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# THREE-TO-ONE PHASE CONVERTER'S CONTROL METHOD FOR POWER FACTOR CORRECTED POWER SUPPLY OF RESISTANCE WELDING MACHINE

**Purpose.** The improvement of control method for a three-to-one phase matrix converter of a power supply of a single-phase resistance welding machine. This allows ensuring electromagnetic compatibility of the power supply with the power network, increasing the converter efficiency by reducing the switching frequency of power switches, and reducing power losses therein.

**Methodology. T**o study the power characteristics of power supplies for resistance welding machines, methods of mathematical and simulation modeling were used. The synthesis of the control system is based on the obtained regularities, characteristics and generalized requirements applicable to power supplies for resistance welding machines.

**Findings.** The control algorithm for the three-to-one phase matrix converter of the power supply of resistance welding machine is developed. A control system based on this algorithm is synthesized. A mathematical simulation of a power supply with a developed control system is performed. A control method of the direct converter of the resistance welding machine power supply is developed. It ensures electromagnetic compatibility with the supply network and improves the converter efficiency.

**Originality.** A method for controlling the three-to-one phase matrix converter, based on one-cycle control algorithm, is proposed. This method consists of sequential switching of bidirectional converter switches. Their switching on depends on the network voltages according to a given algorithm. The commutation moments are determined by the time when the current integral reaches the value of the reference charge, calculated in accordance with the value of the active resistance simulated by the converter with respect to mains.

**Practical value.** The use of a direct matrix converter for supplying the resistance welding machine, which is controlled by the developed algorithm, makes it possible to ensure electromagnetic compatibility of the power supply with the power network and increase its energy efficiency by reducing the switching frequency of power switches, reducing power losses therein.

**Keywords:** *resistance welding machine, power supply, three-to-one phase matrix converter, power factor, control system, one-cycle control*

**Introduction.** Power supplies of resistance welding machines are powerful non-linear consumers of electricity. Most of them use thyristor control circuits. This is due to the high reliability of such power supplies and the ease of controlling the output current. However, such power supplies (PSs) have a number of drawbacks, one of which is the low power factor (PF). This is caused by the high consumption of non-active components of fundamental and non-fundamental power. In addition, the PSs for resistance welding are mainly single-phase, and when they are connected to the line-to-line voltage of a three-phase network, they cause unbalance. Decreasing of power quality (particularly, appearance of voltage deviations and fluctuations), leads to decreasing of the quality of welded joints. All of this characterizes the low energy efficiency of such power supplies.

One of the main ways of developing energy-efficient power supplies for resistance welding machines is to ensure their electromagnetic compatibility with the electric mains. A perspective way for solving this problem is the use of three-to-one-phase matrix converters (TOMC) [1, 2]. Their use makes it possible to obtain an almost unity power factor and symmetrical loading of all three phases of the mains, while ensuring a high quality of the welding process. The low total harmonic distortion (THD) of the input currents makes it possible to ensure compliance with the standards for the emission of harmonics of the input current, established by the valid standards IEC 61000-3-2:2016, IEC 61000-2-12:2013, IEC 61000-3-4:2009.

The main disadvantages of such converters are significant power losses due to high switching frequency of power switches. Absence (or high cost) of the element base, capable of commutating powerful loads at a high frequency, leads to a limited use of TOMC, despite their advantages.

**Unsolved problem.** As noted above, the existing control methods for a direct three-to-one-phase converter imply a high switching frequency. Attempts to reduce the frequency of switching of the converter using existing algorithms lead to saturation of spectrum of the TOMC input currents with higher harmonics.

The converter control system must perform two functions: to generate a specified output voltage (or output current) and provide a high input power factor. At the same time, in the algorithms described in the literature [1] for calculating the duty ratio of the control pulses of the three-to-one phase converter, it is assumed that its output current does not change during the switching period. However, when the switching frequency of the converter decreases, the influence of the increased pulsations of its output current on the shape of the input (mains) current becomes significant. The magnitude of the output current ripple is also influenced, along with the switching frequency, by the inductance of the output circuit of the converter, and the modulation depth. Therefore, the development of TOMC control methods that allow providing a high input PF (accurate formation of input currents) under conditions of a significant level of pulsations in the output current of the converter is actual.

**Analysis of the recent research.** The most common methods for controlling direct matrix converters (MCs) today are: the use of pulse-width (PWM) and space-vector modulation (SVM).

To control the three-to-one phase matrix converter, which is a particular case of the classical MC, the control methods are somewhat different. In [3], a method is described that consists in sequentially switching switch pairs during a period of pulse-density modulation (PDM). In this case, the selection of the switch pair takes place depending on the sector (one of six), on which the voltage period is conventionally divided, and the duration of the enabled switch state is determined according to the pre-calculated templates. The considered control method of the converter [3] provides the formation of a high-frequency voltage on the load. Character of the load is a resonant circuit, which allows for soft switching and a significant reduction in dynamic losses in the converter. Advantages of this method include the possibility of obtaining a close to a unity power factor and small THD of input currents. Such a circuit topology is applicable for resistance welding machines; however, the control algorithm considered is not applicable for RWM power supply.

In [4, 5], a control method involving a low switching frequency (less than 10 kHz) is considered. It consists of the serial switching of switch pairs during the PWM period. The choice of the required pair takes place depending on the voltage sector, similarly to [3]. In this case, the switching times are determined by the moments of equality of the specified voltage and the carrier PWM signal. The simplicity of this method is its advantage. The drawbacks include the presence of significant pulsations of the input current of the TOMC.

In [6], the technique of space-vector modulation for controlling a three-to-one phase matrix converter is described. Several control strategies have been considered. This makes it possible to achieve a minimum THD of output current or a high efficiency of the converter itself.

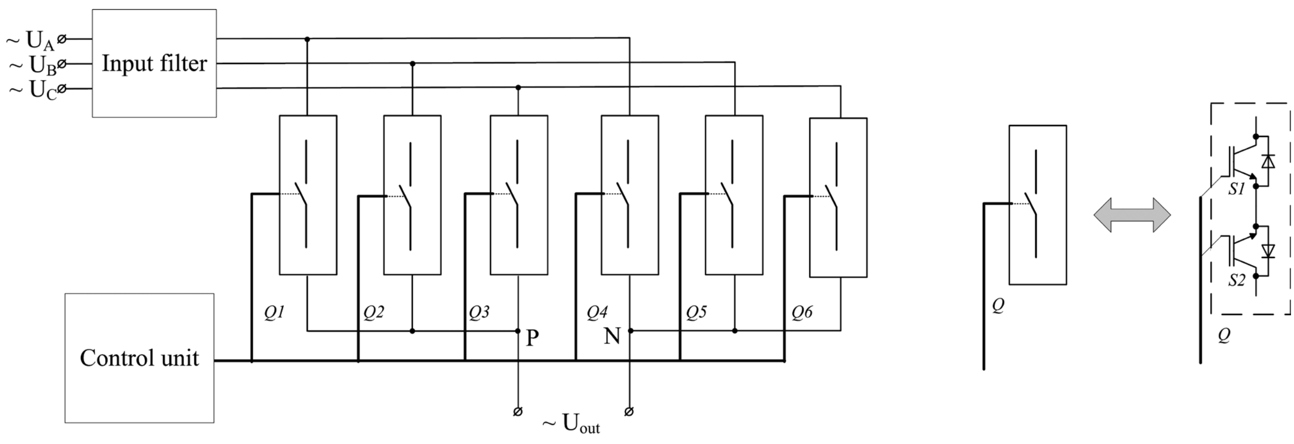
In [1, 2, 7], control methods of converter based on PWM technique are described, and the required switches pair is selected depending on the sector, as well as in [3–5]. In this case, switch control is carried out in such a way as to ensure proportionality between the consumed currents and phase voltages. This will lead to minimization of the power losses in electrical mains during the operation of TOMC. These control methods allow controlling not only the output parameters of the converter, but also the input.

In [8–10], there are described methods for controlling of three-phase MCs, providing stabilization of the input current. In [8], the MC control method was proposed and investigated, the purpose of which is to improve the quality of the input currents for unbalanced input and output conditions. In [9], the approach of using a “virtual matrix converter” is proposed, which makes it possible to simplify the control system and provide more efficient output characteristics in comparison with the standard approach. The disadvantage of this approach is the low voltage transfer ratio of the output voltage and the inability to control the power factor of the converter. In [10], a control method that provides separate control of the “virtual rectifier” and the “virtual inverter” is considered. In this case, the duty cycles of the control pulses are calculated from the condition of minimizing the switching losses in the converter. This approach makes it possible to increase the efficiency of the converter. In [11], a pulsation power compensation system based on prediction modeling is proposed. It can eliminate harmonic components of input currents and voltages.

**Unsolved aspects of the problem.** Disadvantages of the described methods are common – an increased switching frequency of TOMC, causing a significant loss of power in it, and a high THD of input currents.

**Objectives of the article.** The purpose of the work is to develop a control method of TOMC, which will improve the technical and economic performance of the converter by reducing switching losses and ensuring the electromagnetic compatibility of the converter with the electrical network.

**Presentation of the main research.** In the TOMC, which consists of six bidirectional switches [1, 2], the load is alternately connected to the two phases of the network (Fig. 1). Bidirectional switches can be made in




a b

*Fig. 1. Structural scheme of TOMC and its connection to the network (a) and realization of bidirectional switch (b)*

the form of two series-connected transistors with anti-parallel diodes, as shown in Fig. 1, *b*. Control electrodes of transistors are connected to the control system (Fig. 1). The high-frequency components of the input current are filtered by an input filter (Fig. 1).

It is known that the minimum power loss in the network can be obtained by providing proportionality between the consumed currents and the corresponding phase voltages. That is, the converter along with its load must simulate a symmetrical active resistance.

To analyze the principle of TOMC operation, lets consider one network voltage period (Fig. 2). It can be virtually divided into six sectors, in each of which the following condition will be fulfilled: two of the three phase voltages do not change their sign throughout the sector, and the third voltage is in the positive maximum or negative minimum (Fig. 2).

In each sector, there is a switching action only between the three of the six converter switches during the switching period, another switch is permanently on, and two more switches are not turned on at all.

Table 1 shows the distribution of switches by sector when the voltage is positive and negative polarity.

Let us consider any sector, for example, the fifth one, when forming an output voltage with positive polarity. During the time in this sector, there will be a switching between the switches Q1, Q2, Q3, and the Q4 switch will be switched on permanently.

The switches Q5, Q6 in the considered sector do not turn on. Thus, the output voltage of the converter will be formed from the line-to-line voltages $u_{ba}(t)$ (switches Q2, Q4 on) and $u_{ca}(t)$ (switches Q3, Q4 on). The turning on of the switch pair Q1, Q4 will result in a zero voltage at the converter output.

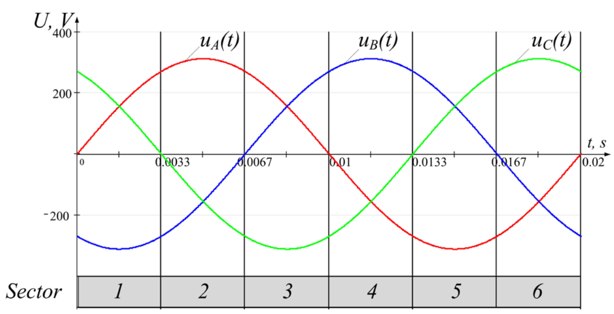



*Fig. 2. Separation of the network period into 6 sectors*

The local average (average over the switching period) output voltage of the converter can be represented by the following expression

$$\begin{aligned}\hat{u}_{out}(t) &= u_{ba}(t)\cdot D_2 + u_{ca}(t)\cdot D_3 = \\ &= u_b(t)\cdot D_2 + u_c(t)\cdot D_3 - u_a(t)\cdot\left(D_2 + D_3\right),\end{aligned} \tag{1}$$

where $D_i$ is duty cycle of the control pulses of the corresponding switch; $u_{ba}(t)$, $u_{ca}(t)$ are phase-to-phase voltages; $u_a(t)$, $u_b(t)$, $u_c(t)$ are phase-to-neutral voltages.

In this case, the local average input currents can be defined as

$$\begin{aligned}\hat{i}_B(t) &= i_{out}(t)\cdot D_2; \\ \hat{i}_C(t) &= i_{out}(t)\cdot D_3; \\ \hat{i}_A(t) &= -i_{out}(t)\cdot\left(D_2 + D_3\right),\end{aligned} \tag{2}$$

where $D_i$ is duty cycle of the control pulses of the corresponding switch; $i_{out}(t)$ is output current.

Expressions (1) and (2) characterize the operating principle of the converter. On the other hand, to simu-

*Table 1*

Distribution of switches by sector when the voltage is positive and negative polarity

| Sector | 1 | 2 | 3 | 4 | 5 | 6 |
|---|---|---|---|---|---|---|
| Positive output voltage | | | | | | |
| Switching | Q3, Q1, Q2 | Q5, Q6, Q4 | Q1, Q2, Q3 | Q6, Q4, Q5 | Q2, Q3, Q1 | Q4, Q5, Q6 |
| Always On | Q5 | Q1 | Q6 | Q2 | Q4 | Q3 |
| Negative output voltage | | | | | | |
| Switching | Q6, Q4, Q5 | Q2, Q3, Q1 | Q4, Q5, Q6 | Q3, Q1, Q2 | Q5, Q6, Q4 | Q1, Q2, Q3 |
| Always On | Q2 | Q4 | Q3 | Q5 | Q1 | Q6 |



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late a symmetrical active load, the following dependencies must be carried out

$$\hat{i}_A(t)=\frac{\hat{u}_A(t)}{R};\quad \hat{i}_B(t)=\frac{\hat{u}_B(t)}{R};\quad \hat{i}_C(t)=\frac{\hat{u}_C(t)}{R},\qquad (3)$$

where $R$ is simulated active resistance; $\hat{u}_A(t),\ \hat{u}_B(t),\ \hat{u}_C(t)$ are local average phase-to-neutral network voltages.

Based on this, it is possible to find expressions for determining the duty cycle of the switch control pulses at a constant switching frequency

$$D_2=\frac{\hat{u}_B(t)}{R\cdot i_{out}(t)};\qquad D_3=\frac{\hat{u}_C(t)}{R\cdot i_{out}(t)}.\qquad (4)$$

For other sectors, the definitions are similar.

Suppose that the output current of the converter during the switching period remains almost unchanged, or changes insignificantly, so it is assumed to be a constant value

$$\hat{i}_{out}(t)=I_{out}.\qquad (5)$$

Let us express the duty ratio of the control pulses through the duration of the on-state of the switches

$$D_2=\frac{t_{on2}}{T};\quad D_3=\frac{t_{on3}}{T},\qquad (6)$$

where $T$ is switching period; $t_{on2},\ t_{on3}$ stand for duration of the on-state of the switches Q2, Q3 during the period $T$.

Then, from the expressions (2, 3, 6), we can write

$$I_{out}\cdot\frac{t_{on2}}{T}=\frac{\hat{u}_B(t)}{R};\quad I_{out}\cdot\frac{t_{on3}}{T}=\frac{\hat{u}_C(t)}{R};\qquad (7)$$

$$I_{out}\cdot t_{on2}=\frac{\hat{u}_B(t)}{R}\cdot T;\quad I_{out}\cdot t_{on3}=\frac{\hat{u}_C(t)}{R}\cdot T.\qquad (8)$$

The right-hand side of equations (8) is an expression for calculating the reference charge, i. e. such a charge that would pass through the simulated active resistance $R$ at the corresponding phase voltage, during the switching period $T$.

However, the assumption (5) leads to the appearance of a current regulation error due to the fact that the product $I_{out}\cdot t_{on}$, where ton – the time of the on-state of the corresponding switch, calculated by the control system, differs from the real value of the charge transferred to the load.

This leads to the fact that the input phase currents are not proportional to the corresponding phase voltages. And the lower the switching frequency is, the greater the mentioned error is.

If the assumption (5) is removed, then (8) can be represented in the following form

$$\int_0^{t_{on2}} i_{out}(t)dt=\frac{\hat{u}_B(t)}{R}\cdot T;\quad \int_{t_{on2}}^{t_{on2}+t_{on3}} i_{out}(t)dt=\frac{\hat{u}_C(t)}{R}\cdot T.\qquad (9)$$

It is possible to generalize expressions (9), assuming that during the switching period $U_{ph}=\hat{u}_{ph}(t)=const$, where $\hat{u}_{ph}(t)$ is corresponding phase-to-neutral voltage

$$\int i_{out}(t)dt=\frac{U_{ph}\cdot T}{R}=Q_{on}.\qquad (10)$$

Expression (10) explains the essence of the developed method for controlling the TOMC in a general form: the duration of the on-state of the corresponding switch is determined by comparing the output current integral with the value of the reference charge $Q_{ref}$, which is calculated in real time.

The proposed method allows reducing the switching frequency of the power switches, while ensuring a reduction in the ripple of the output current, a decrease in THD of the input currents due to the normalization of the charge transferred to the load.

Let us consider the sequence of switching in accordance with the developed method. Switching starts with the switch pair that will connect the highest voltage to the load.

The Q4 switch is turned on throughout the sector all the time. The control system generates the switch-on signal Q2, and the accumulation of the integral $\int_0^{t_{on2}} i_{out}(t)dt$ begins. Turning off of the switch Q2 occurs at the moment when the integral has reached the value $\frac{u_B(t)}{R}\cdot T$. And at the same time, the control system sends a signal to turn on the switch Q3. Then the accumulation of the integral $\int_{t_{on2}}^{t_{on2}+t_{on3}} i_{out}(t)dt,$ begins, until it reaches the value $\frac{u_C(t)}{R}\cdot T$. At this moment, the Q3 switch turns off and the Q1 switch turns on, which will be on-state until the end of the switching period. The switch Q1 will zero the load voltage.

The block diagram of the control system is shown in Fig. 3.

In each switching period, the circuit operates in one of three states. In the "Sector selector" block (Fig. 3, *1*), the phase voltages are compared, the sector is identified and those switches that can be switched during this sector are identified too. Further, in the same block, the switching sequence of the three switches is determined during the switching period.

The first cycle begins. From the "Sector Selector" block (Fig. 3, *1*), a corresponding signal is sent to the selector switch (Fig. 3, *2*). The selector switch (Fig. 3, *2*) connects the required phase voltage. It is filtered from the modulation components at the switching frequency with the low pass filter (LPF) of the filtering unit (Fig. 3, *3*), and multiplied by the factor $T/R$, which is a quotient of the switching period and the simulated resistance. The resulting signal is applied to the inverse input of comparator (Fig. 3, *4*) through the absolute value detection unit. At the same time, the output current signal of the converter is fed to the inverted input of the integrator with reset (Fig. 3, *5*).

The accumulation of the output current integral begins. The output signal of the integrator with a reset falls

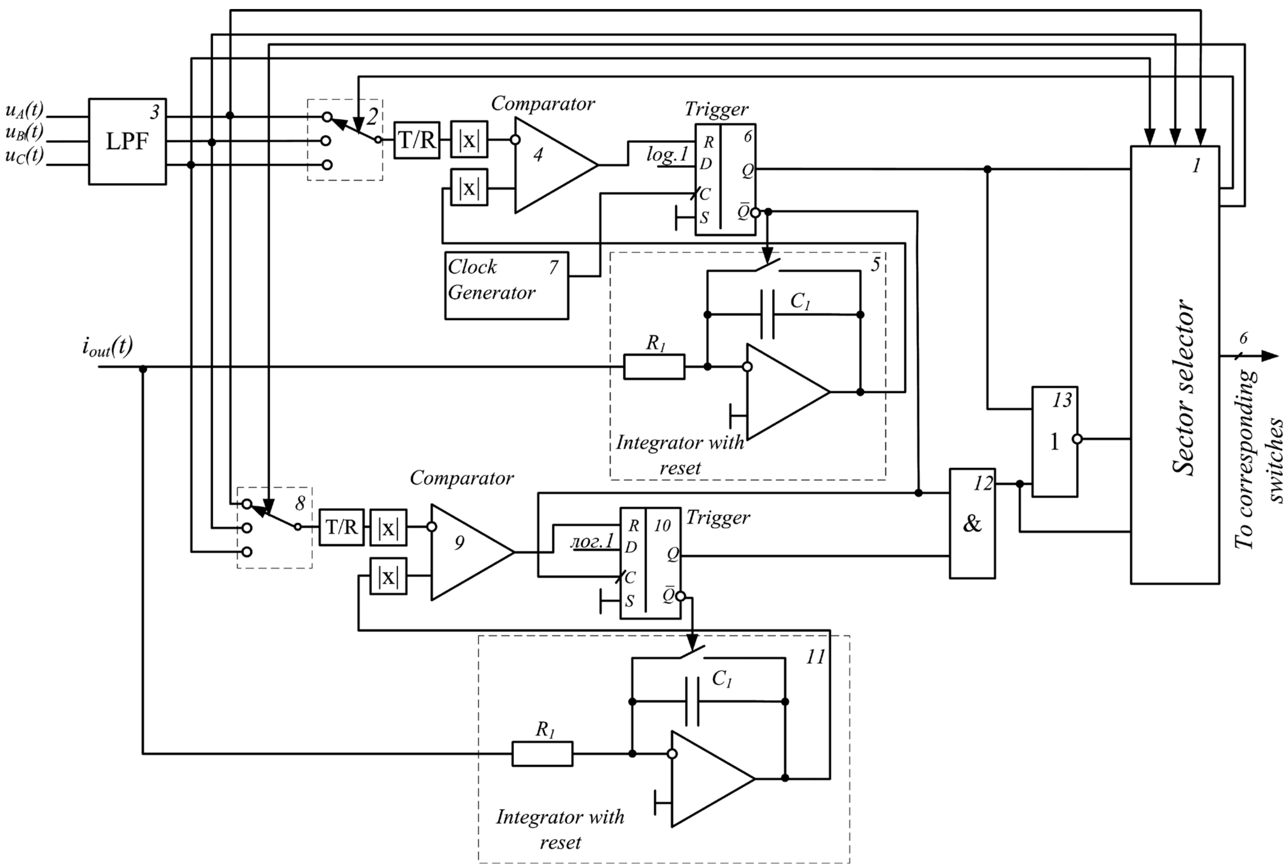


*Fig. 3. Structural diagram of a TOMC control system*

on the direct input of the comparator (Fig. 3, *4*) through the absolute value detection unit.

If this signal is exceeded with respect to the signal at the inverse input, a logic high signal is generated at the output of the comparator (Fig. 3, *4*), which is fed to the asynchronous input R of flip-flop (Fig. 3, *6*). The clock input C of this flip-flop receives a signal from the clock generator (Fig. 3, *7*), and the information input D is connected to logic high. The change in the flip-flop state occurs on the front of the clock signal. From the inverting output of the flip-flop (Fig. 3, *6*), the signal activates the reset of the integrator (Fig. 3, *5*), and from the non-inverting one to the selector of sectors (Fig. 3, *1*). The second cycle begins. The sector selector (Fig. 3, *1*) supplies the corresponding signal to the selector switch (Fig. 3, *8*), and the operation of the second part of the control system is carried out in a manner similar to that described above, except that the clock source for the flip-flop (Fig. 3, *10*) is the inverting output of the flip-flop (Fig. 3, *6*). The signals from the non-inverting output of the flip-flop (Fig. 3, *10*) and the inverting output of the flip-flop (Fig. 3, *6*) are fed to the AND element (Fig. 3, *12*). The logic high on the output of the AND element (Fig. 3, *12*) will appear only when the logic high is at the output of one of the triggers (Fig. 3, *6*) and (Fig. 3, *10*), which corresponds to switching one switch from three in a given sector. From the output of the “AND” element (Fig. 3, *12*), the signal goes to the selector of sectors (Fig. 3, *1*) and to the “NOR” element (Fig. 3, *13*), the other input of which receives a signal from the direct output of flip-flop (Fig. 3, *6*). From the output of element (Fig. 3, *13*), the signal goes to the selector of sectors (Fig. 3, *1*). Use of logic elements (Fig. 3, *12, 13*) leads to ensuring that it is not possible to simultaneously switch on two of the three switches that are switched during the PWM period.

Fig. 4 shows a simulation model of a three-to-one-phase matrix converter in the Matlab/Simulink.

The power part of the converter, made according to the scheme in Fig. 1, is implemented in the “MATRIX CONVERTER” block. The input signals for the subsystem are the pulses coming from the TOMC control system. For its implementation, the “DUTY-CYCLE CALC” subsystem is used, shown in Fig. 5.

Input signals for this subsystem are the phase-to-neutral voltages of the network, depending on which the relevant sectors are calculated and the current pair of voltages inside the sector between which the switching occurs. The definition of sectors is performed in the “Sector_calc” and “Sector_calculation_half” blocks. The reference current *I_ref* is formed in accordance with the expression (9). The “OCC” subsystem, presented in Fig. 5, determines the duration of the enabled states of the respective switch pairs, by supplying control pulses to the selectors S1, S2 (Fig. 5). These selectors are designed to distribute switches by sectors when generating output voltage of positive and negative polarity, according to Table 1. The polarity of the output voltage is determined by the selector S3 (Fig. 5).

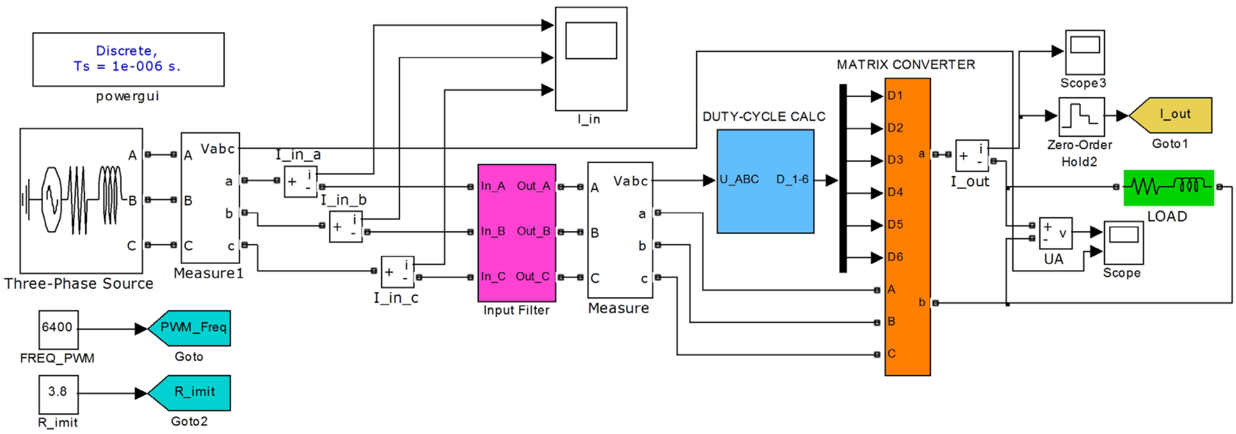


*Fig. 4. Mathematical model of three-to-one phase MC*

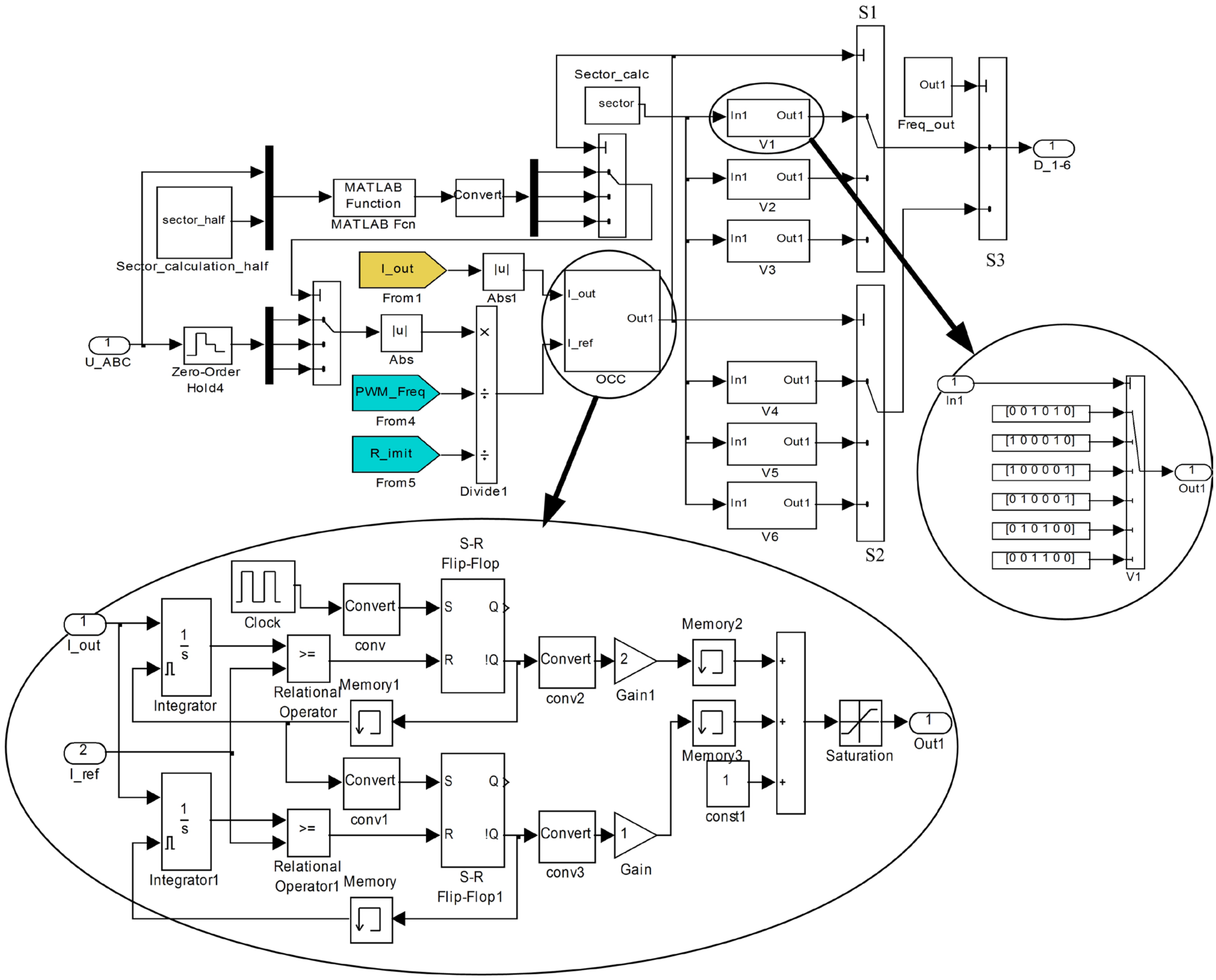


*Fig. 5. The “DUTY-CYCLE CALC” subsystem*

The welding circuit is modeled in the “LOAD” block. Circuit parameters: active resistance 2.5 Ohms, inductance 1.6 mH. The input filter “Input Filter” (Fig. 4) is a first order three-phase LC-filter and serves for filtering the modulation components at the switching frequency.

Electrical mains parameters: short-circuit power 100 MVA, $X$ to $R$ ratio is 5. Switching frequency of the converter is 6.4 kHz. The shape of the output current is rectangular, the output frequency is 37.5 Hz.

Fig. 6 shows the output current diagram of the converter when the converter is operating on the active-inductive load, which is the welding circuit.

Fig. 7 shows the diagrams of input currents and voltages of TOMC when implementing the described control method.

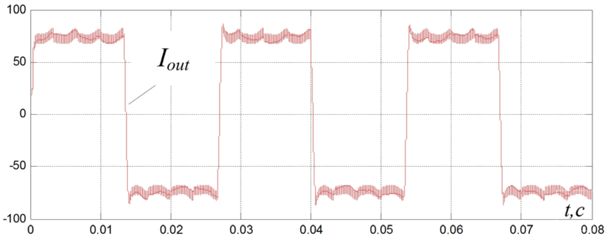


*Fig. 6. Diagrams of output currents of TOMC*

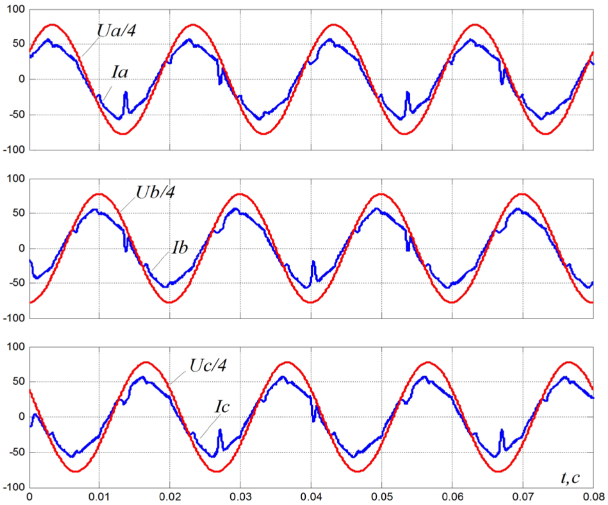


*Fig. 7. Diagrams of TOMC voltages and input currents*

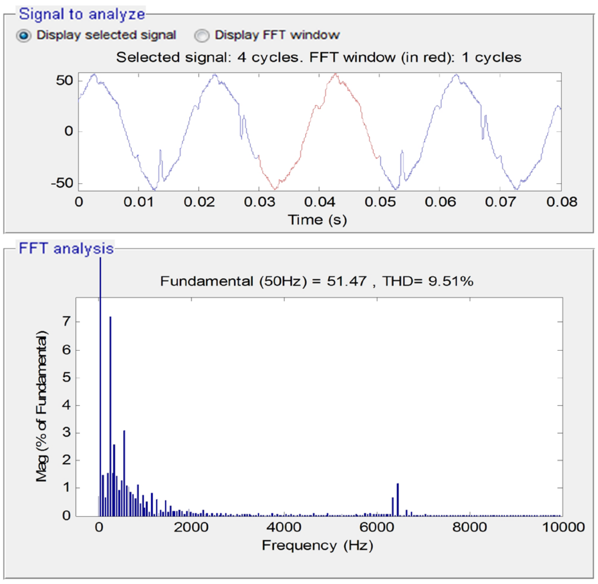


*Fig. 8. Spectrum of the input current of a TOMC*

The spectral composition of the currents consumed by the converter is shown in Fig. 8.

The THD of the input currents after filtering the modulation components at the switching frequency is 9.5–10.5 %. The calculated power factor of the converter is 0.97.

The obtained diagrams show the balanced current consumption from the network and the increased power factor (in comparison with the classical methods for controlling the MC) at a reduced switching frequency.

**Conclusions and recommendations for further research.** A new method for controlling a three-to-one phase direct matrix converter is proposed, in accordance with which the switching instants of the power switches are determined by the equality of the output current integral to the value of the reference charge.

The application of the proposed control method will improve the technical and economic performance of the converter by reducing the switching losses, providing a reduction in the switching frequency in comparison with the existing control methods and avoiding the appearance of ripple of the input current, while ensuring the electromagnetic compatibility of the converter with the electric mains.

A perspective research way is the development of control methods for the converter with the possibility of limited performance of the active filter functions, which will reduce the network voltage harmonics at the point of connection of the TOMC to the network.

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## Спосіб управління трифазно-однофазним перетворювачем джерела живлення машини контактного зварювання з корекцією коефіцієнта потужності

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**Мета.** Удосконалення способу управління трифазно-однофазним матричним перетворювачем джерела живлення однофазної машини контактного зварювання, що дозволяє забезпечити електромагнітну сумісність такого джерела з мережею, підвищити його енергоефективність за рахунок зниження частоти перемикання силових ключів, зменшення втрат потужності у ньому.

**Методика.** Для дослідження енергетичних характеристик джерел живлення машин контактного зварювання використані методи математичного та імітаційного моделювання. Синтез системи управління виконаний на підставі отриманих закономірностей, характеристик і узагальнених вимог до джерел живлення машин контактного зварювання.

**Результати.** Розроблено алгоритм управління трифазно-однофазним матричним перетворювачем джерела живлення машини контактного зварювання. Синтезована система управління на основі цього алгоритму. Проведене математичне моделювання джерела живлення із розробленою системою управління. Розроблено спосіб управління безпосереднім перетворювачем джерела живлення машини контактного зварювання, що забезпечує електромагнітну сумісність із мережею живлення та має високі показники енергоефективності.

**Наукова новизна.** Запропоновано спосіб керування трифазно-однофазним матричним перетворювачем, що полягає в послідовній комутації двоспрямованих ключів. Їх включення здійснюється в залежності від напруги мережі за заданим алгоритмом. Моменти комутації визначаються часом досягнення інтеграла струму величиною опорного заряду, розрахованого відповідно до величини активного опору, імітованого комплексом „перетворювач + навантаження“.

**Практична значимість.** Використання для живлення машини контактного зварювання безпосереднього перетворювача частоти, управління яким здійснюється за розробленим алгоритмом, дає можливість забезпечити електромагнітну сумісність джерела з мережею живлення й підвищити його енергоефективність за рахунок зниження частоти перемикання силових ключів, зменшення втрат потужності у ньому.

**Ключові слова:** *машина контактного зварювання, джерело живлення, трифазно-однофазний матричний перетворювач, коефіцієнт потужності, система управління, одноциклове управління*

## Способ управления трехфазно-однофазным преобразователем источника питания машины контактной сварки с коррекцией коэффициента мощности

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**Цель.** Совершенствование способа управления трехфазно-однофазным матричным преобразователем источника питания однофазной машины контактной сварки, что позволяет обеспечить электромагнитную совместимость такого источника с сетью, повысить его энергоэффективность за счет снижения частоты переключения силовых ключей, уменьшения потерь мощности в нем.

**Методика**. Для исследования энергетических характеристик источников питания машин контактной сварки использованы методы математического и имитационного моделирования. Синтез системы управления выполнен на основании полученных закономерностей, характеристик и обобщенных требований к источникам питания машин контактной сварки.

**Результаты.** Разработан алгоритм управления трехфазно-однофазным матричным преобразователем источника питания машины контактной сварки на основе одноциклового управления. Синтезирована система управления на основе этого алгоритма. Проведено математическое моделирование источника питания с разработанной системой управления. Разработан способ управления непосредственным преобразователем источника питания машины контактной сварки, что обеспечивает электромагнитную совместимость с питающей сетью и обеспечивает высокие показатели энергоэффективности преобразователя.

**Научная новизна.** Предложен способ управления трехфазно-однофазным матричным преобразователем, который заключается в последовательной коммутации двунаправленных ключей. Их включение осуществляется в зависимости от напряжения сети по заданному алгоритму. Моменты коммутации определяются временем достижения интеграла тока величины опорного заряда, рассчитанного в соответствии с величиной активного сопротивления, имитируемого комплексом „преобразователь + нагрузки“.

**Практическая значимость.** Использование для питания машины контактной сварки непосредственного преобразователя частоты, управление которым осуществляется по разработанному алгоритму, позволяет обеспечить электромагнитную совместимость источника с сетью питания и повысить его энергоэффективность за счет снижения частоты переключения силовых ключей, уменьшения потерь мощности в нем.

**Ключевые слова:** *машина контактной сварки, источник питания, трехфазно-однофазный матричный преобразователь, коэффициент мощности, система управления, одноцикловое управление*

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