

A self-oscillating h.f. power generator with a Class E resonant amplifier

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Abstract. Self-oscillating h.f. power generators with a Class E amplifier are used as industrial high-efficiency h.f. power sources. In the paper a modification of the classic Class E generator by applying an additional capacitor in the matching and feedback circuit is presented. The proposed solution increases the degree of freedom in the generator design and can noticeably reduce its cost by eliminating the need for the use of high-value h.f. capacitors in the feedback loop divider. The self-oscillating Class E generator (150 W/6.78 MHz) with the proposed modification has been designed, built and its parameters were measured. Theoretical predictions were in a good agreement with experimental results. Moreover, the burst control method of the output power in the generator has also been presented.

Key words: Class E amplifier, self-oscillating power generator, output power control.

1. Introduction

In many industries h.f. power amplifiers and generators find applications in induction and dielectric heaters, plasma generators, electronic ballasts for fluorescent and HID lamps, energy harvesting circuits as well as resonant dc/dc converters [1–10]. Among possible solutions amplifiers and generators utilizing high-efficiency resonant switching Class DE and Class E amplifiers are particularly attractive. Resonant shaping of current and/or voltage waveforms across transistor switches (ZCS/ZVS switching) in these amplifiers significantly reduces switching losses increasing their power efficiency up 95–98% in practical applications [11, 12]. This decreases the cost of both used transistor switches and their cooling. Moreover, due to high efficiency the circuit miniaturization is also increased.

An essential problem in the application of Class DE and Class E amplifiers in the high-frequency range results from the switching of their transistors (often MOSFETs) which usually require driving of their gates with a rectangular, high-amplitude waveform $v_{GS}(t)$ (a typical amplitude of the gate waveform $V_{GSm} \geq 10$ V). Such a method of transistor switch driving needs charging and discharging of the large MOSFET input capacitance (often of the order of nanofarads) by high-amplitude current pulses and can cause power losses both in the driver circuit and the transistor gate resistance diminishing efficiency. These driving related power losses are proportional to the circuit operation frequency and the input capacitance of the used transistor switches. It becomes a major obstacle in designing switching power amplifiers operating at the frequencies above a few MHz and with output power beyond a few tens of watts. The problem is usually solved by the application of quasi-resonant or resonant gate drivers, which reduces required driving power and thus increases the operation frequency of the amplifiers to many MHz [9]. A disadvantage of

this approach is the complexity of driver circuits caused by the use of additional diodes, transistors and reactive components. Moreover, in quasi-resonant and resonant drivers to compensate for the slow rise and fall of the gate voltage waveform its amplitude is usually higher than in the non-resonant gate drivers with rectangular waveforms. This increases conduction losses in resonant drivers and can breakdown the transistor if the amplitude of its gate voltage waveform exceeds the transistor maximum rating.

A simpler method of resonant driving of transistor switches is used in self-oscillating power generators with a Class DE or Class E amplifier. In these generators the h.f. sinusoidal gate voltage is supplied to the transistor gate from the amplifier resonant output circuit by means of a reactive positive feedback network [11, 13, 14]. Such generators can operate with a high frequency and high efficiency limited only by the parameters of the used transistor switches and the required output power. A disadvantage of the self-oscillating power generators is the sensitivity of the gate voltage waveform to the changes of both the amplifier dc supply voltage and the impedance of the amplifier resonant circuit. These changes have to be controlled to avoid the operation of the amplifier transistors in non-ZVS conditions, which increase switching losses in transistors and lower the circuit efficiency. This phenomenon also makes the control of output power in self-oscillating generators a difficult problem.

Output power control in a h.f. self-oscillating power generator is usually obtained by changing its operating frequency, by regulating its dc supply voltage or by the burst control method. Electronic frequency modulation in a self-oscillating power generator by changing values of capacitors or inductors in a resonant circuit of the generator is difficult to achieve in practice because of high voltages and currents occurring in the components [15]. Therefore, the most often used method of

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output power control is by regulation of the dc supply voltage by means of an additional power supply. This, however, rises the cost and complexity of the generator as well as reduces its total efficiency. In the burst control method the oscillations in the generator are started and stopped periodically at the frequency many times lower than the oscillation frequency. By changing the ratio of the generator turn-on time to its turn-off time the output power can be controlled in a wide range with high efficiency [16, 17].

In the classic Class E self-oscillating power generator [11, 13, 14] its main advantage is high efficiency and simplicity of the resonant and feedback loop circuits. However, in some applications a more complex resonant network in the generator is desirable to optimize its operation as it has been shown in a self-oscillating electronic ballast with a Class E amplifier [3].

In the paper a self-oscillating Class E generator with a modified matching and feedback network is analyzed. The circuit eliminates the need to use expensive high-capacitance h.f. capacitors in the feedback signal divider of the generator. The proposed modification by reducing the values of capacitors in the gate circuit of the transistor switch simplifies also the application of the burst control of output power in the generator. A design method and results of the circuit experimental verification for a 150 W/6.78 MHz self-oscillating generator with a Class E amplifier are also given.

2. A self-oscillating Class E generator

A self-oscillating Class E generator is shown in Fig. 1a. The circuit consists of a Class E amplifier with the bi-directional transistor switch T , supply choke L_{CH} and output resonant circuit $C_R-L_{SR}-C_{SR}-R_L$. The sinusoidal output voltage v_O at the operating frequency f is supplied to the capacitor voltage divider C_1-C_2 . The divider along with inductance L_f comprises a positive feedback loop network driving the gate of the switch T with $v_{GS}(t)$ voltage waveform. The amplitude V_{GSm} of the gate voltage waveform is high enough to switch transistor T and the drain-to-gate phase shift for the fundamental component in the generator network satisfies phase conditions for oscillations. Dc voltage source V_G via resistance R_G sets the dc gate voltage V_{GS0} equal to the switch threshold voltage $V_{GSth} = V_{GS0} = V_G$, which initiates oscillations in the circuit with transistor normalized on-time $D = t_{ON}f = 0.5$. Capacitor C_O is the main reactive component which matches usually standard load resistance R_L to R_{opt} – resistance required in the Class E amplifier to obtain assumed output power at the given value of dc supply voltage V_{DD} . The use of capacitor C_O ensures that the values of capacitors in the divider C_1-C_2 can be small. Bi-directional Zener diode D_Z protects the switch gate from breakdown by a high value of V_{GSm} .

In a classic self-oscillating Class E generator (without capacitor C_O) the feedback signal divider C_1-C_2 is also a matching circuit for load resistance, which causes that even for low output power the value of capacitor C_2 can exceed hundreds of nanofarads [11]. Because C_1 and C_2 usually conduct high-amplitude currents it is necessary to limit conduc-

tion losses in them by applying h.f. low-loss expensive mica or porcelain capacitors as C_1, C_2 . Typical values of available mica or porcelain capacitors do not exceed 10 nanofarads. Therefore, high values of capacitance are obtained by paralleling many capacitors. This can significantly increase the cost of a classic Class E generator.

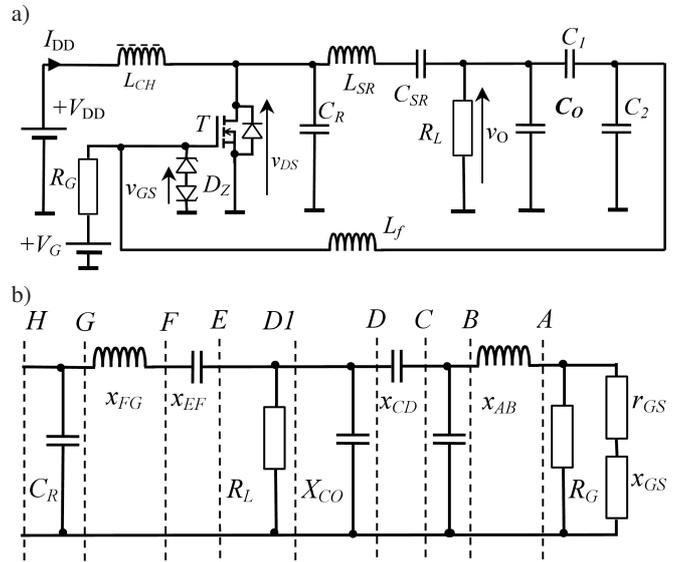


Fig. 1. The proposed self-oscillating Class E generator (a), equivalent circuit of the output and feedback network: C_O – additional matching capacitor, r_{GS}, x_{GS} – series resistance and reactance of the gate circuit, respectively (with taking into account diode D_Z capacitance)

3. Experimental circuit – design and measurements

To test the properties of the proposed modification in the self-oscillating Class E generator, the circuit of Fig. 1a was designed by means of the design procedure for the classic Class E generator given in [11]. In this procedure the resonant circuit and the feedback network are divided into sections (Fig. 1b). For the fundamental component for each section its input impedance (to the right of the section line) and its introduced phase shift are found. To achieve the operation of the generator at the desired frequency with high efficiency and demanded output power level it is necessary to satisfy both Barkhausen criteria [18, 19] and requirements for the nominal or off-nominal operation of the Class E amplifier [11, 12, 14].

In the proposed circuit an additional section DDI is introduced as shown in the equivalent circuit in Fig. 1b to take into account capacitor C_O . A parameter k is defined as a ratio of parallel reactances

$$k = \frac{X_E}{X_{C_O}}, \quad (1)$$

$$1 - k = \frac{X_E}{X_D}, \quad (2)$$

where X_E, X_D each is an input parallel reactance to the right of E and D , respectively, $X_{C_O} = 1/(2\pi f C_O)$ is the

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reactance of capacitor C_O at the operation frequency f . The value of k belongs to the range $0 \leq k < 1$. Thus for $k = 0$ is $X_E = X_D$ and $C_O = 0$, and for $k \rightarrow 1$ is $X_{CO} \rightarrow X_E$, $X_D \rightarrow \infty$ and $C_1 \rightarrow 0$, $C_2 \rightarrow 0$. Therefore using (1), (2) and the design procedure [11] for the assumed k the desired values of capacitors C_1 , C_2 , C_O can be found which are small enough to ensure that the applied capacitors are both small in size and inexpensive.

To illustrate the application of the design method described in [11] to the proposed circuit, a design procedure for the experimental circuit is given. The presented procedure shows how the circuit modifications have been included into the existing design method.

The proposed generator was designed for the following assumptions:

- output power $P_o = 150$ W, operating frequency $f = 6.78$ MHz,
- power efficiency $\eta = 0.91$,
- the supply voltage $V_{DD} = 48$ V, supply choke $L_{CH} = 10$ μ H,
- amplitude of the gate voltage waveform $V_{GSm} = 13$ V,
- loaded quality factor of the Class E amplifier series branch $Q_{SR} = 5$,
- load resistance $R_L = 50$ Ω ,
- series loss resistance in inductances L_{SR} and L_f estimated as $r_{SR} = 0.22$ Ω and $r_f = 0.085$ Ω , respectively,
- parameter $k = 0.95$,
- as the switch T transistor *DE275-501N16A* (IXYS) with Zener diode D_Z (two back to back connected *BZX85/C18* diodes) were used. Measured transistor T threshold voltage was $V_{GSth} = 4.8$ V = V_G . The series gate resistance and reactance measured with diode D_Z connected to the transistor T gate were $r_{GS} = 0.3$ Ω , $x_{GS} = -8.2$ Ω . From the manufacturer data the transistor T on-resistance was $r_{DS} = 0.4$ Ω . The transistor output capacitance was assumed to be a part of capacitor C_R ,
- resistance $R_G = 2$ k Ω was chosen to be high enough not to cause too much power loss in the gate circuit.

The power P_S supplying the generator comprises output power P_O and power losses P_l resulting from losses in the circuit components. Because the value of P_l is unknown at the beginning of calculations therefore P_S is estimated by using the assumed power efficiency η .

$$P_S = P_O + P_l = \frac{P_O}{\eta} = \frac{150}{0.91} = 164.835 \text{ W.} \quad (3)$$

The values of series load resistance R_{opt} , parallel reactance X_{CR} , and capacitance C_R for the nominal operation of the Class E amplifier with the supply power P_S is found from formulas given in [12] as follows:

$$R_{opt} = \frac{8}{\pi^2 + 4} \frac{V_{DD}^2}{P_S} = \frac{8}{\pi^2 + 4} \times \frac{48^2}{164.835} = 8.0623 \text{ } \Omega, \quad (4)$$

$$X_{CR} = -\frac{\pi(\pi^2 + 4)}{8} R_{opt} \quad (5)$$

$$= -\frac{\pi(\pi^2 + 4)}{8} \times 8.0623 = -43.912 \text{ } \Omega,$$

$$C_R = \frac{1}{2\pi f |X_{CR}|} \quad (6)$$

$$= \frac{1}{2\pi \times 6.78 \times 10^6 \times 43.912} = 534.57 \text{ pF.}$$

Resistance R_{opt} is equal to the input series equivalent resistance r_G to the right of the section G . Reactance X_{CR} is the parallel reactance of section GH

$$r_G = R_{opt}, \quad (7)$$

$$X_{GH} = X_{CR}. \quad (8)$$

The reactance factor q_G for the G section of the Class E amplifier in the nominal condition is found as [12]

$$q_G = \frac{x_G}{r_G} = \frac{\pi \times (\pi^2 - 4)}{16} = 1.1525. \quad (9)$$

Thus, the input series reactance x_G is

$$x_G = q_G r_G = 1.1525 \times 8.0623 = 9.2918 \text{ } \Omega. \quad (10)$$

For the assumed value of Q_{SR} and resistance r_G given by (7) the inductance L_{SR} is

$$L_{SR} = \frac{Q_{SR} r_G}{\omega} = \frac{5 \times 8.0623}{2\pi \times 6.78 \times 10^6} = 946.3 \text{ nH.} \quad (11)$$

The reactance of the section FG is equal to the reactance of L_{SR}

$$x_{FG} = \omega L_{SR} = Q_{SR} r_G = 5 \times 8.0623 = 40.312 \text{ } \Omega. \quad (12)$$

The series resistance r_E to the right of the section E is found by subtracting from r_G the loss resistance of inductance L_{SR}

$$r_E = r_G - r_{SR} = 8.0623 - 0.22 = 7.8423 \text{ } \Omega. \quad (13)$$

The input series reactance x_F to the right of the section F equals to

$$x_F = x_G - x_{FG} = 9.2918 - 40.312 = -31.020 \text{ } \Omega. \quad (14)$$

Load resistance R_L equal to the parallel resistance R_{D1E} of the section $D1E$ is shunted by the parallel resistance R_{D1} to the right of section $D1$. The resistance R_{D1} results from the losses in inductance L_f and the transistor T gate circuit (assuming that losses in capacitors C_1 , C_2 and Zener diode D_Z are negligible).

The resultant input parallel resistance R_E to the right of the section E is a parallel connection of R_L and R_{D1}

$$R_E = \frac{R_L R_{D1}}{R_L + R_{D1}} = \frac{R_{D1E} R_{D1}}{R_{D1E} + R_{D1}}. \quad (15)$$

The power P_E lost in R_E comprises output power P_O lost in R_L and power P_{D1} lost in R_{D1} .

$$P_E = P_O + P_{D1}, \quad (16)$$

$$P_O = \frac{V_O^2}{R_L} = \frac{V_O^2}{R_{D1E}}, \quad (17)$$

$$P_{D1} = \frac{V_O^2}{R_{D1}} = P_{Lf} + P_A, \quad (18)$$

where P_{Lf} – power loss in series parasitic resistance r_f of inductance L_f , P_A – power loss in section A caused by resistor R_G and the loss in the transistor gate.

To estimate power losses P_{Lf} and P_A it is necessary first to calculate the amplitude I_{Am} of the current flowing into section A and the input series resistance r_A resulting from losses in the gate circuit.

For the transistor T gate circuit the reactance factor q_{GS} equals

$$q_{GS} = \frac{R_{GS}}{X_{GS}} = \frac{x_{GS}}{r_{GS}} = \frac{-8.2}{0.3} = -27.333 \Omega. \quad (19)$$

The input parallel gate resistance R_{GS} and reactance X_{GS} are found by transforming the input series gate resistance r_{GS}

$$R_{GS} = r_{GS}(1 + q_{GS}^2) = 0.3 \times (1 + (-27.333)^2) = 224.43 \Omega, \quad (20)$$

$$X_{GS} = \frac{R_{GS}}{q_{GS}} = \frac{224.43}{-27.333} = -8.211 \Omega. \quad (21)$$

The input parallel resistance R_A , input parallel reactance X_A and the reactance factor q_A of section A result from the parallel connection of resistance R_G as well as the input parallel resistance R_{GS} and reactance X_{GS} of the transistor gate.

$$R_A = \frac{R_G R_{GS}}{R_G + R_{GS}} = \frac{2 \times 10^3 \times 224.43}{2 \times 10^3 + 224.43} = 201.79 \Omega, \quad (22)$$

$$X_A = X_{GS} = -8.211 \Omega, \quad (23)$$

$$q_A = \frac{R_A}{X_A} = \frac{201.79}{-8.211} = -24.576 \Omega. \quad (24)$$

The input series resistance r_A and reactance x_A for section A are

$$r_A = \frac{R_A}{1 + q_A^2} = \frac{201.79}{1 + (-24.576)^2} = 0.33355 \Omega, \quad (25)$$

$$x_A = q_A r_A = (-24.576) \times 0.33355 = -8.1973 \Omega. \quad (26)$$

The amplitude V_{Am} of the voltage in the node A equals to the amplitude of V_{GSm} . Thus, the amplitude I_{Am} of current flowing into the node A for the assumed amplitude V_{GSm} is given by

$$I_{Am} = \frac{V_{Am}}{\sqrt{r_A^2 + x_A^2}} = \frac{V_{GSm}}{\sqrt{r_A^2 + x_A^2}} = \frac{13}{\sqrt{0.33355^2 + (-8.1973)^2}} = 1.5846 \text{ A}. \quad (27)$$

Hence, power losses P_A and P_{Lf} are

$$P_A = \frac{I_{Am}^2}{2} r_A = \frac{1.5846^2}{2} \times 0.33355 = 0.41877 \text{ W}, \quad (28)$$

$$P_{Lf} = \frac{I_{Am}^2}{2} r_f = \frac{1.5846^2}{2} \times 0.085 = 0.10672 \text{ W}. \quad (29)$$

From (18) power P_{D1} is

$$P_{D1} = P_{Lf} + P_A = 0.41877 + 0.10672 = 0.52549 \text{ W}. \quad (30)$$

Resistance R_{D1} from (17), (18) is equal to

$$R_{D1} = \frac{V_O^2}{P_{D1}} = \frac{P_O R_L}{P_{D1}} = \frac{150 \times 50}{0.52549} = 14272.39 \Omega. \quad (31)$$

Resistance R_E is found from (15) as

$$R_E = \frac{R_L R_{D1}}{R_L + R_{D1}} = \frac{50 \times 14272.39}{50 + 14272.39} = 49.825 \Omega. \quad (32)$$

The input parallel resistance R_E and the input series resistance r_E for section E are related by the equation

$$r_E = \frac{R_E}{1 + q_E^2}, \quad (33)$$

where reactance factor q_E is

$$q_E = \frac{R_E}{X_E} = \frac{x_E}{r_E}. \quad (34)$$

From (13) and (33) the capacitive reactance factor q_E

$$q_E = -\sqrt{\frac{R_E}{r_E} - 1} = -\sqrt{\frac{49.825}{7.8423} - 1} = -2.3137. \quad (35)$$

Hence, the input series reactance x_E to the right of section E is equal to

$$x_E = q_E r_E = -2.3137 \times 7.8423 = -18.145 \Omega. \quad (36)$$

The series reactance of section FE is

$$x_{FE} = x_F - x_E = -31.020 - (-18.145) = -12.875 \Omega. \quad (37)$$

The value of capacitor C_{SR} equals

$$C_{SR} = \frac{1}{2\pi f |x_{FE}|} = \frac{1}{2\pi \times 6.78 \times 10^6 \times 12.875} = 1.823 \text{ nF}. \quad (38)$$

From (35) the input parallel reactance X_E to the right of section E is

$$X_E = \frac{R_E}{q_E} = \frac{49.825}{-2.3137} = -21.535 \Omega. \quad (39)$$

The input parallel reactance X_E is equal to a parallel connection of reactance X_{CO} and the input parallel reactance X_D to the right of section D

$$\frac{1}{X_E} = \frac{1}{X_{CO}} + \frac{1}{X_D}. \quad (40)$$

By substituting (2) to (39) reactance X_{CO} and capacitance C_O are found

$$X_{CO} = \frac{X_E}{k} = \frac{-21.535}{0.95} = -22.668 \Omega, \quad (41)$$

$$C_O = \frac{1}{2\pi f |X_{CO}|} = \frac{1}{2\pi \times 6.78 \times 10^6 \times 22.668} = 1.0356 \text{ nF}. \quad (42)$$

Reactance X_D is from (40)

$$X_D = \frac{X_E \cdot X_{CO}}{X_{CO} - X_E} = \frac{(-21.535) \times (-22.668)}{(-22.668) - (-21.535)} = -430.85 \Omega. \quad (43)$$

Since $R_D = R_{D1}$ thus, the reactance factor q_D of section D equals

$$q_D = \frac{R_D}{X_D} = \frac{R_{D1}}{X_D} = \frac{14272.39}{-430.85} = -33.126. \quad (44)$$

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The input series resistance r_D and input series reactance x_D to the right of section D are

$$r_D = \frac{R_D}{1+q_D^2} = \frac{R_{D1}}{1+q_D^2} = \frac{14272.39}{1+(-33.126)^2} = 12.995 \Omega, \quad (45)$$

$$x_D = q_D r_D = (-33.126) \times 12.995 = -430.47 \Omega. \quad (46)$$

To find C_1 , C_2 and L_f it is necessary to calculate input series resistances r_B , r_C and the reactance factors q_B , q_C for sections B and C

$$r_C = r_D, \quad (47)$$

$$\begin{aligned} r_B &= r_{AB} + r_A = r_{Lf} + r_A \\ &= 0.085 + 0.33355 = 0.41855 \Omega. \end{aligned} \quad (48)$$

As it has been shown in [11] the reactance factors q_B and q_C can be found from (47) and the following relations:

$$R_C = R_B, \quad (49)$$

$$\begin{aligned} -\phi_{AG} &= -\phi_{EG} - \phi_{CD} - \phi_{AB}, \\ 0 &= \phi_{AG} - (\arctan q_E - \arctan q_G) \end{aligned} \quad (50)$$

$$-(\arctan q_C - \arctan q_D) - (\arctan q_A - \arctan q_B),$$

where $-\phi_{AG} = 3.4209$ rad – is equal to the phase shift between the fundamental components of the waveforms $v_{GS}(t)$ and $v_{DS}(t)$ for the nominal operation of the Class E amplifier [11, 12]; ϕ_{EG} , ϕ_{CD} , ϕ_{AB} – are the phase shifts for the fundamental component between the input and output voltages for the respective sections, EG , CD and AB .

The equation (50) is the Barkhausen phase shift criterion for the circuit of Fig. 1a. By assuming

$$\begin{aligned} \theta &= \arctan q_C - \arctan q_B \\ &= \phi_{AG} - \arctan q_A + \arctan q_D - \arctan q_E \\ &\quad + \arctan q_G = -3.4209 - \arctan(-24.576) \\ &\quad + \arctan(-33.126) - \arctan(-2.3137) \\ &\quad + \arctan(1.1525) = -1.4124 \text{ rad} \end{aligned} \quad (51)$$

then from (47), (48), (49) and (50) the reactance factors q_B , q_C are obtained as [11]

$$\begin{aligned} q_C &= \frac{1}{\tan \theta} \left(\sqrt{\frac{r_B}{r_C} (1 + \tan^2 \theta) - 1} \right) \\ &= \frac{1}{\tan(-1.4124)} \left(\sqrt{\frac{0.41855}{12.995} \times (1 + \tan^2(-1.4124)) - 1} \right) \\ &= -0.02222, \end{aligned} \quad (52)$$

$$\begin{aligned} q_B &= \sqrt{\frac{r_C}{r_B} (1 + q_C^2) - 1} \\ &= \sqrt{\frac{12.995}{0.41855} \times (1 + (-0.02222)^2) - 1} = 5.4830. \end{aligned} \quad (53)$$

For section B the series input reactance x_B and reactance x_{AB} are given as

$$x_B = q_B r_B = 5.4830 \times 0.41855 = 2.2949 \Omega, \quad (54)$$

$$\begin{aligned} x_{AB} &= X_{Lf} = x_B - x_A \\ &= 2.2949 - (-8.1973) = 10.4922 \Omega. \end{aligned} \quad (55)$$

Hence, the feedback inductance L_f equals

$$L_f = \frac{X_{Lf}}{2\pi f} = \frac{10.4922}{2\pi \times 6.78 \times 10^6} = 246.3 \text{ nH}. \quad (56)$$

The series input reactance x_C to the right of section C is

$$x_C = q_C r_C = -0.02222 \times 12.995 = -0.28875 \Omega. \quad (57)$$

The reactance X_{C1} of capacitor C_1 is expressed as

$$\begin{aligned} X_{C1} &= x_{CD} = x_D - x_C \\ &= -430.47 - (-0.28875) = -430.18 \Omega. \end{aligned} \quad (58)$$

The value of capacitor C_1 is equal to

$$\begin{aligned} C_1 &= \frac{1}{2\pi f |X_{C1}|} \\ &= \frac{1}{2\pi \times 6.78 \times 10^6 \times 430.18} = 54.57 \text{ pF}. \end{aligned} \quad (59)$$

The input parallel resistance R_B and reactance X_B to the right of section B are

$$\begin{aligned} R_B &= r_B (1 + q_B^2) \\ &= 0.41855 \times (1 + 5.4830^2) = 13.002 \Omega, \end{aligned} \quad (60)$$

$$X_B = \frac{R_B}{q_B} = \frac{13.002}{5.4830} = 2.3713. \quad (61)$$

Because $R_C = R_B$ the input parallel reactance X_C to the right of section C is

$$X_C = \frac{R_C}{q_C} = \frac{13.002}{-0.02222} = -585.15 \Omega. \quad (62)$$

The reactance X_C results from a parallel connection of reactance X_B and reactance X_{C2}

$$\frac{1}{X_C} = \frac{1}{X_B} + \frac{1}{X_{C2}}. \quad (63)$$

Thus, reactance X_{C2} and capacitance C_2 are found as

$$X_{C2} = \frac{X_B X_C}{X_B - X_C} = \frac{2.3713 \times (-585.15)}{2.3713 - (-585.15)} = -2.3617 \Omega, \quad (64)$$

$$\begin{aligned} C_2 &= \frac{1}{2\pi f |X_{C2}|} \\ &= \frac{1}{2\pi \times 6.78 \times 10^6 \times 2.3617} = 9.940 \text{ nF}. \end{aligned} \quad (65)$$

Power losses in the generator components can be found from the following equations [11, 12]:

– dc supply current I_{DD}

$$I_{DD} = \frac{P_S}{V_{DD}} = \frac{164.835}{48} = 3.434 \text{ A}, \quad (66)$$

– amplitude I_m of current in C_{SR} and L_{SR}

$$\begin{aligned} I_m &= \frac{\sqrt{\pi^2 + 4}}{2} I_{DD} \\ &= \frac{\sqrt{\pi^2 + 4}}{2} \times 3.434 = 6.394 \text{ A}, \end{aligned} \quad (67)$$

– dc power loss in choke L_{CH}

$$P_{LCH} = I_{DD}^2 r_{LCH} = 3.434^2 \times 0.062 = 0.7311 \text{ W}, \quad (68)$$

– conduction losses in the transistor T

$$P_{Tcond} = \frac{\pi^2 + 28}{16} r_{DS} I_{DD}^2 \quad (69)$$

$$= \frac{\pi^2 + 28}{16} \times 0.4 \times 3.434^2 = 11.16 \text{ W},$$

– switching losses in the transistor T for the assumed fall time $t_{fall} = 5 \text{ ns}$ of the drain current $i_{DS}(t)$

$$P_{Tswitch} = \frac{(2\pi f t_{fall})^2}{12} P_S \quad (70)$$

$$= \frac{(2\pi \times 6.78 \times 10^6 \times 5 \times 10^{-9})^2}{12} \times 164.835 = 0.6232 \text{ W},$$

– losses P_{GS} in the gate of the transistor T result from series gate resistance r_{GS} and the gate current with the amplitude I_{GSm}

$$P_{GS} = \frac{I_{GSm}^2}{2} r_{GS} = \frac{V_{GSm}^2}{2(r_{GS}^2 + x_{GS}^2)} r_{GS} \quad (71)$$

$$= \frac{13^2}{2 \times (0.3^2 + 8.2^2)} \times 0.3 = 0.3765 \text{ W},$$

– total power losses P_{Ttot} in the transistor T are

$$P_{Ttot} = P_{Tcond} + P_{Tswitch} + P_{GS} \quad (72)$$

$$= 11.16 + 0.6232 + 0.3765 = 12.16 \text{ W},$$

– power losses in inductance L_{SR} is

$$P_{LSR} = \frac{I_m^2}{2} r_{LSR} = \frac{6.394^2}{2} \times 0.22 = 4.497 \text{ W}. \quad (73)$$

Parasitic loss resistances of capacitors are usually low if compared to loss resistances in inductances and the switch. Therefore, the influence of the capacitor losses on waveforms and the network impedance in the generator can be neglected thus simplifying the design procedure. However, even low capacitor losses can significantly raise the temperature of a small-size capacitor degrading its parameters. Therefore, it is necessary to estimate these losses as well. The loss resistances of the used in the experimental circuit SMD mica capacitors were estimated from the manufacturer's catalogue (Cornell Dubilier) as follows $r_{CR} = 0.02 \Omega$, $r_{CSR} = 0.005 \Omega$, $r_{CO} = 0.01 \Omega$, $r_{C1} = 0.1 \Omega$, $r_{C2} = 0.001 \Omega$.

Thus, power losses in capacitors C_R , C_{SR} , C_O , C_1 , C_2 are

$$P_{CR} = \frac{\pi^2 - 4}{16} I_{DD}^2 r_{CR} \quad (74)$$

$$= \frac{\pi^2 - 4}{16} \times 3.434^2 \times 0.02 = 0.0865 \text{ W},$$

$$P_{CSR} = \frac{I_m^2}{2} r_{CSR} = \frac{6.394^2}{2} \times 0.005 = 0.1022 \text{ W}, \quad (75)$$

$$P_{CO} = \left(\frac{V_O}{|X_{CO}|} \right)^2 r_{CO} \quad (76)$$

$$= \frac{P_O R_L}{|X_{CO}|^2} r_{CO} = \frac{150 \times 50}{22.668} \times 0.01 = 0.1460 \text{ W},$$

$$P_{C1} = \frac{V_O^2}{(X_{C1} + x_C)^2 + r_C^2} r_{C1} \quad (77)$$

$$= \frac{P_O R_L}{(X_{C1} + x_C)^2 + r_C^2} r_{C1}$$

$$= \frac{150 \times 50}{(-430.18 - 0.28875)^2 + 12.995^2} \times 0.1 = 0.0040 \text{ W},$$

$$P_{C2} = \left(\frac{V_O}{\sqrt{(X_{C1} + x_C)^2 + r_C^2}} \frac{\sqrt{x_C^2 + r_C^2}}{X_{C2}} \right)^2 r_{C2} \quad (78)$$

$$= \frac{V_O^2}{(X_{C1} + x_C)^2 + r_C^2} \frac{x_C^2 + r_C^2}{X_{C2}^2} r_{C2}$$

$$= \frac{P_O R_L}{(X_{C1} + x_C)^2 + r_C^2} \frac{x_C^2 + r_C^2}{X_{C2}^2} r_{C2}$$

$$= \frac{150 \times 50}{(-430.18 - 0.28875)^2 + 12.995^2} \cdot \frac{(-0.28875)^2 + 12.995^2}{(-2.3617)^2} \times 0.001 = 0.0012 \text{ W}.$$

The estimated total power losses in the generator components equals

$$P_l = P_{D1} + P_{LCH} + P_{Tcond} + P_{Tswitch} + P_{GS} \quad (79)$$

$$+ P_{LSR} + P_{CR} + P_{CSR} + P_{CO} + P_{C1} + P_{C2}$$

$$= 0.52549 + 0.7111 + 11.16 + 0.6232 + 0.3765$$

$$+ 4.497 + 0.0865 + 0.1022 + 0.1460$$

$$+ 0.0040 + 0.0012 = 18.233 \text{ W}.$$

The generator efficiency η obtained from (3) and (79) is

$$\eta = \frac{P_S - P_l}{P_S} = \frac{164.835 - 18.233}{164.835} \approx 0.89. \quad (80)$$

The calculated efficiency (80) differs only about 2% from its assumed value.

For the given assumptions the component values were also calculated for the classic Class E generator ($k = 0$). The obtained results for the classic Class E generator and its modified version of Fig. 1a are presented in Table 1 and compared with the component values used in the experimental generator. The results prove that by the application of capacitor C_O it was possible to reduce by 20 times the values of C_1 and C_2 with slightly increased value of inductance L_f . Comparing to the classic circuit the modification does not change the operating frequency of the generator nor its output power. The current and voltage waveforms of the transistor T also remain unchanged. The differences between calculated and experimental values of the components are caused by the assumed low value of Q_{SR} [20] and by difficulties in the precise estimation of nonlinear capacitances of the transistor T and diode D_Z . Figure 2 and Fig. 3a show the influence of the parameter k on values of C_1 , C_2 , C_O and L_f as well as on normalized current amplitudes in C_1 , C_2 , C_O . By changing k it is possible to optimize the generator with the respect to the

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values of capacitors C_1 , C_2 , C_O or the amplitude of current in them.

The proposed circuit of Fig. 1a was built according to the assumptions using inductances made with toroidal carbonyl cores L_{CH} (T106-2), L_{SR} (T200-6), L_f (T50-6). Measured total efficiency of the generator was $\eta = 0.91$ and resulted mainly from conduction power losses in on-resistance r_{DS} of the transistor switch T as well as losses in inductance L_{SR} . Voltage waveforms measured in the circuit (Fig. 3b) agreed well with the assumed ($V_{GSm} = 13$ V) and calculated values ($V_{DSpeak} = 3.61V_{DD} = 173.3$ V [20], $V_{ORMS} = (P_O R_L)^{1/2} = 86.6$ V).

Table 1

Component values for the Class E generator

Component	classic generator calculated values	values for the proposed circuit of Fig. 1a	
		calculated	experimental
C_R	534.57 pF	534.57 pF	390 pF
L_{SR}	946.3 nH	946.3 nH	944.8 nH
C_{SR}	1.823 nF	1.823 nF	1nF+820 pF
R_L	50 Ω	50 Ω	50 Ω
C_O	–	1.0356 nF	1 nF
C_1	1.095 nF	54.57 pF	51 pF
C_2	199.6 nF	9.940 nF	4*2.2 nF+1nF+3*390 pF
L_f	193.9 nH	246.3 nH	226 nH

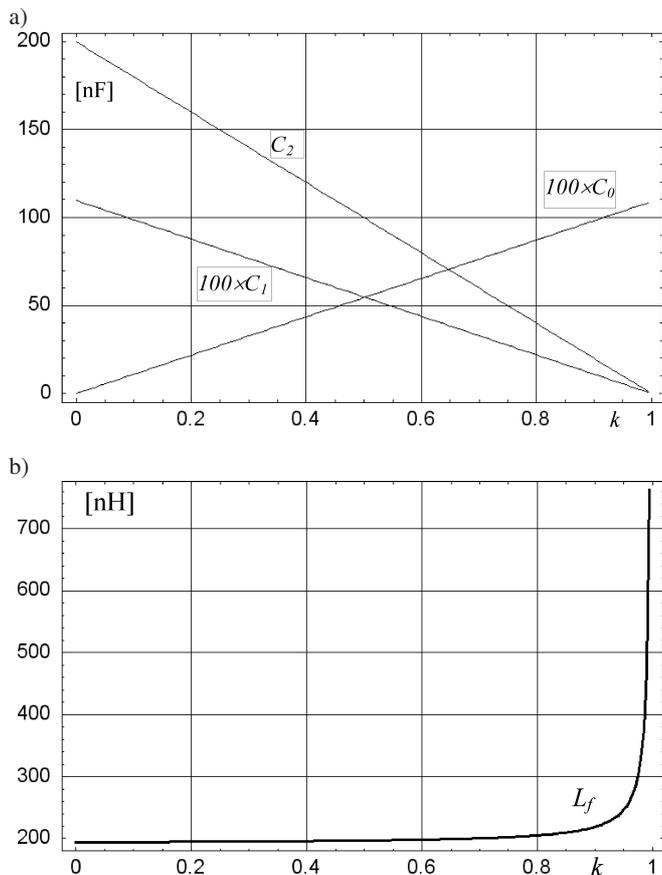


Fig. 2. Calculated values of capacitors C_1 , C_2 , C_O (a) and inductance L_f (b) versus k in the designed generator

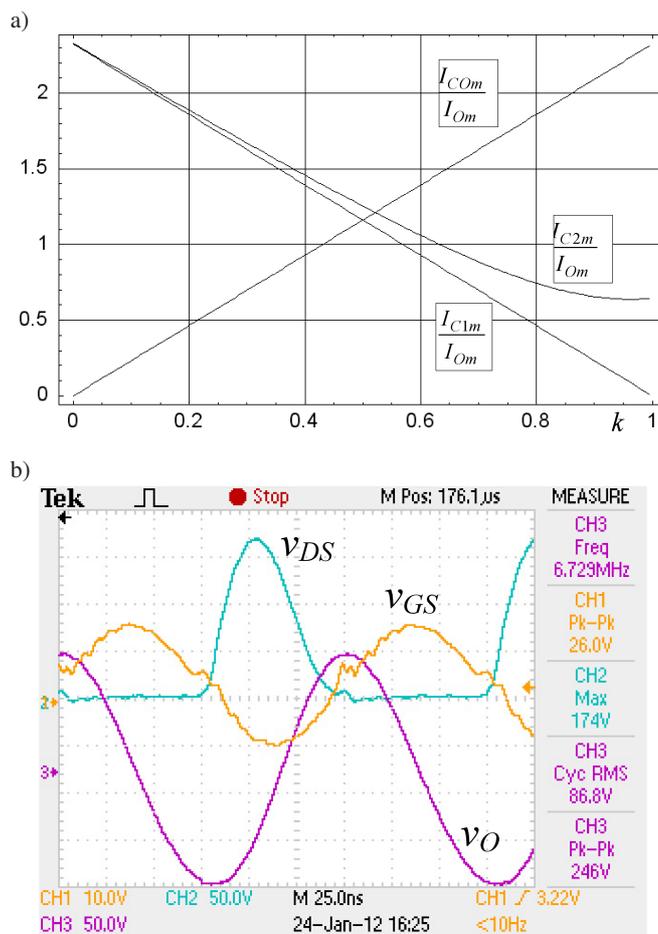


Fig. 3. Calculated normalized amplitudes of current in capacitors C_1 , C_2 , C_O versus k in the designed generator (a) and measured voltage waveforms in the experimental circuit (I_{Om} – amplitude of current in load resistance R_L)

The significant decrease of the capacitor C_2 value in the experimental generator also reduces the time constant $\tau_G \approx R_G(C_{iss} + C_2)$ of the voltage V_{GS0} rise in the gate circuit (C_{iss} – the input capacitance of the transistor switch). Therefore it simplifies the application of the burst control method in the generator (Fig. 4). The burst control of output power is achieved by means of transistor switch S . The voltage v_{CONTR} high state turns on switch S , which short circuits transistor T gate and stops oscillations in the generator. When voltage v_{CONTR} is low the switch S turns off and the voltage V_{GS0} rises with the time constant τ_G till it achieves the threshold value V_{GSth} . Then oscillations are initiated. By regulating the ratio of the switch S turn-on time to its turn-off time the generator output power can be controlled in the whole range with high efficiency. Diodes D_{Z1} , D_1 protect the transistor T from breakdown by the high peak value of v_{DS} voltage which appears when the transistor T is turned off by the switch S . The high value of v_{DS} results from the resonant discharge of energy accumulated in supply choke L_{CH} during the time interval when the generator produces h.f. output power [16, 17].

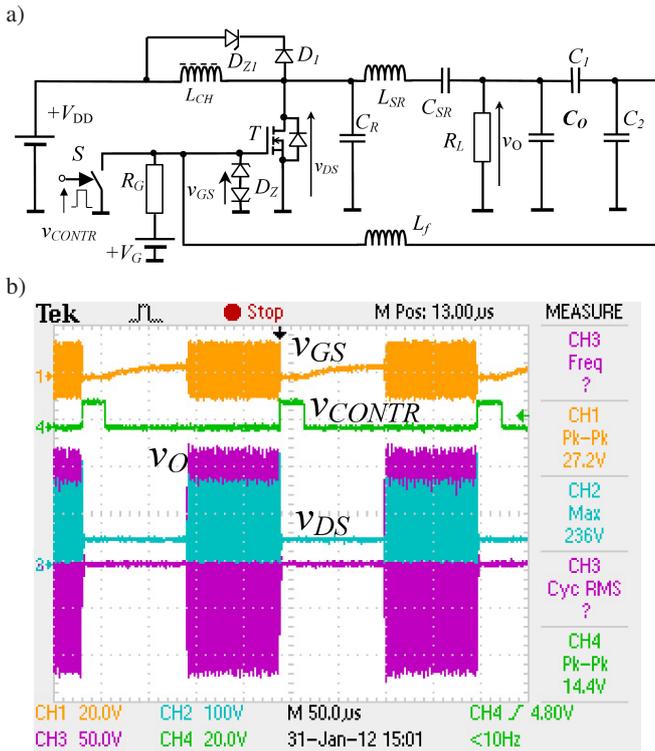


Fig. 4. Burst control of output power in the self-oscillating Class E generator. Circuit diagram (a) and measured voltage waveforms (b)

4. Conclusions

The described modification of the Class E generator increases the degree of freedom in designing the circuit. By decreasing the required values of capacitors in the feedback network a reduction of cost in the generator is possible. Moreover, the presented output power control by the burst control method ensures output power regulation in the generator in the whole range with high efficiency. The generator can find applications in high-efficiency h.f. industrial power generators.

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