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Development of power supply devices for limitations of short circuit on the ship's hull

R S Fediuk and A M Cherneev
Far Eastern Federal University, 8, Sukhanova st., Vladivostok, 690091, Russia

E-mail: roman44@yandex.ru

Abstract. The authors of the paper have analysed the reasons and consequences of single-phase ground faults (hull faults). For all varieties of devices limiting the current single-phase ground faults, the most effective devices were found to be the arc-suppression coils with different switching circuits. In this case the measurement of circuit capacitance takes on a great importance. A number of variants of capacitance measurement is presented in the paper. The authors have had a detailed look at a device, limiting the single-phase short-circuit current. This device was developed on the basis of the Far Eastern Federal University under the direction of Dr. G.E. Kuvshinov. The device is provided with power supply that converts alternating current (AC) to direct current (DC), and is realised due to the use of semiconductor devices - transistors and diodes - in a bridge circuit. The technical outcome of this power supply application consists in the reduction of size and weight parameters (compared to the closest analogues) in order to connect the controlled voltage rectifier to the alternating voltage source, including mass and size of the capacitor bank of the current limiting circuit, and enhancing the dynamic parameters of the stage of uncontrolled charge of the output capacitor of the controlled voltage rectifier.

1. Introduction

Single-phase ground faults are a predominant type of damage in the power grid of medium voltage, which is 6-10 kV (75 to 90% of the total number of electrical damage). Single-phase ground faults are accompanied by an electric arc that can cause a fire and destroy the current conductors at a spot of short-circuit, which is why single-phase short circuit can be coverted into a two- or three-phase one [1-3].

The vast majority of violations of the normal operation of power grids with an isolated neutral wire is connected with local insulation damage. In such grids, even at high insulation resistance heavy current can pass through the short circuit (SC) spot. This occurs due to the large value of the total capacitance of the power grid with respect to the ground surface. The greater the single-phase short-circuit current, the greater the value of the mentioned total capacitance and the higher the voltage in the power grid [4-6].

To ensure the maximum possible reliability of the grid it is necessary that the short-circuit current was so small that for a sufficiently long time (the time required for search and elimination of the damage) it would be possible to dispensed with disconnection of consumers from the power grid. According to the rules of technical maintenance of power plants and grids, the triggering current capacities that do not require an immediate shutdown of such damaged connections are considered to be the currents: in the overhead networks of 6-20 kV on concrete or metal poles and in all networks at a voltage of 35 kV - 10 A; in the overhead networks without metal or concrete supports, at a voltage of

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6 kV - no more than 30 A, at a voltage of 10 kV - no more than 20 A, at a voltage of 15-20 kV - no more than 15 A. If the currents exceed the mentioned values, a compensation of the capacitive earth fault current is required.

The effective value of short-circuit current decreases rapidly up to a small steady-state value, if the neutral point of the power grid is grounded to earth through the arc suppression coil (ASC). The inductive reactance of ASC at a frequency of the voltage source, which supplies the electrical network, is equal to the resistance at this frequency and the total capacity of all three phases of the power grid relatively the earth [7-9].

This capacitance is often changed not only in case of a single-phase circuit, but also during changing of the grid configuration, as well as due to changes in the length of any line during manipulation when networking, or under the influence of temperature, icing, frequency oscillations and for other reasons. Therefore, the inductance of the arc suppression coil must be constantly maintained at a value that resonates with the capacity of the grid. Because of this, the coil is executed in two parts: an adjustable load bearing element and a control unit. The latter affects the load bearing element so that its inductance changed right after the change in the total capacity of the power grid [10-12].

The objective of this work is to develop such power source for the device which will limit the current of the single-phase short ground fault in the power grid.

2. Results and Discussions

In the Far Eastern Federal University under the guidance of prof. G.E. Kuvshinov a device that limits the current of the single-phase ground fault (hull fault) was invented [11, 13].

The electrical block diagram of device 1 for limiting the current of the single-phase ground fault in the power grid with floating neutral 2 is shown in Figure 1. Unit 3 represents the source of three-phase voltage. Unit 9 of capacitors containing capacitances emulates the capacitive susceptance of phases relatively the earth. Resistor R simulates the resistance of the circuit of the single-phase ground fault for phase C. An inductive component of device 1 is unit 4, which includes unit 6 - an electrical filter, unit 7 – a multiplier unit 8 - a voltage-controlled current source, unit 5 - control block.

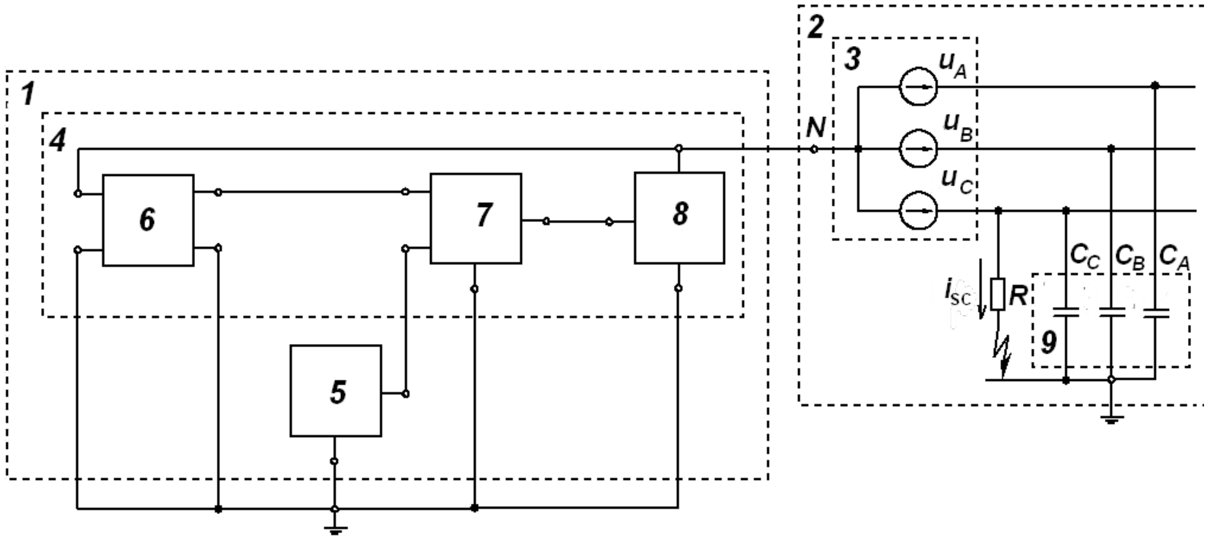

Figure 1. Schematic block diagram of a device limiting the current of the single phase short-circuit

When the SC occurs, voltage appears on the neutral wire of the power grid. This voltage enters the input of electric filter 6. A schematic diagram of the filter realised as a low-pass filter is shown in Figure 2.

The transfer function of the filter shown in Figure 2 has the form:

$$W_{RC}(S) = \frac{U_{F.OUT}}{U_N(S)} = \frac{1}{1+\tau \cdot s}, \tag{1}$$

where $\tau = R_F C_F$ - the time constant of the filter. This time constant is many times greater than the value, which inverse relatively angular frequency ω_1 of source 3. Therefore, at angular frequencies satisfying condition $\omega \geq \omega_1$, the amplitude and phase frequency characteristics of filter 6, which are obtained after substitution of $s=j\omega_1$ into the transfer function of Eq. 1, differ little from the similar characteristics of the ideal integrator with transfer function $1/(\tau \cdot s)$.

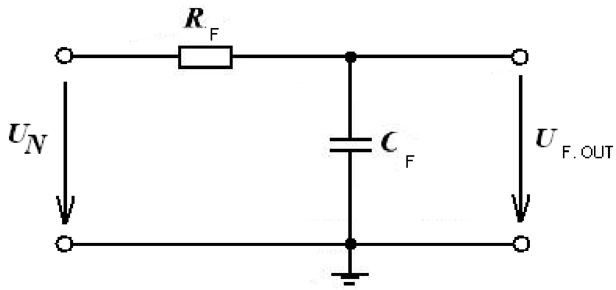

Figure 2. The low pass filter

The output voltage of the filter represented in a complex form:

$$\overline{U}_{F.OUT} = \frac{\overline{U}_N}{1 + j\omega_1\tau} \tag{2}$$

where $\overline{U}_N$ - the voltage on the neutral wire, ω_1- circular frequency.

The instantaneous value of the voltage is supplied to the second input of multiplier 7, on the first input terminal of which the required value of transmissive conductivity G_{TC} is served, which is controlled by current source voltage 8. In this case the output current of the current source is described by:

$$\overline{I}_{TC} = \overline{U}_{F.OUT} \cdot G_{TC} = \frac{\overline{U}_N \cdot G_{TC}}{1 + j\omega_1\tau} \ . \tag{3}$$

From Eq. 3 we can deduce the relation of $\overline{U}_N$ to $\overline{I}_{TC}$ that determines the complex impedance, which is equivalent to inductive component 4:

$$\underline{Z}_{IC} = \frac{1}{G_{TC}} + j\frac{\tau \cdot \omega_1}{G_{TC}} \ . \tag{4}$$

Imaginary component $\underline{Z}_{IC}$ is equivalent to the inductive reactance of the inductive component and is equal to:

$$X_{IC} = \frac{\tau \cdot \omega_1}{G_{TC}} \ . \tag{5}$$

The inductance of the inductive component is:

$$L_{IC} = \frac{\tau}{G_{TC}} \ . \tag{6}$$

Real component $\underline{Z}_{IC}$ is equivalent to the active resistance of the inductive component and is equal to:

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$$R_{IC} = \frac{1}{G_{TC}} \tag{7}$$

The Q-factor of the inductive component is:

$$q_{IC} = \frac{X_{IC}}{R_{IC}} = \frac{\omega_1 \cdot L_{IC}}{R_{IC}} = \omega_1 \cdot \tau \tag{8}$$

Q-factor q_{IC} of the inductance component exceeds the quality factor of the arc suppression coil greatly, since time constant τ can reach one or more seconds. Consequently, the quality factor of the inductive component is 6 or more times higher than the quality factor of the arc suppression coil.

When replacing inductive component 4 with equivalent complex impedance $\underline{Z}_{IC}$, consisting of resistance R_{IC} and inductance L_{IC}, the equivalent circuit (Figure 3) of the system with the earthed neutral wire through the inductive component results. The capacitor with a total capacity of three phases relatively ground (3C) is connected in parallel to the inductive component. The circuit consisting of two parallel connected branches (3C and $\underline{Z}_{IC}$), can be conFigured for current resonance.

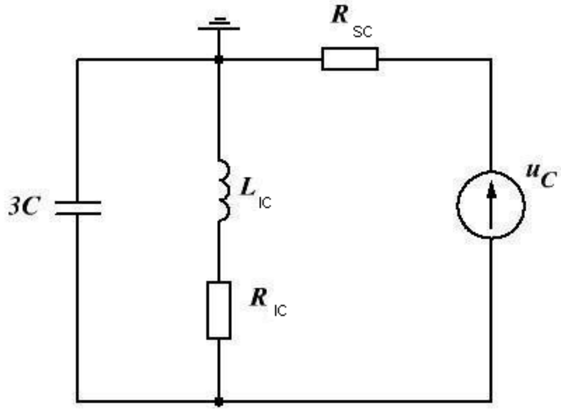

Figure 3. A scheme of replacement of system with the earthed neutral via an inductive component

Then the sum of currents of these branches becomes minimum. This resonance is achieved when susceptance of the indicated parallel connection is equal to zero, that is, condition $X_{IC} = X_{C\Sigma}$ has to be met. From it one derives a formula for the required value of the transfer conduction, which is controlled by the voltage of current source 8:

$$G_{TC} = \tau \cdot \omega_1^2 \cdot C_\Sigma, \tag{9}$$

where C_Σ - the total capacity of phases relatively the earth.

Before connecting inductive component 4 to power supply 2, it is necessary to set the value of transfer conductivity G_{TC} from control block 5, which depends, according to Eq. 9, on the total capacity of all phases relatively the earth. A corresponding signal about value G_{TC} from the output terminal of control block 5 is supplied to the first input terminal of multiplier 7.

Using the electron coverter-source of current it is possible to generate the current at the input marked by the bridge rectifier, which is sinusoidal and is in phase with the input voltage.

The choice of the electronic converter is based on the following assumptions:

- generally, electrical isolation between the input of the load and the output of the power electronic system is unnecessary (e.g., in AC and DC engines) or can be realised provided that the second level converter is applied in the same way as in the mode switch of DC power devices;

- in many cases it is desirable to stabilise DC voltage U_d with a bit of margin of the maximum input AC voltage;
- the input current must be as perfect as it is possible as in the case of a power factor equal to one, so that the interface of the power electronics resistor could emulate a resistor, represented by the source load. This also implies that the electric current is always unidirectional - from the power source to the electrical equipment.
- price, losses and dimensions of the source current should be minimized [14,15].

Based on these rules, the isolation of the linear-frequency transducer is excluded. Also, if it is acceptable, then $U_d > U_s$, where U_s - peak AC input voltage. Thus, the obvious choice for the power supply is a DC-DC upconverter.

This converter is shown in Figure 4, where capacitor C_d is used to minimize fluctuations U_d and to meet the requirements of energy storage systems of the power electronics. Constant current I_{load}, is the power supplied to the rest of the system (a high-frequency component in the output current is actually filtered C_d). For simplicity, the internal inductance Ls of the source is not included into this Figure.

The input current is advisable to supply with sinusoidal and co-phase ones with Us. In practice, the losses in the bridge rectifier and in the DC-DC upconverter are quite small, and in a number of cases they can be neglected. Considering that, $\hat{U}_S = \sqrt{2}U_S$, $\hat{I}_S = \sqrt{2}I_S$ is the input power of AC source:

$$p_{in}(t) = \hat{U}_S |\sin \omega t| \hat{I}_S |\sin \omega t| = U_S I_S - U_S I_S \cos 2\omega t$$

Because of relatively large capacity C_d, voltage U_d can initially be considered permanent: U_d (t) = U_d. Therefore, output power P_d (t) = $U_d I_d$ (t),
where, according to Figure 4:

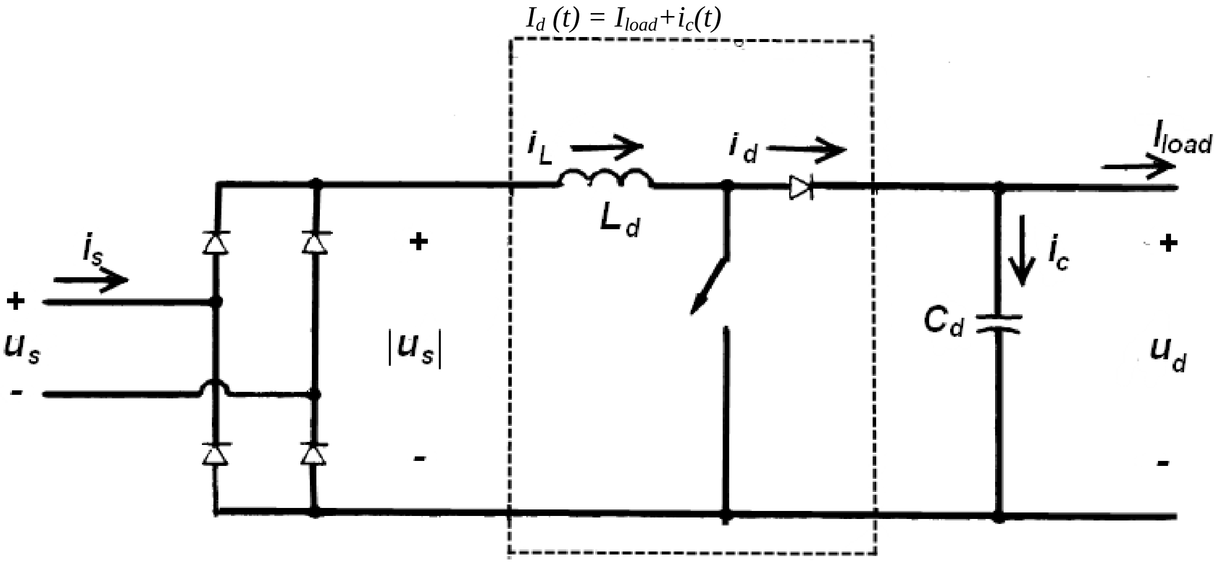

Figure 4. Scheme of the upconverter

If the boost converter in Figure 4 is perfect, it can be assumed that it will operate at a switching frequency approximate to infinity and L_d will be negligibly small. This allows suggesting that $p_{in}(t) = p_d(t)$ at any given time instant.

Therefore, $i_d(t) = i_{load} + i_c(t) = \frac{U_S I_S}{U_d} \cos 2\omega t$, where the average value is $I_d = I_{load} = \frac{U_S I_S}{U_d}$ and the current passing through the capacitor is: $I_c(t) = -\frac{U_S I_S}{U_d} \cos 2\omega t = -I_d \cos 2\omega t$.

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Even if this assumption is satisfied, the capacitor will have DC voltage virtually without any pulsation, and pulsation U_d can be calculated as: $V_{d,ripple}(t) \approx \frac{1}{C_d}\int i_c dt = -\frac{I_d}{2\omega C_d}\sin 2\omega t$, which can be kept at a low level having selected accordingly large value C_d. A series of LC-filters tuned into a doubled AC frequency can be connected in parallel to C_d to minimize the pulsation of DC voltage.

It should be noted that the frequency components of i_d currents and high frequency components of the load currents will also pass through C_d. Since the input current of the upconverter has a certain shape, the upconverter functions in a current control mode, due to conversion from DC to AC.

The feedback control is shown in the block diagram (Figure 5*)*, where i^*_L - the predetermined or desired current value.

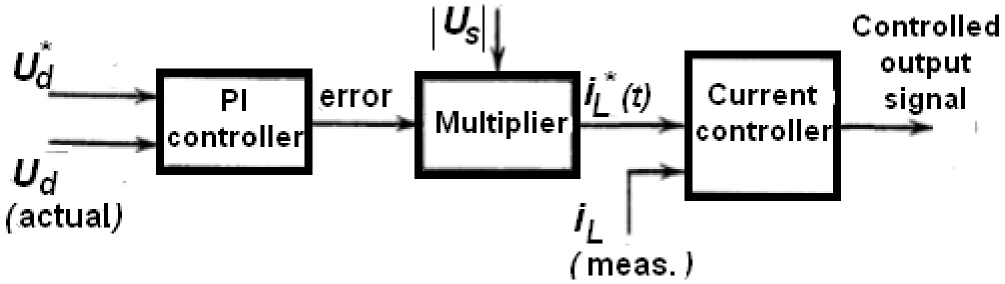

Figure 5. Feedback control of upconverter

Here, i^*_L has the same shape as $|U_S|$ does. The amplitude of i^*_L should be at certain level so as to maintain the output voltage at a predetermined or reference level U^*_d, despite the changes at load and the mains voltage deviation from the nominal value. Waveform of the i^*_L-signal was obtained by measuring $|U_S|$ using resistive voltage divider and having multiplied it with the increased error between initial value U^*_d and actually measured value U_d. Actual current i_L is usually determined by measuring the voltage across a small resistor installed in the direction opposite to i_L. The condition of the switch in the upconverter is controlled by the process of comparing actual current i_L and i^*_L.

Instantaneous value of i^*_L and i_L is determined by various methods of the running control of the mode from the upconverter. These four methods are presented below, where f_s - switching frequency and I_{rip} - peak pulsation of i_L within one period of time of the switching frequency. Determination of the constant frequency control is described in more detail.

1. Controlling the constant frequency. In this case, switching frequency f_s remains constant. When i_L reaches i^*_L, the switch of the upconverter is disconnected. The switch is on for a certain time at fixed frequency f_s, which leads to value i_L, as it is shown in Figure 6, a. The compensation of acceleration tilt is to be performed; otherwise i_L will be uneven in the ratio of switch - debt in excess of 0.5 of normal I_{rip} as shown in Figure 6.

2. The constant of the access to a range of adjustment. Here, current i_L is controlled in a such way that peak pulsation I_{rip} remains constant. With pre-selected values of I_{rip}, value i_L is forced to be within tolerances $i_L^* + \frac{1}{2}I_{rip}$ and $i_L^* - \frac{1}{2}I_{rip}$, and should be adjusted by the position of the switch.

3. Switching to a manual control. In this case, peak pulsation I_{rip} increases in proportion to instantaneous value $|U_S|$.

Otherwise, this approach is similar to the control of the constant of the tolerance range.

4. Controlling intermittent current. In this scheme, the switch is off when i_L reaches doubled value i^*_L.

The key is kept locked until i_L reaches zero. At this moment, the switch is on again. This can be considered as a special case of the variable of the control range allowance.

During the period of the switching frequency it is assumed that the output voltage is constant and input voltage U_d of the upconverter is considered to be constant at this moment of time; I_{rip} - peak pulsation of the current during one period of time of switching frequency.

The following formulas can be derived from Figure 4, during interval t_{on} and interval of switch locking t_{off}.

$$t_{on} = \frac{L_d I_{rip}}{|U_S|} \tag{10}$$

$$t_{off} = \frac{L_d I_{rip}}{U_d - |U_S|} \tag{11}$$

Switching frequency f_s:

$$f_S = \frac{1}{t_{on} + t_{off}} = \frac{(U_d - |U_S|)|U_S|}{L_d I_{rip} U_d} \tag{12}$$

Constant frequency f_s, controlled by the circuit, is constant in this formula and, therefore:

$$I_{rip} = \frac{(U_d - |U_S|)|U_S|}{f_S L_d U_d} \tag{13}$$

Figure 6,*b* shows a graph of normal I_{rip} as a function of $|U_S|/U_d$, taking into account that ratio $|U_S|/U_d$ in the upconverter should be ≤ 1.

The maximum ripple current is set as:

$$I_{rip,\max} = \frac{U_d}{4 f_S L_d} \tag{14}$$

where $|U_S| = \frac{1}{2} V_d$.

In the active loop of shaping current with the use of the DC upconverter the following additional comments have been made:

- Output voltage U_d across capacitor C_d contains 120 Hz of pulsations, which is 2 times higher than the frequency of the grid. A feedback control scheme is used to control U_d at a certain level, and can not compensate this pulsating voltage without distorting the current flow.
- If switching of pulsation frequency i_L is maintained with a small amplitude, it is possible to use a steel core of the inductor coil, which will be smaller in size because of its higher density of saturation flux as compared with high-frequency ferrite materials.
- Higher switching frequency allows for lower value L_d and increase the ease of filtering of high-frequency pulsation. However, the switching frequency is selected as a compromise between the above-mentioned advantages and increasing switching losses.
- U_d voltage exceeding the nominal voltage by more than 10% after reaching the peak of input AC voltage $\hat{U}_S$ will cause a decrease in efficiency.
- To limit the starting current one can use a current-limiting resistor in series with L_d. After the initial transient process, a resistor is bypassed by the contactor or a thyristor connected in parallel to the current-limiting resistor.
- The upconverter topology is well suited for the current input signal, since when the switch is off, the input current directly (via the diode) feeds the output stage. In the current control of the constant frequency, the switch of duty cycle *d* is taken as an example and is shown in Figure 6, *c*, as a function of *ωt*. It is known that a step-up transformer has input voltage $|U_S|$ and output voltage U_d:

  $$(|U_S|/U_d) = 1 - d,$$

  therefore

  $$d = 1 - \frac{|U_S|}{U_d}.$$

From Figure 6, *c* one can see that *d* - the lowest peak of i^*_L. Thus, the maximum value is passing through the switch only during a small part of the switching period.

- A small filter capacitor must be installed at the output of the diode of the bridge rectifier to prevent pulsation i_L. The filter of electromagnetic interferences (EMI) at the input is still required, as well as in the conventional circuit without active current signal.

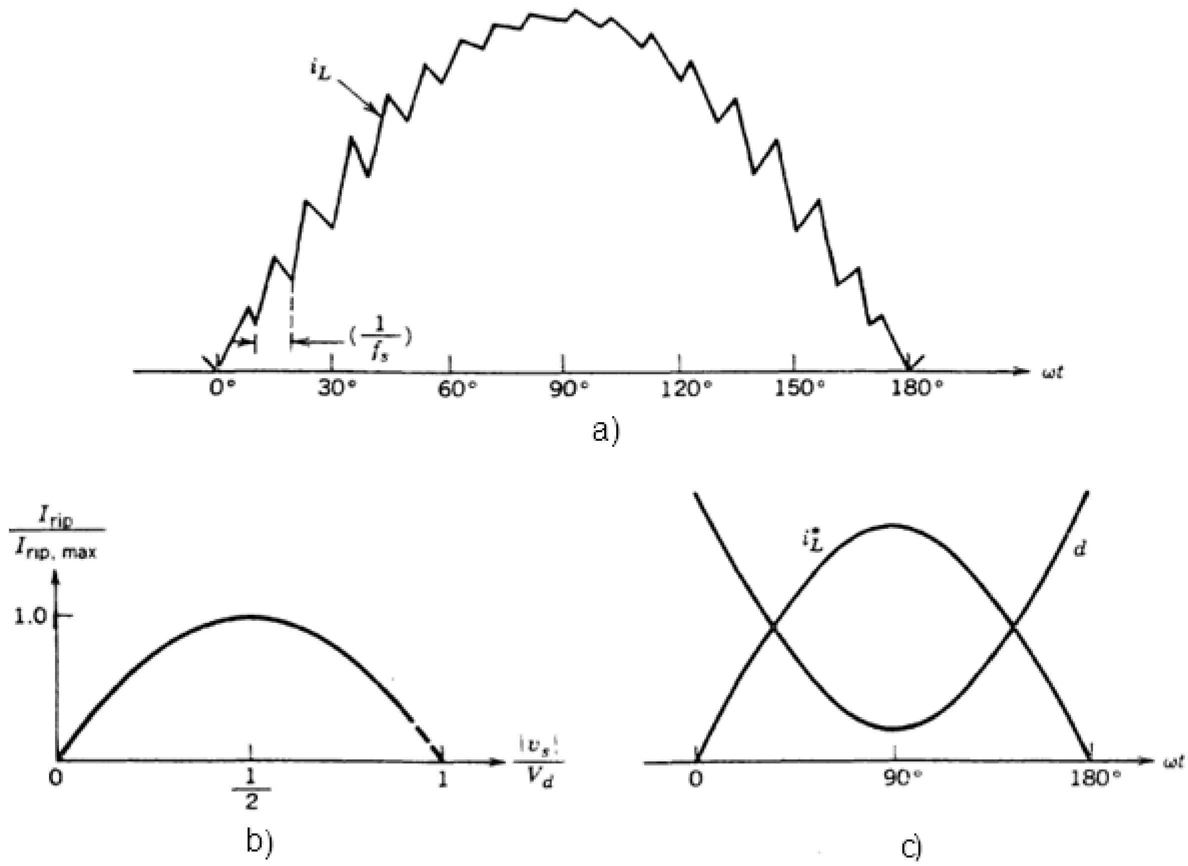

Figure 6. Control of constant frequency
a) "saw-tooth" form of i_L at constant switching frequency f_s; b) steady-state value of peak pulsation I_{rip}; c) dependence of *d* on ωt

3. Conclusion

Thus, the described electrical device can convert AC to DC, and it is realised using semiconductor devices - transistors and diodes - in a bridge circuit. The technical result of the application of this device is to reduce the weight - size parameters (compared to the closest analogues) to connect the controlled rectifier of the voltage to AC power supply, including mass and size of the capacitor bank of the current-limiting circuits, and improving the dynamic parameters of the stage of the uncontrolled charge of the output capacitor of the voltage controlled rectifier.

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