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Optimal Energy Characteristics and Working Parameters of RF Switch Mode Power Amplifier Based on Controllable Current Fed Resonant Inverter

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Abstract: *This article represents us the method for power losses minimizing in transistors of a radio frequency key generator (power amplifying) based on the control method application with overlapping pulses. We see fully formed requirements for the optimal excitation mode, providing maximum efficiency and maximum operating frequency, taking into account the finite times of switching on and off the generator GaN transistors. The article shows the obtained analytical equations for calculating power in a load, loss of efficiency, as well as equations for estimating the maximum operating frequency of the generator depending on the permissible level of losses. The article presents graphics of curves that determine the maximum operating frequency of an RF generator for permissible switching losses on various types of GaN transistors.*

Keywords: *Switch mode power amplifier, generator, power efficiency, pulse-width modulation (PWM), controlled current-fed resonant inverter.*

1. Introduction. The state of the issue

When highly efficient transistor-based power amplifiers (generators with external excitation) of harmonic oscillations are used, circuit design solutions based on resonant inverters are widespread. For high frequency applications (radio frequencies of tens, hundreds of megahertz), the use of nitride-gallium (GaN) transistors in these power amplifiers should be considered an actual task [1, 2]. These transistors have a wide band gap, i.e. they withstand high temperature stresses and have good dynamic characteristics, which allows them to realize switch modes at frequencies up to a few gigahertz.

An important requirement for RF power amplifiers is their ability to linear control of oscillation's amplitude or frequency. This requirement is satisfied by voltage inverters with a series resonant circuit, which are widely used up to the present in powerful radio engineering systems and converter equipment [3, 4]. However, in recent years current inverters with a parallel resonant circuit are becoming more common for high-frequency applications. The advantages of such inverters over voltage inverters are as follows:

1) The voltage jump on the transistor when the current is switched in the voltage inverter is approximately equal to the supply voltage, but in the current inverter at the maximum power mode, this jump is tens or hundreds of times smaller and approximately equal to the residual voltage on the transistor in the stationary mode. As a result, the losses at the fronts, which are fundamental at sufficiently high operating oscillation frequencies, are many times reduced, the frequency properties of the inverter improve by 1–2 orders of magnitude, and, accordingly, the efficiency is increased.

2) In the current inverter, almost constant choke current is commutes, i.e. the choke has a significant filtering effect for the high frequency current, which allows reducing the high-frequency filtering capacity of the power source by tens or hundreds of times.

3) The amplitude of the high-frequency voltage does not depend on the load, it is π/2 = 1.57 times higher than the supply voltage, while in the voltage inverter this amplitude depends on the quality factor of the loaded circuit.

The disadvantage of the classical current fed inverter circuit is the impossibility of operation with excitation pulses of transistors shorter than a half-period of the operating frequency of the oscillations. This is explained by the fact that the choke's current commutes in the pauses between the excitation pulses, is open, the current is broken, the voltage on it and on the transistors theoretically becomes infinite. This makes impossible to control the power (voltage amplitude at the load) using pulse-width modulation (PWM) [6, 7].

In this paper we study a new controlled current-fed resonant inverter circuit, which allows, in particular, power control using PWM and eliminates overvoltage on transistors and on a choke. For this circuit, the optimal excitation mode is considered, which provides maximum efficiency taking into account the finite transistor's times on and off. The maximum possible operating frequency of the power amplifier is estimated depending on the permissible level of losses.

2. Scheme and Principle of Operation

Due to the advantages described above, the current inverter is the most preferred of all switching power amplifiers of harmonic oscillations, including the class E power amplifier known from numerous publications [1, 2, 5], as long as the regulation (modulation) of power using PWM.

When the condition of linearity of the modulation characteristic is satisfied, the current inverter is suitable for creating powerful amplifiers used in radio communication and broadcasting with amplitude (AM), frequency (FM) (phase) and single sideband (SSB) modulations. Such a current inverter, which allows to apply PWM, is proposed at the level of the invention with the priority of 2016 [8]. The circuit of the controlled current-fed resonant inverter is shown in fig. 1, and the voltage and current diagrams on the bridge elements are shown in fig. 2.

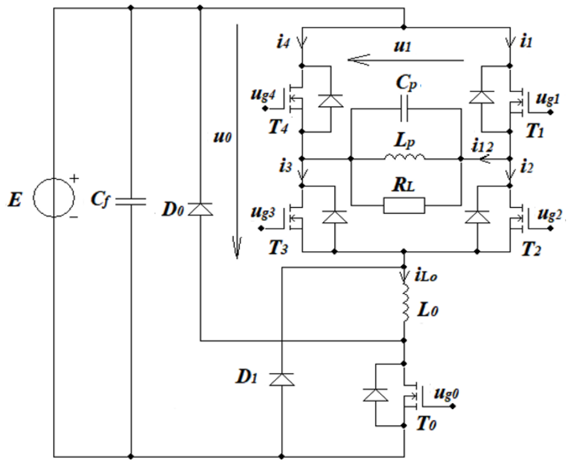


Fig. 1. Scheme of Controlled Current-Fed Resonant Inverter

In this scheme, the choke generator based on the transistor bridge M (T_1–T_4) is driven by high-frequency rectangular oscillations with a duration equal to half the period T of the operating frequency oscillations (fig. 2a, 2b). The current i_{12}, commuted by the bridge, has a shape close to rectangular pulses and excites the harmonic voltage $u_1(t) = U_m \sin(\omega t)$ on the resonant circuit L_P, C_P, R_L (fig. 2d). The voltage u_0 at the input of the bridge M as a result of switching has a double half-wave form (fig. 2c).

The modulation of the oscillations is achieved by using a switching unipolar T_0 amplifier with PWM (class D modulator), the switching frequency of which is chosen much lower than the bridge switching frequency. Due to the presence of a switching modulator and the possibility of regulating the power amplifier supply voltage, in the maximum power mode, the generator is fully utilized by the supply voltage ($U_m = E$).

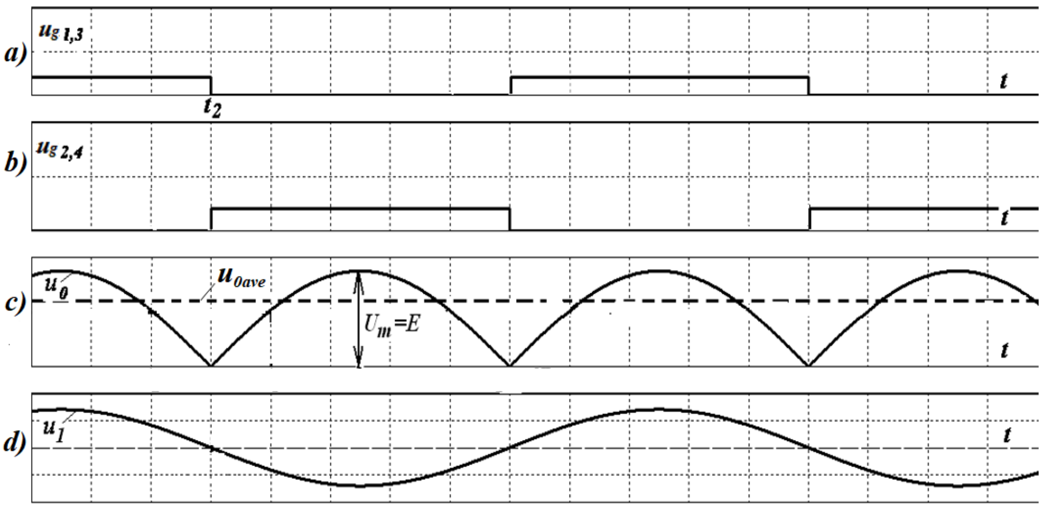


Fig. 2. Time Diagrams of Voltages and Currents in a Controlled Current-Fed Resonant Inverter

The modulation characteristic can be found from the following considerations. Neglecting the small residual voltages of the transistors operating in the switching mode, can find the average voltage at the load of the switching mode class D amplifier in the steady state:

$$U_{0M} = \frac{Et_{on}}{T_M}, \quad (1)$$

where t_{on} – is the duration of a rectangular pulse that triggers the transistor T_0, and T_M – is the repetition period of these pulses. The load of the switching mode class *D* amplifier is a transistor bridge *M* with an oscillatory circuit (fig. 1). Due to the small residual voltages of the bridge transistors *M*, the instantaneous voltage at its input u_0 repeats the voltage on the circuit, and taking into account the commutation, the average voltage U_{0M} (fig. 2d) varies according to the law:

$$U_{0M} = \frac{1}{0{,}5T}\int_0^{0{,}5T} U_M \sin\omega t dt = \frac{2U_M}{\pi}. \quad (2)$$

The average voltage on the choke in the steady state is zero, therefore, algebraically summing (1) and (2) we obtain the modulation characteristic:

$$\frac{U_M}{E} = 0{,}5\pi\frac{t_{on}}{T_M}, \quad (3)$$

when t_{on} varies from 0 to $2T_M/\pi$, the voltage amplitude on the circuit U_m varies linearly from zero to the maximum value of *E*.

A feature of the circuit is also the presence of a recovery diode D_1, which is necessary to eliminate overvoltages on the transistors and return excess energy (through the D_0 L_0 D_1 circuit) to the filter capacitance C_f of the source *E* at times when the choke L_0 current is broken during the switching process when the transistors *M* are locked.

Thus, the insertion of a modulator into the circuit, which introduces only two additional elements (a transistor T_0 and a diode D_0) and a recovery diode D_1 to the circuit, does not significantly complicate the circuit and pays off by improving the energy and frequency properties of the proposed switching choke generator [9].

3. Excitation Mode Optimization of Current-Fed Resonant Inverter at High Frequencies

We consider optimal operation mode of the investigated power amplifier with minimal switching losses. Let us analyze the time diagrams of currents and voltages during the excitation of transistors of the bridge M with overlapping control pulses at the gates. Changes in the output currents of real transistor switches with finite turn-on and turn-off times are approximated by linear relationships when optimizing the mode.

In the initial state ($t \le t_1$) in (fig. 3a-e) with the voltage $u_{g1,3}$ (fig. 3a), transistors T_1 and T_3 are turned on, $i_1 = i_{12}$ (fig. 3e). It is advisable to start switching the bridge M by turning on the transistors T_2, T_4 with the voltage $u_{g2,4}$ (fig. 3d). This inclusion should be made at the time $t = t_1$, when u_1 and $u_0 \approx u_1$ (fig. 3c, d) are positive and sufficiently small in value:

$$u_1(t_1) = U_m \sin[\omega(t_2 - t_1)] = U_m \sin(\omega t_{op}), \quad (4a)$$

$$t_{op} = t_2 - t_1, \quad (4b)$$

i.e., with some lead t_{op} relative to the zero point of voltage u_1. In the time interval $t_1 \le t \le t_2$, the voltage $u_1 \approx u_0$ is applied to the opened transistors T_2 and T_4 in the forward direction, as a result of which the current i_2 begins to flow through them (fig. 3e). Moreover, $i_{Lo}=I_0$=const, the current $i_1 = I_0 - i_2$ of the opened transistors T_1 and T_3 decreases with increasing i_2.

By the time $t = t_2$ $i_1 = i_2 = \frac{1}{2}I_0$, $i_{12} = i_1 - i_2 = 0$. At this moment, by the voltage $u_{g1,3}$ (fig. 3a), the diagonal T_1, T_3 is closed and the current i_1 continues to decrease at $t > t_2$, when the voltage u_1 is already applied to these transistors in the forward direction. The current $i_2 = I_0 - i_1$ increases and at the moment $t = t_3$ $i_1 = 0$, $i_2 = I_0$. On this, the process of switching current i_{12} by the bridge M ends.

In order to avoid excessive losses in transistors T_2 and T_4 when they are turned on, the voltage $u_1(t_1)$ applied to them (fig. 3c) and $u_1(t)$ at $t_1 \le t \le t_3$, as well as the corresponding lead time t_{op}, should be as small as possible, but large enough to support the switching process.

The calculated rate of increase of current i_2 in the interval $t_1 \le t \le t_3$ should be higher than the rate of decrease of current i_1. Otherwise, it may turn out that $i_1 + i_2 < I_0$. This will cause a sharp voltage surge u_0 under the influence of the self-induction EMF in the inductor L_0 to the value L_0, at which the recovery diode D_1 opens and switching losses increase.

Practically possible to take in:

$$t_{op} = t_2 - t_1 = \gamma t_{on}, \quad (5a)$$

$$\gamma = 2 \div 3, \quad (5b)$$

here t_{on} – the time the transistor is turned on at the rated voltage and current growth i_1 from zero to $0{,}5I_0$:

$$t_{on} = 0{,}5I_0/S_{on}, \quad (6a)$$

$$S_{on} = i_{nom}/t_{on.nom}, \quad (6b)$$

$$t_1 \le t \le t_2, \quad (6c)$$

where i_{nom} and $t_{on.nom}$ are the nominal passport values of the current and the on-time of the transistor.

The safety factor γ should be chosen greater than unity, since the transistors T_2, T_4 are turned on at low voltages $u_1(t)$, $t_1 \le t \le t_2$ (fig. 3d), and this voltage is divided in half between the transistors T_2 connected in series, T_4.

If the load circuit L_P, C_P, R_L (fig. 1) is tuned to the resonance at the conversion frequency ω, then the first harmonic of the current i_{12} (fig. 3e) supplied to the circuit must be in phase with the voltage u_1 (fig. 3c) on it. With the accepted linear approximation i_{12}

(fig. 3e), this means that the moments $t_2 + \frac{nT}{2}$ of the passage through zero of the voltage $u_1(t)$ and current $i_{12}(t)$ coincide.

For a fixed switching interval $t_1 \leq t \leq t_3$, the switching losses are minimal, since the voltage values $u_1(t_1) = |u_1(t_3)|$ (fig. 3c) are minimal in absolute value. The shift of this time interval along the t axis leads to an increase in one of these boundary stresses and an increase in losses, since they are proportional to the square of the voltage.

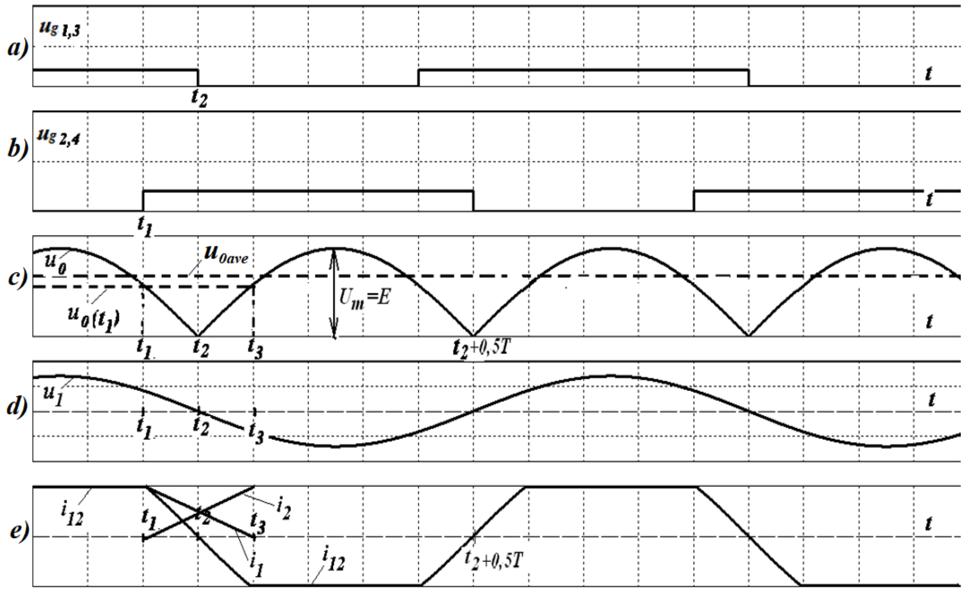


Fig. 3. Time Diagrams of Voltages and Currents in a Controlled Current-Fed Resonant Inverter with optimized excitation mode

From the above consideration it is seen that the optimal mode of operation is possible, in which a current break does not occur. This theoretically allows you to work without a recovery diode D_1 and increase the use of the transistor of the amplifier T_0 voltage. However, practically this should not be done, since during "abnormal" modes (malfunction of the oscillatory circuit, violation of the excitation mode of transistors, etc.), situations may occur with current breaks i_{0L} of the inductor L_0 and the occurrence of sharp voltage surges on it, leading to transistor failure.

It is important to note that in normal mode neither reverse transistor diodes nor the recovery diode D_1 participate in the power amplifier operation. This reduces switching losses and improves the frequency properties of the generator. The influence of the diode D_0 of a unipolar switching amplifier T_0 can be minimal if its conversion frequency is chosen sufficiently low (many times lower than the frequency of the generator).

4. Basic equations. Switching and conduction losses

To determine the output power, it is necessary to determine the first harmonic I_m of the current i_{12} (fig. 3e), applied to the load circuit L_P, C_P, R_L with the amplitude of the voltage U_m=E (fig. 3c). The first harmonic of the periodic oscillation of the trapezoidal form is expressed by the well-known formula, which is applicable to i_{12} (fig. 3e):

$$I_m = \frac{4I_0}{\pi}\frac{\sin\omega\gamma t_{on}}{\omega\gamma t_{on}} \cong \frac{4I_0}{\pi}\left[1 - \frac{(\omega\gamma t_{on})^2}{6}\right], \qquad (7)$$

where, t_{on} is the turn-on time of the transistor at the rated voltage and the rise of the current i_1 flowing through the transistor from zero to $0.5I_0$, γ- safety factor, that should be chosen greater than one, since the switching on of transistors T_2, T_4 occurs at low voltages $u_1(t)$ (Fig. 3d), and the voltage should be divided in half between the series-connected transistors T_2, T_4.

This approximate equality is obtained by expanding the sinusoidal function in a series and gives a high accuracy of calculation, since in practice the phase angle $\omega\gamma t_{on}$ is usually much smaller than unity (one radian). Without considering the correction term in square brackets, formula (4) gives the value of the first harmonic for rectangular oscillations (meanders), i.e. the fronts of trapezoidal oscillations i_{12} (fig. 3e), if they are relatively small, have a very weak effect on the amplitude value of the first harmonic I_m. Thus, the power in the load (tuned to the resonance circuit):

$$P_L = \frac{1}{2}U_m I_m = \frac{1}{2}E\frac{4I_0}{\pi}\frac{\sin\omega\gamma t_{on}}{\omega\gamma t_{on}} \cong \cong \frac{2}{\pi}EI_0\left[1 - \frac{(\omega\gamma t_{on})^2}{6}\right]. \qquad (8)$$

Energy loss when two transistors T_2, T_4 are connected in series, can be estimated according to $i_{12}(T) = I_0\frac{T}{\gamma t_{on}}, u_1(T) = U_m\sin(\omega T) \approx E\omega T, \gamma t_{on} \geq T \geq 0$:

$$W_{on} = \int_0^{\gamma t_{on}} i_{12}(T)u_1(T)dT = \frac{EI_0\omega}{\gamma t_{on}}\int_0^{\gamma t_{on}} t^2\,dT = = EI_0\frac{(\gamma t_{on})^2}{3}\omega. \qquad (9)$$

With the adopted linear symmetric approximation of the current $i_{12}(t)$, the energy loss W_{off} when the pair of transistors T_1, T_3 is turned off is approximately equal to W_{on} (9). Thus, the total power switching losses of four transistors at a frequency $f_o = \frac{\omega}{2\pi}$ can be estimated by the equation:

$$P_{sw} = \left(W_{on} + W_{off}\right)f_o = \frac{4\pi}{3}EI_0(\gamma t_{on})^2 f_o^2. \quad (10)$$

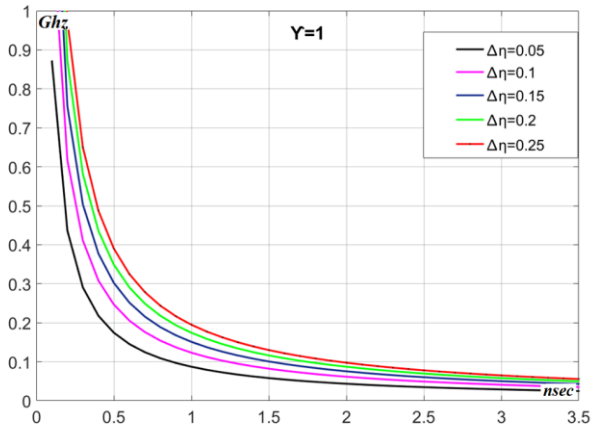
Switching loss of efficiency:

$$\Delta\eta_{sw} \cong \frac{P_{sw}}{P_L} \cong \frac{2\pi^2}{3}(\gamma t_{on})^2 f_o^2. \quad (11)$$

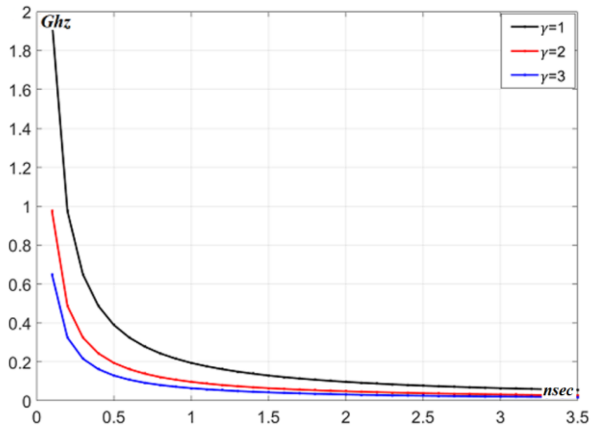
These losses are very small at relatively low frequencies f_o and increase sharply with increasing f_o. If we limit $\Delta\eta_{sw}$ = 0,01Δ (1 %), then can find the limiting operating frequency of the transistor:

$$f_{o.\max} = \sqrt{\frac{0{,}03}{2\pi^2}}\frac{1}{\gamma t_{on}} = \frac{0{,}04}{\gamma t_{on}}. \quad (12)$$

The family of curves is shown in the fig. 4a and fig. 4b to determine the maximum operating frequency of the RF power amplifier. The fig.4a shows the dependence of the maximum operating frequency on the time the transistor is turned on. In the calculation, the safety factor γ was taken equal to unity. The curves shown were obtained at different values of the efficiency loss $\Delta\eta_{sw}$. In fig. 4b the dependence curves are shown for losses of efficiency equal to 25 % and different values of the safety factor γ.



a)



b)

Fig. 4. Dependence of Maximum Frequency on the Turn-on Time for Different Values of a) $\Delta\eta_{sw}$; b) γ

If we assume that the transistors of the bridge M have a residual voltage $U_{res} = I_0 R_{sat}$ in the switch mode, not dependent on current, then in the amplifier there are conduction losses. Energy losses in this situation when two transistors T_2, T_4 are connected in series and switched on, can be estimated according to $i_{12}(t) = I_0\frac{t}{\gamma t_{on}}$, $u_1(t) = U_{res}$, $T/2 \geq t \geq 0$:

$$W_{cond} = 2\int_0^{T/2} i_{12}(t)u_1(t)dt = 2\frac{I_0^2 R_{sat}}{\gamma t_{on}}\int_0^{T/2} t\,dt = \frac{I_0^2 R_{sat}}{\gamma t_{on}}\frac{1}{4f_o^2}, \quad (13)$$

where R_{sat} – internal resistance of bridge transistors in saturation mode.

Energy loss for transistor T_0 can be founded by next equation:

$$W_0 = \int_0^{DT_M} i_{L0}(t)u_1(t)dt = I_0^2 R_{sato}\int_0^{DT_M} dt = \frac{DI_0^2 R_{sato}}{f_M}, \quad (14)$$

where, $i_{L0}(t) = I_0 = \text{const}$ – current flowing through choke and transistor T_0, D-duty cycle of PWM pulses, f_M-modulator frequency, R_{sato} – internal resistance of T_0 transistor in saturation mode.

Then total power loss in the inverter for all transistors will be equal:

$$P_{cond} = 2W_{cond}f_o + W_0 f_M = \frac{I_0^2 R_{sat}}{\gamma t_{on}}\frac{1}{2f_o} + DI_0^2 R_{sato}. \quad (15)$$

Since there is no energy recovery, the power supplied to the generator (bridge M):

$$P_0 = U_{0M}I_0 = \frac{2}{\pi}EI_0\cos\omega\gamma t_{on}. \quad (16)$$

Then, conduction losses of electronic efficiency is determined by:

$$\Delta\eta_{cond} = \frac{P_{cond}}{P_0} = \frac{\pi}{2E}\frac{I_0}{\cos\omega\gamma t_{on}} \times \left(\frac{R_{sat}}{\gamma t_{on}}\frac{1}{2f_o} + DR_{sato}\right). \quad (17)$$

Fig. 5 show the dependence curves for real GaN transistors, the data for the calculation of which are obtained from the manufacturer's technical documentation. The fig. 5a shows the dependence of the maximum operating frequency versus the switching losses. The fig. 5b shows the dependence of the conduction losses versus the operating frequency of the generator.

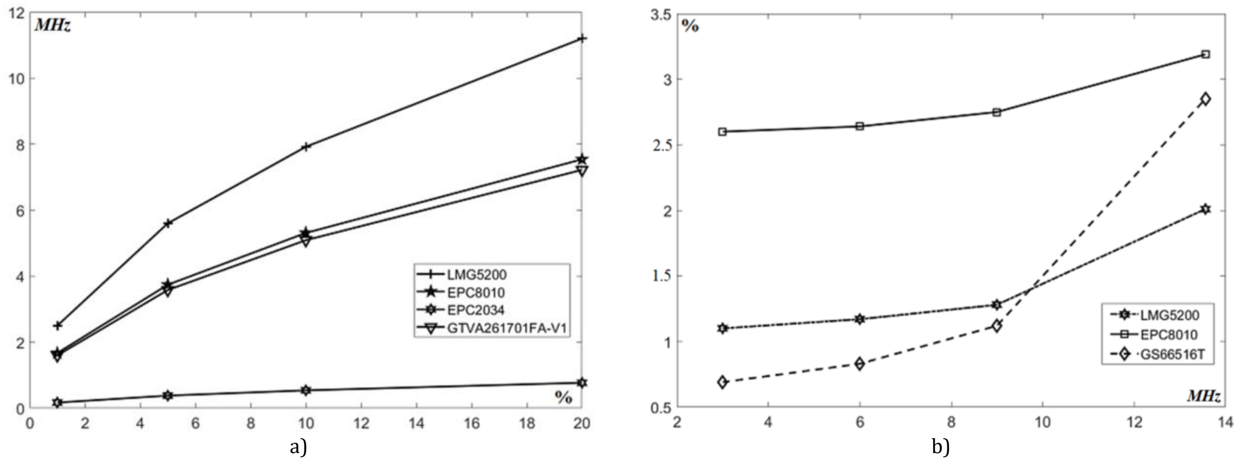


Fig. 5. The Dependence of a) the Maximum Operating Frequency on the Value of Switching Losses; b) the Conduction Losses on the Operating Frequency of RF Power Amplifier

5. Conclusion

Proposed a new, unparalleled controlled current-fed resonant inverter circuit, which allows, in particular, power control using PWM and eliminates overvoltage on transistors and a choke.

In order to minimize power losses in the RF power amplifier, a mode with overlapping control pulses is proposed. For the proposed circuit, the optimal excitation mode is considered, which provides maximum efficiency considering the finite times of transistor on and off.

The analytical calculation of switching and conduction losses is given. An analytical expression is obtained for an approximate calculation of the limiting frequency of a device for given design parameters. The family of curves is shown to determine the maximum operating frequency of the controlled current-fed resonant inverter.

References

1. Sokal N.O. *RF power amplifiers, classes A through S: how the circuits operate, how to design them, and when to use each-short course*. Boston, MA: IEEE IMS Workshop notes; 2000. 102 p.
2. Grebennikov A., Sokal N., Franco M.J. *Switchmode RF and Microwave Power Amplifiers*. Oxford: Elsevier; 2012. 704 p.
3. Artym A.D. *Class D Amplifiers and Key Generators in Radio Communications and Broadcasting*. Moscow: Radio i Sviaz Publ.; 1980. 209 p. (in Russ.)
4. Artym A.D. Key Operation Mode for High Frequency Generators. *Radioengineering*. 1969;24(6):58–64. (in Russ.)
5. Kryzhanovsky V.G. *High Efficiency Transistor Amplifiers*. Donetsk: Apeks, Publ.; 2004. 448 p. (in Russ.)
6. Zhukov S.A., Kozyrev V.B. Push-Pull Key Generator without Commutative Losses. *Telecommunication Semiconductors*. Moscow: Sviaz Publ.; 1975. vol.1. p.95–107. (in Russ.)
7. Popov I.A. Key Mode Single-Ended Transistor. *Telecommunication Semiconductors.* Moscow: Sviaz Publ.; 1970. vol.5. p.15–35. (in Russ.)
8. Filin V.A., Golovin A.N., Smirnov V.S., Ganbaev A.A. *Controlled Resonant Current Inverter*. Patent RF, no. 172407, 07.07.2017. (in Russ.)
9. Ganbaev A.A., Filin V.A. Radio frequency controlled switch mode generator based on current mode resonant inverter. *Computing, Telecommunication and Control (St. Petersburg Polytechnical University Journal. Computer Science. Telecommunication and Control Systems)*. 2019;12(1):21–28. (in Russ.) DOI:10.18721/JCSTCS.12102

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Оптимальные энергетические характеристики и рабочие параметры радиочастотного ключевого усилителя мощности на основе управляемого резонансного инвертора тока

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Аннотация: *Предложен метод минимизации потерь мощности в транзисторах радиочастотного ключевого генератора (усилителя мощности) на основе применения способа управления с перекрывающимися импульсами. Сформулированы требования к оптимальному режиму возбуждения, обеспечивающему максимальный КПД и предельную рабочую частоту с учетом конечных времен включения и выключения GaN-транзисторов генератора. Приведены полученные аналитические уравнения для расчета мощности в нагрузке, потери КПД, а также уравнения для оценки максимальной рабочей частоты генератора в зависимости от допустимого уровня потерь. Представлены графики кривых, определяющих максимальную рабочую частоту ВЧ генератора для допустимых коммутационных потерь на различных типов GaN-транзисторов.*

Ключевые слова: *ключевой усилитель мощности, генератор, энергетическая эффективность, широтно-импульсная модуляция (ШИМ), управляемый резонансный инвертор тока.*

Список используемых источников

1. Sokal N.O. RF power amplifiers, classes A through S: how the circuits operate, how to design them, and when to use each-short course. Boston, MA: IEEE IMS Workshop notes, 12 June 2000. 102 p.
2. Grebennikov A., Sokal N., Franco M.J. Switchmode RF and Microwave Power Amplifiers. Oxford: Elsevier, 2012. 704 p.
3. Артым А.Д. Усилители класса D и ключевые генераторы в радиосвязи и радиовещании. М.: Радио и Связь, 1980. 209 с.
4. Артым А.Д. Ключевой режим работы генераторов высокой частоты // Радиотехника. 1969. Т. 24. № 6. С. 58–64.
5. Крыжановский В.Г. Транзисторные усилители с высоким КПД. Донецк: Апекс, 2004. 448 с.
6. Жуков С.А., Козырев В.Б. Двухтактный ключевой генератор без коммутативных потерь // Полупроводниковые приборы в технике электросвязи. М.: Связь, 1975. Вып. 15. С. 95–107.
7. Попов, И.А. Ключевой режим однотактного транзисторного генератора // Полупроводниковые приборы в технике электросвязи. М.: Связь, 1970. Вып. 5. С. 15–35.
8. Филин В.А., Головин А.Н., Смирнов В.С., Ганбаев А.А. Управляемый резонансный инвертор тока. Патент на полезную модель РФ № 172407. Опубл. 07.07. 2017. Бюл. № 19.
9. Ганбаев А.А., Филин В.А. Управляемый ключевой генератор радиочастотных колебаний на основе резонансного инвертора тока // Научно-технические ведомости СПбГПУ. Информатика. Телекоммуникации. Управление. 2019. Т. 12. № 1. С. 21–28. DOI:10.18721/JCSTCS.12102

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