В монографії викладено теоретичні та конструктивні аспекти розробки високоефективних автогенераторів класу Е високочастотного та надвисокочастотного діапазонів. Розглянуто приклади побудови автогенераторів на різні рівні потужності та різного конструктивного виконання. Наведено відомості про високоефективні автогенератори класів F, EF2 та з маніпуляцією на гармоніках.

Книга допоможе познайомитися з сучасним рівнем та методами розробки автогенераторів та підсилювачів потужності з високим ККД.

Автори: Крижановський В.Г., Макаров Д.Г.,
Чернов Д.В., Крижановський В.В.
Під редакцією В.Г. Крижановского

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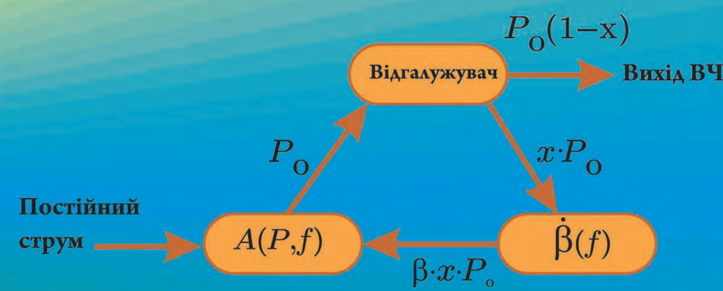

Transmission-Line Load Network Design Technique for Class-E Power Amplifiers

By Denis Makarov, Yulia Rassokhina, Vladimir Krizhanovski, and Andrei Grebennikov

This paper presents the lumped and transmission-line alternatives to Class E.

The switchmode Class-E power amplifiers with shunt capacitance have found widespread application due to their design simplicity and high efficiency of operation [1]. Their load-network configuration shown in Fig. 1(a) consists of a shunt capacitance, a series inductance, and a series fundamentally tuned 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 minimizes the power dissipation and maximizes 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 capacitance and series filter that can provide ideally 100% collector efficiency is not a unique. This paper presents the lumped and transmission-line alternatives to Class E with shunt capacitance and series filter using lumped elements and transmission lines where a shunt filter is used instead of a series filter, with the load-network design parameters, voltage and current waveforms, and design example of a high-efficiency transmission-line 2.15-GHz Class-E GaN HEMT power amplifier.

Class E with shunt filter

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. Figure 1(b) shows the basic circuit configuration of a Class-E power amplifier with shunt capacitance 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 [2]. 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 classical 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 in-stantaneous switching between its on-state and off-state operation conditions.

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

Power Amps

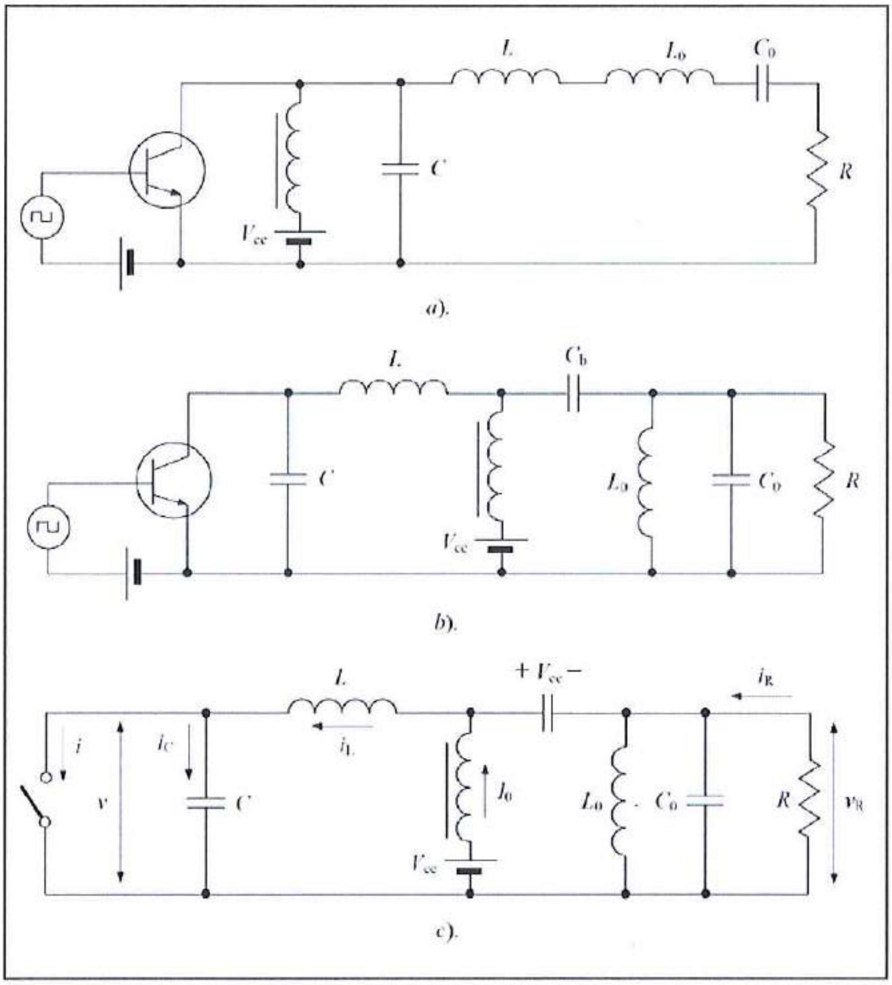

Fig. 1 • Basic circuits of Class-E power amplifier with shunt filter.

- there is no loss in the circuit except the load R
- for simplicity, a 50% duty ratio is used.

For idealized lossless operation mode, it is necessary to provide the following nominal (or idealized optimum) conditions for voltage across the switch (just prior to the start of switch on) at the moment $\omega t = 2\pi$, when transistor is saturated:

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

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

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 [1].

Expressions for the collector current ($0 \le \omega t < \pi$) and voltage ($\pi \le \omega t < 2\pi$) for ideal L_0C_0-circuit tuned to the fundamental frequency when the sinusoidal current $i_R = I_R \sin(\omega t + \varphi)$ is flowing into the load can be written as

$$i(\omega t) = i_L(\omega t) = \frac{V_{cc}}{\omega L}\omega t + \frac{V_R}{\omega L}\left[\cos(\omega t + \varphi) - \cos\varphi\right] \quad (3)$$

$$\omega^2 LC \frac{d^2 v(\omega t)}{d(\omega t)^2} + v(\omega t) - V_{cc} + V_R \sin(\omega t + \varphi) = 0 \quad (4)$$

where φ is the initial phase shift and $V_R = I_R R$ is the voltage amplitude across the load resistance R [2].

In an idealized Class-E operation mode, there is no nonzero voltage and current simultaneously that means a lack of the power losses and gives an idealized collector efficiency of 100%. This implies that the dc power and fundamental output power are equal,

$$I_0 V_{cc} = \frac{V_R^2}{2R}. \quad (5)$$

The optimum normalized series inductance L and shunt capacitance C can be calculated from Eqs. (3) and (4) using optimum switching conditions given by Eqs. (1) and (2) as

$$L = 1.4836\frac{R}{\omega} \quad (6)$$

$$C = \frac{0.261}{\omega R} \quad (7)$$

whereas the optimum load resistance R can be obtained for the given supply voltage V_{cc} and funda-mental output power P_{out} by

$$R = \frac{1}{2}\frac{V_R^2}{P_{out}} = \frac{1}{2}\left(\frac{V_R}{V_{cc}}\right)^2\frac{V_{cc}^2}{P_{out}} = 0.4281\frac{V_{cc}^2}{P_{out}}. \quad (8)$$

Figure 2 shows the normalized collector (a) voltage and (b) current waveforms for idealized optimum Class-E mode with shunt filter during the entire interval $0 \le \omega t \le 2\pi$. 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%.

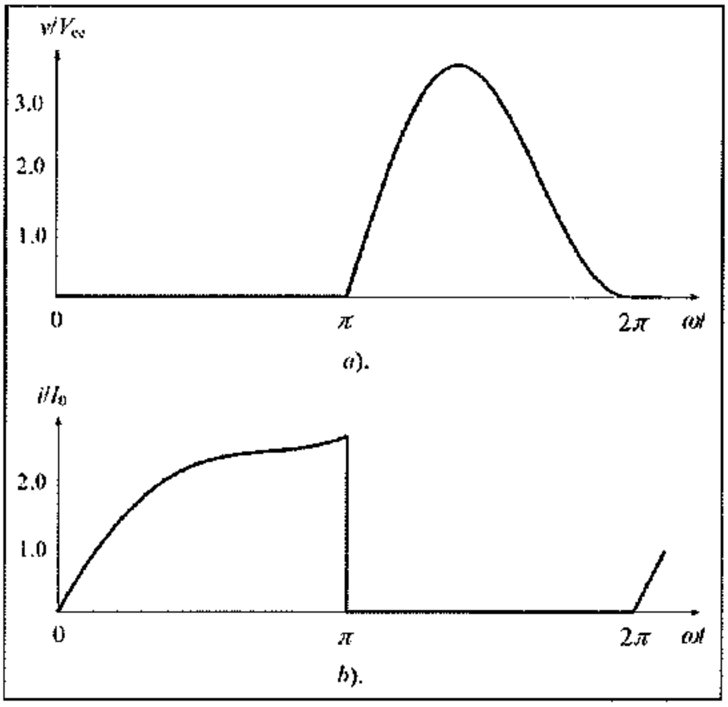

Fig. 2 • Normalized collector (a) voltage and (b) current waveforms for idealized optimum Class E with shunt filter.

The peak collector voltage V_{max} and current I_{max} are defined as

$$\frac{V_{max}}{V_{cc}} = 3.677 \tag{9}$$

$$\frac{I_{max}}{I_0} = 2.768 \tag{10}$$

that shows that the voltage peak factor is as high as in classical Class E with shunt capacitance and series filter [2].

Class E with quarterwave line

The ideal Class-F load network with quarterwave transmission line and series filter tuned to the fundamental frequency can provide a collector efficiency of 100% when the open-circuit conditions for odd-harmonic components and short-circuit conditions for even-harmonic components are realized [1, 3]. However, in practice, the idealized collector rectangular voltage and half-sinusoidal current waveforms corresponding to Class-F operation mode provided by a quarterwave transmission line in the load network can be realized at sufficiently low frequencies when effect of the device output capacitance is negligibly small. Generally, the effect of the device output capacitance contributes to finite switching time, resulting in time periods when the collector voltage and collector current exist at the same time. As a result, such a load network with quarterwave transmission line and shunt capacitor cannot provide the switching-mode operation with an instantaneous transition from the device pinch-off mode to saturation mode. Hence, during the finite time interval, the active device operates in active region as a current source with reverse-biased collector-base junction, and the collector current is provided by this multiharmonic current source. In this case, the required optimum conditions can be provided only for the fundamental frequency and several higher-order harmonic components.

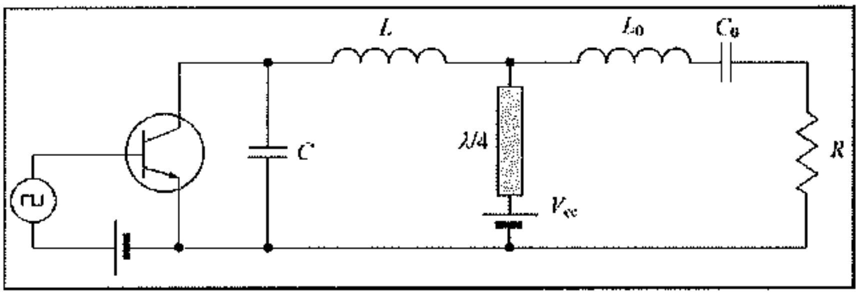

Fig. 3 • Basic circuit of Class-E power amplifier with quarterwave line.

However, the collector efficiency can be increased and effect of the collector capacitance can be compensated with inclusion of a series inductor between the shunt capacitor and the quarterwave transmission line, thus realizing the switching Class-E operation conditions. The obvious advantage of such a load network is a combination of high operating efficiency corresponding to Class-E mode and even-harmonic suppression due to quarterwave transmission line used in Class-F mode. The idealized Class-E load network with shunt capacitance and quarterwave line where the quarterwave transmission line is connected between the series inductance and fundamentally tuned series L_0C_0-filter is shown in Fig. 3.

By using similar analytical approach with zero-voltage and zero-voltage-derivative Class-E switching conditions given by Eqs. (1) and (2), the series inductance L, shunt capacitance C, and load resistance R with an assumption of a high-quality factor for series filter can be properly obtained by

$$L = 1.349\frac{R}{\omega} \tag{11}$$

$$C = \frac{0.2725}{\omega R} \tag{12}$$

$$R = 0.465\frac{V_{cc}^2}{P_{out}} \tag{13}$$

where V_{cc} is the dc-supply voltage and P_{out} is the output power at the fundamental frequency [1, 4].

The peak collector voltage V_{max} and collector current I_{max} are defined as

$$\frac{V_{max}}{V_{cc}} = 3.589 \tag{14}$$

$$\frac{I_{max}}{I_0} = 2.714 \tag{15}$$

which are close to that for Class E with shunt filter given by Eqs. (9) and (10), respectively [1, 4].

The phase angle of the entire load network at the fundamental frequency seen by the switch and required for an idealized nominal Class-E mode with quarterwave line is defined similar to that of a classical Class E with shunt capacitance and can be determined through the load-network parameters using Eqs. (11) and (12) as

$$\phi = \tan^{-1}\left(\frac{\omega L}{R}\right) - \tan^{-1}\left(\frac{\omega CR}{1 - \frac{\omega L}{R}\omega CR}\right) = 30.14°. \tag{16}$$

Figure 4 shows the normalized collector *(a)* voltage and *(b)* current waveforms for idealized optimum Class E with quarterwave line. 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 collector current consisting of the load sinusoidal current and transmission-line current flows through the switch. However, when the transistor is turned off, this current now flows through the shunt capacitance, charging process of which produces the collector voltage. Note that the collector voltage and current waveforms of Class E with quarterwave line are very close to those corresponding to Class E with shunt filter.

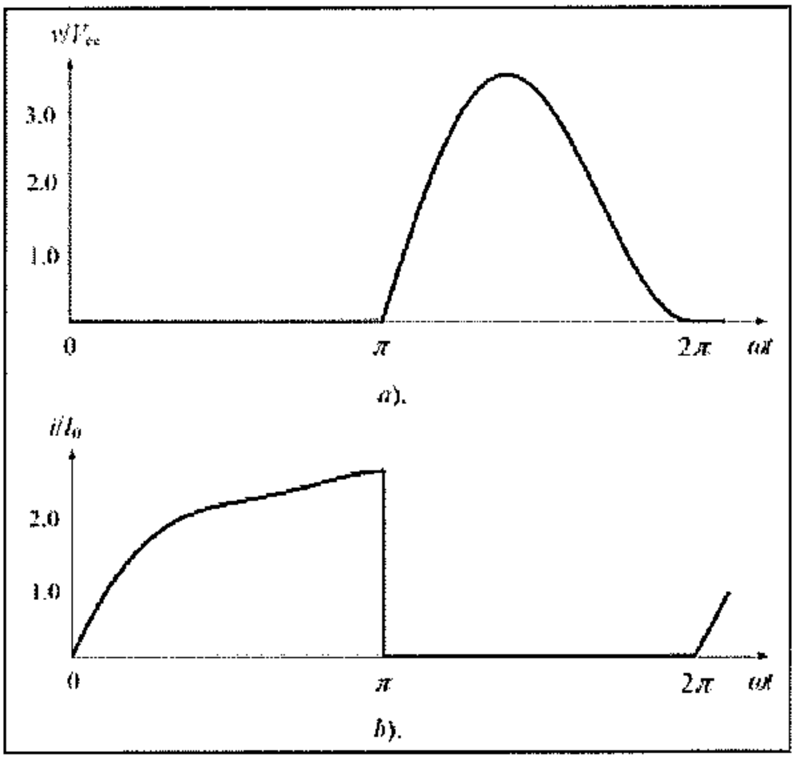

Fig. 4 • Normalized collector (a) voltage and (b) current waveforms for idealized optimum Class E with quarterwave line.

Class E with transmission-line filter

For a microwave power amplifier, all inductances in its load network are usually replaced by the transmission lines to reduce power losses and to make design more predictable and easier in im-plementation. To approximate the idealized Class-E mode with quarterwave line, it is necessary to design the transmission-line load network (instead of the series filter) with optimum impedance at the fundamental frequency and to provide the open-circuit conditions for all harmonics. As it fol-lows from the Fourier analysis, a good approximation to Class E mode with shunt capacitance or Class E with parallel circuit can be obtained with the collector voltage waveform consisting only of the fundamental and second harmonics [5, 6]. In this case, the matching circuit contains the series microstrip line and shunt open-circuit stub, both with electrical lengths of about 45° at the fundamental frequency. An additional increase of the collector efficiency can be provided by the load impedance control at the second and third harmonics simultaneously [7]. Therefore, for Class

E with quarterwave line, it is important to provide the open-circuit conditions at the input of the load net-work (using instead of the series filter) at least at the third harmonic, because the required short-circuit condition at the second harmonic is provided by the RF grounded quarterwave transmission line.

As an alternative, the quarterwave line can be conductively connected into the load network providing a parallel connection of the shunt open- and short-circuit stubs, which provides more compact design and also can add the broader capability of such a Class-E power amplifier with transmission-line filter [8]. Let us analyze these two transmission-line circuits, one representing by a grounded quarterwave line, as shown in Fig. 5*(a)*, and the other representing by a parallel connection of the shunt open- and short-circuit stubs, as shown in Fig. 5*(b)*. The input impedance Z_{net} for the grounded quarterwave line with the characteristic impedance Z_0 and electrical length θ is written as

$$Z_{net} = jZ_0 \tan\theta \tag{17}$$

whereas the input impedance Z_{net} for the parallel connection of a shunt open-circuit stub with the characteristic impedance Z_1 and electrical length $k\theta$ and a shunt short-circuit stub with the characteristic impedance Z_1 and electrical length $(1-k)\theta$, where k is the connecting factor, are obtained by

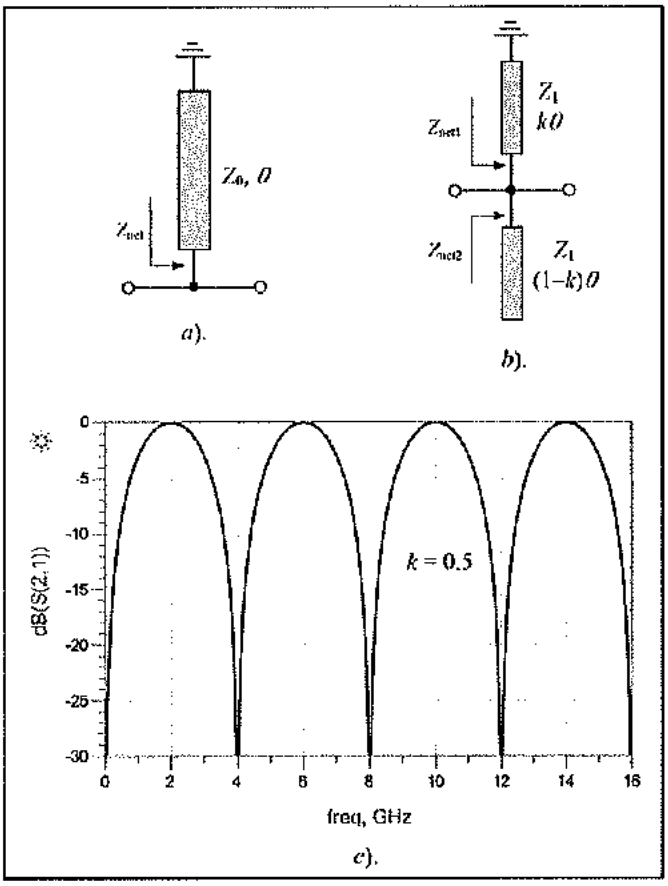

Fig. 5 • Conductive connection of quarterwave line and frequency response.

$$Z_{net} = \frac{Z_{net1} Z_{net2}}{Z_{net1} + Z_{net2}} = jZ_1 \frac{\tan k\theta}{1 - \tan k\theta \tan(1-k)\theta} \tag{18}$$

where

$$Z_{net1} = jZ_1 \tan k\theta \tag{19}$$

$$Z_{net2} = \frac{Z_1}{j\tan(1-k)\theta} . \tag{20}$$

Comparing Eqs. (17) and (18) for identical frequency behavior of the input impedance Z_{net} for both transmission-line circuits results in $k = 0.5$ and $Z_1 = 2Z_0$. This means that, for example at operating frequency of 2 GHz, infinite impedance at the fundamental and odd harmonics and zero impedance at even harmonics will be provided both for the circuit with shunt quarterwave line (θ = 90°) and for the circuit with shunt short-circuit and open-circuit stubs having an electrical length θ = 45° and a characteristic impedance of $2Z_0$ each, as shown in Fig. 5*(c)*.

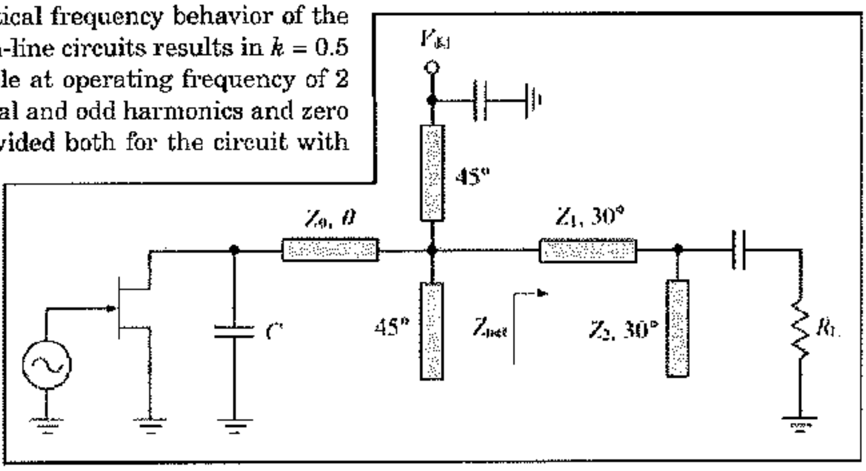

Fig. 6 • Schematic of Class-E power amplifiers with transmission-line load network.

Figure 6 shows the circuit schematic of a high-efficiency Class-E power amplifier with transmission-line filter and output matching circuit, where the nominal load resistance R is matched to the standard load impedance R_L = $\mathrm{Re}Z_{net}(\omega_0) = 50\ \Omega$ at the fundamental frequency using an output transmission-line L-type impedance transformer [1]. Here, the equivalent of a quarterwave line is composed of 45° short-circuit and open-circuit stubs to create short-circuit conditions at even harmonics. To create an open-circuit condition at the third harmonic at the input of the output matching circuit, the corresponding series and shunt open-circuit stubs are used, each having an electrical length of 30°.

To replace the series lumped inductance L by its transmission-line equivalent, a short-length series transmission line with the characteristic impedance Z_0 and electrical length θ of less than 45° can be used. In this case, the required optimum value of θ for Class-E mode with transmission-line filter using Eq. (11) can be approximated for given Z_0 by

$$\theta = \tan^{-1}\left(1.349\frac{R}{Z_0}\right). \tag{21}$$

The output matching circuit is necessary to match the required nominal Class-E resistance R calculated according to Eq. (13) with the standard load resistance of 50 Ω and to provide an open-circuit condition for Z_{net} at the third harmonic. The load-network impedance Z_{net} at the fundamental frequency can be written as

$$Z_{net} = Z_1\frac{R_L\left(Z_2 - Z_1\tan^2 30^\circ\right) + jZ_1Z_2\tan 30^\circ}{Z_1Z_2 + jR_L\left(Z_1 + Z_2\right)\tan 30^\circ} \tag{22}$$

where Z_1 and Z_2 are the characteristic impedances of the series transmission line and shunt open-circuit stub, respectively.

Consequently, the complex-conjugate matching with the load at the fundamental can be provided by proper choice of the characteristic impedances Z_1 and Z_2. Separating Eq. (22) into real and imaginary parts and considering that $\mathrm{Re}Z_{net} = R$ and $\mathrm{Im}Znet = 0$, the system of two equations with two unknown parameters can be written as

$$R\left(4R_L - 3R\right) - Z_1^2 = 0 \tag{23}$$

$$3\left(Z_1^2 - R^2\right)Z_2 - \left(3R^2 + Z_1^2\right)Z_1 = 0 \tag{24}$$

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

$$\frac{Z_1}{R_L} = \frac{\sqrt{4r-3}}{r} \tag{25}$$

$$\frac{Z_1}{Z_2} = 3\left(\frac{r-1}{r}\right). \tag{26}$$

As a result, for the specified value of the parameter r with the required nominal Class-E load resistance R and standard load resistance $R_L = 50\ \Omega$, the characteristic impedance Z_1 is calculated from Eq. (25) and then the characteristic impedance Z_2 is calculated from Eq. (26). For example, if the required load resistance R is equal to 12.5 Ω resulting in $r = 4$, the characteristic impedance of the series transmission line Z_1 is equal to 45 Ω and the characteristic impedance of the shunt open-circuit stub Z_2 is equal to 20 Ω.

Design example

The Class-E load network with a shunt capacitance C, a series inductance L, a series filter tuned to the fundamental frequency f_0, and a load resistance R is shown in Fig. 7*(a)*. For the supply voltage of 28 V and output power of 10 W, the saturation voltage of a selected Cree CGH40010F GaN HEMT transistor can be estimated from the device output *I-V* curves as equal to 2.5 V, assuming the drain efficiency of the Class-E power amplifier to be designed of 80% [1]. This means that the power loss due to the finite device saturation resistance will degrade the drain efficiency by about 10%. In this case, the nominal Class-E inductance L and capacitance C can be obtained by Eqs. (11) and (12),

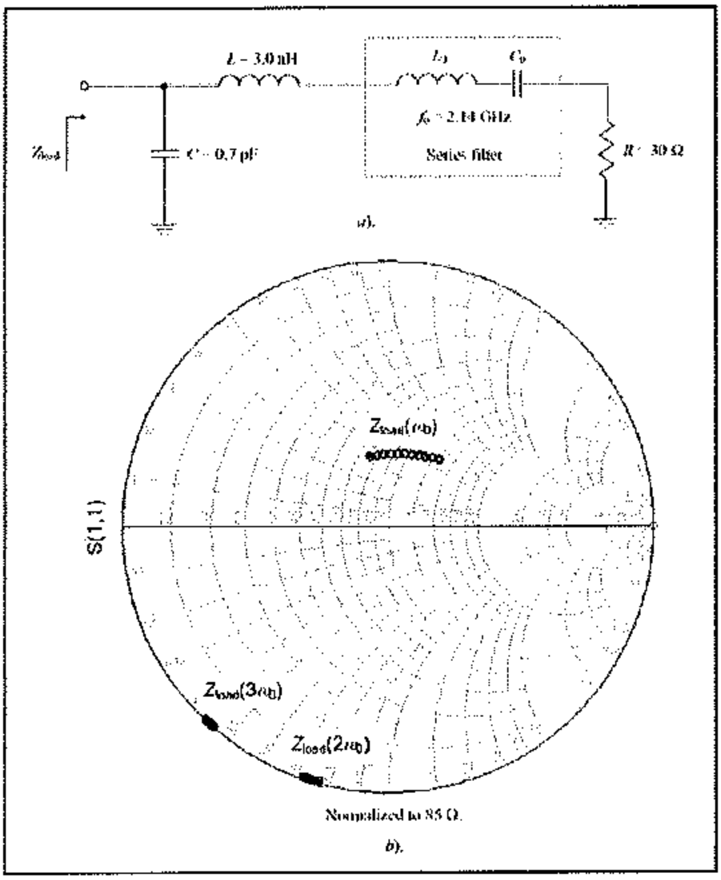

Fig. 7 • Class-E load network with quarterwave line at fundamental and harmonic impedances.

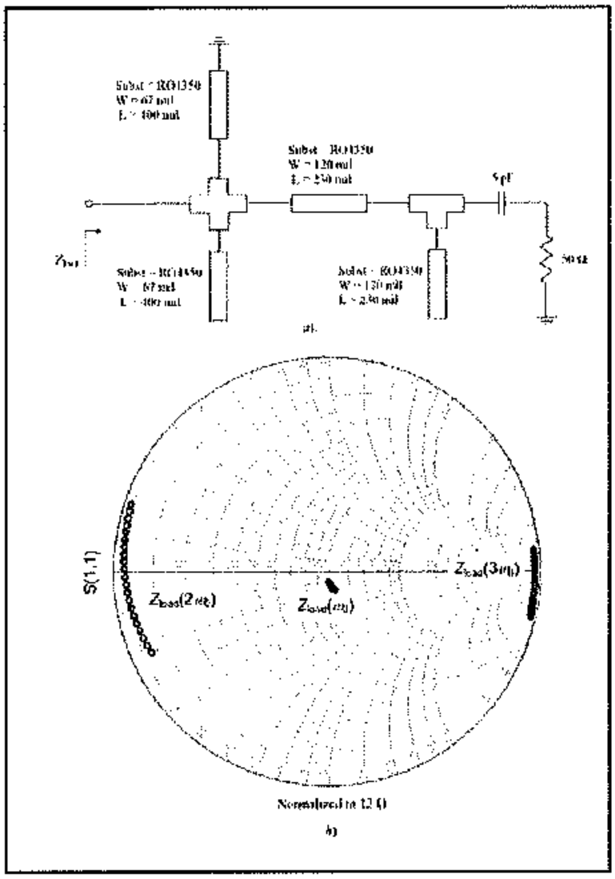

Fig. 8 • Transmission-line load network and harmonic impedances.

respectively, whereas the load resistance R is estimated by Eq. (13), where V_{cc} is replaced by $(V_{dd} - V_{sat})$, thus resulting in $L = 3.0$ nH, $C = 0.7$ pF, and $R = 30\ \Omega$ at the fundamental frequency of 2.15 GHz. By assuming the loaded quality factor $Q_L = 5$, the parameters of the series filter were calculated as $L_0 = 11.5$ nH and $C_0 = 0.5$ pF.

Figure 7*(b)* shows the simulation results in terms of separate hodographs at the Smith chart for the required inductive impedances at the fundamental frequency $Z_{load}(\omega_0)$ within the bandwidth of 2.1-2.2 GHz with a load-network phase angle of around 30° and for the capacitive reactances at the second harmonic $Z_{load}(2\omega_0)$ within the bandwidth of 4.2-4.4 GHz and at the third harmonic $Z_{load}(3\omega_0)$ within the bandwidth of 6.3-6.6 GHz.

Figure 8(a) shows the transmission-line filter consisting of the open- and short-circuit microstrip stubs and an L-type microstrip-line output matching circuit. Because the output drain-source capacitance of a 28-V 10-W Cree GaN HEMT power transistor CGH40010F is equal to 1.3 pF which is almost two times greater than that required for nominal Class-E mode according to Eq. (12), therefore the load resistance Z_{load} was intentionally lowered to 12 Ω, as shown in Fig. 8(b) by hodograph for $Z_{load}(\omega_0)$ at the Smith chart within the bandwidth of 2.1-2.2 GHz, to further approximate the required Class-E inductive impedance at the fundamental frequency seen by the device multiharmonic current source operating as a non-ideal switch. Here, hodograph for the second harmonic $Z_{load}(2\omega_0)$ demonstrates near-zero impedance conditions within the bandwidth of 4.2-4.4 GHz, whereas hodograph for the third harmonic $Z_{load}(3\omega_0)$ shows the high-impedance conditions within the bandwidth of 6.3-6.6 GHz.

The entire Class-E load network with transmission-line filter and L-type transmission-line output matching circuit with a 30° series microstrip-line stub and a 30° shunt open-circuit microstrip-line stub including the device output parameters is shown in Fig. 9*(a)*. Here, the device output shunt capacitance and series inductance with an additional series microstrip line provides the required Class-E inductive impedance at the fundamental frequency and approximate the capacitive reactances at the second and third harmonics, respectively. A comparison between hodographs corresponding to the idealized Class-E load network shown in Fig. 7*(b)* and the entire load network of a 10-W

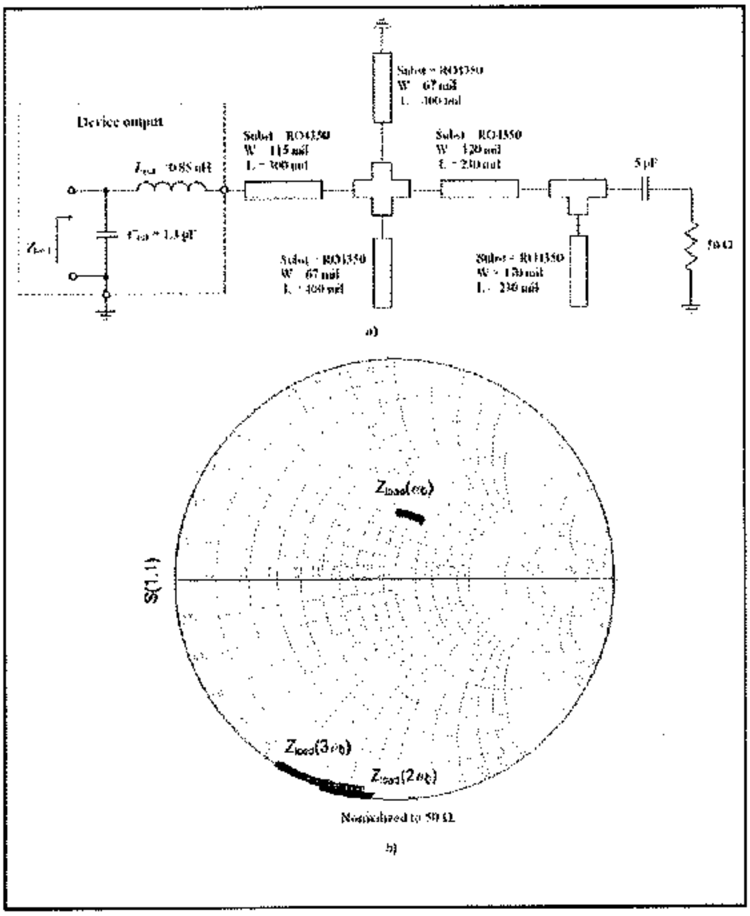

Fig. 9 • Class-E load network with transmission-line filter and harmonic impedances.

GaN HEMT power amplifier shown in Fig. 9*(b)* demonstrates their close proximity to each other. This means that if even the required output device parameters do not properly match the nominal Class-E load-network parameters, the high efficiency can nonetheless be achieved by optimizing the external load-network parameters to provide close approximation of the required impedances for the fundamental, second, and third harmonics.

Figure 10 shows the simulated circuit schematic, which approximates the Class-E power amplifier with transmission-line filter, based on a 28-V 10-W Cree GaN HEMT power transistor CGH40010F and a transmission-line load network shown in Fig. 9*(a)*. Here, the input matching circuit provides a complex-conjugate matching with the standard 50-Ω source with acceptable accuracy and stability of operation. The load network was slightly modified by optimizing the parameters of the series and shunt transmission lines because the device output capacitance C_{out} and series in-ductance L_{out} formed by drain bondwires and package lead do not properly match the required values of C and L for a nominal Class-E mode with transmission-line filter. To maximize the level of harmonic suppression, the lengths of both open- and short-circuit stubs were slightly increased.

For the GaN HEMT Class-E power amplifier with transmission-line filter using a 30-mil RO4350 substrate, the simulated drain efficiency of 85.2% and power gain of 15 dB resulting in a

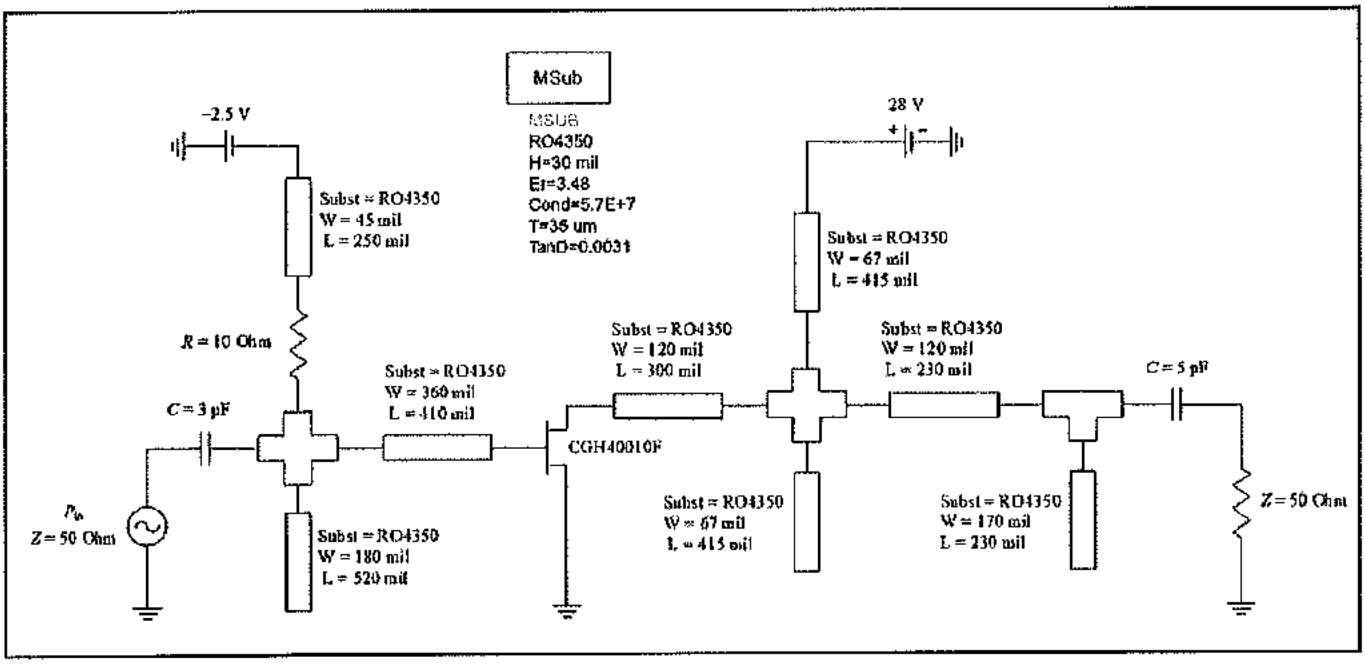

Fig. 10. •Circuit schematic of Class-E GaN HEMT power amplifier with transmission-line filter.

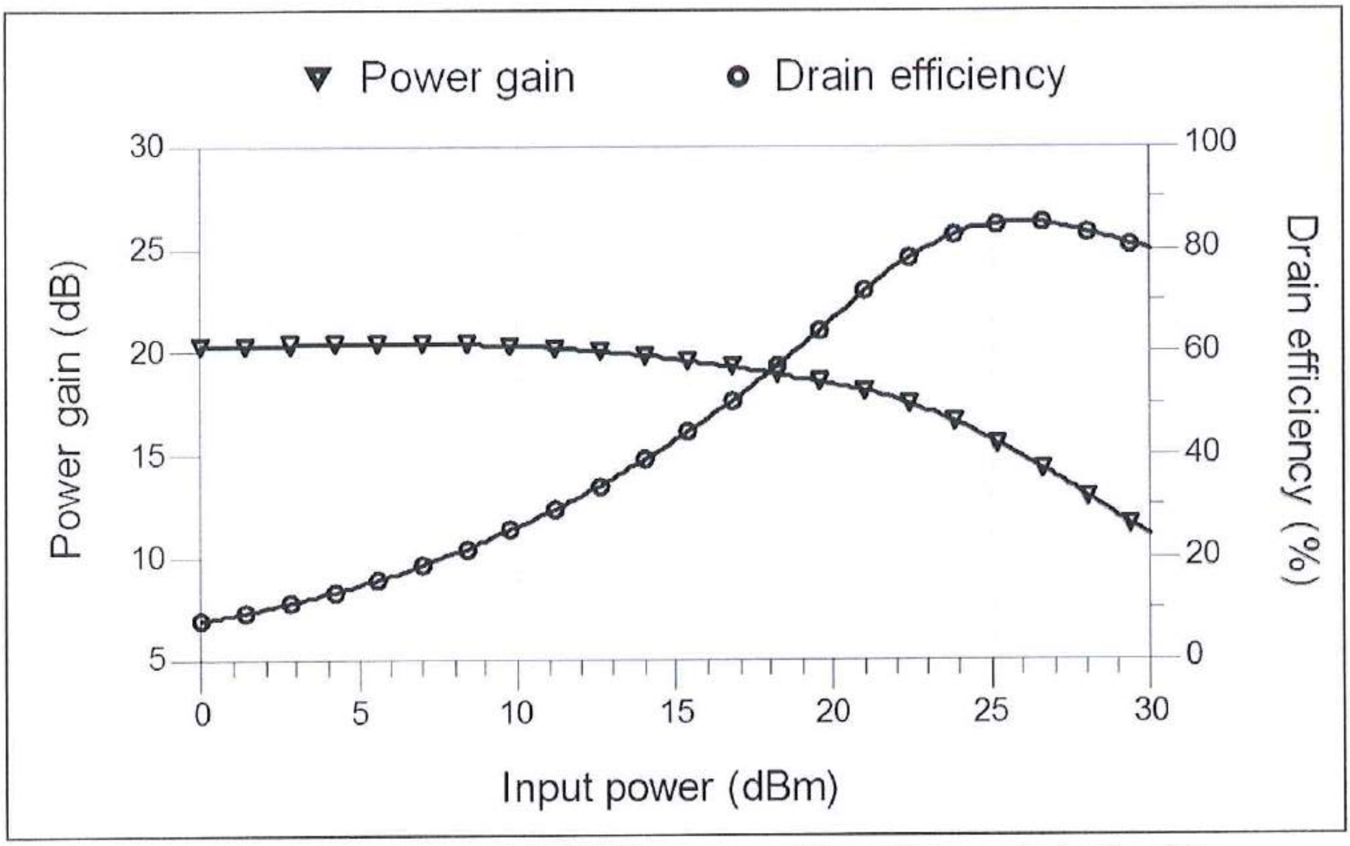

Fig. 11 • Simulated results of Class-E GaN HEMT power amplifier with transmission-line filter.

power-added efficiency *(PAE)* of 82.5% at an output power of 40.8 dBm with a quiescent current of 18 mA are achieved at an operating frequency of 2.15 GHz with a supply voltage of 28 V, as shown in Fig. 11. The second and third harmonics were suppressed by greater than 34 dB. A high-efficiency broadband operation of such a Class-E power amplifier can be achieved by increasing the number of matching sections in the input matching circuit and by lowering the loaded quality factor of the load network [2, 8].

About the Authors

Denis G. Makarov was born in Ukraine in 1982. He received his MSc degree in radio physics and electronics from the Department of Physics, Donetsk National University, Donetsk, Ukraine, in 2005. Currently he is working toward the Ph.D. degree in radio physics at the Donetsk National University, Vinnitsa, Ukraine. His current research interests are in high-frequency, high-efficiency power amplifiers, oscillators, radio frequency integrated circuits, and HF/VHF/UHF power amplifiers.

Yulia V. Rassokhina was born in Donetsk, Ukraine. She received her M.Sc. degree in physics from the Donetsk State University, Donetsk, Ukraine, in 1988 and Ph. D. degree in radio physics from the Institute of Radio Physics & Electronics of the National Academy of Sciences of Ukraine, Kharkov, Ukraine, in 1997. Currently she is working as a senior researcher with the Radio Physics Department, Donetsk National University, Vinnitsa, Ukraine. Her current research interests include the radio frequency integrated circuits, structural analysis based on waveguides and multilayered planar transmission lines, and modeling of the active and passive microwave devices and components. She is an author and coauthor of more than 50 technical papers published in different peer-reviewed journals and international conference proceedings.

Vladimir G. Krizhanovski received his MSc degree in radio physics from the Donetsk State University, Donetsk, Ukraine, in 1974, and Ph.D. degree in physical electronics from the Kharkov State University, Kharkov, Ukraine, in 1987. In 2010, he received his Doctor of Science degree in microwave techniques from Kharkov State University,

Kharkov, Ukraine. Currently he is a Professor in the Radio Physics Department, Donetsk National University, Vinnitsa, Ukraine. His research interests include the UHF electrodynamics and electronics, high-efficiency transistor power amplifiers and oscillators. He is an author and coauthor of more than 150 technical papers including published in international journals and conference proceedings and authored three books. Vladimir Krizhanovski has been a Senior Member of the IEEE since 2002.

Andrei Grebennikov received his Dipl. Eng. degree in radio electronics from the Moscow Institute of Physics and Technology, Moscow, Russia, in 1980, and Ph.D. degree in radio engineering from the Moscow Technical University of Communications and Informatics, Moscow, Russia, in 1991. He obtained a long-term academic and industrial experience working with the Moscow Technical University of Communications and Informatics, Moscow, Russia, the Institute of Microelectronics, Singapore, M/A-COM, Cork, Ireland, Infineon Technologies, Munich, Germany, and Linz, Austria, Bell Labs, Alcatel-Lucent, Dublin, Ireland, and Microsemi, Aliso Viejo, California, as an Engineer, Researcher, Lecturer, and Educator. He has lectured as a Guest Professor with the University of Linz, Linz, Austria, and presented short courses and tutorials as an Invited Speaker at the IEEE Microwave Theory and Techniques Society (MTT-S) International Microwave Symposia (IMS), European and Asia-Pacific Microwave Conferences, the Institute of Microelectronics, Singapore, Motorola Design Centre, Penang, Malaysia, the Tomsk State University of Control Systems and Radioelectronics, Tomsk, Russia, and the Aachen Technical University, Aachen, Germany. He has served as a Co-Chair of the IEEE Topical Conference on Power Amplifiers for Radio and Wireless Applications (PAWR) in 2016-2017. He is an author and coauthor of more than 100 papers, holds 25 European and US patents and patent applications, and authored eight books dedicated to RF and microwave circuit design. Currently he has been with Sumitomo Electric Europe, UK.

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