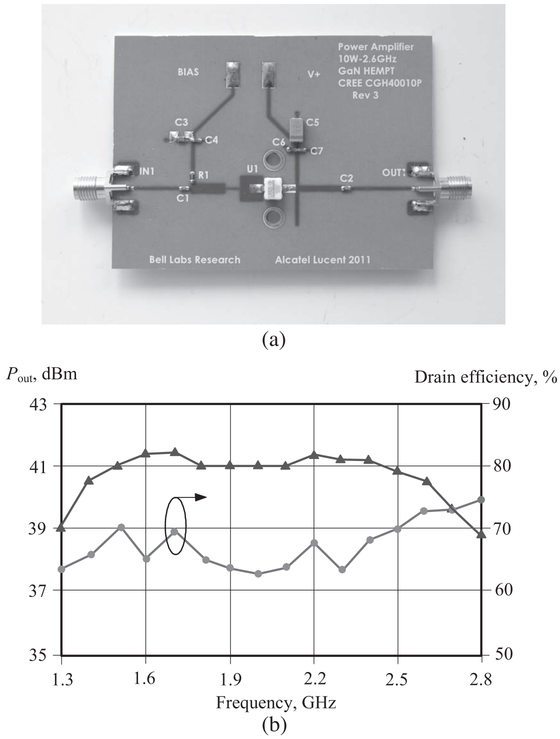


Fig. 17. Test board and measured results of transmission-line broadband Class-E GaN HEMT power amplifier with shunt filter.

VIII. Conclusion

The theoretical analysis of a novel single-ended Class-E mode with shunt capacitance and shunt filter with explicit derivation of the idealized optimum voltage and current waveforms and load-network parameters with their verification by frequency domain simulations with a 50% duty ratio is presented. The ideal collector voltage and current waveforms demonstrate a possibility of 100% efficiency without overlapping between each other. The circuit design with transmission lines at operating frequency of 2.14 GHz is discussed and analyzed. In order to reduce the voltage peak factor, the load network parameters can be rearranged to correspond to Class-E/F_3 mode by providing a short-circuit condition at the third harmonic when the second-harmonic tank is connected in series to the shunt filter. Broadband capability of a Class-E mode with shunt filter using reactance compensation technique has been demonstrated by two examples, one with lumped elements and the other with transmission-line elements. The test board a transmission-line broadband Class-E GaN HEMT power amplifier with shunt filter was measured and high-performance results with the output power of around 41 dBm, average drain efficiency of 68%, and power gain of about 9 dB were achieved across the frequency band from 1.4 to 2.7 GHz.

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