CLASS E WITH PARALLEL CIRCUIT – A NEW CHALLENGE FOR HIGH-EFFICIENCY RF AND MICROWAVE POWER AMPLIFIERS

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Abstract - **In this paper, a new circuit configuration of switched-mode tuned Class E power amplifiers with load network consisting of a parallel capacitance, a parallel inductance and a series resonant circuit tuned on the fundamental is defined using a detailed analytical description with a complete set of the design equations. The ideal collector voltage and current waveforms demonstrate a possibility of 100-percent efficiency. The circuit schematic of a parallel-circuit Class E power amplifier can be realized with lumped or transmission-line elements. Two examples of high power LDMOSFET and low-voltage HBT power amplifiers, utilizing a parallel-circuit Class E circuit configuration, are presented.**

I. INTRODUCTION

The switched-mode Class E tuned power amplifiers with a shunt capacitance have found widespread application due to their design simplicity and high efficiency operation [1]. In the Class E power amplifier, 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 output matching circuit [2].

However, such a circuit schematic when a shunt capacitance and a series inductance can provide ideally 100-percent DC-to-RF efficiency is not a unique. The same results can be achieved using the circuit configuration with parallel circuit consisting of a parallel capacitance and a parallel inductance with an additional series filtering circuit to provide high level of harmonic suppression. The generalized analysis of such a switched-mode power amplifier with calculation of voltage and current waveforms and some graphical results firstly was done by Kozyrev [3]. The circuit schematic, required waveforms, phase angles and values of the circuit elements differ from well-know types of the Class E power amplifiers. Therefore, the presented switched-mode tuned power amplifiers with parallel resonant circuit can be considered as a new subclass of switched-mode tuned Class E power amplifiers.

II. ANALYSIS

A. Basic principles and assumptions

As well as for the Class E power amplifier with shunt capacitance, the transistor in the Class E power amplifier with parallel circuit operates as an on-to-off switch. The basic circuit of a switched-mode Class E power amplifier with parallel circuit is shown in Figure 1(*a*). The load network consists of a parallel inductance L, a parallel capacitance C, a series L_0-C_0 resonant circuit tuned on the fundamental and a load R. In a common case, the parallel capacitance C can represent the intrinsic device output capacitance and external circuit capacitance added by the load network. The active device is considered to be an ideal switch that is driven in such a way in order to provide the device switching between its on-state and of-state operation conditions. In order to simplify an analysis of such a switched-mode power amplifier, it is advisable to introduce the same idealized assumptions as for well-known Class E operation mode [2].

For the theoretical analysis, the ideal switch replaces the active device as it is shown in Figure 1(*b*). Let the moments of switch-on is $t = 0$ and switch-off is $t = \pi/\omega$ with period of repeatability of input driving signal $T = 2\pi$ are determined by the input circuit of the power amplifier. Assume that the losses in the reactive circuit elements are negligible and the quality factor of the loaded L_0-C_0 circuit is sufficiently high. For lossless operation mode, it is necessary to provide the following optimum conditions for voltage across the switch just prior to the start of switch on at the moment $t = 2\pi/\omega$:

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

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

where v_s is the voltage across the switch.

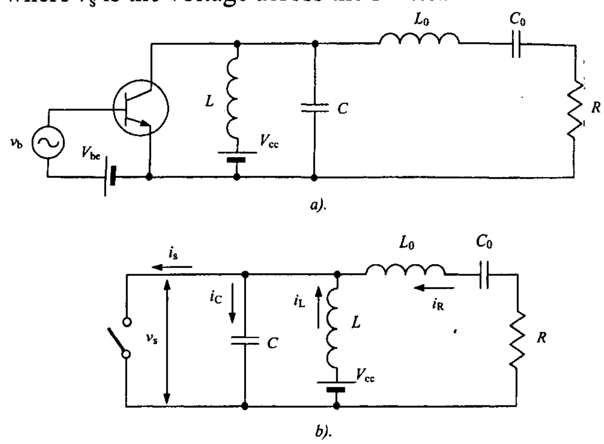

Fig. 1. Equivalent circuits of Class E tuned power amplifiers with parallel circuit

Let the output voltage and current are written as sinusoidal by

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

$$v_R(\omega t) = I_R R \sin(\omega t + \varphi) \quad (4)$$

where φ is the initial phase shift.

B. Voltage and current steady-state waveforms

When switch is on for $0 \le \omega t < \pi$, the voltage $v_s(\omega t) = V_{cc} - v_L(\omega t) = 0$, the current through the capacitance $i_C(\omega t) = 0$ and

$$i_s(\omega t) = \frac{V_{cc}}{\omega L}\omega t + I_R\left[\sin(\omega t + \varphi) - \sin\varphi\right] \quad (5)$$

When switch is off for $\pi \le \omega t < 2\pi$, the current through switch $i_s(\omega t) = 0$ and the current $i_C(\omega t) = i_L(\omega t) + i_R(\omega t)$ flowing through the capacitance C can be rewritten as

$$\omega C\frac{dv_s(\omega t)}{d(\omega t)} = \frac{1}{\omega L}\int_{\pi}^{\omega t}\left[V_{cc} - v_s(\omega t)\right]d(\omega t) + i_L(\pi) + I_R\sin(\omega t + \varphi) \quad (6)$$

under the initial off-state conditions $v_s(\pi) = 0$ and $i_L(\pi) = \frac{V_{cc}\pi}{\omega L} - I_R\sin\varphi$.

Equation (6) can be represented in the form of the second order differential equation given by

$$\omega^2 LC\frac{d^2 v_s(\omega t)}{d(\omega t)^2} + v_s(\omega t) - V_{cc} - \omega L I_R\cos(\omega t + \varphi) = 0 \quad (7)$$

which general solution can be obtained by

$$v_s(\omega t) = C_1\cos(q\omega t) + C_2\sin(q\omega t) + V_{cc} - \frac{q^2}{1 - q^2}\omega L I_R\cos(\omega t + \varphi) \quad (8)$$

where $q = 1/\omega\sqrt{LC}$ and the coefficients C_1 and C_2 are determined from the initial off-state conditions at $\omega t = \pi$.

To solve equation with regard to three unknown parameters, it is necessary to use two optimum conditions given by equations (1)-(2) and to add an additional equation defining the supply voltage V_{cc} from Fourier series expansion as

$$V_{cc} = \frac{1}{2\pi}\int_0^{2\pi} v_s(\omega t)d(\omega t) \quad (9)$$

As a result, solving the system of three equations with three unknown parameters numerically gives the following values:

$$q = 1.412 \quad (10)$$

$$\varphi = 15.155° \quad (11)$$

$$\omega L I_R = 1.21\, V_{cc}. \quad (12)$$

In Figure 2, the normalized (*a*) load current and collector (*b*) voltage and (*c*) current waveforms for idealized optimum parallel-circuit Class E mode are shown. The normalized currents flowing both through load network (*a*) capacitance C and (*b*) inductance L for idealized optimum parallel-circuit Class E mode are given in Figure 3.

C. Load network parameters

The fundamental-frequency current $i_{s1}(\omega t)$ flowing through the switch consists of two quadrature components, active i_R and reactive i_X, which amplitudes can be found using Fourier formulas and equation (5) by

$$I_R = \frac{1}{\pi}\int_0^{2\pi} i_s(\omega t)\sin(\omega t + \varphi)d(\omega t) \quad (13)$$

$$I_X = \frac{1}{\pi}\int_0^{2\pi} i_s(\omega t)\cos(\omega t + \varphi)d(\omega t) \quad (14)$$

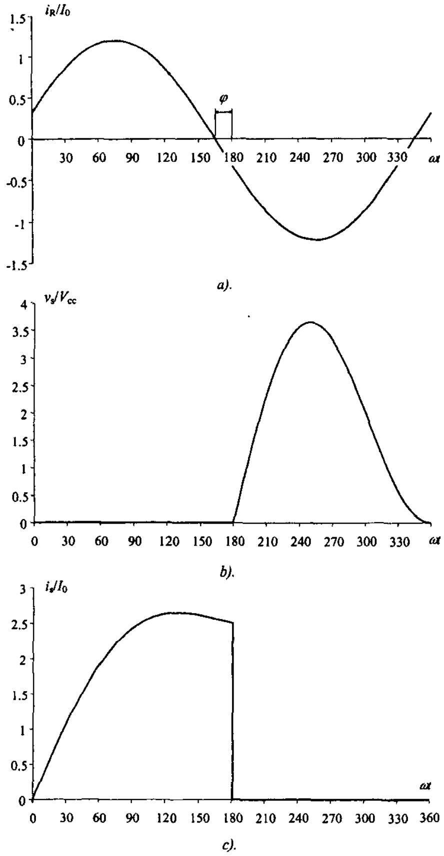

Fig. 2. Normalized (*a*) load current and switch (*b*) voltage and (*c*) current waveforms for optimum parallel-circuit Class E mode

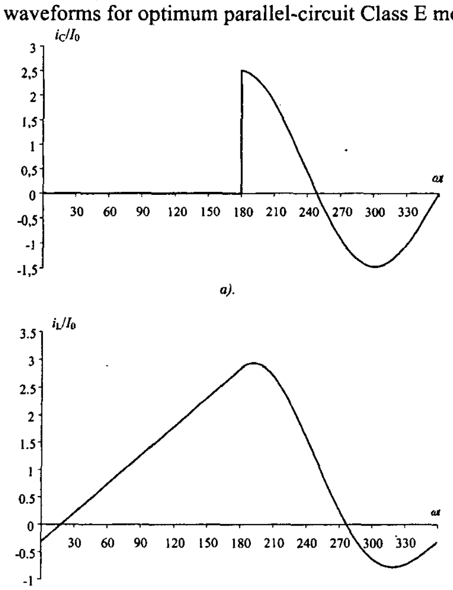

Fig. 3. Normalized currents through (*a*) capacitance and (*b*) inductance for optimum parallel-circuit Class E mode

Consequently, the phase angle ϕ between the fundamental-frequency voltage $v_{s1}(\omega t)$ and current $i_{s1}(\omega t)$ at switch terminal is equal to

$$\phi = \tan^{-1}\left(-\frac{I_X}{I_R}\right) = 34.244°. \qquad (15)$$

From the other hand, the phase angle ϕ can be represented as a function of load network elements as

$$\tan\phi = \frac{R}{\omega L} - \omega RC. \qquad (16)$$

As a result, the optimum series inductance L and parallel capacitance C can be obtained by

$$L = 0.732\frac{R}{\omega} \qquad (17)$$

$$C = \frac{0.685}{\omega R} \qquad (18)$$

The optimum load resistance R for the specified values of supply voltage V_{cc} and output power P_{out} can be calculated from

$$R = 1.365\frac{V_{cc}^2}{P_{out}} \qquad (19)$$

And the parameters of the series resonant circuit depending on the chosen in advance loaded quality factor Q_L, which value should be as high as possible, are calculated by

$$C_0 = \frac{1}{\omega R Q_L},\ L_0 = \frac{1}{\omega^2 C_0} \qquad (20)$$

D. Peak collector voltage and current

The peak collector current I_{smax} and voltage V_{smax} can be determined either by differentiating the appropriate waveforms given by equations (5) and (8), respectively, and setting the results equal to zero or directly from numerical calculation, which gives

$$V_{smax} = 3.647\,V_{cc} \qquad (21)$$

$$I_{smax} = 2.647\,I_0. \qquad (22)$$

E. Maximum operation frequency

When realizing the optimum parallel-circuit Class E operation mode, it is very important to know up to which maximum frequency such an efficient operation mode can be extended. In this case, it is required to establish a relationship between maximum frequency f_{max}, device output capacitance C_{out} and supply voltage V_{cc}. The device output capacitance C_{out} gives the main limitation of the maximum operation frequency. So, using equation (18) when $C = C_{out}$ gives the value of maximum operation frequency of

$$f_{\max} = 0.0798 \frac{P_{\text{out}}}{C_{\text{out}} V_{cc}^2} \quad (23)$$

which is in 1.4 times larger than that one for optimum Class E power amplifier with shunt capacitance[4].

III. RESULTS

In Figure 4, the circuit schematic of the single-stage 500 MHz parallel-circuit Class E high-voltage LDMOSFET power amplifier module with supply voltage V_{cc} = 28 V, output power P_{out} = 42 dBm, linear power gain G_P = 15 dB and power-added efficiency PAE = 67% is shown. The input and output matching is provided by means of the transmission-line T-transformers. The required load network phase angle is provided with the device output capacitance and parallel 50-Ohm transmission line of 25-degree electrical length.

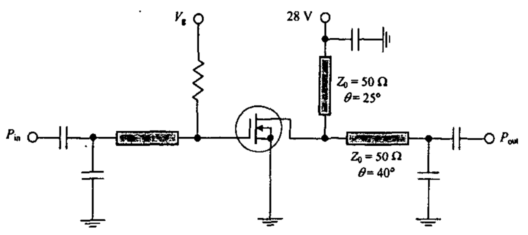

Fig. 4. Schematic of parallel-circuit Class E high-power LDMOSFET power amplifier module

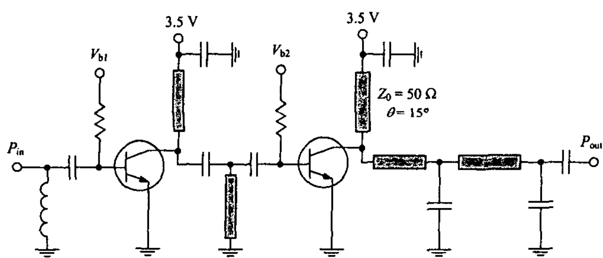

Fig. 5. Schematic of parallel-circuit Class E low-voltage InGaP/GaAs HBT power amplifier for handset application

In Figure 5, the circuit topology of two-stage 1.75 GHz parallel-circuit Class E low-voltage InGaP/GaAs HBT power amplifier for handset application with supply voltage V_{cc} = 3.5 V, simulated output power P_{out} = 34 dBm, linear power gain G_P = 25 dB, collector efficiency of 71% and PAE = 61% is given. The input, interstage and output matching circuits are realized in the form of simple high-pass L-transformers based on the lumped and transmission-line elements. With the increase of the supply voltage up to V_{cc} = 5 V, the collector efficiency increases up to 92% with PAE = 83% and P_{out} = 36 dBm.

IV. CONCLUSIONS

The new subclass of Class E power amplifiers consisting of a parallel capacitance, a parallel inductance and a series resonant circuit tuned on the fundamental is defined. The derived equations were used to determine the load network circuit elements required obtaining a 100-percent efficiency. The appropriate voltage and current waveforms are calculated and shown. Two examples of high power LDMOSFET and low-voltage HBT power amplifiers, utilizing a parallel-circuit Class E circuit configuration, are presented. The switched-mode parallel-circuit Class E power amplifiers offer a new challenge for RF and microwave power amplification providing high-efficiency operation conditions.

V. ACKNOWLEDGMENT

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