Calculation and Analysis of Dynamic Properties of a Soft Switching Converter under Operation on the Arc Load

Eugen Vereshchago
Dept. of Marine Instrument
Admiral Makarov National University of Shipbuilding, NUOS,
Mykolaiv, Ukraine
venmkua@gmail.com

Vitalii Kostiuchenko
Dept. of Marine Electric Power Systems
Admiral Makarov National University of Shipbuilding, NUOS,
Mykolaiv, Ukraine
vikmkua@gmail.com

Andrii Hrieshnov
Dept. of Marine Instrument
Admiral Makarov National University of Shipbuilding, NUOS,
Mykolaiv, Ukraine
hrieshnov@gmail.com

***Abstract*—The processes in the converter with soft switching transistors operating on an arc load are investigated. A comparative analysis of the calculated frequency characteristics of a PWM converter in normal operation with the characteristics of its linear model is presented. The features of calculating the equivalent frequency characteristics of the loop gain of switching devices are considered. The dynamic resistance of a switching power supply is estimated.**

Keywords—switching power supply, frequency characteristics, switching at zero voltage, pulse-width modulation, chaotic modes.

I. Introduction

An analysis of papers [1-2] showed that currently high-frequency voltage converters (HFVC), which implement the soft-switching technology, known as Zero Voltage Switch (ZVS), seem to be the most promising for the implementation of highly reliable sources of medium power for modern electrical technologies [2].

Pulsed DC-DC converters using soft-switching technology are nonlinear closed-loop automatic control systems, prone to chaotic dynamics. In them, with a certain combination of parameters, low-frequency periodic oscillations with a large amplitude can arise, as well as quasiperiodic and chaotic oscillations [5], which are dangerous for the converter power part.

Obviously, the main task at the design stage of these converters is a control system parameters selection for undesired modes to be excluded and the design periodic mode to be stable (period cycle one or 1-cycle) [5].

The aim of the work is to create a simulation model of a converter with soft switching transistors and control devices for conducting their tests, the results of which can be used to implement a new field in the automatic control of objects with uncertain parameters of frequency control [5], and the use of new calculation methods and measurement tools for measurement equivalent frequency characteristics (EFC) of the converter with negative feedback (NFB) for stability, static and dynamic analysis under normal functioning on arc load.

Implementation and application of the “closed loop” methodology for calculating the frequency characteristics of the loop gain function, the frequency dependences of the module and the phase of complex resistance (impedance) of a converter with various types of feedbacks in the mode of their normal functioning, which allows reliable estimation of stability margins, predicting generation modes, and solving the maximization task of the NFB depths of these devices and others become relevant.

To study the dynamics of pulsed energy conversion systems (PECS) in predicting and identifying nonlinear phenomena [2, 4], an independent task is the collection and processing of data from full-scale and numerical experiments. Full automation of such calculations was carried out in the computer program FASTMEAN [4], which is an effective tool for analyzing powerful pulse systems in the time and frequency domains.

There are many options for constructing the topology of the power part of DC-DC converters with soft switching [1, 2]. The best circuitry solution to date, which allows for soft switching (ZVS) of power switches of the power supply at any load, is possible in phase-shifted single-phase voltage inverters [1, 10] with two T-shaped CLC-circuits in the bridge implementation [1]. This particular circuit is examined below - the most promising for the implementation of powerful, highly reliable power supplies with a low level of interference. The FASTMEAN system was used as a working modeling tool [4].

II. Analysis of the Dynamic Modes of Operation of the Power Source

The management object of Power Supply (PS) as an automatic control system is a flow of power transmitted from the power input $U_1(t)$ to an output $U_2(t)$ - in inertial nonlinear load L, so dynamic power processes are determined not only by the properties of the converter, but also the load.

Wherein the structure of closed system PS-arc can be composed of an equivalent circuit elements and cells [2] and used their models to describe the nonlinear discrete energy flow control system. Moreover, in forming such circuit only select elements defining dynamical e properties of the system. The equivalent circuit of the closed-loop control system drawn up taking into account these remarks is shown in Fig. 1. Here, $R_g(i_1)$, $R_g(i_2)$ - nonlinear resistance equivalent circuits of the input and output rectifiers. Through α and β are indicated coefficients of signal amplifier transmission CS (u_{fb}~i_{Load}) and sensor FB on the output voltage VS respectively, as k_{CR} and i_{Load} - the gain of the current regulator (CR) and the additional current load, which can vary regardless of the load parameters, FFT is a discrete Fourier transform block.

978-1-7281-9713-5/20/$31.00 ©2020 IEEE

In the equivalent circuit, the input voltage of the first rectifier is represented by a constant voltage $U'_1 = nU_1$, referred to the secondary transformer winding, and the input filter - by reactive elements $L1C1$ with resistive losses $R1$, $R3$. The output filter is represented by a single-chain $L2C2$ filter with losses, which are taken into account by the resistance $R2$, including the internal resistance of the secondary circuit of the PS.

The correction device (CD) generates from the error signal the required control voltage supplied to the switching function block SF(t).

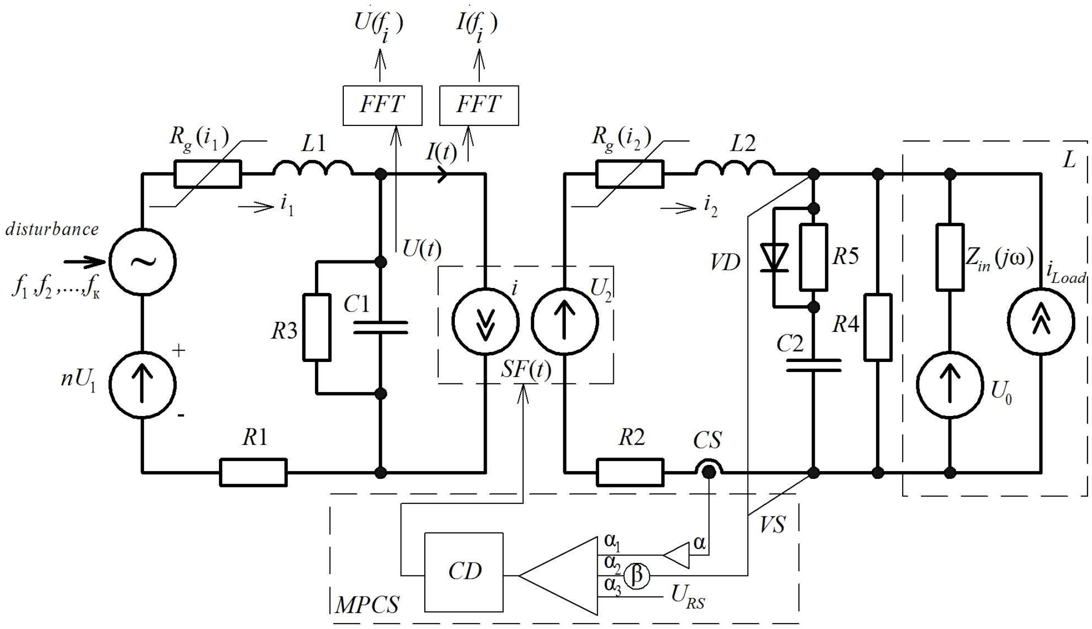


Fig. 1. Equivalent circuit of the closed control system.

The converter is represented by a switching function that connects the input and output circuits equations of the form

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

The putative regulator can be described by the ratio

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

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

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, 5-7]. In this case, the input operational resistance of the arc

$$Z(s) = (R_{st0}\theta s + R_{diff0})/(\theta s + 1)\ ,$$

where R_{st0}, R_{diff0} - are the static and differential resistance of the arc [2, 5, 7], respectively, at the reference point I_0; θ - is the arc time constant.

The equivalent load equivalent circuit of the power supply in the case under consideration can be represented in accordance with [6, 7] with the differential resistance R_{diff0} with a small parasitic inductance connected in series with the active resistance in series with it.

To decrease the effect of the energy output capacitor development and maintenance of arc discharges in technological load in series with the last is entered resistance $R5$ in parallel with the diode VD [6, 8, 9].

The solution of the above tasks was accomplished by conducting computational experiments using the capabilities of the FASTMEAN package for the parameters of the converter studied in [1], shown in Fig. 1: $U_1 = 540$ V; $n = 0.46$; $I_{Load} = 100$ A; $L = 300$ μH; $C = 0.5$ μF; $k_i R_{CS} = 0.01$ Ω; $\beta = 0.5$; $k_i = 15$; $R_{diff0} = -0.49$ Ω; $U_0 = 170$ V; $U_m = 1.0$ V; $U_{RS} = 1.0$ V; $k_u = 0.001$.

III. Static Operating Modes of the Converter

For the system shown in Fig. 1, in the initial analysis, we compose a graph (Fig. 2, a) covering the static relationships between the variables. Here $R_{eq} = R \parallel (-R_{diff0})$, R_Σ - includes all active losses in the circuit.

The stabilization of the output current, based on control only by the deviation from the set value, for traditional regulators does not give the desired quality. Therefore, for example, it is possible to compensate for the component of the deviation caused by a change in the supply voltage by implementing control over the disturbing effect. To do this, it is enough 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 gain of the converter with respect to the constant component remains constant when U_1 changes.

Another and more dangerous disturbing effect is a EMF change of the load U_0. The absolute invariance condition of even a linearized stabilization system to the disturbing action U_0 cannot be practically realized, because due to the non-minimum phase properties of the system [3, 4], the coupling circuit is unstable due to the disturbing effect. However, it is quite possible to exclude the influence of u_0 on i_{Load} in the static mode ($\tilde{u}_0 = \text{const}$) and in the absence of an NFB loop ($K(s) = 0$). It's enough to accept

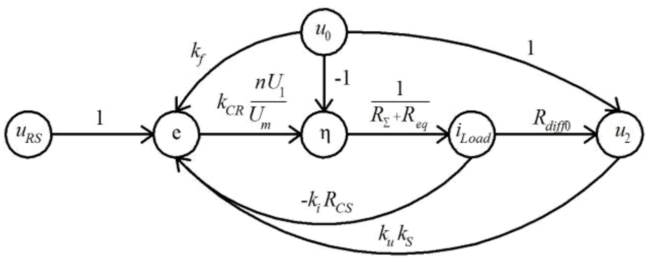
$$k_f = U_m / (k_{CR} n U_1) .$$

Indeed, in this case we obtain a zero result at the vertex η under the action of two channels, which ensures $\tilde{i}_{Load} = 0$ for $\tilde{u}_0 = \text{const}$ in the steady state. In this case, the control action is applied to the converter immediately at the time of the disturbance.

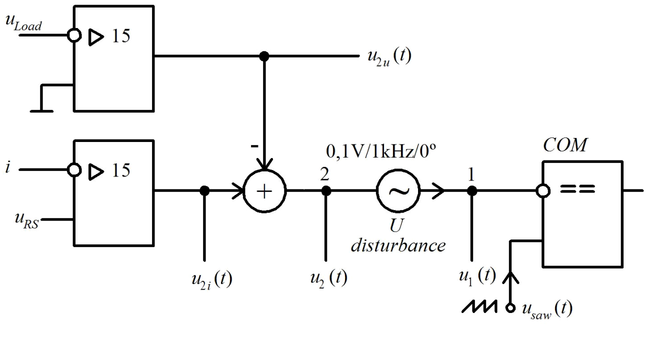
The maximum depth (theoretical limit) of the NFB of a converter with a PWM, taking into account the nonlinear properties of the latter in a given frequency band, can be determined from the expression [5]:

$$A_{0\max} = 6\ln 2 - 2\ln\frac{\omega_{cut}}{\omega_0} - 2\ln\pi - 2 , \qquad (1)$$

where $A_0 = A_{0\max}$, $0 \le \omega \le \omega_{cut}$ is the maximum depth of the NFB, constant in the working range; ω_{cut} is the cutoff frequency of the correction circuit; ω_0 is the PWM clock frequency.



a



b

Fig. 2. The graph of the equivalent circuit a closed-loop control system (a) and a system with output and voltage control (b)

Since in the case under consideration $\omega_{cut} / \omega_0 = 0.25$, in accordance with (1) we have

$$A_{0\max} \cong 23 \text{ dB}.$$

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

$$k_{CR.\lim} < \frac{L_2 / \left|R_{eq}\right| - C_2 R_\Sigma}{k_i R_{CS} C_2} \cdot \frac{U_m}{n U_1} = 16 \text{ (24 dB)},$$

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

The experimental value of the boundary gain factor obtained on the simulation model of a physical object is 30 (29.5 dB). At $k_{CR} > k_{CR.\lim}$ subharmonic half-frequency self-oscillations are established, or further the process is accompanied by significantly increased pulsations of a chaotic nature.

Accepting a minimum margin of stability of 6 dB, we obtain the maximum possible gain of the regulator in deviation

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

IV. Calculation of Frequency Characteristics Through Analysis in the Time Domain

All the advantages of this approach are manifested in the analysis of nonlinear systems, including pulsed ones [3, 4]. Calculation of frequency characteristics (FC) through analysis in the time domain does not require any changes to the studied system and takes into account real processes occurring in a nonlinear (pulse) circuit. To obtain reliable results using this method in FASTMEAN, two procedures are required.

Let it be required to obtain the frequency response of the input-output transfer coefficient for the current stabilization system under consideration (Fig. 1). Construction of the frequency dependence of the transfer function (TF) of a closed stabilization system with a disturbance from the input voltage

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

necessary for evaluating the smoothing properties of a pulsed current stabilizer.

First, we calculate the transient process for the output current (Fig. 3 a). From fig. 3 a, it can be seen that at the end of the 16th period, the transition process enters the linear zone and, starting from the 17th period, a steady-state operation mode is observed.

From the analysis of the curve (Fig. 3 a) it follows that the steady-state process corresponds to a stable single-cycle mode of the converter. The amplitude of the pulsations of a clock frequency of 52 kHz is 4 A. The starting process, i.e. the process of establishing the output current is a monotonous curve without surges and oscillations and with an accuracy of 4% error ends in a time less than 0.31 ms.

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

The output of an object excited by such an effect will change according to the law

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

where the function $A_c(\omega)$ is the amplitude-frequency characteristic of a closed system. Setting the generator to different frequencies, we get the frequency response and phase response of a closed system, combined into the main complex transfer function:

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

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

$$k_{sm} = I_{out0}U_{in\sim}/(U_{in0}I_{out\sim}),$$

where $U_{in\sim}$, $I_{out\sim}$ - is the amplitude of the variable component, respectively, of the voltage at the input and current at the output of the stabilizer. The value of k_{sm} is determined for the sinusoidal form of ripple of the input voltage [2].

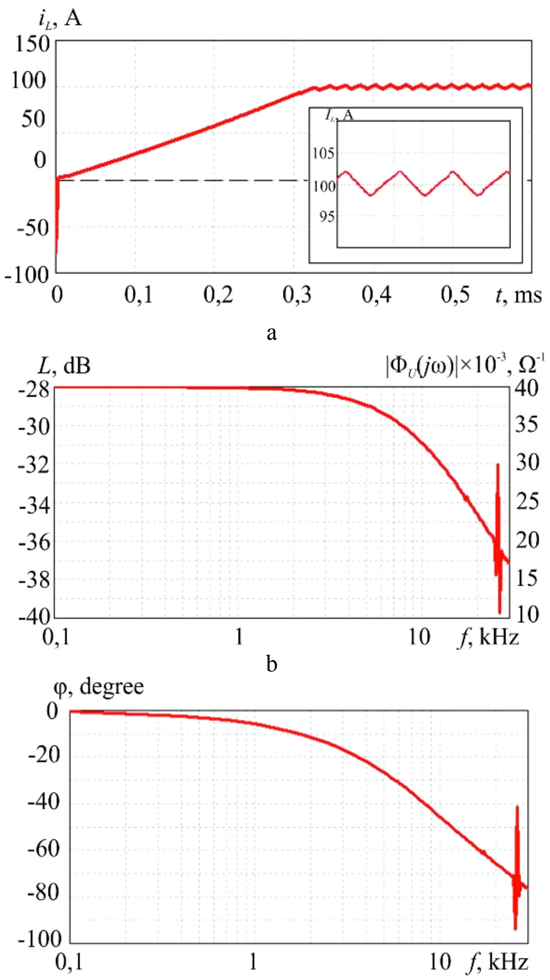


Fig. 3. Oscillograms of the transient and steady-state conditions for the current in the load when working out the set-point jump U_{RS} = 1 V from zero initial conditions (a) and the frequency dependence for the exact pulse model of the converter with output-only control (b, c)

For assess the smoothing properties of the considered PCS, it is necessary to have the frequency dependence of the input-output transfer coefficient for a closed-loop system with PWM, i.e. characteristic of attenuation of input low-frequency pulsations q ($q = 1/k_{sm}$).

Such a calculated frequency dependence for an accurate pulse model of a converter with only output control, which shows how its output reacts to a change in input voltage, is shown in Fig. 3, b, c. Frequency response shown in Fig. 3 b, c, are constructed for $k(s) = 15$ under the same parameters and operating conditions of the stabilizer. As follows from Fig. 3, b, c, the ordinate level of LAFC is less than zero not only at low frequencies corresponding to long transition times, i.e. the converter in question has a wide range of suppression of ripples of the input voltage, and the absence of a resonant peak indicates the monotonicity of the transient process.

V. Features of the Calculation EFC of Loop Gain of Switching Devices

To determine the transfer coefficient of the loop amplification PCS while maintaining the nature of the processes occurring in the system with NFB, we use the “closed loop” method [3, 4], known in foreign measurement practice as a means of maintaining the operating point by direct current. Analysis of the AFC and PFC of the transmission function of the NFB loop (loop gain) allows us to evaluate the achieved NFB depth, stability margins, stability of the main system indicators, and the degree of suppression of distortion and interference.

In the general case, the frequency characteristics of the loop gain function can adequately reflect the stability conditions and the possible generation modes of a PWM system provided that they are measured in a normally functioning system, i.e. with closed feedback. This method is based on the introduction of a functioning closed-loop system of harmonic disturbance into the ring of the FB and the subsequent measurement of the steady-state reaction to this disturbance in a wide frequency range. To ensure high measurement accuracy, the “injection” of the disturbance signal must be chosen between the nodes of the circuit having a significantly different input Z_{in} and output resistance Z_{out}. Of the two possible methods, preference is given to the method in which a perturbation source is introduced between the error signal amplifier and the comparator (Fig. 2 b), since it is this method that allows one to find the total loop gain function in the presence of several FB circuits in the system.

The loop gain was calculated using the formulas:

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

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

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

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. 4 b.

The equivalent PFC (Fig. 4 a) differs significantly in the high-frequency region from the similar characteristic obtained for the linearized converter model. EFC reveals a nonlinear resonance in the vicinity of the second subharmonic ($f_T/2$), the maximum emissions of which reach 10 dB in amplitude and 50-60º in phase. An analysis of the characteristics in this frequency range allows us to conclude that the generation conditions can be fulfilled with a further increase in the loop gain module.

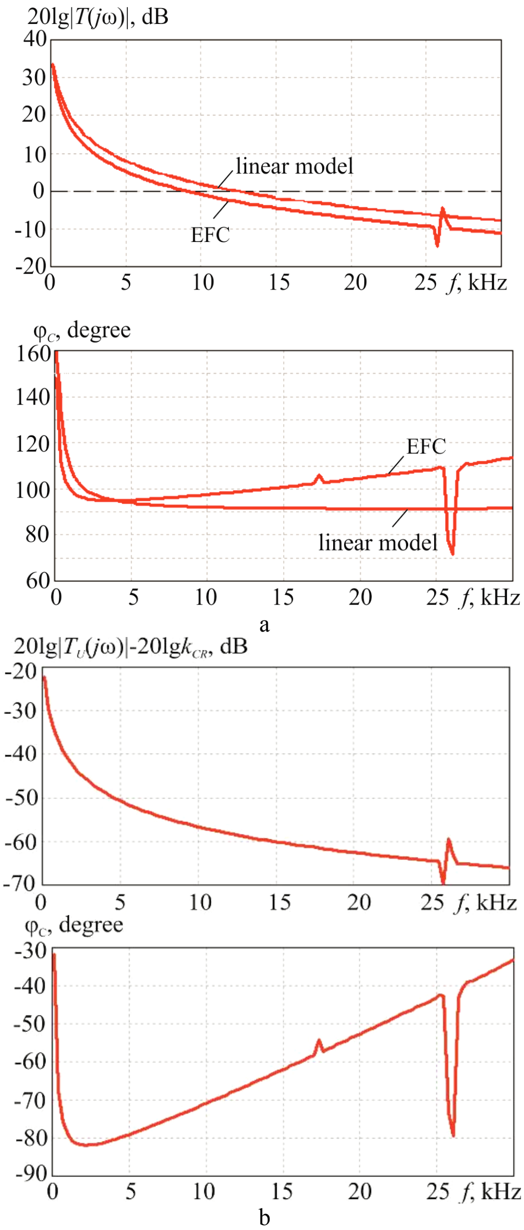


Fig. 4. Frequency dependences for the pulse model of the converter (a) and calculated frequency dependences (b)

VI. Assessment of the Dynamic Resistance of a Switching Power Supply

Of all the dynamic parameters of voltage converters, the most important is its output resistance [2, 4].

The output resistance of the SPS with open circuits of the FB (Fig. 5 a)

$$Z_{00}(s)=\frac{\widetilde{U}_{Load}(s)}{\widetilde{i}_{Load}(s)}=$$

$$=\frac{R_\Sigma+sL_2}{L_2C_2s^2+\left(R_\Sigma C_2+\dfrac{L_2}{R_{diff\,0}}\right)s+1+R_\Sigma/R_{diff\,0}}$$

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

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

where T_V is the loop gain of the FB circuit loop in voltage, numerically equal to the TF of the open-source system of SPS. In the output impedance curve of a pulsed power supply, in the general case, there is a resonant maximum at the resonant frequency of the output filter

$$\omega_0=1/\sqrt{LC} \qquad (f_0\approx 13\text{ kHz}).$$

Due to the action of the FB on the output voltage, the resulting resistance increases significantly. As follows from a comparison of the results (Fig. 5 b), 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 the absence of such a connection. In this case, $Z_0(j\omega)$ is flatter, therefore, the frequency dependence $Z_0(j\omega)$ for a two-loop control system is more favorable with respect to dynamics than for the case of applying only one feedback on the output current.

The input resistance of the PCS (Fig. 5 c) without an input filter with open FB, loaded on the load resistance R_{diff0}:

$$Z_{i0}(s)=\left.\frac{\widetilde{U}_{in}(s)}{\widetilde{I}(s)}\right|_{\widetilde{d}(s)=0}=\frac{R_{eq}+R_\Sigma}{(nD)^2}\times$$

$$\times\frac{1+s\left[\dfrac{1}{R_{eq}+R_\Sigma}+\left(R_{eq}\|R_\Sigma\right)C_2\right]+s^2L_2C_2\dfrac{R_{eq}}{R_{eq}+R_\Sigma}}{1+sR_{eq}C_2},$$

and the input resistance of the device (Fig. 1), for example, with closed current feedback, is defined as

$$Z_\text{i}(s)=Z_{i0}(s)(1+T_i(s)),$$

where $R_{eq} = R_4 \,||\, R_{diff0}$; $T_i(s)$ is the loop gain circuit FB of current, which is equal to the TF of the open system PCS $T_i(s) = W_r(s)$.

Consider the frequency dependence of the phase of the input resistance of a closed converter. From Fig. 5 c, it can be seen that at low compared with the clock frequencies ($f/f_\text{T} < 0.01$), the input resistance phase is close to -180º. At $f/f_\text{T} > 0.01$, a smooth phase change occurs. Thus, in the frequency range $0.001 < f/f_\text{T} < 0.5$, the phase of the input resistance of the converter arg $Z_{in}(j\omega)$ < -90º.

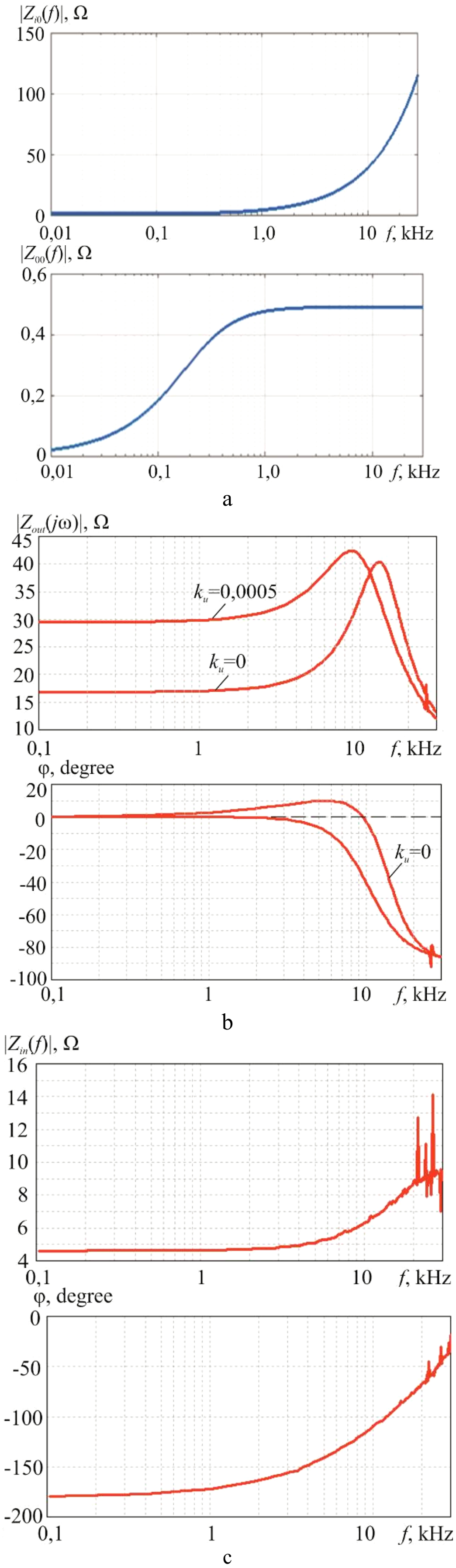


Fig. 5. The dynamic resistance of a switching power supply: a - output impedance with open circuits of the FB; b - output impedance with a closed circuit FB; c - input resistance without an input filter with open FB loaded with load resistance

As follows from the theory of circuits [4], in the indicated frequency range, the real part of the complex input resistance $U(\omega) = \mathrm{Re}\, Z_{in}(j\omega) < 0$, i.e. the active component of the input resistance is negative. It should be noted that it is in this frequency range that instability is possible, for example, a network filter - converter system and a leading source - slave system [3, 4].

VII. Conclusions

1. It is the EFC loop frequency response that describes the relationship of the frequency properties with the dynamics of voltage converters with PWM and reliably determines the real stability margins, predicts the generation modes and opens up the possibility of obtaining the maximum NFB in a given frequency band of switching devices.

2. The non-minimum phase properties of the converter significantly complicate the achievement of a high quality stabilization of the PCS output current when using only the deviation control principle. A successful solution to the problem gives the use of the principle of combined control.

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 clock subharmonics.

References

[1] E.N. Vereshchago, I.F. Fel'dsher, V.I. Kostiuchenko, “Quasiresonant inverters in power supply devices for air plasma cutting”, Technical Electrodynamics, Power Electronics and Energy Efficiency Issue, 2007, ch. 4, pp. 8-11. (Rus).

[2] U. Nasir, M. Rivera, S. Toledo, A. Costabeber, P. Wheeler, A review of power converter topologies with medium/high frequency transformers for grid interconnection systems. 2016 IEEE International Conference on Automatica (ICA-ACCA). doi: 10.1109/ica-acca.2016.7778471.

[3] A.G. Aleksandrov, Frequency theory of automatic control (frequency control). Book 1. Elektrostal': EPI MISiS Publ., 2010. 320 p. (Rus).

[4] V.P. Bakalov, V.F. Dmitrikov, B.I. Kruk, Fundamentals of circuit theory: a textbook for universities. Moscow: Goriachaia liniia – Telekom Publ., 2007. 597 p. (Rus).

[5] V.N. Sidorets, I.V. Pentegov, Deterministic chaos in nonlinear circuits with an electric arc. Kiev: Mezhdunarodnaia assotsiatsiia «Svarka» Publ., 2013. 272 p. (Rus).

[6] E.N. Vereschago, V.I. Kostiuchenko, Sensitivity of characteristics of model of the electric arc to change of parameters of elements. Electrical engineering and power engineering, 2019, no 1, pp. 22-31. doi: 10.15588/1607-6761-2019-1-2. (Rus).

[7] E.N. Vereshhago, V.F. Kvasnickij, L.N. Miroshnichenko, I.V. Pentegov, Circuitry of inverter power sources for arc load. Nikolaev: UGMTU Publ., 2000. 283 p. (Rus).

[8] F. Bordry, Power converters: definitions, classification and converter topologies. In Proc. CAS–CERN Accelerator School and CLRC Daresbury Laboratory: Specialized CAS Courseon Power Converters, Warrington, UK, 2004, p. 13-41. DOI:10.5170/CERN-2006-010.13.

[9] G. Zulauf, Z. Tong, J.D. Plummer, J.M. Rivas Davila, Active Power Device Selection in High- and Very-High-Frequency Power Converters. 2018. IEEE Transactions on Power Electronics, 1–1. doi:10.1109/tpel.2018.2874420.

[10] M.M. Jovanovic, Y. Jang. State-of-the-art, single-phase, active power factor correction techniques for high-power applications-an overview. IEEE Trans. Ind. Electron., 2005, vol. 52, no. 3, pp. 701-708.