THREE-PHASE INVERTER POWER SOURCE WITH DIRECT CONVERSION AND INCREASED POWER FACTOR

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Topology of a converter with high-frequency transformer decoupling and three-phase input without intermediate rectification of input voltage is proposed. Application of direct conversion principle allows reducing the number of elements in the inverter power circuit, thus increasing its efficiency. In addition, application of a special algorithm of switch control allows achievement of input power factor close to a unity.

Keywords: *arc welding, inverter power source, power factor, modelling, direct conversion*

Modern tendencies of development of welding power sources determine increased requirements to such characteristics as conversion efficiency, power per a unit of volume, power factor (PF), quality and dynamics of stabilization of output parameters (current or voltage). Power sources with high-frequency conversion meet such a set of requirements to the greatest degree.

The majority of modern inverter power sources are made by double conversion schematic [1]: mains voltage is rectified by noncontrolled, controlled or active rectifier, smoothed, and then applied to DC-DC converter, made by the single-step [2], half-bridge or bridge circuit.

The disadvantages of such sources include the non-sinusoidal nature of input current (sources with PF active corrector are an exception, where 2–3 % lower efficiency is the price to pay for the low harmonic factor of input currents), presence of high-voltage high-capacity electrolytic capacitor in DC circuit, that creates problems of its initial charge at the source switching on and increases the overall dimensions and weight of the source.

In study [3] a variant of single-phase welding source is proposed, in which the function of input voltage rectification is eliminated (four-transistor AC voltage chopper and low-frequency (50 Hz) transformer with low scattering are applied). The source shows good results on efficiency and PF, but application of low-frequency transformer leads to deterioration of weight and dimensional characteristics of the devices using such a regulation principle. In addition, if it is necessary to perform DC welding, energy storage has to be used in single-phase sources in any case. It ensures arcing at the moments of mains voltage going through zero. This can be filter capacitor or output choke. Work [3] also outlines the long-term goal of development of three-phase sources with isolating high-frequency transformer and direct conversion.

This work suggests circuit implementation of a device, using current achievements in the field of circuit engineering of matrix direct frequency converters [4] with high-frequency transformer decoupling.

The source (Figure 1) consists of input *LC*-filter (*L1–L3*, *C1–C3*), six bilateral controlled switches *S1–S6*, high-frequency isolating transformer *T1*, output bridge rectifier *VD1*, *VD2* and smoothing choke *L4* [5]. The circuit is a matrix converter with three-phase input and two-phase output.

At converter operation primary winding of transformer *T1* is alternatively connected with a high frequency to mains phases, with just one switch from *S1–S3* group and one from *S4–S6* group being open at each moment of time to prevent interphase short-circuit. Capacitors *C1–C3* smooth pulsed voltage surges at the moments of switch switching. Sequence of switching and off-duty ratio are selected so that during the switching period average voltage in *T1* primary winding was equal to zero:

$$\int_0^{T_{sw}} u_{T1} dt = 0, \tag{1}$$

where u_{T1} is the voltage on *T1* primary winding; T_{sw} is the switching period.

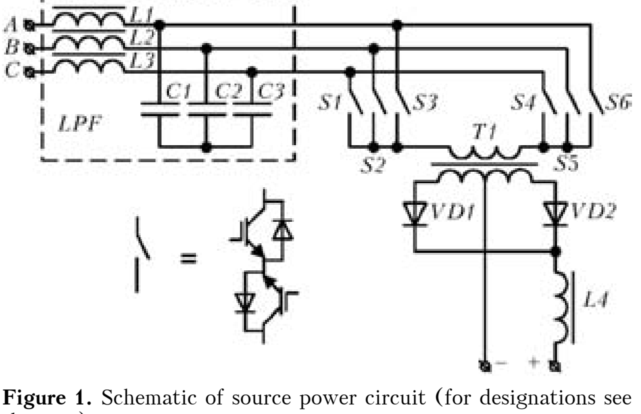

Figure 1. Schematic of source power circuit (for designations see the text)

This condition is necessary to prevent biasing and saturation of *T1* magnet core. Here, the voltage at bridge rectifier output is equal to $|u_{T1}/K_{T1}|$, where K_{T1} is the *T1* transformation ratio. We will determine average (over switching period) output voltage allowing for output filter *L4*:

$$U = \frac{1}{T_{sw}} \int_0^{T_{sw}} \left| \frac{u_{T1}}{K_{T1}} \right| dt. \qquad (2)$$

Thus, changing the sequence of *T1* connection to mains phases (while observing condition (1)) allows control of output voltage and, what is quite important, input current shape.

Let us denote the time of transformer connection to *A*, *B*, *C* phases as t_a, t_b, t_c, and off-duty ratios relative to *A*, *B*, *C* phases as $D_a = t_a/T_{sw}$, $D_b = t_b/T_{sw}$, $D_c = t_c/T_{sw}$, respectively. Then for input currents of the considered converter we can write:

$$\begin{aligned} i_a &= I_l D_a \operatorname{sign}(u_a), \\ i_b &= I_l D_b \operatorname{sign}(u_b), \\ i_c &= I_l D_c \operatorname{sign}(u_c), \end{aligned} \qquad (3)$$

where I_l is the load current reduced to primary side; u_a, u_b, u_c are the phase voltages of the circuit.

To ensure PF close to a unity, it is necessary for input current in each phase to be proportional to the respective phase voltage. This can be achieved by selection of off-duty ratio as follows:

$$D_a = \gamma|u_a|, \quad D_b = \gamma|u_b|, \quad D_c = \gamma|u_c|, \qquad (4)$$

where γ is the coefficient determining the output voltage.

Average voltage on transformer primary winding during time T_{sw} is defined as

$$U_{T1} = u_a D_a + u_b D_b + u_c D_c, \qquad (5)$$

this voltage sign being determined by numbers of switched-on switches. Substituting (4) into (5), we get:

$$U_{T1} = \gamma(u_a^2 + u_b^2 + u_c^2) = 1.5\gamma U_{ph}^2, \qquad (6)$$

where U_{ph} is the amplitude of mains phase voltage.

Thus, at observation of condition (4) it is possible to achieve source PF close to a unity. Moreover, another important conclusion follows from (6): ripple at mains frequency is absent at source output. This allows considerably increasing the quality of converter voltage and lowering the requirements to output filter.

We have conducted modelling of the proposed source in Mathcad environment. Modelling parameters are as follows: $L1 = L2 = L3 = 330$ μH, $C1 = C2 = C3 = 4.7$ μF, switching frequency of 20 kHz, $\gamma = 1/U_{ph}$. Results show that fundamental harmonics (50 Hz) and harmonics with frequencies, which are a multiple of switching frequency (20 kHz), are present in the input current of switch matrix. After filtering with application of second order input filter, components with frequencies close to the filter resonance frequency appear in the spectrum. In the given example this is 4.04 kHz. The total coefficient of input current harmonics is equal to 4.4 %. Electric diagram of power components of the source, corresponding to Figure 1, is shown in Figure 2. It has the obvious drawback of a large number of power switches equal to 12, and of complexity of their control. To simplify the control circuit and reduce the number of power switches, it is rational to apply the method of conditional division of the source power components into the rectifier and the converter. Such a procedure is often used for analysis of the process in matrix frequency converters.

In the optimized circuit, given in Figure 3, the three-phase input is made as a non-reversible rectifier with current output, transformer «excitation» is performed by bridge *VT4*–*VT7*, while added elements *VT8*, *C4* are used for limiting the transistor voltage during no-current pauses between switching.

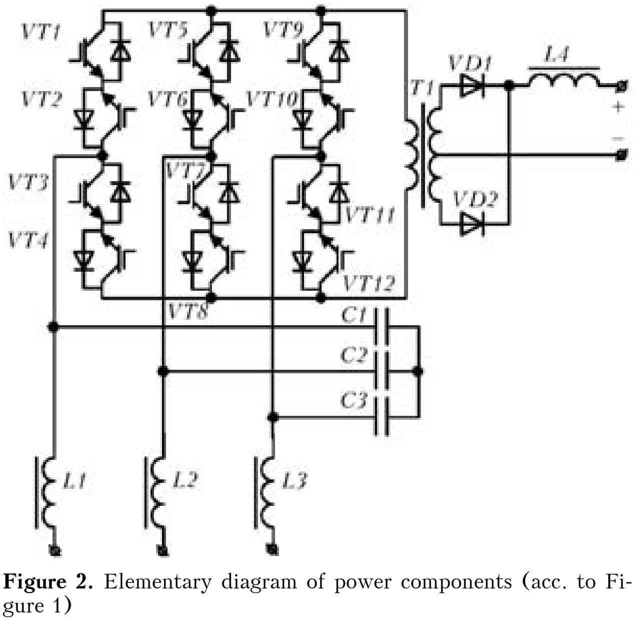

Figure 2. Elementary diagram of power components (acc. to Figure 1)

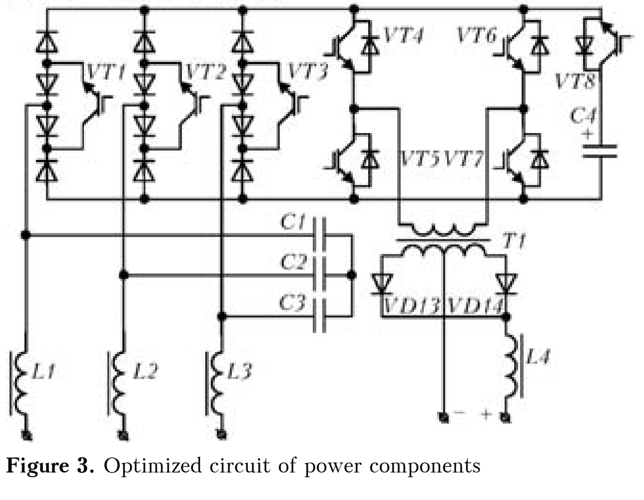

Figure 3. Optimized circuit of power components

To check the theoretical results, a test low-power mock-up of the source was assembled by the circuit shown in Figure 3. Oscillographing of its consumed current, phase supply voltage and PF measurement showed the low level of nonlinear distortions of the current curve and high PF — more than 0.95. These experimental results fully confirm the serviceability of circuit designs, and correctness of theoretical principles presented in the work.

Schematic in Figure 3 can be implemented using mass-produced power units for frequency converters, accommodating power transistors and control and protection circuits in one case. So, elements *VT1*–*VT3* and *VD1*–*VD12* can be replaced by three specialized power modules VUI3012N1 (IXYS), and bridge *VT4*–*VT7* can be of MKI50-12E7 or MKI100-12F8 type. MEK600-04DA assembly can be used as *VD13* and *VD14* diodes. Application of the above power units will enable achieving up to 15 kW power in the load at power supply from the mains with 380–460 V line voltage.

If it is necessary to increase power, higher-power components can be used in the power circuit. However, increase of output current can be also achieved by parallel joining of several sources and respective synchronizing of their control systems. Summing of output currents of several converters is often more cost-effective, than construction of one powerful source. So, three inverters, made by the circuit in Figure 3, can provide output current of up to 1000–1200 A at working voltage of 30–36 V, thus allowing their application in automatic welding technologies. Detuning of switching frequencies of individual sources within narrow limits will allow lowering spectral density of emitted electromagnetic noise.

Control signals can be formatted using specialized DSP processors (for instance, ADSP21xx of Analog Devices) or single-crystal microcontrollers (for instance, AVR ATMega or ARM Cortex STM32).

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INVESTIGATION OF FATIGUE RESISTANCE OF WELDED JOINTS ON ALUMINIUM ALLOYS MADE BY MODERN WELDING METHODS

The research work on the above subject was completed in 2011 by the E.O. Paton Electric Welding Institute (supervisor – Prof. V.I. Kyrian)

Quantitative estimation of factors (stress concentration, residual stresses, etc.) caused by the modern processes of welding of thin-sheet aluminium alloys (TIG, MIG, FSW), which affect service properties of the welded joints, was performed. It was established that the residual stressed state can be simulated with comparatively narrow (80–100 mm width) specimens in fatigue tests of the welded joints on 1–3 mm thick aluminium alloys. Fatigue resistance of the welded joints made by the above methods was investigated. Optimal parameters were identified for high-frequency mechanical peening (HFMP) of the welded joints on thin-sheet aluminium alloys to improve their fatigue resistance by bringing it closer to the level of the base metal. It was proved that HFMP is an efficient method for reducing the concentration of stresses caused not only by convexity of the weld but also by angular deformation. The investigations completed showed a high potential of widening of a range of thicknesses of aluminium alloys with different alloying systems from 1 to 3 mm for the high-productivity MIG welding technology (in contrast to requirements of GOST 14806–80) to manufacture transport-application structures operating under alternating loading conditions. The values of fatigue limits of the welded joints required for design and evaluation of service life of the transport-application structures were calculated