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CHARACTERISTICS OF CLASS E POWER AMPLIFIER WITH COMPLEX IMPEDANCE LOAD

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Abstract. Unlike the well-known publications focused on the analysis of the characteristics of a Class E power amplifier (PA), in which the authors limit themselves to considering a particular case of a real load, this paper presents the results of calculating the characteristics of a Class E PA with a complex impedance load. It is especially relevant when operating in a frequency band or amplifying broadband signals. The relations given in this paper can be used to solve two types of problems. In the first case, related to "soft-switching" mode Class E PAs characteristics can be determined, when the voltage on the transistor and its derivative at the moment of turn-on are equal to zero, which eliminates switching losses. In the second case, the problem of synthesizing a matching circuit that ensures the operation of the PA in the extended frequency band can be solved. The matching circuit synthesis can be carried out under the limitations on the acceptable change of output power and voltage drop on transistor just before switching.

Keywords: power amplifier, efficiency, Class E, complex impedance load, analytical model, simulation modeling, harmonic balance

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ХАРАКТЕРИСТИКИ УСИЛИТЕЛЯ МОЩНОСТИ КЛАССА Е ПРИ РАБОТЕ НА КОМПЛЕКСНУЮ НАГРУЗКУ

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Аннотация. В отличие от известных публикаций, посвященных анализу характеристик усилителя мощности (УМ) класса Е, в которых авторы ограничиваются рассмотрением частного случая вещественной нагрузки, в настоящей работе представлены результаты расчета характеристик УМ класса Е при работе на комплексную нагрузку, что особенно актуально при работе в полосе частот или усилении широкополосных сигналов. Приведенные в работе соотношения могут быть использованы для решения задач двух типов. В первом случае может быть осуществлен расчет характеристик УМ в «гладком» режиме, когда напряжение на транзисторе и его производная в момент коммутации электронного прибора равны нулю, что позволяет устранить коммутационные потери. Во втором случае может быть решена задача синтеза согласующей цепи, обеспечивающей работу УМ в расширенной полосе частот. Решение задачи синтеза согласующей цепи предусматривает учет ограничений на допустимое изменение выходной мощности и напряжения на транзисторе в момент коммутации.

Ключевые слова: усилитель мощности, КПД, класс Е, комплексная нагрузка, аналитическая модель, имитационное моделирование, гармонический баланс

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Introduction

Along with the increase of information transmission rate, the most significant tendencies of radio communication and telecommunications equipment characteristics improvement include efficiency increase. This allows not only better use of batteries, but also opens up the possibility to reduce mass-size characteristics of devices at the expense of reduction or even full exclusion of cooling elements.

This problem can be solved by modifying power amplifiers (PAs) into the switched-mode operation, where, as is known, the efficiency can reach values of 90% and higher [1–3]. However, taking into account that switching losses due to overcharging of the output capacitance of transistors increase with increasing frequency, the most attractive mode in terms of PAs efficiency improvement is the use of the class E mode. In it, due to the use of the forming LC-circuit, switching at zero voltage (ZVS) and zero current (ZVDS) can be realized [1–3, 5, 11].

Even though the Class E mode of operation has been known for quite a long time, due to its obvious advantages, it still arouses the interest of specialists in the field of radio- and telecommunications. This has been reflected not only in a number of monographs by well-known specialists in switched-mode PAs [1–3, 6–8], but also in numerous publications that have appeared in recent years [9, 10, 12, 14 etc.].

Unfortunately, the authors of these publications limited themselves to considering only one of the possible cases, when the amplifier operates on a resistive load at a fixed frequency. At the same time, it should be taken into account that antennas used in wireless communication systems usually have VSWR ≤ 2. It follows that the PA load in general is a complex impedance load and the use of the relations obtained just for resistive load can lead to an error.

The relations given in this paper can be used for solving two types of problems. In the first case, using known parameters of the transistor (maximum drain-to-source voltage and drain current), it is possible to determine the values of the real and imaginary parts of the complex impedance load, which are necessary for the transistor commutation in the "soft-switching" mode, when drain-to-source voltage and its derivative are simultaneously equal to zero, when the transistor is turned on [1–3, 5–9, 11]. Implementation of these conditions leads to an increase of PAs efficiency due to the elimination of switching power losses in the transistor, but it should be noted that a rigorous solution of this problem can be realized only at a fixed frequency.

No less actual for practice is the task of the second type, when it is necessary to ensure the operation of Class E PAs in the frequency band provided that the reduction of efficiency does not exceed the permissible values. This approach can be useful when using Class E PAs for amplification of signals with high peak-to-average power ratio (PAPR). In other words, the solution of the second type of problem does not guarantee the complete elimination of switching power losses at each of the operating frequencies or when changing the output power level. At the same time, this solution can lead to a significant reduction of switching losses in a certain frequency band within permissible change in the PA output power.

Taking into account the above, the paper goals are as follows:

1) determination of Class E PA characteristics when operating on a complex load;

2) estimation of the load impedance real and imaginary parts variation limits proceeding from allowable deviations from the nominal values of PA output power and transistor turn-on voltage, which is relevant for PA operation in the frequency band.

Analysis of class E power amplifier with a complex impedance load

The schematic of Class E PA is shown in Fig. 1, *a*. The analytical model was developed using several assumptions, including:

- output capacitance of the transistor does not depend upon the drain-to-source voltage and has a constant value;
- transistor turn-on and turn-off times are negligibly small compared to the duration of the output waveform period;
- on-state resistance of the transistor can be neglected;
- blocking capacitance C_b is large enough, so that the voltage across it can be considered constant and equal to the supply voltage V_{dd}. This gives a reason to replace it with a constant voltage source during the analysis.

Since in transmitters at the PA output frequency-selective circuits are used in order to attenuate higher harmonics, we will assume that the voltage across the load $Z(j\omega) = R + jX$ (Fig. 1, *a*) varies according to the harmonic law. This opens the possibility to apply the harmonic balance method when creating the analytical PA model, replacing the complex load by a voltage source $V_m \sin(\omega t + \varphi)$ [4, 14], with unknown amplitude V_m and initial phase φ to be determined as a result of the analysis (Fig. 1, *b*).

Considering the adopted assumptions, transistor can be represented in the equivalent circuit (Fig. 1, *b*) as an ideal switch in parallel with a capacitance. This capacitance is equal to the combined output capacitance of the transistor and stray capacitances.

Let us examine the steady-state operation of the circuit, assuming that during the time interval $0 < \omega t \leq \pi$, the transistor is in the off-state (switch S open). In the most interesting "soft-switching" mode of operation of Class E PA, two conditions must be satisfied [1–3, 11]:

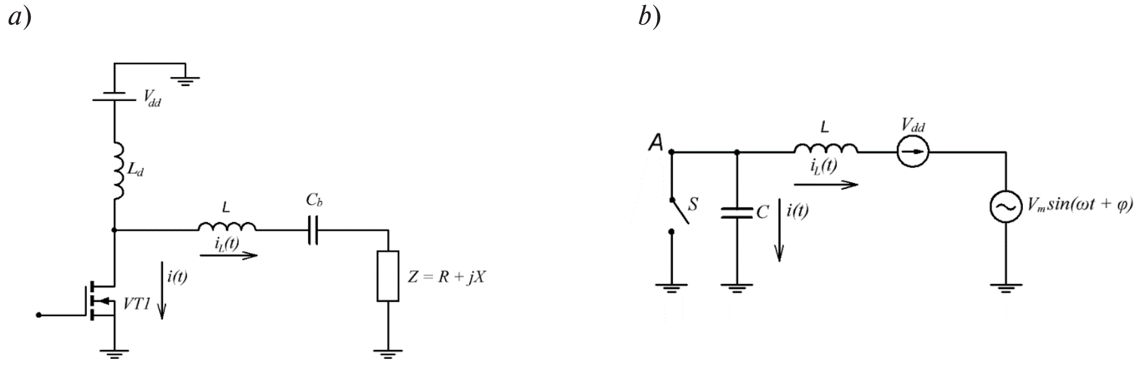

Fig. 1. *a*) functional diagram of a Class E PA; *b*) Class E PA equivalent substitution diagram

$$\begin{cases} \upsilon_C(t)\Big|_{t=\frac{\pi}{\omega}} = 0, & (1) \\ \dfrac{d\upsilon_C(t)}{dt}\Big|_{t=\frac{\pi}{\omega}} = 0. & (2) \end{cases}$$

According to Kirchhoff's laws for currents and voltages during the time interval $0 < \omega t \leq \pi$ (stage 1), the following system of equations can be created for the circuit in Fig. 1, *b*:

$$\begin{cases} L\dfrac{di_L(t)}{dt} + \upsilon_C(t) - V_m \sin(\omega t + \varphi) - V_{dd} = 0, & (3) \\ C\dfrac{d\upsilon_C(t)}{dt} - i_L(t) = 0, & (4) \end{cases}$$

where υ_C(t) is the voltage stress across the switch, $i_1(t)$ is the current in inductance of forming circuit (FC).

The voltage across the FC capacitance can be found solving the system of equations (3) and (4):

$$\upsilon_C(t) = k_1 \sin(\omega_0 t) + k_2 \cos(\omega_0 t) + 1 + \frac{\omega_0^2}{\omega_0^2 - \omega^2} V_m \sin(\omega t + \varphi), \tag{5}$$

where $\omega_0 = 1 \big/ \sqrt{LC}$, $k_{1,2}$ is the constants, which are determined using the initial conditions.

To facilitate further calculations, it is advisable to express the voltage across the capacitance (5) in normalized form:

$$\upsilon(\theta) = k_1 \sin(\nu\theta) + k_2 \cos(\nu\theta) + 1 + \frac{\nu^2}{\nu^2 - 1} q \sin(\theta + \varphi), \tag{6}$$

where

$$\upsilon(\theta) = \upsilon_C(t) \big/ V_{dd}, \; q = V_m \big/ V_{dd}, \; \nu = \omega_0 \big/ \omega, \; \theta = \omega t. \tag{7}$$

At the time interval $\pi < \omega t \le 2\pi$ (stage 2), with the switch S in the closed state, the current flowing through the inductance L is given by:

$$i_L(\theta) = \frac{1}{\omega L}\int_{\pi}^{\theta}\left[V_{dd} + V_m \sin(\theta+\varphi)\right]d\theta + I_{02}, \tag{8}$$

where I_{02} is the current through the inductance at the time $\omega t = \pi$.

After normalization, expression (8) will be modified as follows:

$$i(\theta) = \nu(\theta-\pi) - \nu q\left[\cos(\theta+\varphi)+\cos\varphi\right] + \frac{I_{02}}{V_{dd}/\rho}, \tag{9}$$

where $i(\theta) = \dfrac{i_L(\theta)}{V_{dd}/\rho}$, $\rho = \sqrt{L/C}$.

Relations (6) and (9), which describe the behavior of current and voltage across the FC capacitance, depend on six unknown parameters: v, q, φ, k_1, k_2, I_{02}.

To find the value of I_{02} using equation (9), let us apply the following condition:

$$I_{02} = i_L(\theta)\Big|_{\theta=\pi} = i_C(\theta)\Big|_{\theta=\pi} = \omega C\frac{d\upsilon(\theta)}{d\theta}\Big|_{\theta=\pi} = 0. \tag{10}$$

At the time $\theta = 0$, the transistor was in the on-state and the capacitance C was discharged. Using equation (6), we can derive the equation to determine k_2:

$$k_2 + 1 + \frac{\nu^2}{\nu^2-1}q\sin\varphi = 0. \tag{11}$$

The equation used to determine k_1 is derived from equation (6), while also considering condition (1):

$$k_1\sin(\pi\nu) + k_2\cos(\pi\nu) + 1 - \frac{\nu^2}{\nu^2-1}q\sin\varphi = 0. \tag{12}$$

By solving equations (11) and (12), we can express the constants k_1, k_2 in terms of the unknowns q and φ:

$$k_1 = -\mathrm{ctg}(\pi\nu)k_2 - \left(1-\frac{\nu^2}{\nu^2-1}q\sin\varphi\right)\Big/\sin(\pi\nu); \tag{13}$$

$$k_2 = -1 - \frac{\nu^2}{\nu^2-1}q\sin\varphi. \tag{14}$$

Taking into account that the DC resistance of the inductor L_d is zero, the average voltage at point A (Fig. 1, b) is equal to the supply voltage V_{dd}. Considering this, we can express it using equation (6):

$$\frac{\upsilon_{cp}}{V_{dd}} = \frac{1}{2\pi}\int_0^{2\pi}\upsilon(\theta)d\theta = 1. \tag{15}$$

After performing the necessary calculations, we get the equation that includes the unknown variables q and φ:

$$k_1\left[\cos(\pi\nu)-1\right]-k_2\sin(\pi\nu)-\pi\nu-\frac{2\nu^2}{\nu^2-1}q\sin\varphi=0. \tag{16}$$

The second equation for finding these unknowns is obtained by differentiating (6) and equating the resulting expression, in accordance with condition (2), to zero:

$$k_1\cos(\pi\nu)-k_2\sin(\pi\nu)-\frac{\nu^2}{\nu^2-1}q\cos\varphi=0. \tag{17}$$

By substituting relations (13) and (14) into equations (16) and (17), we finally get a system of two equations to determine the unknowns q and φ. The normalized frequency ν present in equations (16) and (17) can be treated as an independent variable.

Solving (16) and (17) bring us to:

$$\varphi=\frac{1}{2}\arcsin(x); \tag{18}$$

$$q=\pm\sqrt{\frac{1+\sqrt{1-x^2}}{2}}\frac{\nu^2}{\nu^2-1}(A-B), \tag{19}$$

where,

$$A=k_1\cos(\pi\nu); \tag{20}$$

$$B=k_2\sin(\pi\nu); \tag{21}$$

$$x=1-\frac{\pi\nu}{A-B}. \tag{22}$$

The time diagrams of the normalized voltage on FC capacitance $\upsilon(\theta)$ and the normalized current $i(\theta)$ referring to various values of the parameter ν are shown in Fig. 2.

As can be seen in Fig. 2, the behavior of the voltage across the capacitance C at $\theta = \pi$ satisfies conditions (1) and (2). In this case, the change of the relative frequency in the range from $\nu = 1.3$ to $\nu = 2.1$ has a comparatively weak effect on the value of the maximum voltage across the FC capacitance, the deviation of which does not exceed 5%. As for the maximum current through the transistor, here the influence of the parameter ν is much stronger and is accompanied by its decrease by approximately four times.

The load impedance to provide a "soft-switching" operation

The harmonic balance method enables to determine the characteristics of Class E PAs in a general form, eliminating the need to consider the structure and parameters of the load circuit. For this purpose, the complex load is replaced by an equivalent harmonic voltage source. When addressing the issues related to the operation of the PAs in a frequency band, it is advisable to shift from representing the load as a voltage source to using complex impedance Z. This will allow us to determine the behavior of the real and imaginary parts of Z, which are necessary to provide a "soft-switching" mode while the relative frequency

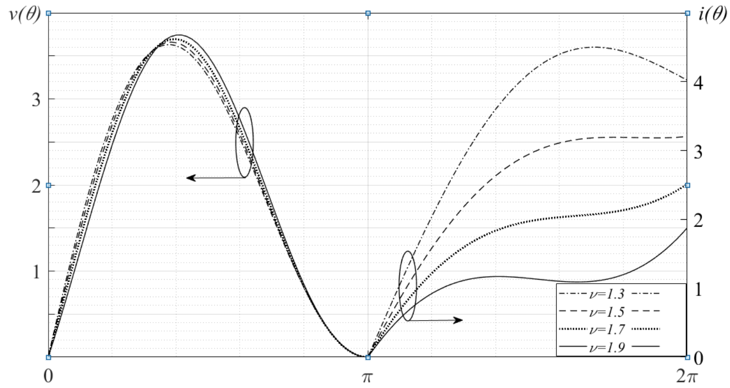

Fig. 2. Time-normalized diagrams of the voltage across the FC capacitance and current through the transistor at various values of the parameter ν

ν changes. It will open the possibility of solving the problem of synthesizing a matching circuit, which enables the operation of the PA in the frequency band.

To determine the real and imaginary parts of Z, we use equation (9) to find the amplitude and phase of the first harmonic in the load current:

$$\left|I_1\right| = \sqrt{a_1^2 + b_1^2}; \tag{23}$$

$$\varphi_I = \operatorname{arctg} \frac{b_1}{a_1}, \tag{24}$$

where a_1 and b_1 are the coefficients of the Fourier series:

$$a_1 = \frac{1}{\pi} \int_0^{2\pi} i_L(\theta) \sin\theta d\theta = \frac{V_{dd}\nu}{\pi\rho}\left(\pi + \frac{1}{2} q\pi \sin\varphi + 2q\cos\varphi\right); \tag{25}$$

$$b_1 = \frac{1}{\pi} \int_0^{2\pi} i_L(\theta) \cos\theta d\theta = \frac{V_{dd}\nu}{\pi\rho}\left(2 + \frac{1}{2} q\pi \cos\varphi\right). \tag{26}$$

The modulus and phase of the complex impedance load are equal to:

$$\left|Z\right| = \frac{\left|U_1\right|}{\left|I_1\right|} = \frac{V_m}{\sqrt{a_1^2 + b_1^2}}; \tag{27}$$

$$\varphi_z = \varphi - \varphi_I. \tag{28}$$

Using relations (27) and (28), it is straightforward to determine both the real and imaginary parts of the load impedance:

$$R = \left|Z\right| \cos\varphi_Z; \tag{29}$$

$$X = \left|Z\right| \sin\varphi_Z. \tag{30}$$

The current consumed from the power supply is equal to:

$$I_0 = \frac{1}{2\pi}\int_0^{2\pi} i(\theta)\,d\theta = \frac{V_{dd}}{2\pi\rho}\left[\frac{\pi^2\nu}{2} - \nu q\left(2\sin\varphi + \pi\cos\varphi\right)\right]. \tag{31}$$

Fig. 3 shows the behavior of the normalized characteristics of the PA as the parameter ν is varied. It includes the normalized power of the first harmonic P_{1n}, the normalized real R and imaginary X parts of the load impedance, the current stress on transistor I_{max}, the maximum voltage V_{max} across the FC capacitance, and the current consumed from the power supply I_0.

Based on the analysis of the curves shown in Fig. 3, it can be seen that the real part of the load impedance decreases rapidly to zero when the parameter ν exceeds 2. Concurrently, there is a noticeable reduction in both the load power and the current consumed from the power supply. As the parameter ν decreases, the current stress I_{max} increases, particularly when ν is less than 1.2. This trend can lead to some complications in selecting the appropriate transistor.

The features outlined above lead to the following conclusion. When the PA operates within the specified frequency band, it is reasonable to limit the variation of the relative frequency dispersion to the range of $1.3 \le \nu \le 2.0$. The values of the normalized parameters that correspond to this condition, obtained from analytical calculations and ensuring the realization of the "soft-switching" mode in the PA, are given in Table 1.

Table 1

Results of the analytical calculations

ν	1.3	1.4	1.5	1.6	1.7	1.8	1.9	2.0
k_1	3.97	3.17	2.60	2.15	1.77	1.42	1.10	0.79
k_2	2.94	1.47	0.60	0.03	–0.36	–0.64	–0.85	–1.00
q	1.61	1.24	1.02	0.92	0.93	0.99	1.08	1.18
φ, deg	–89.1	–76.6	–60.5	–42.8	–26.7	–14.4	–5.7	0
φ_I, deg	–70.8	–70.9	–71.1	–71.4	–72.0	–73.0	–75.0	–79.2
φ_z, deg	–18.3	–5.7	10.6	28.6	45.7	58.6	69.3	79.2

To evaluate the reliability of the analytical model, one can compare the data presented in Table 2. This table lists the main characteristics of the PA at various relative frequency values, obtained through both analytical calculations (A) and simulation (S). The discrepancy between these results can be assessed using the relative error (Δδ).

Table 2

Comparison of analytical and simulation results

ν	$V_s(\max)/V_{dd}$			q			φ, deg		
	A	S	Δδ, %	A	S	Δδ, %	A	S	Δδ
1.40	3.63	3.60	0.8	1.24	1.29	3.9	–76.6	–70.1	5.5
1.60	3.68	3.65	0.8	0.94	0.99	5.1	–42.8	–40.4	2.4
1.80	3.70	3.66	1.1	0.99	1.13	12.4	–14.4	–18.5	–4.1
2.00	3.78	3.70	2.2	1.17	1.30	10.0	0.10	–5.0	–5.1

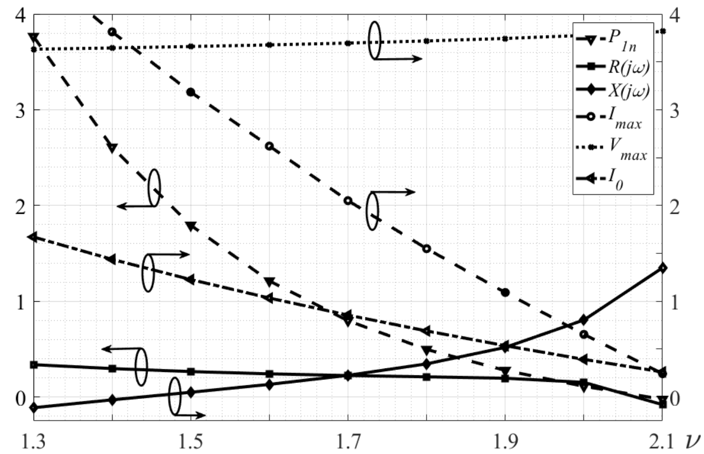

Fig. 3. Normalized characteristics of the PA as the parameter ν is varied

The table shows that the error in determining the voltage amplitude across the load does not exceed 10.0...12.4%. The error in determining transistor voltage stress is no more than 2.2%, and the initial phase φ is 5.5 degrees, respectively.

Equal values of output power and switching losses lines diagram

The dependencies of the real and imaginary parts of the load impedance, identified earlier, enable to formulate the problem of designing a matching circuit that ensures the operation of PA in the frequency band under specified permissible deviation of output power and efficiency reduction on account of transistor switching losses.

When amplifying variable-envelope signals, as well as when operating within the frequency band, the conditions for "soft-switching" (1) and (2) cannot be always met [15]. We assume that the voltage across the transistor at turn-on is equal to $\delta\upsilon$ ($0 \le |\delta\upsilon| < 1$):

$$\upsilon_{Cs}(\theta)\Big|_{\theta=\pi} = \delta\upsilon. \tag{32}$$

Using (6), we write (27) in the following form:

$$k_1^* \sin(\pi\nu) + k_2^* \cos(\pi\nu) + 1 - \frac{\nu^2}{\nu^2 - 1} q^* \sin\varphi^* = \delta\upsilon. \tag{33}$$

In a similar manner to equations (13) and (14), we can express:

$$k_1^* = -\cot(\pi\nu) k_2^* - \csc(\pi\nu)\left(1 - \frac{\nu^2}{\nu^2 - 1} q^* \sin\varphi^*\right) + \csc(\pi\nu)\delta\upsilon; \tag{34}$$

$$k_2^* = -1 - \frac{\nu^2}{\nu^2 - 1} q^* \sin\varphi^*. \tag{35}$$

To find two unknown parameters q^* and φ^* we use relations (2) and (16), respectively:

$$k_1^* \left[\cos(\pi\nu) - 1\right] - k_2^* \sin(\pi\nu) - \pi\nu - \frac{2\nu^2}{\nu^2 - 1} q^* \sin\varphi^* = 0; \tag{36}$$

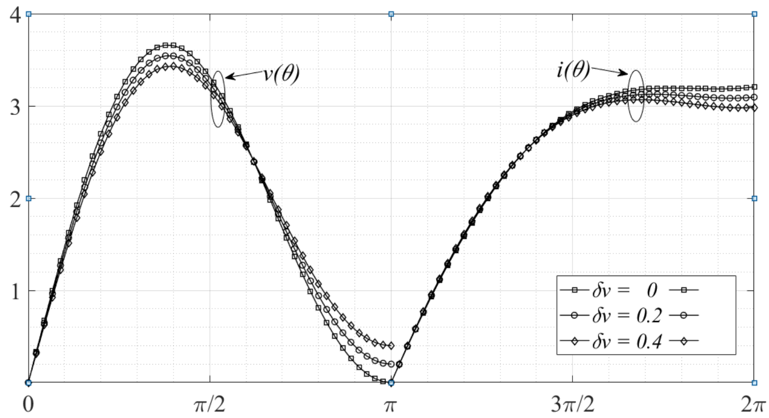

Fig. 4. Normalized voltage and current curves at different δυ

$$k_1^* \cos(\pi\nu) - k_2^* \sin(\pi\nu) - \frac{\nu^2}{\nu^2 - 1} q^* \cos\varphi^* = 0. \tag{37}$$

By solving equations (32), (33), (36), and (37), one can determine the parameters q^*, φ^*, k_1^* and k_2^*, which correspond to the PA mode with non-zero voltage across the transistor at turn-on. Time diagrams illustrating the behavior of the voltage across the FC capacitance and transistor current for different values of $\delta\upsilon$ are shown in Fig. 4.

The sensitivity of the parameter $\delta\upsilon$ to the deviation of the real and imaginary parts of the load impedance from their nominal values determined by relations (24) and (25) can be evaluated using Fig. 5, *a*, *b*.

The analysis of these graphs shows that the voltage across the transistor at turn-on is rather sensitive to $\Delta R(j\omega)$ and $\Delta X(j\omega)$ variations. Specifically, if we assume that turn-on voltage does not exceed a value of $\delta\upsilon = 0.2$, then for $\nu = 1.5$, the imaginary part of the load impedance can deviate from the nominal value by 23%, while the real part can only deviate by 4.3%. Conversely, when $\delta\upsilon = 0.2$ and $\nu = 1.7$, the relative deviation of the imaginary part must not exceed 3%, while the deviation of the real part can be no more than 5.3%.

Taking into account such an ambiguous character of the sensitivity of $\delta\upsilon$ to deviations in $\Delta R(j\omega)$ and $\Delta X(j\omega)$, it is prudent to utilize a series of curves representing equal power values (continuous lines) alongside a series of curves for equal values of the parameter $\delta\upsilon$ (dashed lines). Both series of curves are presented in the plane [$ReZ(j\omega)$, $ImZ(j\omega)$] (Fig. 6).

The use of these series of curves opens up the approach for solving the problem of synthesizing a matching circuit based on the condition of achieving the maximum frequency bandwidth:

$$\Delta\nu = (\nu_{\max} - \nu_{\min}) \to \max, \tag{38}$$

at the specified allowable limits for output power variation and transistor turn-on voltage:

$$\left[P(\vec{x}) - P_n\right] / P_n \le \Lambda; \tag{39}$$

$$\upsilon_C(\vec{x})\big|_{\omega t = \pi} \le (\upsilon_C)_{al.}, \tag{40}$$

where $\vec{x}$ is the vector of variable parameters and Λ, $(\upsilon_C)_{al.}$ are the allowable limits for output power variation and transistor turn-on voltage.

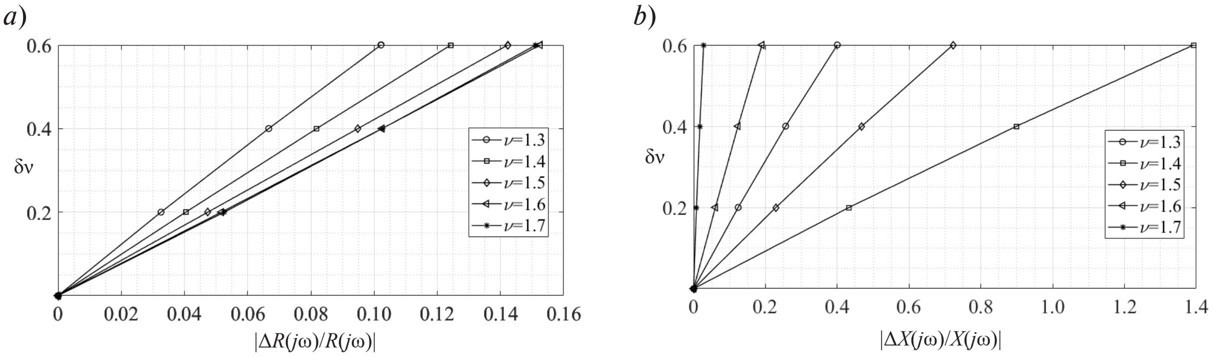

Fig. 5. The sensitivity of the parameter δυ to the deviation of the real (*a*) and imaginary (*b*) parts of the load resistance from their nominal values

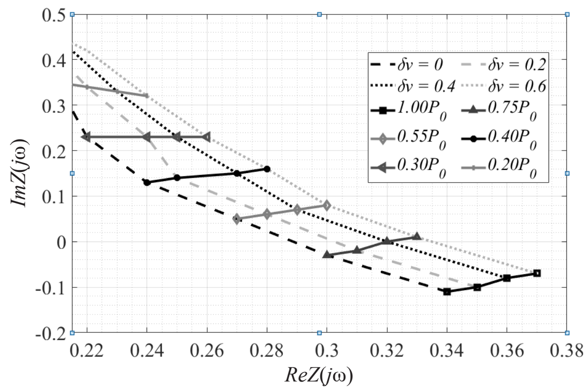

Fig. 6. Series of curves of equal output power values and equal δυ values in [*ReZ*(*j*ω), *ImZ*(*j*ω)] plane

It is important to note that based on the behavior of both curves series presented in Fig. 6, it is impossible to increase the operating frequency bandwidth of a Class E PA without negatively impacting at least one of its characteristics – either the relative output power or the switching losses in the transistor. To illustrate this, one can fix either the output power or the voltage across the transistor during turn-on. The relative frequency of the signal ν being varied, the trajectory of the corresponding point of load impedance *Z* will intersect the lines of equal levels associated with the other parameter.

Conclusion

Summarizing the results presented in this paper, we highlight the following:

1. Based on the harmonic balance method, the analysis of Class E PA when operating under complex load was carried out. The relationships have been obtained that allow one to determine the real and imaginary parts of the load impedance for different values of the relative frequency $\nu = \omega_0/\omega$, necessary for the implementation of the “soft-switching” mode, that means elimination of switching power losses in transistors.
2. The assessment of the adequacy of the proposed analytical model has been carried out which confirmed that the error in determining the voltage amplitude across the load does not exceed 10.0...12.4%. The error in determining the transistor voltage stress is no more than 2.2%, while the initial phase φ has an error of approximately 5.5 degrees.
3. An approach is proposed to solve the problem of matching a Class E PA with antenna when operating in a frequency band proceeding from the allowable output power deviation and efficiency reduction under the frequency change.

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