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CONSIDERATION OF SOLUTION FOR ENHANCEMENT OF FREQUENCY CONVERTER SUPPLY POWER PARAMETERS

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The paper presents results of analysis of a possible solution for enhancement of frequency converter (FC) AC supply power parameters. The method proposed is based on switched stabilisation of DC current of FC front-end rectifier unit. Such stabilisation allows obtaining rather good AC supply power parameters, i.e., its power factor and total harmonics distortion (THD) indicator for phase current. The paper also demonstrates a possible realisation scheme, a simplified mathematical description of processes in the scheme, as well as methods for consideration of its parameters, accounting the rated power of FC and the appointed level of the rectifier DC current continuity ratio. The results of computer simulation, to a great extent, testify a possible enhancement level.

Keywords: *converter, frequency, inductor, inverter, load, power, rectifier, ripple, stabilisation, switch, switching*

1. INTRODUCTION

Traditionally, frequency converters have the three main blocks. The first one, the front-end, is an uncontrolled 6-pulse bridge rectifier. The second one is capacitor bank that is necessary for smoothing of DC voltage at the output stage of the rectifier and for absorbing a reverse current in the braking regime of FC load – induction motor. The third one is the voltage source inverter that converts DC voltage over the capacitor plates to single or three-phase AC voltage, which is applied to the motor. The capacitor can be connected directly to output DC clamps of the rectifier (usually, a 3-phase one), as well as through an inductor coil. In the second case, the inductor together with the capacitor bank forms a passive LC mode DC smoothing filter, which allows improving the quality of DC voltage at the input of the inverter, as well as improving the quality and electromagnetic compatibility of FC with regard to the AC supply network. Smoothing effect depends on the volume of inductor inductance, which is limited in size, as well as there are price restrictions. At admissible inductance restrictions, a smoothing effect is relatively weak and power parameters are far from achievable at full smoothing of current.

Naturally, in order to decrease size and price of the FC, very often a case without inductor, which is rather bulky, is applied. Electrical current supplied from AC network is greatly distorted and has large content of high frequency harmonics, worsening the THD factor and the power factor, too [1], which in the case really depends on a leakage inductance of a supply network, only. For instance, Fig. 1.a depicts an experimentally obtained diagram of supply phase current of FC of 7.5 kW power, when volume of capacitor is 1 mF, smoothing circuit is without an inductor, and leak inductance of the supply phase is 0.5 mH. The THD factor for this very distorted current is 1.075, but power factor is 0.7297. When an extra inductor of 3 mH is introduced in the circuit, forming the rectifier output current smoothing LC filter, then the shape of supply current is less distorted and THD factor at full loading of FC can be improved to 0.45, but power factor – to a value of 0.917. Naturally, further increase in volume of inductance improves a THD factor, and at absolutely smoothed current and large volume inductor (Fig. 1. b), the harmonic distortion factor THD should be reduced to a value of 0.31.

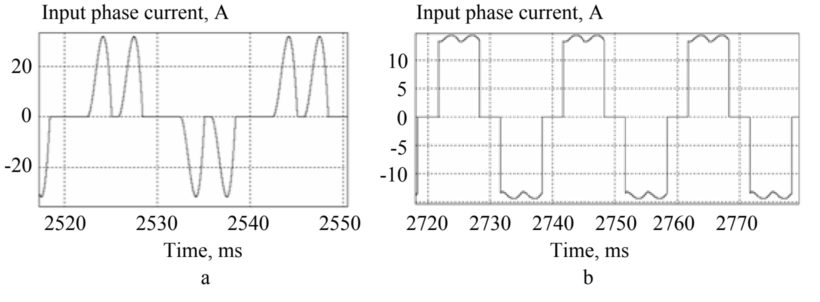

Fig. 1. Diagram of network phase current of FC 7.5 kW at network leak inductance L=0.5 mH without smoothing inductor; THD factor is 1.075; power factor is 0.7297 (a); high quality smoothing LC filter; THD factor is 0.31 (b).

The above-mentioned considerations show that enhancement of FC input DC stage is a topical issue, which asks for efficient solutions. With regard to the input stage, some other improvements should be introduced, too: an electronical limitation of FC turn-on current spike and provision of bi-directionality, which could enable reversion of energy flow at braking processes of induction motors supplied through FC. The paper discusses one method of smoothing of input stage current, realisation of which is relatively simple and based upon power electronics principles, as well as asks for small volume inductance of an inductor, only.

2. ANALYSIS OF TOTAL HARMONIC DISTORTION OF THE DISCUSSED SOLUTIONS

Dependence of power factor (bold lines) and THD (thin lines) for the FC of rated power 7.5 kW, with three different realisation methods of input junction, on the realised power (Fig. 2) is presented: with C-filter at C = 1 mF, inductance L of supply line 0.5 mH; with LC filter at L = 3 mH, L of supply line zero; with an ideal smoothing filter.

It should be taken into account that the quality, characterising parameters, to a great extent, depend on loading power of a frequency converter, and if load is lighter, then parameters are worse.

Taking into account problems related to the introduction of a sufficiently large volume inductor, which allows obtaining well smoothed DC current of the rectifier, it is necessary to find solutions, which could provide a sufficiently good shape of supply phase current at acceptable small volumes of filter element parameters. One solution should be related to electronically controlled smoothing of the inductor current, at which it should be possible to reduce volume of inductance and size of input junction of FC, thus providing sufficiently good distortion factor for the supply AC current.

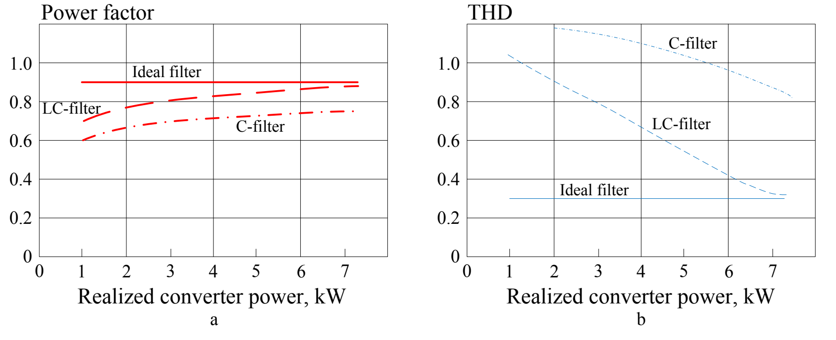

Fig. 2. Dependence of power factor (a) and THD (b) of FC 7.5 kW on a loading level of the device.

3. CONTROL SYSTEM OF THE OFFERED SOLUTION

To realise the DC current smoothing feature, it is necessary in the intermediate block of FC to introduce elements of well-known BOOST DC/DC converter topology – an inductor L1, electronic switch Q1, separating diode D1, resistor R1 as a sensor of the DC circuit current (Fig. 3.a).

Periodically **on-off** switching of the switch Q1 is supporting almost constant DC component of current through the inductor L1, i.e., current of inductor is smoothed at the reference level I_L which corresponds to the appointed value by equality of powers at output of the rectifier and at input of the inverter:

$$I_L = \frac{V_{ld} \cdot I_{ld}}{V_1}, \mathrm{A}, \tag{1}$$

where $V_{ld} = V_C$ – the average value of capacitor C_1 voltage;

I_{ld} – the average value of the DC input current of the inverter;

V_1 – the output DC voltage of the rectifier.

Specificity of the system lies in the fact that instantaneous values v_1 of the output voltage of the rectifier v_1 are periodically variating over 1/6 of total cycle of supply AC value V_{AC} (Fig. 3.b). The maximal value of v_1 is $V_{1,\max} = \sqrt{2} \cdot V_{AC}$, the minimal value is $V_{1,\min} = \sqrt{1.5} \cdot V_{AC}$, but voltage V_1 is the average value of output voltage of the rectifier, which for the 3-phase bridge mode rectifier, without accounting for inductance of AC supply lines, keeps value $1.35 \cdot V_{AC}$, where the V_{AC} is RMS value of the AC interphase voltage.

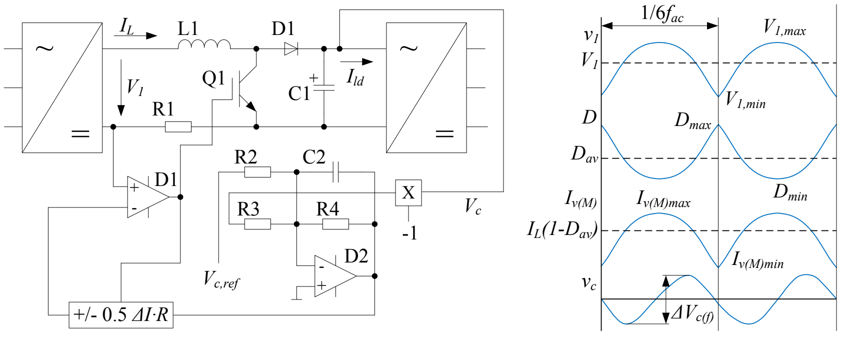

Fig. 3. Realisation scheme of rectifier DC output current electronic stabilisation (a) and rippling waves of rectifier output voltage v_1, variations of duty ratio D indicator, variations of diode V1 average current I_v, rippling waves of capacitor C1 voltage due to the variations of D (b).

The automated control of the system is provided using the proportional regulator, which generates negative polarity signal $-I_L{\cdot}R$ proportional to difference between real capacitor plate voltage V_C and the reference value $V_{C,ref}$, and prescribes the automatically supported DC value I_L of the inductor L1 current.

Taking into account the described principle, a control system must provide the comparison of voltage signal across a measure resistor R1, i.e., signal $v_R = i_L{\cdot}R$, with the prescribed by control feed-back reference signal $V_L = I_L{\cdot}R$ +/- $0.5{\cdot}\Delta I{\cdot}R$. When v_R is less than V_L, the switch must be **on**, but when situation is opposite, then switch is in position **off**. To define switching quality, in the comparison system it is necessary to introduce dead gap signal $\Delta v = \Delta I_L{\cdot}R$, where ΔI_L is a ripple range of inductor current. When switch stands in its **on** position, gap correction signal minus $0.5\Delta I_L{\cdot}R$ is added to the negative polarity signal $I_L{\cdot}R$, but when switch stands in off position, then signal $+0.5\Delta I_L{\cdot}R$ is added.

Applying the main relations for the BOOST converter, it is possible to state that ripple range of the inductor current, at zero values of AC supply line leakage inductances, can be found as follows:

$$\Delta I_L = \frac{v_1 \cdot D}{L \cdot f}, \mathrm{A}, \tag{2}$$

where v_1 – the instantaneous value of the rectifier output voltage;

D – the duty ratio of the electronic switch;

L – the inductance of the coil.

The input voltage of the inverter, DC voltage across the capacitor C_1, can be defined [2] as follows:

$$V_C = \frac{v_1}{1-D}, \mathrm{V}, \tag{3}$$

but module of the current i_L ripple range, at **off**-position of the switch and continuous current of the inductor, is:

$$\Delta I_L = \frac{(V_C - v_1)\cdot(1-D)}{L\cdot f}, \mathrm{A}. \tag{4}$$

Taking into account the both expressions of the ripple range, a duty ratio for each instantaneous value of the rectifier output voltage can be expressed as follows:

$$D = 1 - \frac{v_1}{V_C}. \tag{5}$$

As a result, diode V_1 current average value in each modulation cycle of switch depends on the instantaneous value of v_1 (see Fig. 3.b):

$$I_{v(M)} = \frac{v_1}{V_C}\cdot I_L, \mathrm{A}. \tag{6}$$

As consequence, an average value of capacitor current over each modulation interval is formed with similar shape to the voltage v_1:

$$I_{C(M)} = I_{v(M)} - I_{ld} = I_{ld}\cdot\left(\frac{v_1}{V_{1av}} - 1\right), \mathrm{A}. \tag{7}$$

Rippling of capacitor voltage over 1/6 of AC supply cycle $1/f$ due to the instantaneous variations of supply DC voltage can be simplified as follows:

$$\Delta V_{C(f)} = \frac{0.6\cdot 0.0476\cdot I_{ld}}{12\cdot f\cdot C}, \tag{8}$$

where it is accepted that this voltage ripple is formed by a positive part of the capacitor average current over each modulation interval. At the same time, rippling of the capacitor voltage is raised due to the impact of modulation processes with frequency f_M. These ripples are with higher frequency and spread upon relatively low frequency ripples discussed before. Range of the capacitor voltage ripples can be calculated as follows:

$$\Delta V_{C(M)} = \frac{I_{ld} \cdot D}{C \cdot f_M}. \tag{9}$$

These ripples are much smaller than the created ones due to variations of DC supply voltage. For instance, if load DC current is 15 A, an average duty ratio is 0.1, modulation frequency 20 kHz, but capacitor is with $C = 1$ mF, then voltage ripple range due to the modulation reason is 0.0075 V, but due to the voltage variation reason it is 0.714 V. Anyway, the capacitor voltage ripple range in the scheme is much smaller than that at the application of a passive LC-filter, when the parameter stands for 5.4 V. It means that in the proposed scheme the volume of capacitor C_1 should be lowered substantially.

With regard to the switching frequency, its value also depends on an instantaneous value of the DC supply voltage v_1:

$$f_M = \frac{V_C \cdot v_1 - v_1^2}{L \cdot \Delta I_L \cdot U_C}, \mathrm{Hz} \tag{10}$$

and its maximum value is at the minimal value of v_1:

$$f_{M(\max)} = \frac{V_C \cdot \sqrt{1.5} \cdot V_{AC} - 2.25 \cdot V_{AC}^2}{L \cdot \Delta I_L \cdot V_C}, \mathrm{Hz}. \tag{11}$$

Its minimum value stands at the maximal value of v_1:

$$f_{M(\max)} = \frac{V_C \cdot \sqrt{2} \cdot V_{AC} - 2 \cdot V_{AC}^2}{L \cdot \Delta I_L \cdot V_C}, \mathrm{Hz}. \tag{12}$$

The average value of switching frequency can be found as follows:

$$f_{M(av)} = \frac{V_1 \cdot (V_C - V_1)}{V_C \cdot L \cdot \Delta I_L} = \frac{V_1 \cdot (I_L - I_{ld})}{I_L \cdot L \cdot \Delta I_L}, \mathrm{Hz}, \tag{13}$$

where I_L – the average current of the inductor.

As it can be seen, at other constant parameters, switching frequency is increasing along with an increase in capacitor C_1 voltage level, which must be above the value of $V_{1,max}$. For instance, if V^*_C (with regard to V_1) is raised from 1.05 up to 1.1, then frequency is raised by 1.82 times. Conclusion can be made that parameter D should be kept at small values, which corresponds to the admissible low values of V_C. The last expression (13) should be rewritten as follows:

$$\frac{V_C - V_1}{V_C} = 1 - \frac{I_{ld}}{I_L} = D. \qquad (14)$$

It can be stated that organising the control system with stabilisation of voltage V_C, at first it is necessary to take into account parameter V^*_C , then product of comparison of real capacitor voltage with the reference one can produce the reference value for providing automated stabilisation of inductor L1 current, which will be proportional to load current as $I_L = I_{ld}/(1 - D_{av})$. At such realisation of a control system, the capacitor voltage will be kept at level $V_C = V_I/(1 - D_{av})$.

The efficient realisation of control can be provided only at continuity of the inductor current i_L. It can be provided if a half of prescribed current ripple range $0.5\Delta I_L$ is not greater than average current I_L of the inductor. At the boundary case $I_{LB} = 0.5\Delta I_L$, but at $I_L > I_{LB}$ current will be continuous.

Due to the fact that load current $I_{ld} = (1 - D_{av})\cdot I_L$, it can be stated that boundary value of the load current of an input junction of FC is:

$$I_{ldB} = (1 - D)\cdot 0.5 \cdot \Delta I_L, \mathrm{A}. \qquad (15)$$

Taking into account dependence of the factor D_{av} on switching frequency, all parameters of the stabilisation system can be generalised into expression:

$$f_{Mav} = \frac{V_1 \cdot D_{av} \cdot (1 - D_{av})}{2 \cdot L \cdot I_{ldB}} = \frac{V_1^2 \cdot D_{av}}{2 \cdot L \cdot I^*_{ldB} \cdot P_N}, \mathrm{Hz}, \qquad (16)$$

where P_N – the rated power of the converter;
I^*_{ldB} – minimal rated load current at continuity of inductor current.

If, for instance, V_I = 540 V, L = 0.5 mH, I_{ldB} = 2 A, then, for operation under rectifier output voltage ripple, switching frequency must be f_{Mav}=30 kHz, parameter D_{av} = 0.127. This means that the prescribed ripple range ΔI_L = 4.58 A, capacitor voltage stands for 618.5 V, but at the average duty ratio D_{av} = 0.127 on-duty interval of a switch is about 4.24 µs. As it can be stated from expression (16), switching frequency depends on accepted parameters D_{av}, L and I_{ldB}: if L and I_{ldB} are accepted smaller, then an average frequency is higher, which provokes an increase in switching power losses. Estimating switching frequency changes over a characteristic rectifier voltage interval with time duration 3.33 ms (at frequency of supply 50 Hz), instantaneous change in rectifier voltage level and duty ratio variations have to be considered.

It should be mentioned that real frequency of switching also depends on leak inductances of AC supply network lines. Since these parameters are occasional by their value, zero values of the parameter can be applied, only.

4. EXPERIMENTAL INVESTIGATION OF THE SYSTEM

Experimental investigations have been done in computer simulation system, where parameters of elements were suited to the case calculated above at V_{int} = 400V and for FC of rated power 7.5 kW. In Fig. 5.a, there is a diagram of coil with L = 0.5 mH and supply phase current at continuous current mode of the coil and with zero value of supply line leak inductance. The ripple range was accepted 4.48 A, and control system was arranged in accordance with principles presented in Fig. 2, applying as sensor current resistor R = 0.01 Ω. The reference level of capacitor voltage is applied as U_{Cref} = 615 V. The main parameters of the system at some load levels for continuous mode of current are presented in Table 1.

Table 1

Operational Parameters of the FC with Rated Power 7.5 kW at Different Loadings (computer simulation results)

Parameters	Load			
	7.6 kW	2.53 kW	1.31 kW	0.97 kW
Load DC current, A	12.45	4.13	2.136	1.581
Inductor DC current, A	14.08	4.71	2.44	1.84
Inductor RMS current, A	14.15	4.9	2.79	2.42
AC supply phase current, A	11.57	3.96	2.28	1.98
Capacitor DC voltage, V	610.4	612.9	613.4	613.5
AC supply app. power, VA	8015.9	2743.5	1579.6	1370.2
Power factor P/S	0.948	0.922	0.829	0.708
THD factor for phase current	0.345	0.447	0.66	0.68
Av switching freq., Hz	27,027	29,129	28,059	28,070
Max switching freq., Hz	40,000	42,735	42,427	42,500
Min switching freq. , Hz	18,367	20,134	22,222	22,345

As it can be seen from Table 1, power factor for the AC supply input at continuous current of the inductor L1 (the two left side columns) is 0.9, but if inductor current is at discontinuous shape, then power factor is smaller. Similarly, THD factors for continuous shape of inductor current are rather good and close to the values at ideally smoothed inductor current (Fig. 4). Again, at a discontinuous shape of inductor current, THD factor is worse, but anyway it is better as for the case of application of a passive LC smoothing filter.

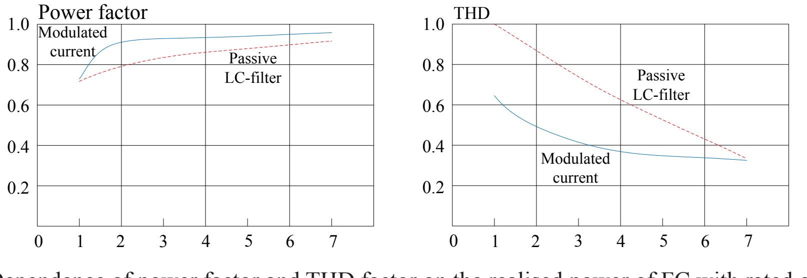

Fig. 4. Dependence of power factor and THD factor on the realised power of FC with rated one 7.5 kW for the cases with modulated current of the inductor (solid lines) and for the case with a passive LC-filter (dashed lines).

For a better effect, modulation of inductor current has to be done with a smaller ripple range and smaller load current for the boundary case. Then, if a switching frequency is to be supported on a sufficiently low level, factor D_{av} must be kept smaller, i.e., excess of capacitor voltage level has to be lowered. If, for instance, factor D_{av} is applied at a value 0.05, and voltage of the capacitor at a level 570 V, then, for the case when average frequency 30 kHz is applied, the accepted ripple range must be installed as ΔI_L = 1.8 A. The minimal value of load current will be 1.71 A, or load power of the FC at a level 974 W. This means that a range of realised powers, at continuous current of the inductor and at good values of THD and PF, will be widened compared to the above-mentioned case with ΔI_L = 4.58 A. The application cases have been considered for a zero value of AC phase leakage inductances. Really such inductances, though eventually defined, could be accounted, and then switching frequency and ripple range will be lesser. The switching frequency at the same ripple range ΔI_L = 4.58 A is 3-fold smaller now, because the total loop of current way in each 1/6 of the supply voltage cycle content in summary is 1.5 mH, instead of 0.5 mH at zero value of the leak inductance. It means that expression for estimation of switching frequency should be, more correctly, applied as follows:

$$f = \frac{V_1^2 \cdot K}{2 \cdot (L + 2 \cdot L_n) \cdot I_{ldB}^* \cdot P \cdot P_N}, \text{Hz}. \tag{17}$$

where L_n is leakage inductance for the phase of AC supply.

If this inductance should be sufficiently large, then the discussed scheme could operate without any extra inductor in the circuit of rectifier output. At that shape of AC phase current is even better as with an extra inductor (Fig. 5.b). At full loading of FC 7.5 kW, power factor with regard to supply is 0.96, but THD factor for the presented wave of current is 0.283. Average switching frequency is 15916 Hz. The obtained parameters prove the possibility to provide stabilisation of the rectifier output DC current even without the application of some extra inductor in the circuit, if leakage inductance is sufficiently great. Real value of leakage inductance should be considered using a loading experiment for supply network with regard to an input AC terminal of FC.

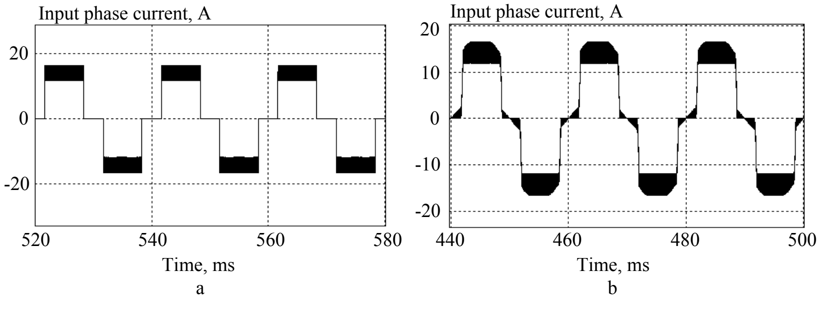

Fig. 5. Simulated diagram of AC phase current at modulated stabilisation of inductor L = 0.5 mH DC current and full loading of FC 7.5 kW (a) and at leak inductance of power supply lines at the level of 0.5 mH, excluding an extra inductor at the intermediate stage of FC (b).

5. CONCLUSIONS

Stabilisation (smoothing) effect of application of a passive LC filter is restricted by admissible volume of inductance of inductor applied to the filter and, therefore, it is not possible to obtain good AC power parameters within the entire operation range of FC. AC power parameters (power factor and THD) of frequency converter, as well as volumes of applied L and C reactive elements can be enhanced providing stabilisation (smoothing) of the front-end device of FC – rectifier – DC output current. Stabilisation of rectifier output current can be organised applying relatively small, with regard to the passive filter case, volume of inductor and, at sufficiently great leakage inductance of the network lines, without any extra inductor, using electronic switch supported modulation of rectifier output current in a scheme based on BOOST converter topology, which defines the level of filter capacitor DC voltage above an amplitude value of voltage of rectifier output. Value of an applied ripple range of the stabilised current, for providing continuous shape of the current, at which enhanced power parameters could be provided, depends both on minimal accepted load DC current and on the ratio between the output voltage levels of capacitor and that of rectifier.

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FREKVENČU PĀRVEIDOTĀJA BAROŠANAS JAUDAS PARAMETRU UZLABOŠANAS RISINĀJUMU APSVĒRUMS

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Kopsavilkums

Šajā pētījumā ir parādīts viens no iespējamiem frekvenču pārveidotāja (FP) maiņstrāvas barošanas avota jaudas parametru uzlabošanas risinājumiem. Pielietotā metode balstās uz FP ieejas moduļa – taisngrieža - izejas strāvas komutējamas stabilizācijas. Šāda stabilizācija ļauj iegūt diezgan labus maiņstrāvas barošanas parametrus – tīkla fāzes jaudas koeficientu un strāvas harmoniskā kropļojuma THD faktoru. Tiek parādīta iespējamā īstenošanas shēma, dots vienkāršots matemātiskais apraksts par procesiem un parametru noteikšanas metodēm atkarībā no pārveidotāja nominālās jaudas, kā arī minimālās slodzes strāvas, pie kuras induktora momentānā strāva ir nepārtraukta. Veikto datorsimulāciju rezultāti lielā mērā apliecina risinājuma iespējamo efektivitāti.

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