

A simulation model of a converter with soft switching of transistors has been created. New calculation methods and means of measuring the equivalent frequency characteristics of converting devices with negative feedback to the problems of stability analysis have been applied. A comparative analysis of the calculated frequency characteristics of the converter in the normal operation mode with the characteristics of its linear model is presented. The features of calculating the equivalent frequency characteristics of the loop amplification of key devices are considered. It is noted that it is the equivalent frequency characteristics of the loop amplification that describe the relationship of frequency properties with the dynamics of voltage converters with pulse-width modulation and make it possible to reliably determine the real stability margins, predict the generation modes and open up the possibility of obtaining maximum negative feedback in a given frequency band of key devices. The non-minimal phase properties of the converter significantly complicate the achievement of high-quality stabilization of the output current (stabilization coefficient) of pulse stabilizers using only the deflection control principle. A successful solution to the problem is provided by the use of the principle of combined control. Compensation of the disturbance effect (changes in the supply voltage) gives a much better result than counteracting it through the feedback loop. It is quite realistic to exclude the influence of the input voltage $u(t)$ on the output current i_2 in a static mode ($u=const$) and in the absence of a negative feedback loop, which provides $\dot{i}=0$ for $u=const$ in a steady state, and to provide better dynamic properties of the converter. The influence of the nonlinear properties of pulse width modulation on stability is manifested in the considered characteristics in the form of an additional phase shift and resonant bursts in the vicinity of the clock frequency subharmonics. Additional feedbacks introduced into the control loop make it possible to solve synthesis problems using the most simple technical means

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ANALYSIS OF DYNAMIC CHARACTERISTICS OF THE INVERTER OPERATING ON A COMPLEX LOAD

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1. Introduction

Analysis of information materials [1–8] showed that at present high-frequency voltage converters (HVC) implementing the technology of “soft” switching, known as Zero Voltage Switch (ZVS), seem to be the most promising for the implementation of highly reliable medium-power power supplies for modern electrical technologies (plasma – cutting, welding, surfacing and spraying) [2]. Pulse DC voltage converters using the technology of “soft” switching based on pulse width modulation (PWM) provide high efficiency (at least 90–95 %), corresponding to the world standards “80 PLUS, GOLD”, acceptable weight and dimensions (up to 1.5 kW/kg) and reliability [3, 4]. They are nonlinear closed-loop automatic control systems prone to chaotic dynamics. In them, with a certain combination of parameters, low-frequency periodic oscillations with large amplitude, as well as quasiperiodic and chaotic oscillations [11], which are dangerous for the power section of the converter, can occur.

Obviously, the main task at the design stage of these devices is such a selection of control system parameters in which the specified modes are excluded and the design periodic mode (one cycle or 1-cycle) will be stable [11]. Achiev-

ing high-quality stabilization of the output current remains an important direction in the construction of power supplies for modern electrical technologies.

2. Literature review and problem statement

In [2, 3], it is noted that the best circuitry solution for the implementation of soft switching is possible in phase-shifted single-phase voltage inverters with the introduction of additional reactive elements, which allows adaptively changing the operating mode of the converter, while maintaining the soft switching mode of power switches at any load. In the presence of two T-shaped CLC-circuits in the bridge implementation, it is possible to further increase the efficiency, output power and specific indicators of a single device and reduce electromagnetic interference [2, 3].

Fig. 1 shows a schematic diagram of the power part of the converter with soft switching of MIS transistors (PS-ZVS-FB-PWM Converter) [2], the development was awarded with a diploma for winning the Ukrainian competition-exhibition “The Best Domestic Product of 2008”. The main parameters of the investigated power supply were:

- rated output power of the source – 15 kW;
- load current regulation range $D_I=5$ ($I_n=20-100$ A);
- rated operating voltage – 70–200 V;
- maximum no-load voltage – 250 V.

The development of methods for modeling and studying the dynamics of pulse converters (PC) is one of the most important problems of power electronics [12]. This is explained by the widespread use of PC in energy-saving power systems of technological installations, an increase in requirements for the parameters of PC and an increase in the order of differential equations describing the dynamics, which leads to the complication of their analysis and synthesis [13].

To simplify the design process of such systems, quasi-continuous representations of PC are often used [6, 9]. Designing on the basis of a continuous model of a PC requires a significant removal of the cutoff frequency ω_c and switching ω_s ($\omega_c/\omega_s \geq 10$) [6, 9, 10]. An increase in the ratio ω_c/ω_s with an attempt to reduce the possibility of undesirable dynamic modes is not an optimal solution, since it leads to a deterioration in the quality of regulation ($\omega_c \downarrow$), or to an increase in losses in powerful elements ($\omega_s \uparrow$) [5, 9].

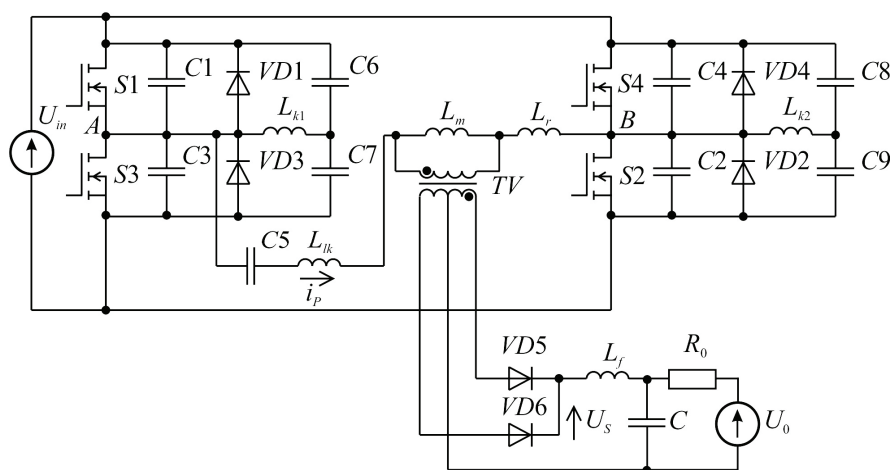


Fig. 1. Power circuit diagram of the FB-ZVS-PS DC-DC converter

In [7, 8], the possibility of operating a converter for a resistive or complex load is considered. At the same time, the work of a modern PC on a more complex arc load is of interest [5, 9]. When working, for example, on an arc load with a capacitance, the frequency characteristic of the continuous part of the direct transmission channel of a switching power supply changes and the stability of its operation, as well as its dynamic characteristics, can be violated.

In linear or linearized systems, to find the frequency characteristics in accordance with the classical approach in a certain section, a negative feedback (NF) circuit is opened and a harmonic source is connected to the input [10]. In pulse systems with output variable regulation, such a technique for measuring frequency characteristics cannot give a reliable estimate of the stability of a closed system [10, 11].

To determine the frequency dependences of the module and the phase of the loop gain, the input and output impedances of powerful PC in conditions when the feedback loop is closed, it is required to introduce a source of harmonic disturbance into the feedback circuit. Repetition of such an analysis for different values of the disturbance frequency makes it possible to reveal the frequency dependences of the studied characteristics. Performing such calculations is a

time-consuming process and requires a lot of computer time. Therefore, the development of an approach that makes it possible to level the above-mentioned drawbacks of the methods of analysis and synthesis of regulators in the frequency domain seems promising. In particular, the results obtained in this case can be used:

- when implementing a new direction in automatic control of objects with undefined parameters – frequency control [13],
- when applying new calculation methods and means for measuring the equivalent frequency response (EFR) of converting devices with negative feedback [11, 14] to the problems of analyzing stability, static and dynamic characteristics in their normal operation on an arc load.

The considered methods of analysis and synthesis of PWM systems make it possible, from a unified position, using well-tested methods, to solve with most of the problems that arise in this case high accuracy:

- to explore sustainability in both small and large;
- to study the dynamic characteristics of the system, taking into account the inertia of the PWM;
- to assess the increase or decrease in stability margins when the system operates in nonlinear modes, to explain the reason for the occurrence of multifrequency oscillations in the system;
- to synthesize corrective devices that provide the necessary stability margins.

3. The aim and objectives of the study

The aim of the work is to create a simulation model of a converter with soft switching of transistors and control devices for analyzing the characteristics of the converter under consideration.

To achieve the goal, the following objectives were set:

- analysis of the dynamic model of the closed-loop current stabilization system;
- analysis of static operating modes of the converter;
- calculation of frequency characteristics and EFR of loop gain of switching devices;
- assessment of the dynamic resistances of the switching power supply.

4. Analysis of the dynamic model of the closed-loop current stabilization system

The structure of the closed-loop PC circuit can be assembled from equivalent circuits of certain elements and cells and use their models to describe a nonlinear discrete energy flow control system. At the same time, when forming such a scheme, it is important to select only the elements that determine the dynamic properties of the system, since often the influence of elements that are insignificant from this point of view can be taken into account in the parameters of the main links of the structural scheme. When these conditions are met, we get the equivalent circuit shown in Fig. 2.

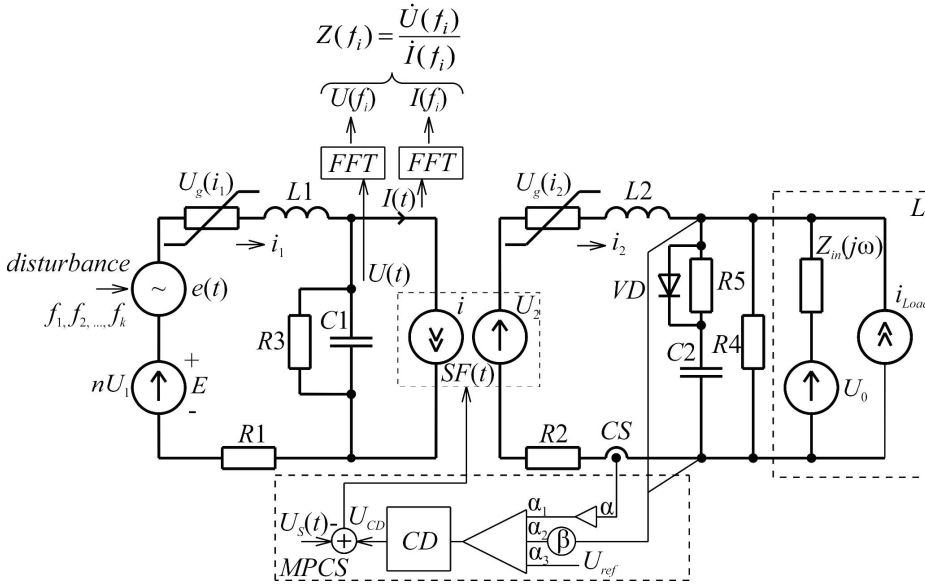


Fig. 2. Equivalent circuit of a pulse converter with feedback on the choke current and output voltage

Here $R_g(i_1)$, $R_g(i_2)$ are the nonlinear resistances of the equivalent circuits of the input and output rectifiers, FFT is the discrete Fourier transform unit, i_{Load} is the additional load current that can vary arbitrarily regardless of the main load.

In the equivalent circuit, the output voltage of the first rectifier is represented by a constant voltage $E=nU_1$ and a source of a variable component $e(t)$, and the input filter is represented by reactive elements $L1C1$ with losses $R1, R3$. The output filter is represented by a single-link $L2C2$ filter with losses, which are taken into account by the resistance $R2$, which also includes the internal resistance of the secondary circuit of the PC.

The control object of the PC as an automatic control system is the energy flow transmitted from the energy input $U_1(t)$ to the output $U_2(t)$ – to the inertial nonlinear load L , therefore, the dynamic processes of the power source are determined not only by the properties of the converter itself, but also by the load.

Feedback is provided by the current i_2 and voltage U_{Load} sensors and microprocessor control scheme (MPCS). The control scheme is made double-circuit and includes a source of the reference signal U_{ref} , a comparison element, two correcting devices with gains α_1, α_2 , and a pulse-width modulator (PWM). The transistor control signals are designated by the switching break function SF .

The correcting device CD forms the required control voltage from the error signal, which is supplied to the block of the switching function $SF(t)$.

The switching converter itself is represented here by a switching function connecting the input and output circuits with equations of the form (information ξ and energy U_1 inputs with output U_2)

$$U_2(t) = SF(t)U_{C1}(t), \quad i(t) = SF(t)i_2(t).$$

The switching function SF for controlling switching elements in the most general form is determined through the feedback equation $\xi(X, t) = U_{CD} - U_s(t) = 0$, in which the unfolding voltage $U_s(t)$ is formed according to the law

$$U_s(t) = U_m \left[\frac{t}{\tau} - E_1 \left(\frac{t}{\tau} \right) \right],$$

where U_m is the peak value of the unfolding voltage; τ is the PWM quantization period; E_1 is the Antje integer function.

The variable resistances of rectifier diodes for this case can be represented by the following expressions:

$$\begin{aligned} R_g(i) &= r_T + \\ &+ R_R \cdot \frac{1}{2} [1 - \text{sign}(i)] = \\ &= \begin{cases} r_T & \text{at } i > 0; \\ r_T + R_R & \text{at } i < 0. \end{cases} \end{aligned}$$

If we neglect the relatively small values of the leakage current i_R , then we can assume that $R_R \rightarrow \infty$ and u_R coincides with the abscissa axis up to the value $u_R = U_{BR}$.

The supposed regulator can be described by the relation

$$u_{CR} = -k_{CR} (U_{ref} - k_i R_{CS} I_{Load} + \beta k_u U_{Load}),$$

where $k_i = \alpha\alpha_1 / \alpha_3$, $k_u = \alpha_2 / \alpha_3$, $k_{CR} = \alpha_3$.

The dynamic properties of the load – the electric arc of the power source are determined by the dependence of its input resistance on the frequency $Z_{in}(j\omega)$ [2, 15, 16]. In this case, the input operator arc resistance

$$Z_m(s) = \frac{R_{st0}\theta s + R_{diff0}}{\theta s + 1},$$

where R_{st0} , R_{diff0} – static and differential arc resistance [2, 15, 16], respectively, at the point of attachment I_0 ; θ is the arc time constant.

The resulting dynamic model simulates the converter well in both steady state and transient modes. To get closer, if necessary, to real electromagnetic processes, for example, the inductance and resistance of the magnetizing circuit of the transformer, etc. can additionally be presented in the form of separate elements.

The solution of the above problems was carried out by performing computational experiments with the following parameters of the mathematical model of the converter shown in Fig. 2: $U_1=540$ V; $n=0.46$; $I_{Load}=100$ A; $L_2=300$ μ H; $C_2=0.5$ μ F; $k_i R_{CS}=0.01$ Ω ; $\beta=0.5$; $k_u=15$; $R_{diff0}=-0.49$ Ω ; $U_0=170$ V; $U_m=1.0$ V; $U_{ref}=1.0$ V $k_u=0.001$; $r_T=4.7$ m Ω ; $U_{T0}=1.13$ V; $t_{rr}=35$ ns; $C_1=1,000$ μ F; $L_1=0$; $f=52$ kHz.

For a pulse converter (Fig. 2) operating on a complex load, it is easy to obtain a linearized circuit model of the power unit for the continuous current mode of the choke [12] and according to it transfer functions (TF) without taking into account the dynamic properties of the arc for the control action:

$$W_{i/d}(s) = \frac{\tilde{I}_L(s)}{d(s)} = \frac{(b_0 s + b_1) n U_m / R_{eq}}{Q(s)}, \quad (1)$$

and for disturbing influences:

$$W_{i_L/u_m}(s) = \frac{\tilde{I}_L(s)}{\tilde{U}_{in}(s)} = \frac{(b_0s+1)nD/R_{eq}}{Q(s)},$$

$$W_{i_L/u_0}(s) = \frac{\tilde{I}_L(s)}{\tilde{U}_0(s)} = -\frac{1/R_{eq}}{Q(s)}, \quad W_{i_L/i}(s) = \frac{\tilde{I}_L(s)}{\tilde{I}(s)} = \frac{1}{Q(s)},$$

where $Q(s) = a_0s^2 + a_1s + a_2$; $a_0 = L_2C_2$; $a_1 = L_2/R_{eq} + C_2R_\Sigma$; $a_2 = R_\Sigma/R_{eq} + 1$; $R_{eq} = R \parallel R_{diff0}$; $b_0 = C_2R_{eq}$; $b_1 = 1$; R_Σ – internal resistance of the converter (this also includes the active resistance of the choke); the sign “-” indicates an infinitesimal change in the variable relative to the value in the periodic mode; D is the value of the duty cycle in the steady state.

The differential arc resistance R_{diff0} is negative. In what follows, we will understand by R_{diff0}^* its module. Taking this into account, relation (1) takes the form

$$W_{i_L/d}(s) = \frac{-C_2R_{eq}^*s + 1}{L_2C_2s^2 - (L_2/R_{eq}^* - C_2R_\Sigma)s + 1 - R_\Sigma/R_{eq}^*} \times \left(-\frac{nU_{in}}{R_{eq}^*}\right). \quad (2)$$

This system has TF (2) in the open state and, therefore, zero in the right half-plane. In this case, it is impossible to obtain an arbitrarily high performance and at the same time the limiting error (equal to zero) is not attainable [13].

In this case, the stability conditions obtained on the basis of the static current-voltage characteristic of the arc are as follows:

$$\frac{L_2}{R_{eq}} + C_2R_\Sigma > 0; \quad \frac{R_\Sigma}{R_{eq}} + 1 > 0,$$

whence follows

$$L_2 > C_2R_\Sigma R_{eq}^* \left(R_{diff0}^* < \frac{L_2}{L_2/R_{eq} + CR_\Sigma} \right);$$

$$R_\Sigma > R_{eq}^* \left(R_{diff0}^* < R_\Sigma \parallel R_A \right). \quad (3)$$

In the second inequality (3), usually $R \gg R_\Sigma$ (in particular, R is absent, i.e., $R \rightarrow \infty$).

For convenience, we introduce dimensionless variables:

$$\bar{R} = R_\Sigma / R_{eq}^*; \quad \tau_1 = L / (R_{eq}^* \theta); \quad \tau_2 = R_\Sigma C_2 / \theta.$$

Then the stability conditions will be rewritten in a simpler form

$$\bar{R} > 1, \quad \tau_2 < \tau_1.$$

For the characteristic of an arc of the form $u = P_0/i$ [9], we obtain

$$\frac{P_0}{i^2} < R_\Sigma.$$

Consequently, in this case, only the steepness of the I-V characteristic of the arc is of decisive importance for stability.

5. Analysis of static operating modes of the converter

For the system shown in Fig. 2, in the initial analysis we will compose a graph (Fig. 3, a), covering the static connections between the variables. Here $R_{eq} = R \parallel (-R_{diff0})$, R_Σ – includes all active losses in the circuit.

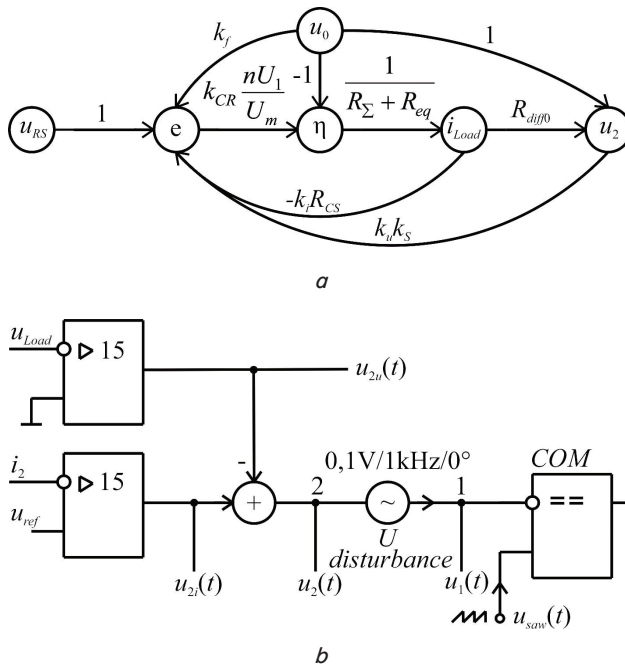


Fig. 3. Control system: a – graph of the equivalent circuit of the closed-loop control system; b – system with output and voltage control

It is possible to compensate for the component of the deviation caused by a change in the supply voltage by realizing control by the disturbance. For this, it is sufficient to change the relative duration (t_p) of the components of the period (T) of switching the system structure ($d = 2t_p/T$) depending on the supply voltage U_1 . In this case, the DC gain of the converter remains constant when U_1 changes.

Another and more dangerous disturbing effect is the change in the EMF of the load U_0 . The condition of absolute invariance of even a linearized stabilization system to the disturbing action U_0 cannot be realized in practice, since due to the non-minimal phase properties of the system [17], the disturbing action communication chain turns out to be unstable. However, it is quite realistic to exclude the influence of u_0 on i_{Load} in the static mode ($\dot{u}_0 = \text{const}$) and in the absence of the feedback loop ($K(s) = 0$). For this, it is enough to accept

$$k_f = \frac{1}{k_{CR}} \cdot \frac{U_m}{nU_1}.$$

Indeed, in this case, we obtain a zero result at the vertex η under the action of two channels, which ensures $\dot{i}_{Load} = 0$ at $\dot{u}_0 = \text{const}$ in the steady state.

The maximum depth (theoretical limit) of the NF of the PWM converter, taking into account the nonlinear properties of the latter in a given frequency band, can be determined from the expression [18]:

$$A_{0max} = 6 \ln 2 - 2 \ln \frac{\omega_{cut}}{\omega_0} - 2 \ln \pi - 2 \text{ [Np]}, \quad (4)$$

where $A_0=A_{0max}$, $0 \leq \omega \leq \omega_{cut}$ – maximum depth of the NF, constant in the operating range; ω_{cut} is the cutoff frequency of the correcting circuit; ω_0 is the PWM clock frequency.

Since in the case under consideration $\omega_{cut}/\omega_0=0.25$, in accordance with (4) we have

$$A_{0max}=2.6428 \text{ Np}; A_{0max} \cong 23 \text{ dB}.$$

In the linear continuous representation of the system, the boundary gain of the feedback amplifier is determined from the relation

$$k_{CR.lim} < \frac{L_2/|R_{eq}| - C_2 R_\Sigma}{k_1 R_{CS} C_2} \cdot \frac{U_m}{nU_1} = 16 (24 \text{ dB}),$$

where $R_{eq}=R_4||(-R_{diff0})$.

The experimental value of the boundary gain obtained on the simulation model of a physical object is 30 (29.5 dB).

Taking the minimum stability margin of 6 dB, we obtain the maximum possible gain of the regulator by deviation

$$k=k_{CR.lim}/2=15.0.$$

At $k_{CR}>k_{CR.lim}$, subharmonic self-oscillations of half-frequency are established, or further the process is accompanied by significantly increased pulsations of a chaotic nature.

6. Calculation of frequency characteristics and EFR of loop gain of switching devices

All the advantages of this approach are manifested in the analysis of nonlinear systems, including impulse systems [11]. Calculation of frequency characteristics (FC) through analysis in the time domain does not require any changes in the system under study and takes into account real processes occurring in a nonlinear (pulse) circuit. To obtain reliable results using this method, two procedures are required.

Let it be required to obtain the FC of the “input-output” transfer coefficient for the considered current stabilization system (Fig. 2). Construction of the frequency dependence of the TF of the closed-loop stabilization system with a disturbance from the input supply voltage

$$\Phi_U(j\omega) = \tilde{I}(j\omega) / \tilde{U}_1(j\omega)$$

is necessary to assess the smoothing properties of a pulsed current stabilizer.

First, we calculate the transient process for the output current (Fig. 4, a). Fig. 4, a shows that at the end of the 16th period, the transient process enters the linear zone and, starting from the 17th period, a steady operating mode is observed.

From the analysis of the curve (Fig. 4, a) it follows that the steady-state process corresponds to a stable one-cycle mode of the converter. The 52 kHz ripple clock is 4 A.

Then we apply a harmonic disturbance with an amplitude of 15 V to the input. The calculation interval is divided into two parts: the first – from 0 to 0.31 ms – free component; the second – from 0.31 ms to 20 periods of disturbance after 0.31 ms – the harmonic part of the response.

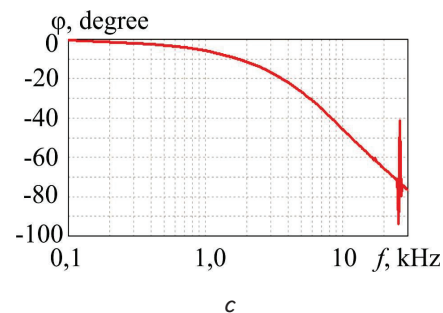
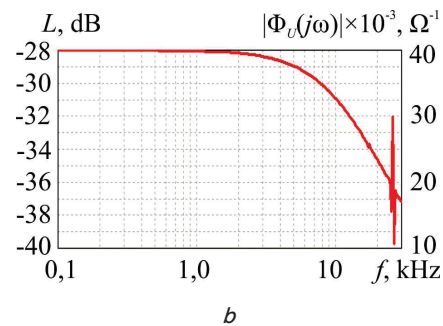
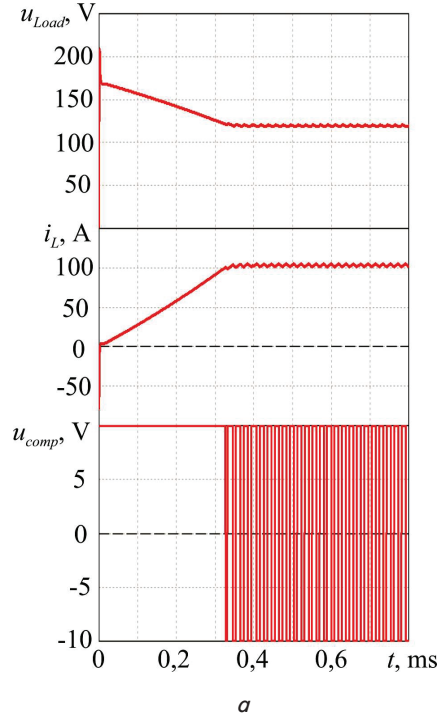


Fig. 4. Oscillograms of the transient and steady-state modes: a – for the current in the load when working out the jump of the reference action $U_{ref}=1 \text{ V}$ from zero initial conditions; b, c – frequency dependence for the exact pulse model of the converter with control only by the output

The output of an object excited by such an action will (after the damping of transient processes) change according to the law

$$y(t) = A_c(\omega) \sin[\omega t + \phi_c(\omega)],$$

where the function $A_c(\omega)$ is the amplitude-frequency characteristic of the closed-loop system. By setting the generator to different frequencies, we get the AFC and the PFC of the

closed-loop system, combined into the main complex transfer function:

$$\Phi_U(j\omega) = A_c(\omega) \exp j\phi_c(\omega).$$

Important characteristics of the quality of pulse current stabilizers (PCS) are their stability and ability to suppress the ripple of the input supply voltage, especially when powered from network sources; the latter is characterized by the smoothing factor

$$k_{sm} = \frac{I_{out0} \cdot U_{in-}}{U_{in0} \cdot I_{out-}},$$

where U_{in-} , I_{out-} – the amplitude of the variable component, respectively, of the voltage at the input and the current at the output of the stabilizer. The value k_{sm} is determined for a sinusoidal ripple of the input supply voltage.

To assess the smoothing properties of the considered PC, it is necessary to have the frequency dependence of the input-to-output transmission coefficient for the closed-loop system with PWM, i.e. the characteristics of attenuation of the input low-frequency ripple q ($q=1/k_{sm}$).

Such a calculated frequency dependence for an accurate pulse model of an output-only converter, which shows how its output responds to a change in input voltage, is shown in Fig. 4, *b, c*. FC shown in Fig. 4, *b, c* are constructed for $k(s)=15$ and with the same parameters and operating conditions of the stabilizer. As follows from Fig. 4, *b, c*, the level of the LAFC ordinates is less than zero not only at low frequencies corresponding to long transient times, i.e., the considered converter has a wide range of suppression of input voltage ripples, and the absence of a resonant peak indicates the monotonicity of the transient process.

To determine the transfer coefficient of the loop gain, we use the “closed loop” method [11], which is known in the measurement practice as a means of maintaining a DC operating point. Analysis of the frequency response and phase response of the feedback loop transmission function allows one to estimate the achieved feedback depth, the amplitude and phase margins, the stability of the main system quality indicators, and the degree of distortion and interference suppression.

Loop gain was calculated using the formulas:

$$T_U(j\omega) = \frac{\dot{U}_{2v}(j\omega)}{\dot{U}_1(j\omega)}, \quad T_I(j\omega) = \frac{\dot{U}_{2i}(j\omega)}{\dot{U}_1(j\omega)},$$

$$T(j\omega) = \frac{\dot{U}_2(j\omega)}{\dot{U}_1(j\omega)}.$$

The found frequency dependences $T_I(j\omega)$ for the pulse model of the converter (Fig. 4, *a*) are similar to the frequency dependences $T(j\omega)$. The calculated frequency dependences $T_U(j\omega)$ are shown in Fig. 5, *b*.

For comparison, the investigated converter (Fig. 1) is compared with an equivalent linear circuit [11]. For this purpose, we will use the replacement of the nonlinear elements of the converter with a broadband amplifier [11]. In this case, the magnitude of the gain of this amplifier is chosen so that the values of the frequency response and phase response of the loop gain of the pulse and linear models coincide at the lowest measured frequency of 100 Hz.

Equivalent phase response (Fig. 5, *a, b*) significantly differs in the high-frequency region from the analogous

characteristic obtained for the linearized model of the converter.

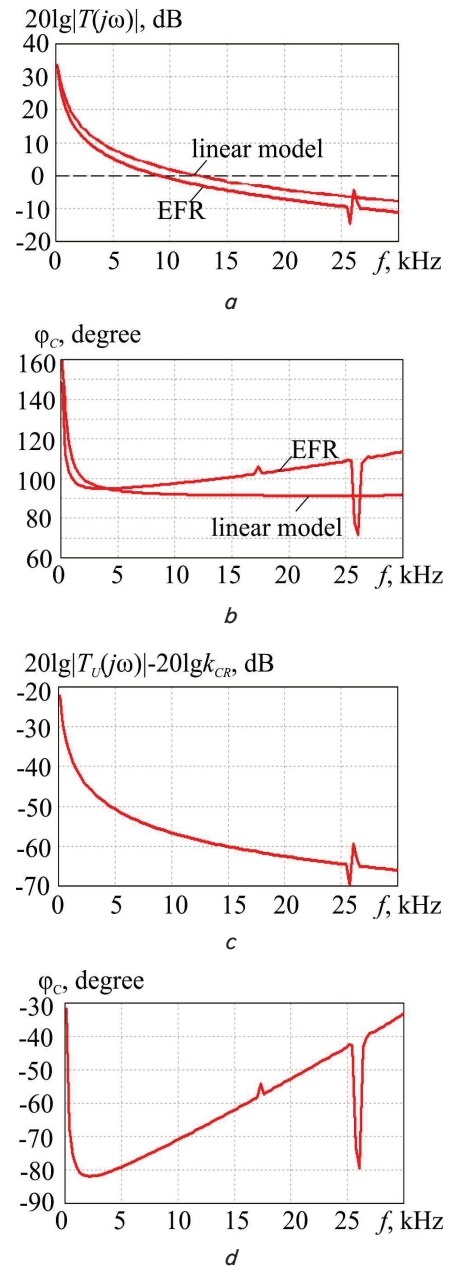


Fig. 5. Frequency dependences: *a, b* – for the pulse model of the converter; *c, d* – calculated frequency dependences

EFR reveals nonlinear resonance in the vicinity of the second subharmonic ($f_T/2$), the maximum peaks of which reach 10 dB in amplitude and 50–60° in phase. Analysis of the characteristics in this frequency range allows us to conclude that it is possible to fulfill the generation conditions with a further increase in the transmission coefficient of one of the blocks of the forward path.

7. Evaluation of the dynamic resistances of a switching power supply

Of all the dynamic parameters of voltage converters, the most important is its output impedance [19, 20]. An increased

value of the latter leads to a decrease in emissions and dips in the output current during the transient process in the case of a sharp change in load parameters, which is highly desirable.

A radical way to increase the output resistance and improve the parameters of the transient process of a pulsed secondary power supply (PPS) is the use of a positive feedback on the disturbance – on the output voltage.

Output impedance of PPS with open circuits of the FB (Fig. 6, *a, b*)

$$Z_{00}(s) = \frac{\tilde{U}_{Load}(s)}{\tilde{i}_{Load}(s)} = \frac{R_{\Sigma} + sL_2}{L_2 C_2 s^2 + (RC_2 + L_2 / R_{diff0})s + 1 + R_{\Sigma} / R_{diff0}},$$

and the output impedance of the PPS with a closed circuit of the FB, for example, by voltage

$$Z_0(s) = \frac{Z_{00}(s)}{1 - T_V(s)},$$

where T_V is the loop voltage amplification factor of the feedback circuit, which is numerically equal to the PF of the open-loop PPS system.

Due to the action of the feedback on the output voltage, the resulting resistance increases significantly. As follows from the comparison of the results (Fig. 6, *c, d*), the average value of the output resistance in the presence of such a connection ($k_u=0.0005$) is ~ 1.8 times greater than in its absence. In this case, $Z_0(j\omega)$ is flatter, therefore, the frequency dependence $Z_0(j\omega)$ for the two-loop control system is more favorable in terms of dynamics than for the case of using only one output current feedback.

As is known [14], the stability of the “converter – complex load” system can also be considered from the point of view of a cascade connection of four-terminal networks: the

first four-terminal network is a power supply, the second is a complex load.

From this position, the system is stable if the condition [14] is not met in the entire frequency range:

$$\frac{Z_{out}^{(1)}(j\omega)}{Z_{in}^{(2)}(j\omega)} = -1 \Leftrightarrow \begin{cases} |Z_{out}^{(1)}(j\omega)| = |Z_{in}^{(2)}(j\omega)| \\ \arg(Z_{out}^{(1)}(j\omega)) - \\ -\arg(Z_{in}^{(2)}(j\omega)) = 180^\circ, \end{cases} \quad (5)$$

where $Z_{out}^{(1)}(j\omega)$ is the output resistance of the converter; $Z_{in}^{(2)}(j\omega) = Z_l(j\omega)$ – complex load resistance.

Thus, for the fulfillment of condition (5) in the “converter – complex load” system, it is necessary that the argument of the PC output resistance exceeds 90° . With a further increase in the conversion coefficient, we obtain a frequency range of 10...30 kHz, in which the argument of the PC output impedance is greater than 90° . In this frequency range, it becomes possible to fulfill condition (5) and a self-oscillating mode of PC operation can occur.

Fig. 6 it follows that, for example, for $k_u > 0.0005$, the argument of the output resistance varies in the range from 0° to 90° (i.e., the PC degenerates into a circuit with one reactive element – a capacitor), and since the argument of the resistance of any linear complex two-terminal is in the range from $+90^\circ$ to -90° , then condition (5) for autogeneration is not met. The figure below illustrates the main effect – the feedback on the output voltage with the P-correction element in its circuit leads to the equalization of the frequency response.

The main condition (criterion) of the static (aperiodic) stability of the energy system of the PC-arc is, as is known [1, 15], the ratio (parametric criterion of stability)

$$k_s = k_r > 0, \quad (6)$$

where k_s, k_r – the stability and resistance coefficients, respectively.

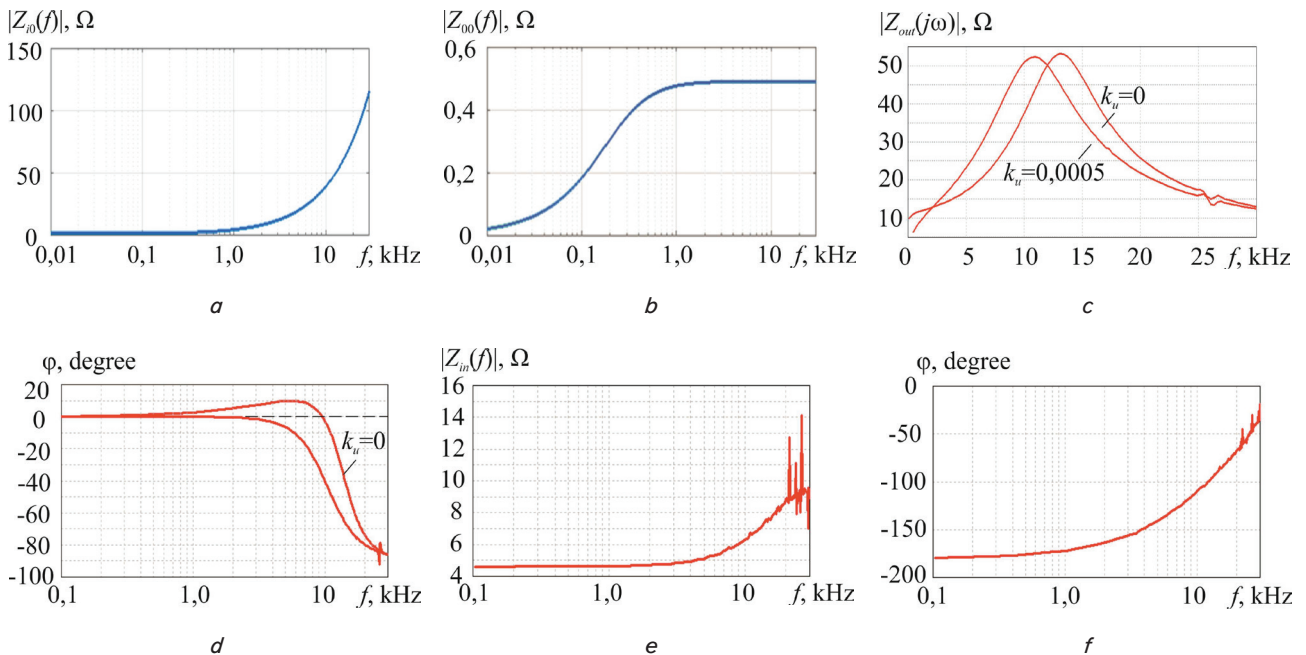


Fig. 6. Dynamic resistances of a pulsed power supply: *a, b* – output resistance with open circuits of the FB; *c, d* – output resistance with a closed circuit FB; *e, f* – input resistance without an input filter with open FB, loaded on a load resistance

The fulfillment of condition (6) usually directly follows from the nature of the intersection of the static characteristics of the PC and the arc in the vicinity of the considered equilibrium position, and in this case, the calculation of k_c from the equations is not required at all. Of course, with a steep external characteristic of the PC, only one regime point is possible (the point of static equilibrium). In this case, the more steeply dipping the external characteristic of the PC, the smaller the amplitude of the current fluctuations at a given value of the voltage fluctuations on the arc and the more effective the use of the plasmatron will be without reducing the reliability of its operation.

Input impedance of PCS (Fig. 6, e, f) without an input filter with open feedback, loaded on the load resistance R_{diff0} :

$$Z_{i0}(s) = \left. \frac{\tilde{U}_m(s)}{\tilde{I}(s)} \right|_{d(s)=0} = \frac{R_{eq} + R_y}{(nD)^2} \times \frac{1 + s \left[\frac{1}{R_{eq} + R_y} + (R_{eq} \parallel R_y) C_2 \right] + s^2 L_2 C_2 \frac{R_{eq}}{R_{eq} + R_y}}{1 + s R_{eq} C_2},$$

and the input impedance of the device (Fig. 2), for example, with a closed current feedback is defined as (serial feedback)

$$Z_i(s) = Z_{i0}(s)(1 + T_i(s)),$$

where $R_{eq} = R_4 \parallel R_{diff0}$; $T_i(s)$ – loop current amplification factor of the feedback circuit, equal to the TF of the open-loop PCS system $T_i(s) = W_r(s)$.

Consider the frequency dependence of the phase of the input resistance of a closed converter. Fig. 6, e, f shows that at low compared to the clock frequencies ($f/f_t < 0.01$), the phase of the input resistance is close to -180° . At $f/f_t > 0.01$, a smooth phase change occurs. Thus, in the frequency range $0.001 < f/f_t < 0.5$ the phase of the input resistance of the converter $\arg Z_{in}(j\omega) < -90^\circ$. As follows from the theory of circuits [19], in the indicated frequency range, the real part of the complex input resistance $U(\omega) = \text{Re } Z_{in}(j\omega) < 0$, i.e. the active component of the input resistance is negative. Note that it is in this frequency range that instability is possible, for example, of the line filter – converter and system the master source – slave system.

8. Discussion of the results of the study of the dynamic characteristics of the converter operating on a complex load

In the presented work, the peculiarity of the operation of a promising PC for an arc load is considered and it is shown that in a PC that stably operates on a resistive load, a self-oscillating mode can occur when operating on an arc.

From the frequency characteristics of the loop amplification of the PC operating on an arc load (Fig. 5), it follows that it is stable. The phase stability margin is 38° , and the gain stability margin is 8 dB, which guarantees reliable operation and the required quality of transients. However, as the loop gain increases, it can become agitated. The self-oscillating mode is also indicated by the ratio (5) of the PC output resistance and the complex load resistance.

Using the frequency characteristics of the PC loop gain, as well as using the frequency characteristics of the PC out-

put impedance and the complex load impedance, it is possible to determine whether the converter will work stably for an electric arc. Moreover, the frequency characteristics of the PC loop amplification make it possible to determine whether the transducer will be absolutely stable or conditionally stable.

The equivalent frequency characteristics of the power supply system substantially depend on the mode of its operation. Therefore, when designing a stable impulse system, it is necessary to take larger stability margins in the frequency domain in modulus (by 3–6 dB) and phase (by $10\text{--}20^\circ$) than when designing linear systems.

From the dependences $Z(f)$ shown in Fig. 6, c, it follows that in the presence of an FB for the output voltage, the output resistance of the converter in the range of $0\text{--}10^4$ Hz increases almost 5 times, and there is no resonance maximum.

The research results can be used in the implementation of a new direction in the automatic control of objects with undefined parameters – frequency control. Additional feedbacks introduced into the control loop make it possible to solve synthesis problems using the most simple technical means.

Although these techniques already allow solving a large number of problems associated with power converters with PWM, they require development to solve more complex problems, for example, when building a system that is optimal in terms of speed.

9. Conclusions

1. The EFR of the loop amplification really more successfully describes the relationship of frequency properties with the dynamics of voltage converters with PWM and makes it possible to reliably estimate the real stability margins in amplitude and phase, predict the generation modes and open up the possibility of obtaining a simpler way of maximum NFB in a given frequency band of the converting devices.

2. A successful solution to the problem of achieving high-quality stabilization of the output current – the experimental value of the boundary gain can be reduced, and the transient processes are significantly improved – gives the use of the principle of combined control. In the above-mentioned converter, the compensation of the disturbing effect (changes in the supply voltage) gives a much better result (coordinate-parametric invariance is ensured) than its parrying along the feedback loop.

3. The influence of the nonlinear properties of PWM on stability is manifested in the considered characteristics in the form of an additional phase shift and resonant bursts in the vicinity of the clock frequency subharmonics.

4. A radical way to increase the output resistance of the PC is the use of a positive FB for the output voltage. The average value of the output impedance can be increased by 4–5 times, and at the same time its frequency dependence is more favorable in relation to dynamics.

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