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RESONANT CONVERTER AS A TRANSFORMER FOR VARYING THE RATIO BETWEEN INPUT AND OUTPUT CURRENTS

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Abstract. Resonant AC-DC converters of 50(60) Hz are described. In these converters, at each half-cycle of network voltage, the capacitor and the inductor of an oscillatory circuit are switched from series into parallel and vice versa. The duration of series and parallel connection and the transformer ratio are parametrically dependent on the load. In the case of short circuit, only the parallel oscillating circuit operates. This restricts the output current sharply. The reactive power of the capacitor and that of the inductor compensate each other both in the case of series and parallel connection. Therefore the power factor is very high, ranging from no-load to short-circuit. This converter suits very well for power supply of arc furnaces.

Key words: resonant converter, power factor, electrical arc, flicker, higher harmonics.

1. INTRODUCTION

As consumers, arc furnaces have specific requirements for the network. To ensure stability of the arc, power supply voltage has to be in simultaneous correspondence with arc parameters that vary fast and irregularly in time. In case supply voltage is higher than needed for the arc at that moment, current will increase rapidly, since the dynamic resistance of the arc is close to zero. Current increase could be avoided by means of a thyristor rectifier. By controlling thyristors, arc current can be kept approximately constant in all modes, including short circuit. If the voltage of the arc varies, the active power will also vary. If the active component of the line-side current decreases, the reactive component will increase, because ordinary rectifiers have approximately invariable line current when the output current is constant. In spite of stabilizing the line current, fluctuations of the reactive component of the current will occur, causing voltage fluctuations. To avoid voltage fluctuations, it is often necessary to use a fast reactive power compensator together with a higher harmonics filter.

In his graduation thesis, in 1978, Janson suggested that an AC-DC converter with parametric reactive power compensation could solve the problems of power supply for arc furnaces. This converter was built on the principle of a "loss-free resistor". If such a hypothetical "loss-free resistor", connected in series with the arc were available, the arc could easily be supplied from the AC network. There are no reactive components of the current, nor are there any fluctuations of the reactive component. This idea could probably be realized if the circuit connected in series with the arc includes inductive as well as capacitive reactances. For example, if we had two separate resistances instead of the electric arc, a simple circuit in Fig. 1*a* could be used.

In case the inductive and capacitive reactances are equal, $x_L = x_C$, and resistors R_1 and R_2 vary, such that at each instant, their resistances are equal, $r_1 = r_2$, then the current of the inductive branch $\dot{I}_L$ and that of the capacitive branch $\dot{I}_C$ are always equal in size. The phase shift angles φ_L and φ_C of these

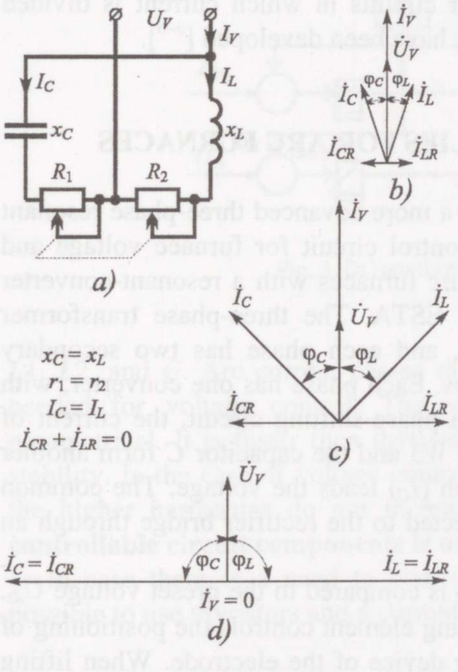

Fig. 1. The circuit limiting short-circuit current compensating the reactive component of the current (*a*) and its vector diagrams for three different loads: *b*, $r = 3x$; *c*, $r = x$; *d*, $r = 0$.

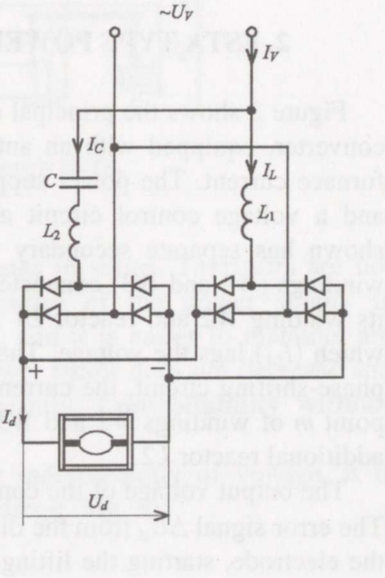

Fig. 2. Power supply converter for an arc furnace.

currents on the vector diagram (Fig. 1*b*, *c*, and *d*) are always equal in size, but with different signs. In the case of load variations, the reactive components of the branch current $\dot{I}_{CR}$ and $\dot{I}_{LR}$ vary within substantial limits, but their sum is always equal to zero and there is no reactive component in the mains current.

In case the currents of load resistors R_1 and R_2 are rectified and the resistors are switched in parallel, only one load resistor can be used instead of two.

If a variable resistor is replaced by an electric arc, the higher harmonics and the reactive component will increase. In addition, higher harmonics occur in the capacitive branch. This drawback can be reduced significantly when an additional inductance L_2 is connected in series with the capacitance C (Fig. 2).

Such a circuit, patented by Brandli and Dick [1] and used for power supply for high-pressure lamps, provides a satisfactor operation in every aspect. However, in this circuit, the installed power of reactive components is still relatively high (capacitor battery about 1.5 kvar/kW and total inductors' power about 1.1 kvar/kW). As a result of testing the first converter with parametric reactive power compensation at Tallinn Technical University (TTU) in 1978, a circuit with less installed power was invented (capacitor battery 0.7 kvar/kW and the inductor 0.5 kvar/W). This circuit, incorporating a transformer with two secondary windings connected in series, was granted the USSR Authorship Certificate [2] in 1984. Numerous similar circuits in which current is divided between inductive and capacitive branches have been developed [3–6].

2. ESTA TYPE POWER SUPPLIES FOR ARC FURNACES

Figure 3 shows the principal circuit of a more advanced three-phase resonant converter, equipped with an automatic control circuit for furnace voltage and furnace current. The power supplies for arc furnaces with a resonant-converter and a voltage control circuit are called ESTA. The three-phase transformer shown has separate secondary windings, and each phase has two secondary windings (*W*2 and *W*3) connected in series. Each phase has one converter, with its winding *W*2 and reactor *L*1 forming a phase-shifting circuit, the current of which (I_{L1}) lags the voltage. The winding *W*3 and the capacitor *C* form another phase-shifting circuit, the current of which (I_C) leads the voltage. The common point *m* of windings *W*2 and *W*3 is connected to the rectifier bridge through an additional reactor *L*2.

The output voltage of the converter U_d is compared to the preset voltage U_{ds}. The error signal ΔU_d from the discriminating element controls the positioning of the electrode, starting the lifting-lowering device of the electrode. When lifting the electrode, the arc will become longer. Arc voltage is approximately proportional to its length. Therefore it is possible to control arc voltage by lifting-lowering it. The output voltage of the converter changes automatically according to arc voltage due to voltage drop changes on the reactive elements

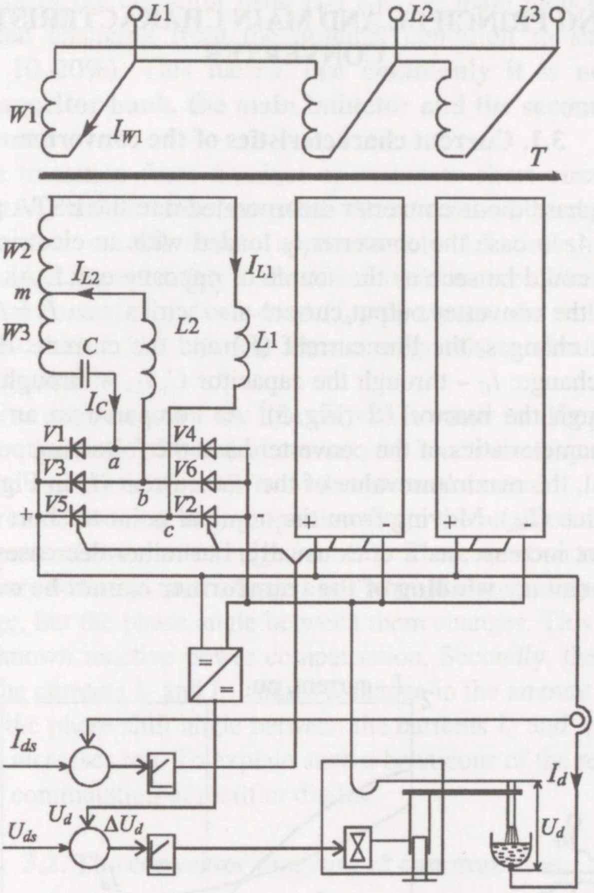

Fig. 3. DC arc furnace power supply ESTA.

$L1$, $L2$, and C. Arc current passes these elements in series. Thyristors are not needed for voltage control. The voltage control of the power supply is parametrical. It is faster than thyristor control, and it is easier to maintain arc stability. In the case of voltage control, the power factor does not decrease and the higher harmonics do not increase significantly. **Controllability without controllable circuit components is unusual**.

In case there is a need to control current independently of voltage, it is possible to use thyristors and a current control circuit (Fig. 3).

3. OPERATING PRINCIPLE AND MAIN CHARACTERISTICS OF THE CONVERTER

3.1. Current characteristics of the converter

The single-phase diode converter incorporated into the ESTA power source is shown in Fig. 4. In case the converter is loaded with an electric arc varying in length, the arc could be seen as the source of opposite e.m.f. As the electric arc voltage varies, the converter output current also varies, i.e., $I_d = f(U_d)$. When the opposite e.m.f. changes, the line current (I_V) and the currents through reactive elements also change: I_C – through the capacitor C, I_{L1} – through the reactor $L1$, and I_{L2} – through the reactor $L2$ (Fig. 5). As compared to an ordinary diode rectifier, the characteristics of the converter have the following peculiarities.

1. In general, the maximum value of the line current (I_V in Fig. 5) is the same as the rated value (I_{Vn}). Moving from the nominal point to short circuit, the line current does not increase, as it does usually, but rather decreases. **By changing the load, the primary winding of the transformer cannot be overloaded**.

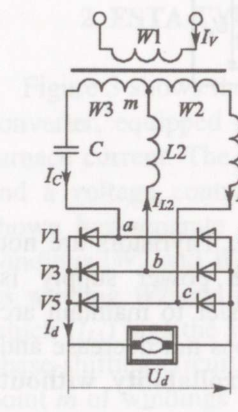

Fig. 4. Single-phase converter with two secondary windings.

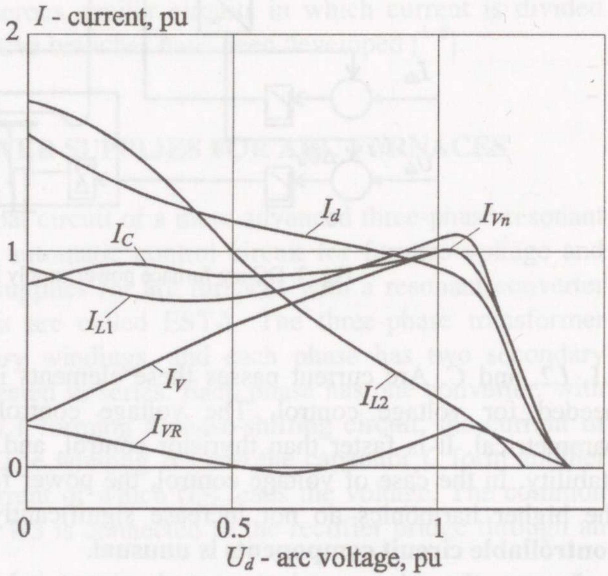

Fig. 5. The output current (I_d), network current (I_V), capacitor bank current (I_C), current of the main inductor (I_{L1}) and additional inductor (I_{L2}) as a function of the arc voltage U_d of the converter with a single-phase diode-rectifier used in the ESTA power supply.

2. The current of the capacitor bank (I_C) and the main inductor (I_{L1}) do not vary much when changing from the nominal condition to short circuit (in general, up to 10–20%). This means, that **commonly it is not possible to overload the capacitor bank, the main inductor and the secondary windings of the transformer**.

3. When the transition from nominal operation to short circuit occurs, the rectified current (I_d) will increase more than the currents of the secondary circuit I_C and I_{L1}, but this is still small (approximately 1.7 times). Such change corresponds to about the operating mode of a current source. When the transition from the nominal current (I_{dn}) to no-load ($I_d = 0$) occurs, the voltage (U_d) does not change more than 1.3–1.7 times. This corresponds to about the operating mode of a voltage source.

4. The reactive component of the line current is small in all operation modes (I_{VR} in Fig. 5). In the range of nominal current, the reactive component is usually inductive and the amount is 10–40% from the nominal current.

The line current I_V does not vary proportionally to the rectified current I_d like in the case of a conventional diode bridge rectifier. There are two reasons for this. Firstly, the primary current I_V will change even when the secondary currents I_C and I_{L1} do not change, but the phase angle between them changes. This corresponds to the commonly known reactive power compensation. Secondly, the change of the phase angle of the currents I_C and I_{L1} causes a change in the amount of the rectified current. In case the phase shift angle between the currents I_C and I_{L1} increases, the rectified current increases too. To explain such a behaviour of the rectified current, let us follow the commutation of rectifier diodes.

3.2. The converter diagram of commutation

In a conventional rectifier, during commutation period, the transformer windings participating in the commutation process are short-circuited. In a resonant converter, reactive elements *C*, *L*1 and *L*2 are connected between transformer windings and rectifying bridge input and, as a result, for the transformer during commutation period, short circuit does not exist. The converter in Fig. 3 has a three-phase rectifier in one phase, inputs *a*, *b* and *c*, which are supplied by different reactive elements. That is why three different commutation stages and commutation angles should be distinguished. If points *a* and *c* are connected through diodes *V*1, *V*5 (or *V*2, *V*4), the commutation angle is γ_p. In the same way, if points *a* and *b* are connected, the angle is γ_C and if points *b* and *c* are connected, the angle is γ_L.

For the converter shown in Fig. 4, the relationship between the commutation instants and the relative output voltage U_d is given in Fig. 6. The curves of turn-on and turn-off instants (or commutational instants) of diodes are given in electrical degrees of the supply voltage (angle υ in the range –30 to +210°). In addition to commutational curves, curves of instant values of the capacitor

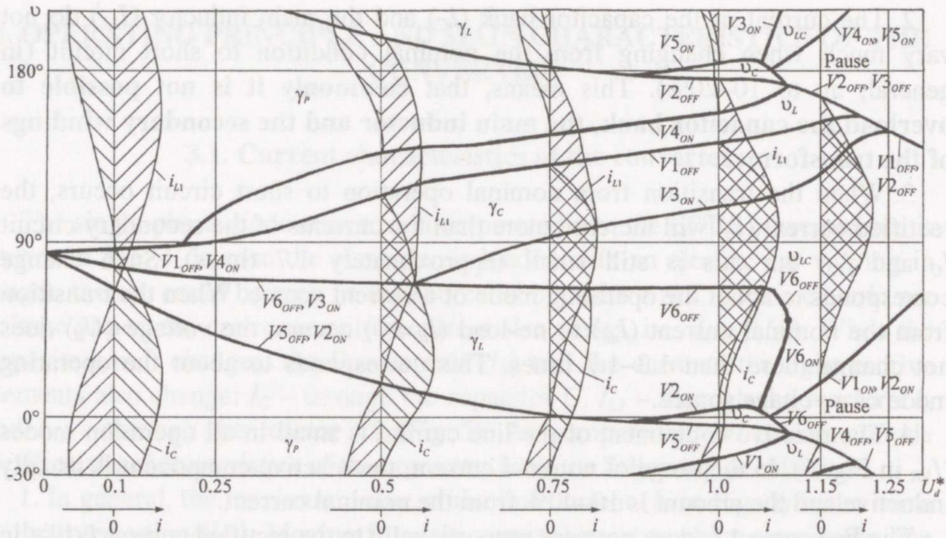

Fig. 6. Curves of turn-on and turn-off instants of the converter diode rectifier bridge of one-phase according to the relative output voltage of the converter (U_d^*) and curves of the instant value of the current for the capacitive branch (i_C) and inductive branch (i_{L1}) in the case of five output voltages.

current (i_C) and the main inductor current (i_L) are shown. The current curves are given for the relative output voltage U_d^* values 0.1, 0.5, 0.75, 1.0, and 1.15. Between the commutations there are three different intervals denoted by υ_L, υ_C, and υ_{LC}. During the interval υ_L, current passes inductors $L1$, $L2$ and the winding $W2$. During the interval υ_C, current passes the capacitor C, the winding $W3$, and the inductor $L2$. During the interval υ_{LC}, current passes the inductor $L1$, windings $W2$, $W3$, and the capacitor C.

As shown in Fig. 6, currents i_C and i_{L1} are of opposite direction during the commutation angle γ_p. These currents also pass the secondary windings of the transformer in the opposite direction. The sum of currents goes through the common point m of the windings to the inductor $L2$. Therefore transformer windings $W2$ and $W3$ are switched in parallel during γ_p. If the current is zero in one winding, then it can be regarded as a special case of parallel operation of the transformer windings. The current in one of the secondary windings is equal to zero in the intervals υ_L and υ_C between the commutations. During the interval υ_{LC} between commutations, currents i_C and i_{L1} are of the same direction and equal. In addition, the currents in the transformer windings $W2$ and $W3$ are of the same direction and equal. Transformer windings are then operating in series. During commutation intervals γ_L and γ_C, currents i_C and i_{L1} are of the same direction, although they are not equal in size. Within these intervals, transformer windings are operating in series, but loaded unequally.

From Fig. 6 follows that during intervals υ_L, γ_p, and υ_C, the transformer windings are operating in parallel and during intervals γ_L, υ_{LC}, and γ_C, these windings are operating in series.

3.3. Variation of the transformer ratio

Let us explain the claim that the secondary windings of the transformer are switched from series to parallel. The well-known switch-over circuit of the transformer windings from series to parallel is shown in Fig. 7*a*. When switching over, the switch shown first disconnects contacts 3 and 4 which are normally closed, connecting the beginning of one winding and the end of another. Then, the beginnings of the windings are connected by the help of contacts 1 and 2. As a result, the beginning and the end of the winding *W*2 are changed (before the switch-over the common point *m* was connected to the end of the winding *W*2 and after the switch-over to the beginning). The switch-over of the beginning and the end of windings is equivalent to the phase shift of 180°. The common point is not disconnected in the converter. The phase shift of 180° is created so that the direction of the current vector of the inductive branch *W*2 (I_{L1} in Fig. 7) changes 90° in the lagging direction and the current vector direction of the capacitive branch *W*3 (I_C in Fig. 7) changes 90° in the leading direction.

The switch-over circuit shown in Fig. 7*a* incorporates contacts 5 and 6 connecting the ends of windings in the case of parallel connection. The converter does not include these contacts and they are replaced by diodes. The common point *n* of the simultaneously opened diodes *V*1 and *V*5 (Fig. 7*b*) is equivalent to the connection of contacts 5 and 6 in Fig. 7*a*. In fact, all the operations are done in the converter, similarly to the well-known switch-over circuit in Fig. 7*a*. However, the converter uses other means for this purpose.

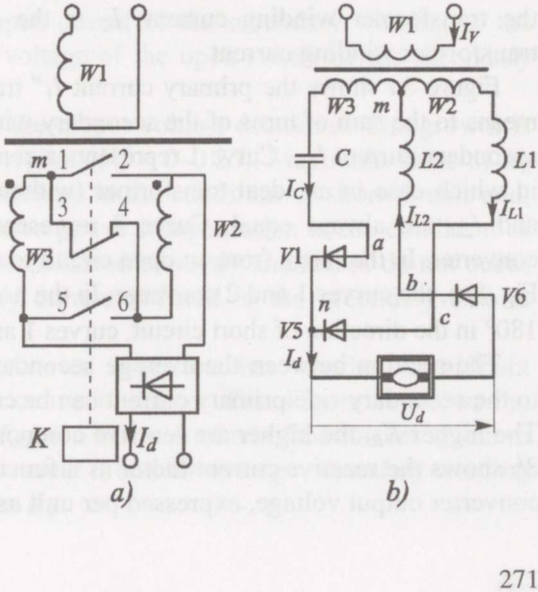

Fig. 7. Switch-over of transformer windings from series to parallel by the help of relay contacts (*a*) and by reactive elements and rectifier bridge diodes (*b*).

Switch-over from series operation to parallel takes place each half-cycle. When the duration of this operation increases, the average transformation ratio of one half-cycle also increases. **Such phenomena of smooth variation of the transformer ratio are unusual and original**.

In a conventional diode rectifier with the matching transformer and bridge connection, the line current I_V in the active load condition is proportional to the rectified current I_d

$$I_V = K_{TR} \cdot K_f \cdot I_d, \tag{1}$$

where K_{TR} is matching transformer transformation ratio; $K_{TR} = w1/w2$, where $w1$ and $w2$ are numbers of turns of transformer primary and secondary windings, respectively; K_f is the factor which takes into account the relation between the current RMS value (in the case of I_V) and the average value (in the case of I_d).

The relation between primary and secondary currents can be expressed as

$$I_2 = K_{TR} \cdot I_1, \tag{2}$$

where I_2 is the transformer secondary current, I_1 is the transformer primary current equal to the line current $I_1 = I_V$.

The parametrically compensated converter has two secondary windings with different currents and voltages. To describe the current transformation, it is convenient to use the average secondary current defined as follows:

$$I_{2M} = 0.5(K_{UL}I_L + K_{UC}I_C), \tag{3}$$

where K_{UL} is the voltage factor of the inductive branch of the transformer winding, $K_{UL} = 2w_L/(w_L+w_C)$; K_{UC} is the voltage factor of the capacitive branch of the transformer winding, $K_{UC} = 2w_C/(w_L+w_C)$; w_L is the number of turns of the inductive branch of the transformer winding; w_C is the number of turns of the capacitive branch of transformer winding; I_L is the inductive branch, containing the transformer winding current; I_C is the capacitive branch, containing the transformer winding current.

Figure 8*a* shows the primary current I_1'' transformed to the secondary side (it means to the sum of turns of the secondary windings) as a function of the average secondary current I_{2M}. Curve 1 represents a conventional two winding transformer, in which case in an ideal transformer (without magnetizing current) currents I_1'' and I_{2M} are always equal. Curve 2 represents the parametrically compensated converter. In the range from an open circuit to a rated load (currents from 0 to 1 in Fig. 8*a*), the curves 1 and 2 converge. In the nominal point, curve 2 turns for about 180° in the direction of short circuit, curves 1 and 2 differ.

The relation between the average secondary current I_{2M} and the transformed to the secondary side primary current can be called the reactive current factor K_R. The higher K_R, the higher are reactive components in secondary currents. Figure 8*b* shows the reactive current factor as a function of parametrically compensated converter output voltage, expressed per unit as follows:

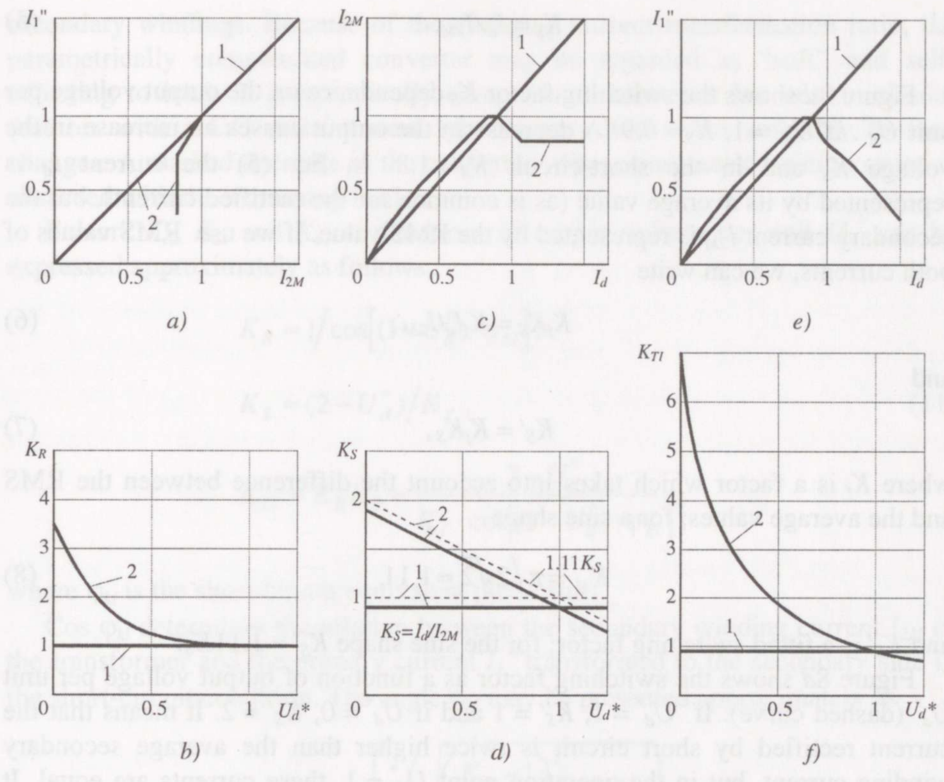

Fig. 8. Dependence of I_1'' on I_{2M}, of I_{2M} and I_1'' on I_d, and of K_R, K_S, and K_{TI} on U_d^* for a simple single-phase diode rectifier (curve 1) and load adapting resonant converter (curve 2).

$$U_d^* = U_d/(U_{WC0} + U_{WL0}), \tag{4}$$

where U_{WC0} is the voltage of the open circuit of the secondary winding of the capacitive branch, and U_{WL0} is the voltage of the open circuit of the secondary winding of the inductive branch.

The value of K_R increases when the output voltage decreases. The growth is especially sharp in the short-circuit region ($U_d^* = 0$–0.25). Maximum value of K_R depends first, on the losses in the scheme in the condition of short circuit and, second, on the accuracy of the reactive power compensation in the condition of short circuit. If reactive power is compensated correctly and losses do not occur (in an ideal case), then the primary current reduced to the secondary side is $I_1'' = 0$.

Parametrically compensated converter operation has another significant difference as compared to that of a conventional rectifier. Figure 8*c* shows the average secondary winding current I_{2M} as a function of rectified current I_d (curve 2). The relation between the output current I_d and the average secondary winding current can be called switching factor K_S

$$K_S = I_d/I_{2M}. \tag{5}$$

Figure 8*d* shows the switching factor K_S dependence on the output voltage per unit U_d^*. If $U_d^* = 1$, $K_S \approx 0.9$. A decrease in the output causes an increase in the voltage K_S and in the short-circuit $K_S \approx 1.8$. In Eq. (5) the current I_d is represented by its average value (as is common for the rectified current), but the secondary current I_{2M} is represented by the RMS value. If we use RMS values of both currents, we can write

$$K_f K_S = K_f I_d/I_{2M}, \tag{6}$$

and

$$K_S' = K_f K_S, \tag{7}$$

where K_f is a factor which takes into account the difference between the RMS and the average values; for a sine shape

$$K_f = \pi/2\sqrt{2} \approx 1.11, \tag{8}$$

and K_S' is a fitted switching factor; for the sine shape $K_S' \approx 1.11 K_S$.

Figure 8*d* shows the switching factor as a function of output voltage per unit U_d^* (dashed curve). If $U_d^* = 1$, $K_S' = 1$ and if $U_d^* = 0$, $K_S' = 2$. It means that the current rectified by short circuit is twice higher than the average secondary winding current, but in the operation point $U_d^* = 1$, these currents are equal. It becomes possible because in the short circuit, both secondary winding currents are rectified. In the operation point $U_d^* = 1$, both secondary windings draw seriously rectified current.

Figure 8*e* shows the dependence of the primary current I_1'' of the transformer transformed to the secondary side on the rectified current I_d (curve 2). The relation of the output current I_d and the current transformed to the secondary side line current (without magnetizing current) can be called the transformation ratio of the converter current, expressed as

$$K_{TI} = I_d / I_1. \tag{9}$$

The current transformation ratio can be also written as

$$K_{TI} = K_R \cdot K_S. \tag{10}$$

Figure 8*f* shows the current transformation ratio K_{TI} as a function of U_d^* (curve 2). When compared to a conventional diode bridge rectifier (curve 1) in which the relation between the supplied current and the rectified current is constant, a marked difference is obvious. In transition from the open circuit to the short-circuit condition, the current transformation ratio changes for about ten times. The changes of the current transformation ratio K_{TI} are caused by the reactive power compensation in the converter and by the parallel operation of the secondary windings. Because of the changing current transformation ratio, the parametrically compensated converter may be regarded as "soft" and self-adjusting to load. A parametrically compensated converter has certain similar features with a DC main current type engine. Both react smoothly to load changes. A marked decrease in the converter voltage and motor rotation speed is observed with the increasing load.

Factors K_R, K_S, and K_{TI} as functions of output voltage per unit U_d^* can be expressed approximately as follows:

$$K_R \approx 1 \big/ \cos\left[(1-U_d^*)\cdot\varphi_{kl}\right],$$

$$K_S \approx (2-U_d^*) \big/ K_f, \tag{11}$$

$$K_{TI} \approx K_R K_S \approx \frac{2-U_d^*}{K_f \cdot \cos\left[(1-U_d^*)\cdot\varphi_{kl}\right]},$$

where φ_{kl} is the short-circuit equivalent phase shift.

Cos φ_{kl} determines the relation between the secondary winding current I_{2M} of the transformer and the primary current I_1'' transformed to the secondary side in the short-circuit condition. The angle φ_{kl} may be presented approximately as

$$\varphi_{kl} = \arccos\left(1 \Bigg/ \sqrt{\left(\frac{K_Q-1}{K_Q+1}\right)^2 + K_{\eta K}^2}\right), \tag{12}$$

where K_Q is the reactive power compensation factor for the short circuit, $K_Q = \Sigma Q_{LK}/Q_{CK}$; ΣQ_{LK} is the sum of reactive powers of the reactors $L1$ and $L2$ and reactive powers of transformer leakage inductance in the short circuit; Q_{CK} is the reactive power of the capacitor bank in the short circuit; $K_{\eta K}$ is the active losses factor for short circuit, $K_{\eta K} = P_K/P_N$; P_K is the power losses of the converter in the short circuit; P_N is the converter-rated active power.

The relation of the line current I_V and the rectified current I_d can approximately be determined by the transformation factor of the converter current as follows:

$$I_V = \frac{K_{TR}}{K_{TI}} \cdot I_d \approx \frac{K_{TR} \cdot K_f \cdot \cos\left[(1-U_d^*)\cdot\varphi_{kl}\right]}{2-U_d^*} \cdot I_d, \tag{13}$$

where K_{TR} is the transformer turns ratio, $K_{TR} = w_1/(w_C+w_L)$, and w_1 is the number of turns of the primary winding.

3.4. Vector diagram for the circuit in the parallel mode

In the case of short-circuit and very low rectified voltage, the converter operation can be simplified and shown such that the series connection of the windings does not arise, and that the currents are sinusoidal. Then the vector diagram can be used as shown in Fig. 9. This vector diagram is composed for the common point *m* of the transformer secondary windings (Fig. 4).

The voltage vectors $\dot{U}_{W2}$ and $\dot{U}_{W3}$ of the secondary windings are in opposite directions relative to point *m* (Fig. 9) because in this point, the beginning of one winding and the end of the other one are connected. The current vector $\dot{I}_{L1}$ of the inductive branch is lagging relative to the voltage vector $\dot{U}_{W2}$ of the same branch and the current vector $\dot{I}_C$ of the capacitive branch is leading with respect to the voltage vector $\dot{U}_{W3}$ of its branch. Currents $\dot{I}_C$ and $\dot{I}_{L1}$ join in the auxiliary inductor *L*2.

The secondary winding currents $\dot{I}'_C$ and $\dot{I}'_L$ reduced to the primary winding *W*1 are almost in the opposite phase in the case of short circuit. The reactive components of these currents compensate each other. The primary current is approximately in phase with voltage.

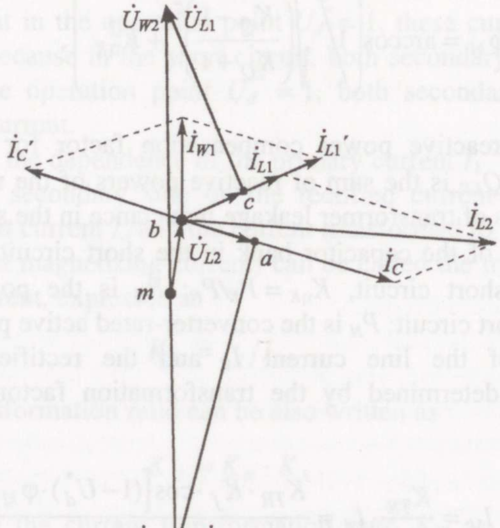

Fig. 9. Vector diagram of currents and voltages for the mode close to short circuit, for the common point *m* of the secondary windings.

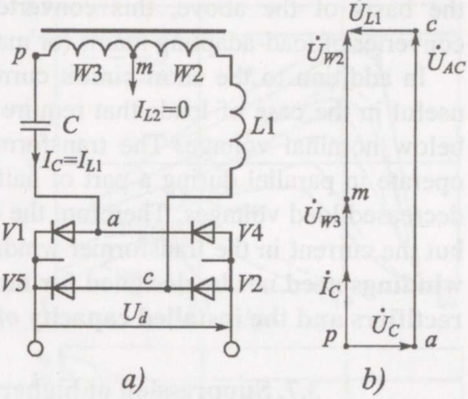

Fig. 10. Equivalent circuit for small load currents (*a*) and the corresponding vector diagram (*b*).

3.5. Vector diagram for the circuit in series mode

In the case of loads less than the nominal mode ($U_d^* > 1.0$), the converter operation can be simplified and shown as there is no parallel connection of windings in this mode and because the currents are sinusoidal. Then the equivalent circuit and the vector diagram are as shown in Fig. 10.

The total current flows through both secondary windings of the transformer and the reactive elements *C* and *L*1 in series. After rectifying, the load draws the same current. Voltage drops on the capacitor *C* and the inductor *L*1 mutually compensate each other. This is shown on the vector diagram in Fig. 10.

Approximately the whole secondary voltage is applied to the rectifying bridge and the load. The current is in phase with voltage as a result of the mutual compensation of reactive power of reactive elements.

3.6. Self-adjustment to load

Switch-over of the reactive elements from series connection to parallel and back occurs together with the switch-over of transformer windings. In most operating modes, including nominal, such switch-over occurs each half-cycle. The parallel oscillatory circuit is current restricting during short circuit. For the supply network, its impedance is high and low current is drawn from the network. Thereby the current in the oscillatory circuit is much higher, and this higher current is rectified. On the contrary, the previous impedance of the series oscillatory circuit is very low. In this case, the current drawn from the network is determined by the load (arc) impedance. The circuit operates then as an ordinary rectifier. When arc voltage changes, the ratio between parallel connection and series connection changes as well. The operating mode of the converter is self-adjusting to the arc voltage. The range of self-adjustment is from no-load to short circuit. The converter characteristics are opposite at the ends of this range. On the basis of the above, this converter may be called a frequency resonant converter of load-adapting mains (or mains frequency LAR-converter).

In addition to the short circuit current limitation, self-adjustment to load is useful in the case of loads that require more current when voltage is decreased below nominal voltage. The transformer windings of the LAR-converter will operate in parallel during a part of half-cycle of network voltage in the case of decreased load voltages. Therefore the output current of the converter increases, but the current in the transformer windings does not increase. **The transformer windings need not be designed for maximum load current like in traditional rectifiers and the installed capacity of the transformer will be less.**

3.7. Suppression of higher harmonics in the converter

Reactive elements C and $L1$ form a series oscillatory circuit tuned approximately to network frequency. During the interval υ_{LC} (Fig. 6), the oscillatory circuit $L1$, C is switched in series with the rectifier bridge. This oscillatory circuit is a filter for higher harmonics. Due to long commutational angles (if $U_d < 1.0$, then $\gamma_p = 18°$, $\gamma_C = 26°$ and $\gamma_L = 63°$), the speed of current variations is also low. This also decreases the amount of higher harmonics in the network. Maximal values of higher harmonics in the line current compared to the rated current are approximately the following: 5th harmonic – 6%, 7th harmonic – 4%, 11th harmonic – 1.5%.

3.8. Partial thyristor control

Control of the current by replacing diodes $V2$ and $V5$ (Fig. 3) by thyristors is possible only partly, because the diodes $V1$, $V3$, $V4$, $V6$ do not change the rectified voltage. If the arc voltage U_d is higher than half of the nominal voltage U_{dn} ($U_d > 0.5U_{dn}$), the current may be controlled freely without restrictions (Fig. 11*d*). In the case of lower arc voltages ($0 < U_d < 0.5U_{dn}$), it is only possible to reduce the current up to half of the nominal.

Increasing the firing angle α, the apparent power S and the active power P from the network decrease. Moreover, in most modes, the apparent power is a little higher than the active power (Fig. 11*b*).

By control, the reactive power Q (Fig. 11*c*) increases. The reactive power is maximal when in the case of nominal arc voltage, the firing angle $\alpha = 75°$ is used. But the maximal values of Q are still about three times less than the rated power P_n. This is considerably less than in a usual thyristor rectifier, where maximal reactive power exceeds active power.

Controlling current with thyristors, the harmonics in the network current also change. The changes of the 5th harmonic are shown in Fig. 11*a*. A slight increase in the 5th harmonic (from 6% to 7.7%) occurs only in some modes. Regarding other harmonics, no considerable increases occur.

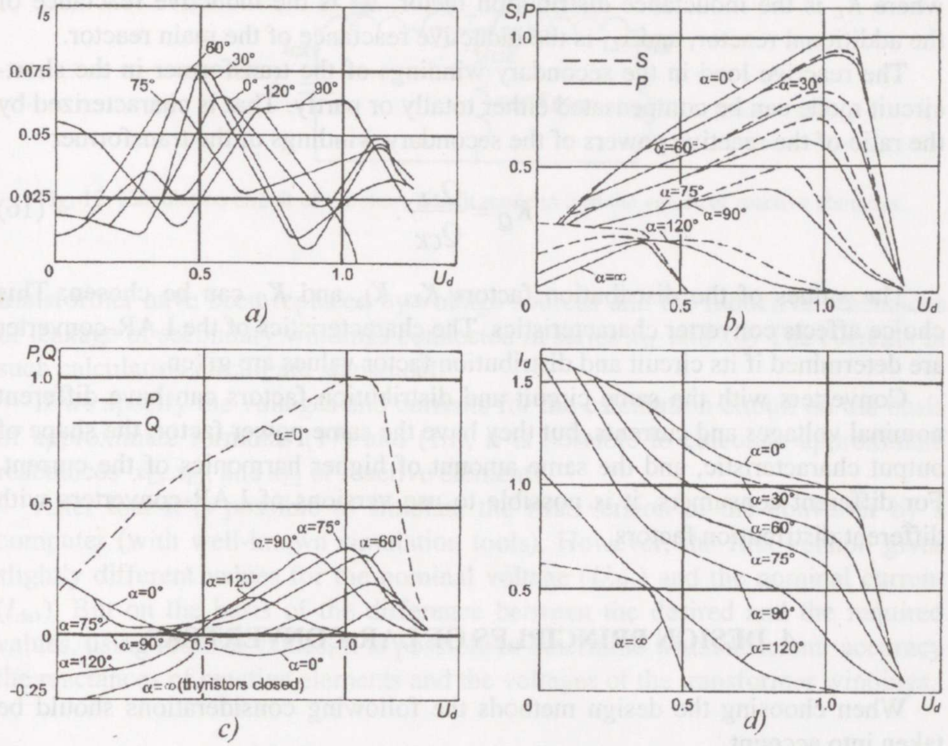

Fig. 11. Variation of the ESTA power supply parameters as a function of the arc voltage U_d in the case of different thyristor firing angles α; *a* – 5th harmonic in the network current, *b* – apparent power *S* and active power *P*, *c* – active power *P* and reactive power *Q*, *d* – rectified current I_d.

3.9. Variability of converter characteristics and parameter distribution factors

One phase of the converter transformer includes two secondary windings (*W*2 and *W*3 in Fig. 3). Voltages of these windings can be chosen equal or different. This choice is characterized by the ratio of no-load voltages

$$K_E = \frac{U_{WC0}}{U_{WL0}}. \tag{14}$$

Factor K_E may be called the distribution factor of converter voltage.

One phase of the converter includes two reactors (*L*1 and *L*2 in Fig. 3). The ratio between the inductive reactances of the additional reactor (*L*2) and the main reactor (*L*1) is expressed as

$$K_L = \frac{x_{L2}}{x_{L1}}, \tag{15}$$

where K_L is the inductance distribution factor, x_{L2} is the inductive reactance of the additional reactor, and x_{L1} is the inductive reactance of the main reactor.

The reactive load in the secondary windings of the transformer in the short-circuit mode can be compensated either totally or partly. That is characterized by the ratio of the reactive powers of the secondary windings of the transformer

$$K_Q = \frac{Q_{LK}}{Q_{CK}}. \tag{16}$$

The values of the distribution factors K_E, K_L, and K_Q can be chosen. This choice affects converter characteristics. The characteristics of the LAR-converter are determined if its circuit and distribution factor values are given.

Converters with the same circuit and distribution factors can have different nominal voltages and currents, but they have the same power factor, the shape of output characteristic, and the same amount of higher harmonics of the current. For different consumers, it is possible to use versions of LAR-converters with different distribution factors.

4. DESIGN PRINCIPLES OF LAR-CONVERTERS

When choosing the design methods the following considerations should be taken into account.

1. Nominal voltage of the converter U_{dn} is approximately equal to the sum of voltages of the one-phase secondary windings of the transformer

$$U_{dn} \approx U_{WC0} + U_{WL0}. \tag{17}$$

2. The level of the current in the secondary windings of the transformer does not vary significantly when the transition from nominal operation to short circuit occurs (Fig. 5). In the case of single-phase converter, this current of secondary windings is also approximately equal to the nominal current of the converter I_{dn}

$$I_{dn} \approx I_{WCK} \approx I_{WLK}, \tag{18}$$

where I_{WCK} and I_{WLK} are currents of secondary windings in the case of short circuit.

3. For the short-circuit mode, an equivalent calculation circuit shown in Fig. 12 may be used. The rectifier bridge is not shown. The voltage drop at the short-circuited rectifier bridge is approximately $2V$. This is much smaller than voltage drops on reactive elements, and such voltage drop does not affect significantly the currents of the reactive elements. As the sum of the secondary currents reduced to the primary winding of the transformer is in the case of short circuit very small (equal to zero in an ideal case), the primary winding and the magnetic core have also been ignored. The secondary windings of the

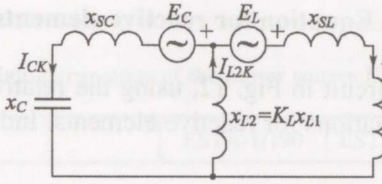

Fig. 12. Calculation circuit of the short-circuit mode to find the values of reactive elements.

transformer have been replaced by voltage sources and the inductive reactances of leakage of secondary windings connected in series x_{SC} and x_{SL}. The currents in such calculation circuit are sinusoidal.

If we specify the voltages and currents for the calculation circuit on the basis of approximate formulas (17) and (18), it is possible to calculate approximate reactances x_C, x_{L1}, and x_{L2} of reactive elements.

After that it is possible to simulate the first version of the converter by a computer (with well-known simulation tools). However, the first version gives slightly different values for the nominal voltage (U_{dn1}) and the nominal current (I_{dn1}). But on the basis of the difference between the desired and the required values, using relative values, it is possible to determine with sufficient accuracy the reactances of reactive elements and the voltages of the transformer windings.

4.1. Special system of relative values

Usually the system of relative values is based on the nominal current and the nominal voltage. For the above calculation method, it is still more suitable to use a system where the basis of voltage is the sum of voltages of two single-phase secondary windings in the case of no-load

$$U_B = U_{WC0} + U_{WL0}, \tag{19}$$

where U_{WC0} is no-load voltage of the capacitive branch, U_{WL0} is no-load voltage of the inductive branch, and for the basic power of one phase, the sum of apparent powers of secondary windings in short-circuit apparent powers is defined as

$$S_{BP} = U_{WC0} I_{WCK} + U_{WL0} I_{WLK}. \tag{20}$$

The basic current of one phase I_{BP} and the basic resistance z_{BP} can be written as

$$I_{BP} = \frac{S_{BP}}{U_B}, \tag{21}$$

$$z_{BP} = \frac{U_B^2}{S_{BP}}. \tag{22}$$

4.2. Equation for reactive elements

From the equivalent circuit in Fig. 12, using the relative values given above, it is possible to derive equations for reactive elements. Inductive reactance of the main reactance is

$$x_{L1} = \frac{z_{BP}\left[K_E(K_Q+1) - x_{sL}^* \cdot K_E \cdot K_Q(K_E+1)^2\right]}{(K_E+1)^2\left[K_E \cdot K_Q(1+K_L)+K_L\right]}. \tag{23}$$

The capacitive reactance of the capacitor is

$$x_C = z_{BP}\left[x_{sC}^* + x_{L1}^* \cdot K_L(1+K_E \cdot K_Q) + \frac{K_E^2(K_Q+1)}{(K_E+1)^2}\right]. \tag{24}$$

From (15), we get the inductive reactance of the additional reactance

$$x_{L2} = K_L \cdot x_{L1}. \tag{25}$$

4.3. Determination of reactive elements for the first version of the converter

The basic resistance z_{BP} can be determined from formulas (23) to (25) from the ratio of specified nominal voltage U_{dn} and nominal current I_{dn}

$$z_{BP1} = \frac{U_{dn}}{I_{dn}}. \tag{26}$$

The distribution factors K_E, K_L, and K_Q should be specified. In the case of no experience, it is suggested to start from the values $K_E = 1.0$, $K_L = 0$, and $K_Q = 1.0$. The inductive reactances of the transformer leakage should also be specified.

5. PRACTICAL RESULTS

Two ESTA power sources have been erected, operating in Chelyabinsk, Russia. Power source ESTA-1/190 has been used to modify a AC-supplied furnace. The ESTA-8.5/450 power source was built to replace the steel melting arc-furnace DC thyristor power source. The former furnace transformer and its voltage tap-changer have also been used, but the secondary windings of the transformer are reapplied. The scheme shown in Fig. 3 (without thyristors) is used.

The main parameters of power sources ESTA-1/190 and ESTA-8.5/450 are shown in Table 1. The same table gives the parameters of ESTA-94/750-T2, which has been designed but not built yet.

Table 1

Main parameters of the power source ESTA

Parameter	ESTA-1/190	ESTA-8.5/450	ESTA-94/750-T2
Rated power, MV·A	1.0	9.2	98
Nominal arc voltage, V	190	446	750
Nominal arc current, kA	4.8	19	125
Short-circuit current, kA	8.6	32	125
Nominal power factor	0.99	0.995	0.99
Mains current THD in nominal conditions	8	6	6
Short-circuit power of minimal permissible supply network, MV·A	20	120	2000
Installation power of capacitor banks, MV·A	0.70	6.9	84
Total installation power of AC reactances, MV·A	0.54	5.7	62
Method of current stabilization and control	parametric + tap-changer	parametric + tap-changer	parametric with thyristors

The main aim in the ESTA power source design was to minimize the negative influence of the arc furnace to the supply network. In addition to the main result, some other positive results have been acquired. Furnace operation noise is lower than that with a thyristor source. Before modifying, the supplied power was deficient. By making use of the effect of reactive power compensation and self-adaption of the load, the DC power increased by 44%. The fact that the transformer has not been changed refers to a more efficient use.

6. CONCLUSIONS

Some power consumers need constant power. An example is electrical arc, in which case low current at high voltage and high current at low voltage is required. A transformer with a smoothly varying ratio suits well for such consumers. On this condition, power supply does not need to exceed the maximum power of the consumer.

The features of a transformer with a smoothly varying ratio are incorporated in the resonant converter, where the capacitor and reactor of an oscillatory circuit are switched from series to parallel and vice versa in each half-period of the network voltage. At the same time, the two secondary windings of the transformer are also switched from series to parallel and vice versa. The switch-over takes place with the help of the rectifier diodes and it is parametrical. The switch-over is caused by the phase shift of the reactive components. Variation of the load current causes variation of the ratio between the duration of series and parallel connection and also changes the average ratio of the transformer up to two times.

Besides the transformer, the ratio of the converter input current to the output current is altered by the switch-over of reactive components. In the case of parallel circuit, the current in the circuit is higher than the current supplying the oscillatory circuit. The parallel oscillatory circuit operates as a transformer. The current is not transformed in the case of series oscillatory circuit. If the ratio between the durations of series and parallel oscillatory circuits varies, the ratio between the current supplying reactive components and that supplying the rectifier bridge by the reactive components also varies. This ratio can be 1 : 5 or even higher at high load currents.

With the transition from no-load to short circuit, the ratio between the input and output current of the converter varies considerably (about ten times). The operating mode of the converter is self-adjusting to the load. In short circuit the output current is strictly limited and the input current is much less than the nominal input current. In the range of the nominal mode, the power supplied to the load will be nearly constant in the case of varying load (this is the effect of parametric stabilization of power).

The power factor of the converter is very high in all modes (approximately 0.99 in the nominal mode). The form (curve) of input current is good. Maximal values of higher harmonics in the line current compared to the rated current are approximately: 5th – 6%, 7th – 4%, 11th – 1.5%.

The resonant converter with self-adjustment to load is suitable for deeply and steeply fluctuating loads and for large powers (arc furnaces). The converter solves the problems with reactive power compensation, flicker and higher harmonics. The power factor is very high (in the nominal mode approximately 0.99) and the reactive power increases very little when controlling or varying the load.

When calculating parameters of the converter components, it is recommended to use the step-by-step approach and a special system of relative values based on short-circuit currents and no-load voltages. In the first approach, component values can be calculated on the basis of the short-circuit mode.

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RESONANTSMUUNDUR KUI SISEND- JA VÄLJUNDVOOLU SUHET MUUTEV REGULEERITAV TRANSFORMAATOR

Kuno JANSON ja Jaan JÄRVIK

Mõningate tarbijate, näiteks elektrikaare, toitmiseks on sobiv konstantne võimsus. Sel juhul on soodne kasutada sujuvalt muutuva ülekandesuhtega transformaatorit. Nii ei pea toiteallika võimsus olema suurem kui tarbija maksimaalvõimsus.

Muutliku ülekandesuhtega transformaatori omadused on resonantsmuunduril, kus võnkeringi kondensaator ja drossel lülitatakse igal võrgupinge poolperioodil järjestikühendusest paralleelühendusse ja tagasi. Sealjuures lülitatakse ka transformaatori kaks sekundaarmähist järjestikühendusest paralleelühendusse ja tagasi. Ümberlülitus toimub alaldussilla ventiilide abil; see on parameetriline ning selle põhjustab reaktiivelementide voolude faasinihe. Koormusvoolu suurenemine muudab paralleel- ja järjestikühenduse kestuse suhet ja sellega transformaatori keskmist ülekandetegurit kuni kaks korda.

Muunduri sisend- ja väljundvoolu suhet muudab lisaks transformaatorile veel reaktiivelementide ümberlülitamine. Paralleelvõnkeringi puhul on voolutugevus võnkeringi sees suurem kui võnkeringi toitev vool. Paralleelvõnkering töötab transformaatorina, järjestikvõnkering aga mitte. Kui järjestik- ja paralleelvõnkeringi kestuse suhe muutub, siis muutub ka reaktiivelemente toitva ja reaktiivelementide poolt alaldussillale antava voolu suhe.

Üleminekul tühijooksult lühisele muutub muunduri sisend- ja väljundvoolu suhe suures ulatuses (umbes 10 korda). Lühisrežiimis on väljundvoolu suurus järsult piiratud ja sisendvool on nimivoolust palju väiksem. Nimirežiimi piirkonnas jääb koormustakistuse muutuse korral sellele takistusele antav võimsus ligikaudu konstantseks (võimsuse parameetrilise stabiliseerimise efekt). Muunduri võimsustegur on kõikides režiimides väga kõrge (nimirežiimis umbes 0,99). Sisendvoolu kõvera kuju on hea. Sisendvoolu kõrgemate harmooniliste maksimaalväärtused nimivoolu suhtes on järgmised: 5. harmooniline – 6%, 7. – 4% ja 11. – 1,5%.

Koormusega isekohastuv resonantsmuundur on sobiv järsult ja sügavalt muutliku koormuse puhul ka suurtel võimsustel (kaarleekahjud võimsusega 100 MW ja rohkem). See muundur lahendab suurte ahjude puhul tavaliselt esinevad probleemid (reaktiivvõimsuse kompenseerimine, vilkumine, kõrgemad harmoonilised).