ARPN Journal of Engineering and Applied Sciences

www.arpnjournals.com

ANALOG AND DISCRETE-ANALOG MATHEMATICAL MODELS OF DOWN DC-DC-CONVERTERS

D. Yu. Denisenko, M. E. Denisenko, Yu. I. Ivanov, V. V. Ignatyev, V. I. Finaev and O. B. Spiridonov
Southern Federal University, Rostov-on-Don, Russia
E-Mail: iskobersi@gmail.com

ABSTRACT

This article contains the results of modelling and researching the parameterization of DC-DC-converters, in particular, a converter output stage as a controlled object of the automatic control system. This research is focused on deriving the mathematical models of DC-DC-converters and their analysis. The operational analysis of a down DC-DC-converter frequency domain was made. The main part of a converter is a low pass filter (LPF). Unipolar pulse sequence is supplied in the LPF input. Output stage parameters are chosen by analyzing the signal spectrum components and the analytic form of a LPF transfer function. The research was made on the basis of various filter circuits: RC-circuit and RLC-circuit. There also made an operational analysis of a down DC-DC-converter in the time domain. Electronic switches were used for deriving a voltage source. Electronic switches are controlled by two pulse patterns. The research deals with organizing control modes of electronic switches. Difference equation systems, which describe a DC-DC converter operation, are given. This converter was modelled by means of the Miro-Cap circuit analysis program. The derived voltage and current waveforms of a DC-DC-converter with diode and transistor switches are given.

Keywords: DC-DC-converter, undervoltage, modeling, transfer function, low pass filter, electronic switch, regulating characteristic, mutual characteristic.

INTRODUCTION

DC-DC-converters modelling ("DC" -direct current) deals with a solution to the problem about choosing the parameter parts of a converter output stage [1], and the parameters of feedback circuits. Feedback circuits provide the defined transient performances and marginal stability. Any DC-DC-converter is an autostabilizer and a regulation [3, 4, 6, 11-13], which has a definite marginal stability. If this system has a big marginal stability, it solves the problem of electrical value transformation exactly. For automatical stabilization and regulation, it is important to know what marginal stability it possesses. Therefore, the high-power output stage of a DC-DC-converter is considered as a controlled object in the automatic control system [1, 2, 14]. When designing the automatic control system, the problem of object parameter identification is solved. If solving the problem of object parameter identification allows defining the coefficients of its transfer function exactly enough, one can also choose reactor parameters exactly enough. Solving this problem successfully allows providing the defined quality indices of the automatic control system with enough high fidelity, which is a vital task. A solution to the problem of analyzing the direct voltage pulse converter is given in this article. Also, its analog and discrete transfer functions were obtained [5, 9, 10].

THE ANALYSIS OF A DOWN DC-DC-CONVERTER FREQUENCY DOMAIN

We shall present the block-scheme of a converter in the form of two series -connected blocks as it is shown in Figure-1. A down DC-DC-converter consists of voltage source *e(t)* and low pass analog filter – LPF [7, 8, 10]. Voltage source generates an output voltage, which is applied in the LPF input in the form of unipolar pulse sequence with the amplitude$E$, theinterval τandtheperiod T as it is shown in Figure-2.

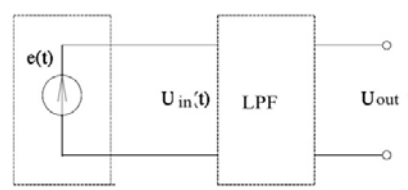


Figure-1. The block-scheme of a down DC/DC-converter.

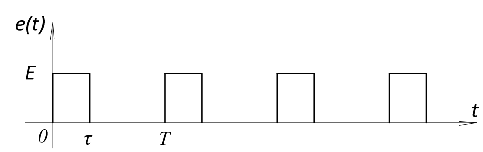


Figure-2. Voltage waveform in the LPF output.

The model of rectangular waves is analytically defined in Fourier's series [15, 16]:

$$e(t) = E\frac{\tau}{T} + E\frac{2\tau}{T}\sum_{k=1}^{\infty}\frac{\sin\left(k\frac{\pi\tau}{T}\right)}{k\frac{\pi\tau}{T}}\cos\left(k\frac{\pi\tau}{T}t\right) = Eq + 2qE\sum_{k=1}^{\infty}\frac{\sin(k\pi q)}{k\pi q}\cos(k\omega_1 t) \quad (1)$$

where $q=\tau/T$ is a pulse duty factor; $Q=1/q$ is a relative pulse duration; $\omega_l=2\pi/T$ is a fundamental frequency.

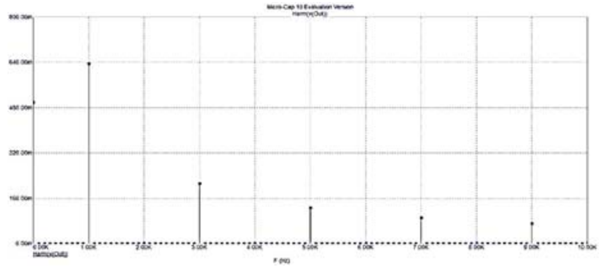
The analysis of this formula (1) shows that the signal spectrum contains the constant component $E\frac{\tau}{T}$ and

ISSN 1819-6608
ARPN Journal of Engineering and Applied Sciences

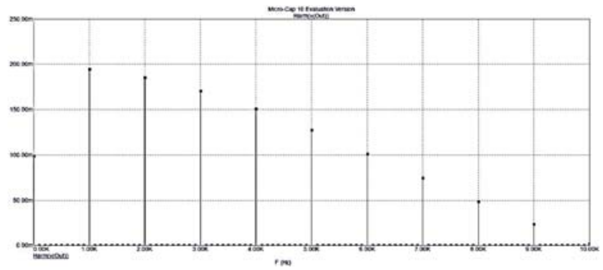
www.arpnjournals.com

also the infinite number of harmonic components. The fundamental frequency ω_1.

As an example there shown spectra of unipolar pulse sequence with the amplitude E=*1 B*, the fundamental frequency *f=1000 Hz*, the relative pulse duration *Q=2* (Figure-3a) and the relative pulse duration *Q=10* (Figure-3b) in Figure-3. It was modelled in the Micro-Cap.



(a)



(b)

Figure-3. Unipolar pulse sequence spectra with the relative pulse duration *Q=2* (a) and *Q=10* (b).

The analysis of the pulse spectra shows that the signal constant component changes in inverse proportion to the change of the relative pulse duration. There observed the growth of higher harmonic amplitudes with the growth of the relative pulse duration in the signal spectrum. If all harmonic components are taken away out of this signal spectrum, there is a constant component left. Such work can be performed by using the LPF. The LPF cutoff must be much less than the fundamental frequency. The attenuation of its amplitude-frequency response on a fundamental frequency must be sufficient for applying a small level of output voltage ripple at the output of a DC-DC-converter. The simplest circuit of the LPF is a first order circuit, based on the integrating the RC-circuit [17, 18], shown in Figure-4.

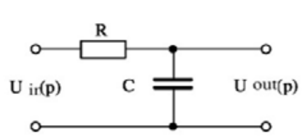


Figure-4. The LPF of the first order.

AFR (amplitude-frequency response) and LPF behavior depends on a choice (task) of its transfer function coefficients. The circuit transfer function, given in Figure-4, is defined by the following formula:

$$F(p) = \frac{U_{out}(p)}{U_{in}(p)} = \frac{1}{pRC+1} = \frac{\frac{1}{RC}}{p+\frac{1}{RC}} = \frac{\omega_p}{p+\omega_p}, \qquad (2)$$

where ω_P is a pole frequency.

The AFR type of the considered LPF is defined by the only transfer function coefficient that determines a cutoff. For the transfer function of the first order a cutoff coincides with a pole frequency. Let us define 5 Hz cutoff and choose 100 ufd (microfarad) capacitance value and find the resistor resistance value from the following formula:

$$R = \frac{1}{2\pi f_p C} = 328 \text{ Ohm}. \qquad (3)$$

The AFR of the LPF with the chosen part parameters is shown in Figure-5. The AFR attenuation on a cutoff of this LPF comes to 3dB, on a frequency of 1 kHz comes to 46 dB. This neutralization of the fundamental frequency can turn out to be insufficient for applying a small level of pulses at the output of a DC-DC-converter, so let us decrease the LPF cutoff by 1 Hz, for instance, increasing a capacitance by 500 ufd. The AFR with the new filter parameters is shown in Figure-6.

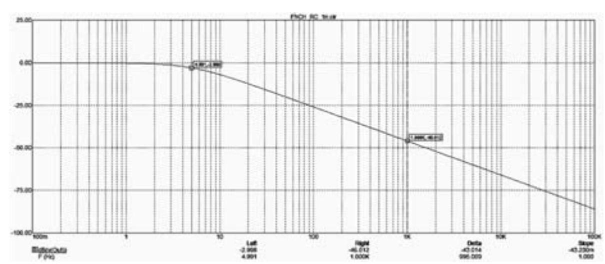


Figure-5. The AFR's low pass filter of the first order with a cutoff, equal to 5 Hz.

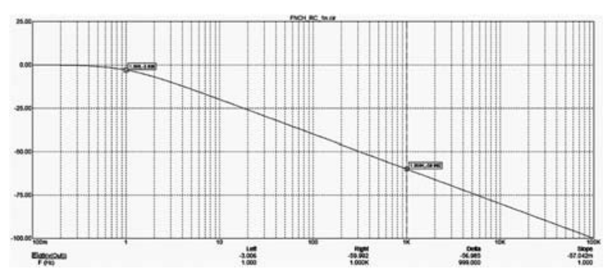


Figure-6. The AFR low pass filter of the first order with a cutoff, equal to 1Hz.

The attenuation of the derived AFR filter on a frequency of 1 kHz comes to 60 dB that corresponds to the fundamental frequency neutralization of the input pulse pattern more than 1000 times.

VOL. 12, NO. 2, JANUARY 2017 ISSN 1819-6608
ARPN Journal of Engineering and Applied Sciences

www.arpnjournals.com

If to apply a voltage with the form, shown in Figure 2 at the inlet of the LPF, the voltage that is equal to the constant component, in proportion to the relative pulse duration, as it is shown in Figure-7, will be adjusted in its output some time later. Figure-7 also shows the level of ripples in voltage output, whose value is defined by the final level of attenuating the AFR's low pass filter. Voltage ripples amplitude in the filter output, measured "peak" to "peak", made approximately 0,5 mV (millivolt) with 1B input pulse and 10 pulse ratio amplitude.

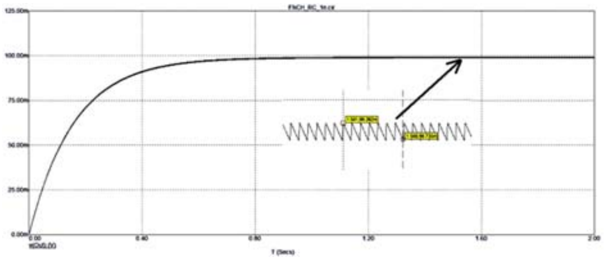


Figure-7. Adjusting voltage output in the LPF output.

DC/DC-converter must work for an effective load, so it is necessary to connect resistance to the LPF output as a load (Figure-8).

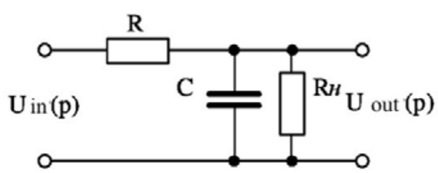


Figure-8. The LPF with the resistance load *Rн*.

The LPF transfer function with a resistance load is defined by the following formula:

$$F(p)=\frac{U_{out}(p)}{U_{in}(p)}=\frac{\frac{1}{RC}}{p+\frac{1}{RC}+\frac{1}{R_H C}}=\frac{\omega_p}{p+\omega_p}=$$

$$=\frac{R_H}{R+R_H}\frac{\frac{1}{C}\frac{R+R_H}{RR_H}}{p+\frac{1}{C}\frac{R+R_H}{RR_H}}=M\frac{\omega_p}{p+\omega_p} \quad (4)$$

where $M=\frac{R_н}{R+R_н}$ is a transfer constant of the LPF on DC, $\omega_p=\frac{1}{C}\frac{R+R_н}{RR_н}$ is the LPF's pole frequency.

The transfer constant was changed to a zero frequency under the load resistance in the filter that is an undesirable factor, as it leads to stepping down the voltage output of a DC/DC - converter. The maximum possible voltage output is defined by the following formula:

$$U_{\text{out max}}=E\frac{R_н}{R_н+R}, \quad (5)$$

where E is voltage amplitude of input pulses.

The pole frequency is changed in the filter under the load resistance. As load resistance reduces, pole frequency increases. Another disadvantage of using the considered filter in a DC-DC-converter is a low efficiency output (EO), as *R* (resistance) is used in the filter on which power dissipates. One had better use RLC-circuit as a low pass filter (Figure-9).

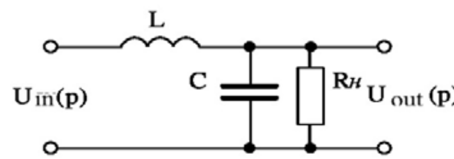


Figure-9. LPF circuit on a RLC-circuit basis.

The LPF transfer function on a *RLC*-circuit basis is defined by the following formula:

$$F(p)=\frac{U_{out}(p)}{U_{in}(p)}=\frac{\frac{1}{LC}}{p^2+p\frac{1}{LR_н}+\frac{1}{LC}}=M\frac{\omega_p^2}{p^2+pd_p\omega_p+\omega_p^2} \quad (6)$$

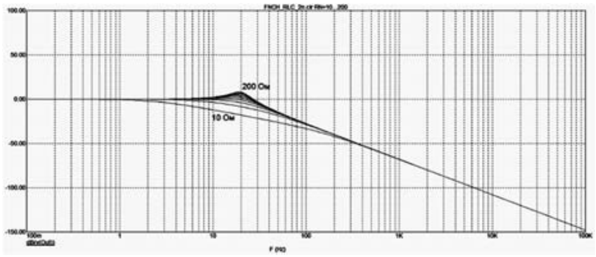
where $\omega_p=\sqrt{\frac{1}{LC}}$ - LPF pole frequency; $d_p=\frac{1}{R_н}\sqrt{\frac{C}{L}}$ - pole attenuation, M=1 - filter gain on a zero frequency.

The analysis of the *RLC*-circuit transfer function (6) shows that the transfer constant on a zero frequency does not depend on part parameters and is always equal to 1 that is important for creating a DC/DC-converter. The analysis of the formula (6) showed that the transfer function had a second order and the filter pole did not depend on a load resistance. When varying a load resistance, pole attenuation changes.

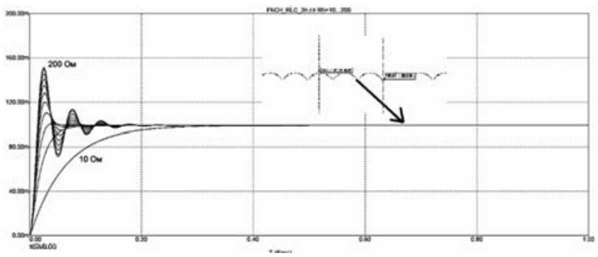
Figure-10 shows the family of the AFR low pass filter of the second order in varying a load resistance from 1 to 200 Ohm at a pitch of 20 Ohm, 20 Hz pole frequency, 100 ufd capacitance and the induction coefficient equal to

$$L=\frac{1}{(2\pi f_p)^2C}=0{,}634\text{ H} \quad (7)$$

VOL. 12, NO. 2, JANUARY 2017 ISSN 1819-6608
ARPN Journal of Engineering and Applied Sciences



(a)



(b)

Figure-10. The family of the AFR (a) and the LPF second order mutual characteristic (b) in varying a load resistance from 1 to 200 Ohm at a pitch of 20 Ohm.

The filter of the second order on a pole frequency of 20 Hz provides both approximately the same attenuation of the AFR on a frequency of 1 kHz and the filter of the first order on a pole frequency of 1 Hz. If to apply pulse pattern with a voltage waveform, as it is shown in Figure-2, many mutual characteristics will be obtained (Figure-10b).

A voltage level in steady-state at the output of the filter is defined by the relative pulse duration and their amplitude. Eruption value of the mutual characteristic when connected is inversely proportional to the pole attenuation that is also in inverse proportion to the load resistance. Voltage ripples amplitude in the filter of the second order output, shown in Figure 10b scaled-up and measured "peak-to-peak", made approximately 0,5 mV with 1B input pulse and 10 pulse ratio amplitude.

Filter output voltage in steady-state does not depend on a load resistance. It was determined by the fact that the filter gain on a zero frequency was equal to 1. However, it is just when the resistance of an induction coefficient is equal to 1.

THE ANALYSIS OF A DOWN DC-DC-CONVERTER IN THE TIME DOMAIN

Direct current is the voltage input of a DC-DC-converter. Electronic switches are used for deriving a voltage source with a voltage output in rectangle waves. For instance, connecting two electronic switches to a constant voltage source, which are controlled by the antiphased pulsed signals, one can receive one of the possible circuits of a down DC-DC-converter [1-5], shown in Figure-11.

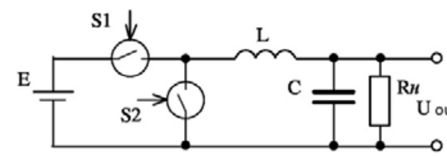


Figure-11. The circuit of a down pulse DC-DC-converter based on two electronic switches.

The two switches are controlled by two pulse patterns, shown in Figure-12.

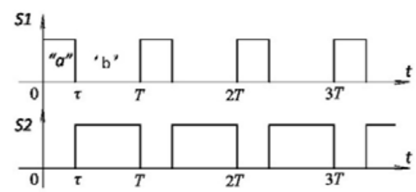


Figure-12. Time diagrams of the directing pulses of electric switches.

The regulating characteristic in the constant voltage pulse regulators is the dependence of a voltage output regulator on a ratio of directing pulses. The regulating characteristic is obtained for a DC-DC-converter from the formula (1), when all spectrum components are equal to 0, except for that component, not depending on the frequency:

$$Uout = \frac{\tau}{T} E = qE \tag{8}$$

Figure-13 shows the graph of the regulating characteristic of a down pulse DC-DC-converter.

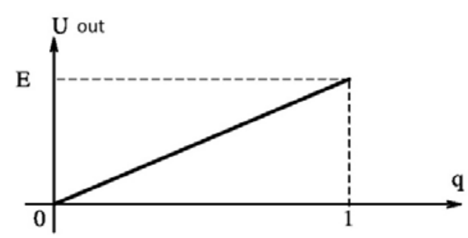


Figure-13. The regulating characteristic of a down DC-DC-converter.

The analysis of the regulating characteristic shows that at any value of pulse ratio ranging from 0 to 1, a converter output voltage cannot be more than the input voltage *E.* Therefore, it is called a down converter.

In real circuits of a down converter the regulating characteristic can have modifications from that given in Figure 1.13, as it was obtained for elements with idealized parameters by assuming that the resistance of electronic switches in the open position is equal to eternity and it is equal to 0 in the closed condition. Switches are on and off

VOL. 12, NO. 2, JANUARY 2017 ISSN 1819-6608
ARPN Journal of Engineering and Applied Sciences

www.arpnjournals.com

instantly, the resistance of an inductance coefficient is equal to zero and there are no losses in a condenser.

The regulating characteristic (8) of a down DC-DC-converter is found through a spectral representation of pulse input sequence. It is possible when the low-pass postfilter can be explicitly extracted in the circuit of a down converter. However, there is no opportunity to find the regulating characteristics for inverting boost converters, as their circuits do not contain low-pass filters. The regulating characteristic can also be found from the transfer function, according to the works [14, 16]. For finding the transfer function of the down converter (see Figure-11) we shall locate an input source in its input and consider the operation of the circuit for two switches.

According to Figure-11, the electronic switch *S1* is closed, and the one *S2* is opened in phase "a". For this phase of electronic switches wiring the down convertor circuit considering the initial conditions (capacitor voltage and inductance current in phase "a") is shown in Figure-14.

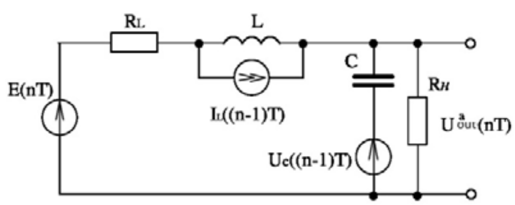


Figure-14. Circuit condition in phase "a".

The electronic switch *S2* is closedand the one *S1B* is opened in phase "b". Figure-15 shows circuit condition after transferring from the phase "a" to the phase "b":

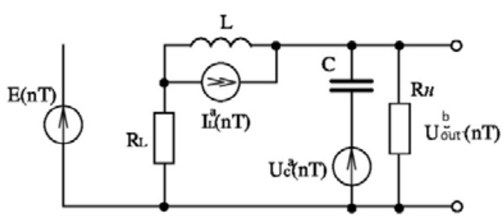


Figure-15. Circuit condition in phase "b".

The symbols are as follows in these figures: *n*- period number; $E(nT)$ - input voltage in *n*- period; $I_L((n-1)T)$ - inductance current at the end of the preceding period $(n-1)T$; $U_C((n-1)T)$ - capacitor voltage at the end of the preceding period; $U^a_{out}(nT)$ - circuit output voltage at the end of the commutating phase "a"; $U^{б}_{out}(nT)$ - circuit output voltage at the end of the commutating phase "b"; $I^a_L(nT)$ - inductance current at the end of the commutating phase "a"; $U^a_C(nT)$ - capacitor voltage current at the end of the commutating phase "a"; R_L- inductance loss resistance; R_H- load resistance.

The transfer function can be found by generating simultaneous equations, which connect current and voltage for each commutating phase. The condenser current $i_C(t)$ and its voltage drop $u_C(t)$, related by

$$u_C(t) = \frac{1}{T}\int_{t_0}^{\infty} i_C(t)dt + u_C(t_0) \tag{9}$$

where $u_c(t_0)$ - capacitor voltage at the moment of t_0, current flow (initial conditions).

In real circuits of voltage switching regulators the switching frequency of electronic switches is chosen rather high. It is usually ranged from 10 kHz and above; it can be higher than 1 MHz in modern voltage stabilizers. Therefore, the commutating period T turns out to be rather small. As the commutating period is small, one can consider that the condenser current does not change practically at the time of the circuit being in phase "a"; it remains constant. According to Equation (9) the capacitor voltage by the time of the phase "a" terminating will have been equal to

$$u^a_c(t) = \frac{1}{C}\int_{t_0}^{\tau} i^a_c(t)dt + u_C(0) = \frac{1}{C}\int_{t_0}^{\tau} I^a_c(t)dt + u_C(0) = \frac{\tau I^a_c}{C} + u_C(0) \tag{10}$$

The last expression can be written as a finite-difference equation:

$$U^a_C(nT) = \frac{\tau I^a_c(nT)}{C} + u_C((n-1)T) \tag{11}$$

The voltage applied to the inductance coil and current flowing in it, related by $u(t) = -e(t) = L\frac{di(t)}{dt}$. Solving this equation for the current and considering the initial current range in the coil before commutating, we find current by formula:

$$i_L(t) = \frac{1}{L}\int_{t_0}^{\infty} u_L(t)dt + i_L(t_0) \tag{12}$$

By analogy with the condenser, we shall consider that the voltage across inductance remains unchangeable practically at the time of the circuit being in the phase "a". The current in the coil by the time of the phase "a" terminating will have been equal to:

$$i^a_L(t) = \frac{1}{L}\int_0^{\tau} u^a_L(t)dt + i_L(0) = \frac{1}{L}\int_0^{\tau} U^a_L(t)dt + i_L(0) = \frac{\tau U^a_L}{L} + i_L(0) \tag{13}$$

Expression (13) can be written as a finite-difference equation.

 ISSN 1819-6608
ARPN Journal of Engineering and Applied Sciences

$$I_C^a(nT) = \frac{\tau U_c^a(nT)}{L} + I_L((n-1)T) \tag{14}$$

Considering the accepted symbols in Figure-3.40 and ratios (11) and (14), we shall find the full difference equation system, describing the circuit condition in phase "a"

$$\left.\begin{aligned}
&I_C^a(nT) = -U_C((n-1)T)\frac{1}{R_H} + I_L((n-1)T);\\
&U_C^a(nT) = U_C((n-1)T) + \frac{\tau}{C}\left[I_L((n-1)T) - U_C((n-1)T)\frac{1}{R_H}\right];\\
&U_C^a(nT) = E(nT) - U_C((n-1)T) + R_L I_L((n-1)T);\\
&I_C^a(nT) = I_L((n-1)T) + \frac{\tau}{L}\left[E(nT) - U_C((n-1)T) - R_L I_L((n-1)T)\right],
\end{aligned}\right\} \tag{15}$$

and in phase "b"

$$\left.\begin{aligned}
&I_C^б(nT) = -U_c^a(nT)\frac{1}{R_H} + I_L^a(nT);\\
&U_C^б(nT) = U_c^a(n)T + \frac{T-\tau}{C}\left[I_L^a(nT) - U_C^a(n)T\frac{1}{R_H}\right];\\
&U_L^б(nT) = -U_c^a(nT) - R_L I_L^a(nT);\\
&I_L^б(nT) = I_L^a(nT) + \frac{T-\tau}{L}\left[R_L I_L^a(nT) - U_C^a(nT)\right]
\end{aligned}\right\} \tag{16}$$

For simplifying records of equations in variables, depending on period number n (n=0, 1, 2, 3….), this period T is not usually written. Then the above-mentioned equations are given in the form:
- for the phase "a":

$$\left.\begin{aligned}
&I_C^a(n) = -U_C(n-1)\frac{1}{R_H} + I_L(n-1));\\
&U_C^a(n) = U_C(n-1) + \frac{\tau}{C}\left[I_L(n-1) - U_C(n-1)\frac{1}{R_H}\right];\\
&U_C^a(n) = E(n) - U_C(n-1)R_L I_L(n-1));\\
&I_C^a(n) = I_L(n-1) + \frac{\tau}{L}\left[E(n) - U_C(n-1) - R_L I_L(n-1)T\right],
\end{aligned}\right\} \tag{17}$$

- for the phase "b":

$$\left.\begin{aligned}
&I_C^б(n) = -U_c^a(n)\frac{1}{R_H} + I_L^a(n);\\
&U_C^б(n) = U_c^a(n) + \frac{T-\tau}{C}\left[I_L^a(n) - U_C^a(n)\frac{1}{R_H}\right];\\
&U_L^б(n) = -U_c^a(n) - R_L I_L^a(n);\\
&I_L^б(n) = I_L^a(n) + \frac{T-\tau}{L}\left[R_L I_L^a(n) - U_C^a(n)\right]
\end{aligned}\right\} \tag{18}$$

Solving the simultaneous equations (17) and (18) about input $E(n)$ and output voltage $U_{out}(n)$, and considering that the circuit voltage is equal to the capacitor voltage $U_{out}(n)=U_c(n)$, and neglecting the values of the second order of infinitesimals such as $\tau(T-\tau)\rightarrow 0$, we find the difference equation system:

$$\left.\begin{aligned}
&U_{out}(n) = U_C^б(n) = U_C(n-1) + \frac{\tau}{C}I_L(n-1) -\\
&-\frac{\tau}{CR_H}U_C(n-1) + \frac{T-\tau}{C}I_L(n-1);\\
&I_L^б(n) = I_L(n-1) + \frac{\tau}{L}E(n) - \frac{\tau}{L}U_C(n-1) - \frac{\tau}{L}R_L I_L(n-1) -\\
&-\frac{T-\tau}{L}R_L I_L(n-1) - \frac{T-\tau}{L}U_C(n-1).
\end{aligned}\right\} \tag{19}$$

Applying the cycling method to the finite-difference equations that defines a connection between variables in different periods (for instance, capacitor voltage at the end of the preceding $U_c(n$-1)) and succeeding periods $U_C^б(n)$ and considering the connection between the variables of the finite-difference equation and the Laplace discrete transformation, t.e.

$$\left.\begin{aligned}
&S(n) \Rightarrow S(z);\\
&S(n-1) \Rightarrow z^{-1}S(z);\\
&S(n-2) \Rightarrow z^{-2}S(z);\\
&\dots\dots\dots\ \dots\dots\dots\ \dots\dots\\
&S(n-m) \Rightarrow z^{-m}S(z),
\end{aligned}\right\} \tag{20}$$

we find

$$\left.\begin{aligned}
&U_{out}(z) = U_C^б(z) = z^{-1}U_C(z) + \frac{\tau}{C}z^{-1}I_L(z) -\\
&-\frac{\tau}{CR_H}z^{-1}U_C(z) + \frac{T-\tau}{C}z^{-1}I_L(z);\\
&I_L^б(z) = z^{-1}I_L(z) + \frac{\tau}{L}E(z) - \frac{\tau}{L}z^{-1}U_C(z) - \frac{\tau}{L}R_L z^{-1}I_L(z) -\\
&-\frac{T-\tau}{L}R_L z^{-1}I_L(z) - \frac{T-\tau}{L}z^{-1}U_C(z).
\end{aligned}\right\} \tag{21}$$

Solving equation (21) about input $E(z)$ and output voltage $U_{out}(z)$, we find the transfer function of a down converter

$$F(z) = \frac{U_{out}(z)}{E(z)} = \frac{z^{-1}\frac{T\tau}{CL}}{1-A+B}, \tag{22}$$

where

$$A = z^{-1}\left(1 - R_L\frac{T}{L} - \frac{\tau}{R_H C}\right),$$

$$B = z^{-2}\left(1 - R_L\frac{T}{L} - \frac{\tau}{R_H C} + \frac{T\tau}{R_H CL}R_L - \frac{T^2}{CL}\right).$$

The regulating characteristic of the converter is by its transfer function and is defined as a transfer constant

ARPN Journal of Engineering and Applied Sciences

on a zero frequency of a loop input signal, t.e. on a DC voltage.

Considering that the variable $z=e^{i\Theta}$, and $\Theta=\omega\Theta$ is a complex frequency on the zero ω=0, the variable z is equal to 1. Inserting this value in Formula (22), we find the transfer constant of a DC-DC-converter

$$F(1)=\frac{\frac{\tau}{T}}{1+\frac{\tau}{T}\frac{R_L}{R_H}} \quad (23)$$

If the constant voltage source E has been adjusted at the input of a converter, voltage is equal to q in its output according to Formula (23). If to neglect the impact of inductance coil resistance, we have the regulating characteristic from Formula (23)

$$Uout=\frac{\tau}{T}E=qE \quad (24)$$

that coincides with the earlier found expression (8) through the spectrum analysis.

At a charge coefficient equal to 1 (q=1), we have a maximum voltage at the output of a down converter from Formula (23)

$$U_{\text{out}}=E\frac{1}{1+\frac{R_L}{R_H}} \quad (25)$$

The analog circuit of a converter is shown in Figure-16. The inductance coil in each commuting phase of switches is either hooked into a voltage source or a global bus. An electric switch is on in each phase in series with inductance. If this switch has been realized on n-channel MOS transistor, it is easy to consider the impact of its resistance channel on the regulating characteristic.

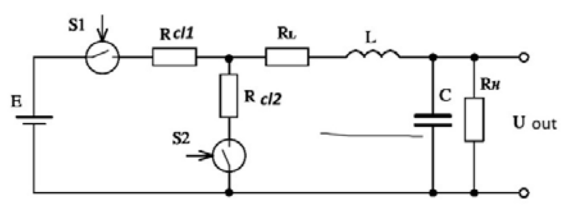


Figure-16. The analog circuit of a converter.

Considering the resistance channels of switches in the closed condition to be equal and symbolizing them as R_{cl} (cableline), we find

$$U_{\text{out}}=E\frac{q}{1+q\frac{R_L+R_{\text{CL}}}{R_H}} \quad (26)$$

It follows from Formula (25) that the inductance coil resistance and the final value of electronic switch resistance in the closed condition limit the maximum value of a converter output voltage.

In phase "b" the current is maintained by the condenser and by the coil inductance. The current in the coil inductance in this phase is opposite in sign. For the purpose of simplifying the circuit, in many cases diode is used as a switch instead the second transistor (Figure-17).

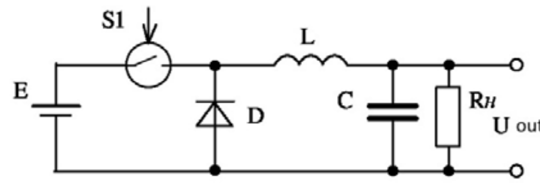


Figure-17. Converter circuit with a diode switch.

Using the diode in the circuit allows simplifying the control circuit of a converter, as it is enough to form one sequence of directing pulses for one switch. The efficiency output of a converter with a diode switch is lower than with a switch made on a fieldistor. It deals with rather a big loss of voltage across the diode (0,2 - 1V). The loss of efficiency is considerably shown in undervoltage converters.

If the duration of directing pulses changes, LPF parameters change too. Let us convert the discrete transfer function (22) to its analog equivalent for evaluating these changes. This transformation will be made by replacing the variable z in the transfer function. The variables of the Laplace discrete transformation z and the continuous transformation p, related by $z=e^{pT}$. Transforming e^p in the Laurent series

$$e^{-pT}=1-pT+\frac{(pT)^2}{2}-.... \quad (27)$$

and being limited to two series terms, we find

$$z^{-1}=e^{-pT}\approx 1-pT \quad (28)$$

Inserting Formula (28) in Formula (22) and neglecting the coefficients of the second order of infinitesimals, we find

$$F(p)=\frac{U_{\text{out}}(p)}{E(p)}=\frac{\frac{\tau}{T}\frac{1}{CL}}{p_2+p\left(\frac{R_L}{L}+\frac{\tau}{T}\frac{1}{R_H L}\right)+\frac{1}{CL}\left(1+\frac{\tau}{T}\frac{R_L}{R_H}\right)} \quad (29)$$

Under $T=\tau$ (the switch S1 in the circuit is being on all the while) and also the expressions of the transfer functions (6) and (29) fully coincide in the inductance coil resistance equal to 0 (R_L=0). By analogue with the RLC-circuit transfer function, we find the pole frequency.

ISSN 1819-6608
ARPN Journal of Engineering and Applied Sciences

www.arpnjournals.com

$$\omega_p = \sqrt{\frac{1}{CL}\left(1+\frac{\tau}{T}\frac{R_L}{R_H}\right)} \tag{30}$$

the pole attenuation

$$d_p = \left.\frac{\frac{R_L}{L}+\frac{\tau}{T}\frac{1}{R_H L}}{\sqrt{\frac{1}{CL}\left(1+\frac{\tau}{T}\frac{R_L}{R_H}\right)}}\right|_{R_l=0} \approx \frac{\tau}{T}\frac{1}{R_H}\sqrt{\frac{C}{L}} \tag{31}$$

and the transfer constant on a zero frequency

$$M = F(0) = \frac{\frac{\tau}{T}}{1+\frac{\tau}{T}\frac{R_L}{R_H}} = \frac{q}{1+q\frac{R_L}{R_H}} \tag{32}$$

The analysis of the last three formulas shows that all parameters of the filter depend on the charge coefficient q. The less the value of the coil resistance is, and the closer the value q to1 is, the less the deviations of these parameters from the circuit with ideal elements are.

While designing the circuit, it is difficult to consider all nonidealities of elements, as bulky analytical formulas are obtained. It is limited to finding the idealized characteristics for the purpose of studying the principles of a device operation. The nonidealities of those elements, which have the most significant impact on the output parameters of a device, are just considered in calculating formulas.

Schematic design programs allow studying the principle of the circuit operation and obtaining the characteristics, which are the closest to the real circuits. These programs allow identifying the deficiencies of calculations before modelling and wiring up, and eliminating them before modelling.

Figure-18 shows the circuit modeling of a converter in the Miro-Cap circuit analysis program [19].

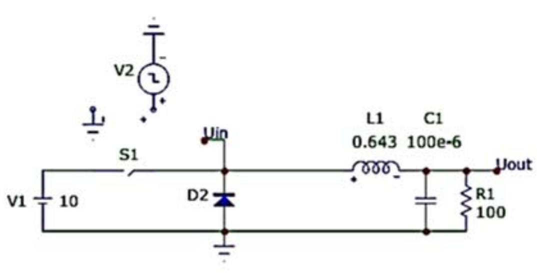


Figure-18. The circuit modeling of a down DC-DC-converter with a diode switch.

A voltage source is shown as the battery *V1* in the circuit (Figure-18). The sequence of directing pulses for the electronic switch *S1* is formed by the rectangular pulse source *V2* as a pulse pattern with the period *T=1ms* and the charge coefficient *q=0,5*. The coil resistance of the electronic switch in the model is given equal to 0,001 Ohm. For the purpose of reducing losses in the circuit, the Schottky diode was used as a diode.

The results of modelling the circuit are shown in Figure-19. The graphs of the diode voltage U_{in} (in the RLC-filter input), the converter output voltage U_{out} and the current inductance I_{L1} in this figure.

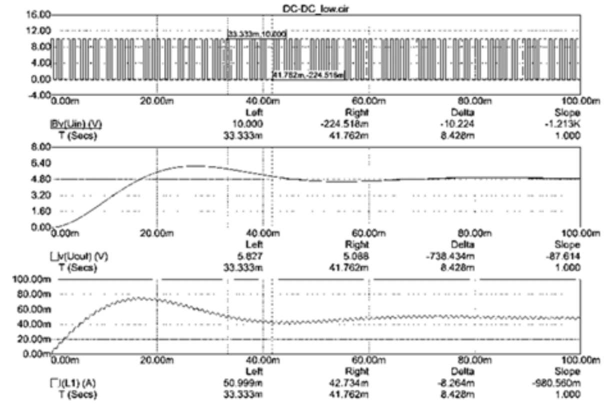


Figure-19. Voltage and current waveforms in the circuit of a down DC-DC-converter with a diode switch.

On completing the transient phenomenon, the calculated value of the output voltage under the chosen parameters must be 5V. The modeling showed that the value of the output voltage is adjusted at the level of 4,885 V. It deals with the fact that the diode was used as the second electronic switch and the voltage of 224 mV falls on it (see the graph in Figure-19), and the diode parameters were not considered in the formulas.

The nature of the current change in the inductance coil, the ripple amplitude and its flow are shown in the bottom graph (Figure-19).

The graph analysis of the output voltage shows that the circuit output voltage can exceed the final value greatly after switching on the circuit before completing the transient phenomena. Exceeding the set value of an output voltage can lead to the failure of the electronics, connected to a converter. The special circuits for “soft start”, which allow avoiding an overvoltage in the course of turning on the converter, are used for eliminating this effect in control circuits of stabilizers. Figure-20 gives the circuit modeling of a down DC-DC-converter with electronic switches transistorized. The sequence of directing pulses for the second switch *S2* is formed by the impulsive energy source *V3* in the circuit.

ARPN Journal of Engineering and Applied Sciences

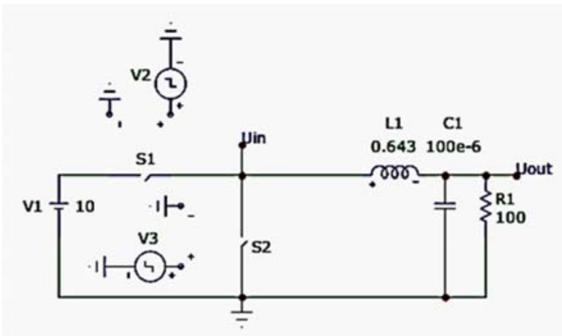


Figure-20. The circuit modeling of a down DC-DC-converter with transistor switches.

Figure-21 shows the results of modeling a down DC-DC-converter with transistor switches:

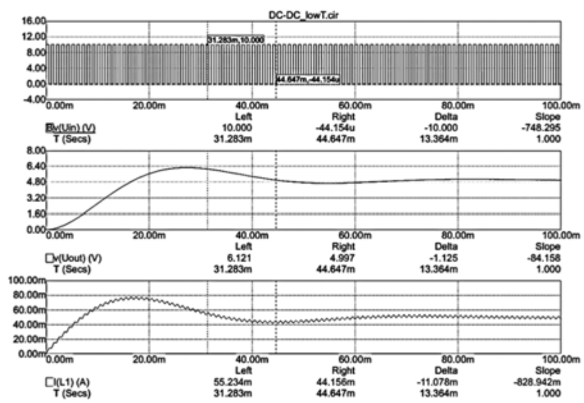


Figure-21. Voltage and current waveforms of a down DC-DC-converter with transistor switches.

An output voltage in the steady conditions comes up to 0,5 V, and the voltage U_{in} is equal to 44 мkV under the second switch being on in Figure-21.

If to use a channel resistance on-state voltage instead the diode in the controlled electronic switch of the fieldistor with a low value, one can reduce losses in a converter and upgrade its efficiency.

CONCLUSIONS

The made research and modeling results showed that when designing down DC-DC-converters, the vital task is providing a stability factor of the automatic stabilization and regulating system. The research of the article is aimed at substantiating the parametrization of a converter output stage. While analyzing the DC-DC-converter frequency-domain, it was shown that using the RLC-circuit ensured the efficient operation of the low pass filter. While analyzing the DC-DC- converter in the time domain it was shown that using the controlled electronic switch instead the diode on the fieldistor with the low value of a channel resistance on- state voltage, allowed reducing losses in a converter and upgrading its efficiency.

This article was prepared according to the research plan No.213.01-07-2014/02 SGC (Secondary Group Converter) Developing the methods of polyoptimizing the parameters of the hybrid adaptive intelligent controllers of the ill-formalized technical objects.

REFERENCES

[1] Meleshin, V.I. 2005. Transistor Converter Equipment. Moscow. Technosphere. p.632.

[2] Reddy, S.Rama. 2013. Fundamentals of Power Electronics, Город: Издательство. p. 230.

[3] Ivanov A.G., Belov G.A. and Sergeyev A.G. 2010. Control Systems of Semiconductor Converters. Cheboksary: Chuvash University Press. p. 448.

[4] Mack Raymond. 2005. Demystifying Switching Power Supplies, Munich-Amsterdam-London: Elsevier Science. p. 323.

[5] Meleshin V.I. and Ovchinnikov D.A. 2011. Transistor Power Converter Control. Moscow: Technosphere. p. 576.

[6] Rozanov Yu.K. 1992. Fundamentals of Power Electronics. Moscow:Energoavtomizdat. p. 296.

[7] Designing a TMS320F280x Based Digitally Controlled DC-DC Switching Power Supply. Application Report. Texas Instruments, Inc. 2005.

[8] Mohan N., Undeland T.M. and Robbins W.P. 2003. Power Electronics: Converters, Applications, and Design. Third Edition. John Wiley & Sons, Inc. p. 802.

[9] Perspectives in Control Engineering: Technologies, Applications, and New Directions (Edited by T.Samad). N.Y.: IEEE Press, 2001, pp. 502.

[10] Components for power sources URL: http://www.symmetron.ru/articles/brochures/SMPS.pdf (accessed date 12.08.16).

[11] Dorf R. and Bishop R. 2002. Modern Control Systems (translated from English by B.I. Kopylov). Moscow: Basic Knowledge Laboratory. p. 832.

[12] Gaiduk A.R. 2010. Automatic Control Theory: A text-book. Moscow: Vysshaya shkola. p. 415.

[13] Olsson G. and Piani G. 2001. Automation and Control Digital Systems. Saint Petersburg: Nevsky Dialect. p. 557.

ARPN Journal of Engineering and Applied Sciences

[14] Denisova G.V. and Ivanov Yu.I. 2013. Mathematical Model Development of a DC-DC-Converter, Information technologies, system analysis and control. ITSA&C-2013/ Collected Works of the XI Russian Scientific Conference for Young Researchers, Postgrads and Students. Taganrog: Southern Federal University Press. 2: 167-169.

[15] Gonorovky I.S. 2006. Radiotechnical circuits and signals: A university manual (5th edition). Moscow: Drofa p. 719.

[16] DenisenkoD.Yu., Ivanov Yu.I. and Finaev V.I. 2015. Fundamentals of Power Converter Equipment. Part I: A manual. Taganrog: Southern Federal University Press. p. 147.

[17] Lam Harry. 1982. Analog and Digital Filters. Design and Realization (translated from English). Moscow: Mir. p. 592.

[18] Denisenko D.Yu., Denisenko M.E., Ivanov Yu.I. and Finaev V.I. 2015. Analog and Discrete-Analog Retunable Integrators, Don Engineering Bulletin, No. 4. URL: ivdon.ru/ru/magazine/archive/n4y2015/3388 (accessed date 12.08.16).

[19] Micro-Cap Evaluation/Student Version URL: http://www.spectrum-soft.com/download.shtm (accessed date 11.08.16).