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NEW ACTIVE DIODE-SWITCHED TRANSFORMER FILTERS TO IMPROVE POWER FACTOR OF THREE-PHASE DIODE RECTIFIERS WITH CAPACITIVE SMOOTHING

Tiiu SAKKOS and Vello SARV

Estonian Energy Research Institute, Paldiski mnt 1, EE-0001 Tallinn, Estonia

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Abstract. This paper proposes and analyzes new schemes of the low-distortion three-phase diode rectifiers with capacitive smoothing, using active diode-switched transformer filters. The power factor correction technique proposed is based on the third-harmonic power conversion into an additional dc power in the active filter.

Key words: active filter, diode-switched transformer, power factor correction, three-phase rectifier, capacitive smoothing.

1. INTRODUCTION

Contemporary power electronic converters, using switches and diodes as basic conversion elements, represent an important link between power sources and loads for transforming and adapting the parameters of electrical energy to consumers' needs. Unfortunately, switches and diodes inevitably cause the generation of distortion harmonics. Therefore, harmonic reduction as well as power quality problems in general, become increasingly important. Among the major sources of harmonic distortion, diode-bridge rectifiers, particularly those with capacitive smoothing, are widely used [1,2]. Active filters (AF), which use fully controlled switches, facilitate provision of both approximately sinusoidal line currents and controlled dc output voltage of diode rectifiers [1,3]. Nevertheless, in the case of uncontrolled output voltage, various less expensive filter topologies can be efficiently used for power factor correction [2,4,5]. Considerable harmonic reduction and the corresponding power factor improvement can be achieved by means of the second-harmonic power conversion in single-phase rectifiers and the third-harmonic power conversion in three-phase rectifiers into an additional portion of dc power. By using various techniques, this trend seems promising [4,6–9]. In this paper, first, some new efficient diode-switched transformer AFs for power factor correction of the three-phase diode-bridge rectifiers with capacitive smoothing of the output voltage are proposed and then analyzed. Second, the possible filter configurations are compared to achieve the optimum schemes.

2. OPERATING PRINCIPLES

With regard to power flow, the line-current harmonic reduction is based on the improvement of the power conversion process. Let us consider power conversion in the three-phase diode rectifier shown in Fig. 1. The circuit contains a diode bridge (DB), a current-shaping AF, a resistive load R, two smoothing capacitors C in series connection to provide the midpoint n for the output voltage V_o, and a zero-sequence filter (ZSF) to provide the zero-potential terminal O of the symmetrical sinusoidal supply voltages v_A, v_B, and v_C with $v_A = V_m \sin\omega t$, where V_m is the amplitude value of supply voltage. The passive ZSF can be implemented, for example, by a zigzag autotransformer [3], a star-delta transformer [5], or an autotransformer in the Scott connection [6,8]. To simplify the analysis, assume that the DB and ZSF are ideal, output capacitances are equal to infinity, and the AF ensures the appropriate waveforms of the rectified currents i_d, i_g and the zero-sequence current i_z, needed for the power factor correction.

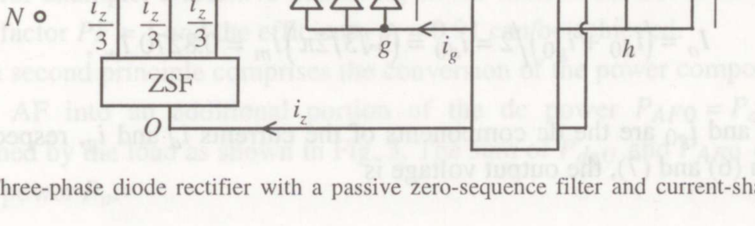

Fig. 1. Three-phase diode rectifier with a passive zero-sequence filter and current-shaping active filter.

2.1. Operation mode of sinusoidal supply current

From the analysis of an ideal case with sinusoidal supply currents i_A, i_B, and i_C with $i_A = I_m \sin \omega t$ [8], where I_m is the amplitude value of current, we obtain

$$\frac{i_z}{3} = \frac{3\sqrt{3}}{2\pi} I_m \left(-\frac{1}{2} \sin 3\omega t + \frac{1}{20} \sin 9\omega t - \dots \right), \quad (1)$$

$$i_d = \max\{i_A, i_B, i_C\} + i_z/3 = i_{d0} + i_z/2 + i_{d6} + i_{d12} + \dots$$

$$= \frac{3\sqrt{3}}{2\pi} I_m \left(1 - \frac{3}{4} \sin 3\omega t + \frac{2}{35} \sin 6\omega t + \frac{3}{40} \sin 9\omega t + \dots \right), \quad (2)$$

$$i_g = -\min\{i_A, i_B, i_C\} - i_z/3 = i_{g0} - i_z/2 + i_{d6} + i_{d12} + \dots$$

$$= \frac{3\sqrt{3}}{2\pi} I_m \left(1 + \frac{3}{4} \sin 3\omega t + \frac{2}{35} \sin 6\omega t - \frac{3}{40} \sin 9\omega t + \dots \right), \quad (3)$$

$$v_{dN} = \max\{v_A, v_B, v_C\}$$

$$= \frac{3\sqrt{3}}{2\pi} V_m \left(1 - \frac{1}{4} \sin 3\omega t + \frac{2}{35} \sin 6\omega t + \frac{1}{40} \sin 9\omega t + \dots \right), \quad (4)$$

$$v_{gN} = \min\{v_A, v_B, v_C\}$$

$$= -\frac{3\sqrt{3}}{2\pi} V_m \left(1 + \frac{1}{4} \sin 3\omega t + \frac{2}{35} \sin 6\omega t - \frac{1}{40} \sin 9\omega t + \dots \right). \quad (5)$$

As in the ideal case assumed, the average supply power P_s equals the output power

$$P_o = V_o I_o = P_s = 1.5\, V_m I_m , \quad (6)$$

and from (2) and (3), the output current is

$$I_o = (i_{d0} + i_{g0})/2 = i_{d0} = (3\sqrt{3}/2\pi)\, I_m = 0.8270\, I_m , \quad (7)$$

where i_{d0} and i_{g0} are the dc components of the currents i_d and i_g, respectively. Then from (6) and (7), the output voltage is

$$V_o = P_o / I_o = \left(\pi/\sqrt{3}\right) V_m = 1.8138\ V_m \,. \tag{8}$$

At the same time, the dc component of the rectified voltage V_{dg0} is

$$V_{dg0} = \left(3\sqrt{3}/\pi\right) V_m = 1.6540\ V_m < V_o \,, \tag{9}$$

and, consequently, the dc power at the output of the DB, P_{dg0} can be written as

$$P_{dg0} = V_{dg0}\ I_o = \left(27/2\pi^2\right) V_m\ I_m = 0.91189\ P_s < P_s = P_o \,. \tag{10}$$

Thus, with regard to the energy flow, in the case of undistorted operation mode, the DB consumes the power P_s at the supply frequency and converts it to the dc power $P_{dg0} = 0.91189\, P_s$, the third-harmonic power $P_{dg3} = 0.08549\, P_s$, the sixth-harmonic power $P_{dg6} = 0.00149\, P_s$, the ninth-harmonic power $P_{dg9} = 0.00085\, P_s$, etc. Since $P_{dg0} + P_{dg3} = 0.99738\ P_s$, the supply power P_s is converted mainly to the dc and third-harmonic power as shown in Fig. 2.

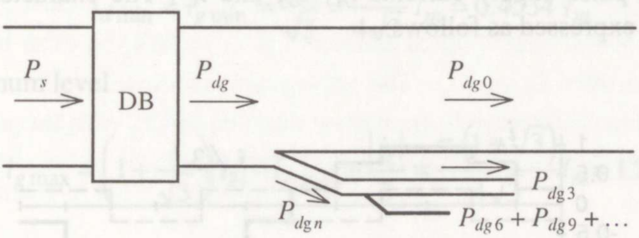

Fig. 2. Power conversion in the diode bridge DB under undistorted operation mode.

To ensure the undistorted operation mode of the DB with the sinusoidal supply current, two principles can be used, either separately or in a combination. The first one involves dissipation of the third-harmonic power component P_{dg3}, using, for example, a resistive ballast R_b in the filter structure. In this case, the power factor $PF = 1$ and the efficiency $\eta = 0.91$ can be achieved.

The second principle comprises the conversion of the power component P_{dg3} in the AF into an additional portion of the dc power $P_{AF0} = P_{dg3}$ to be consumed by the load as shown in Fig. 3. The sum of P_{dg0} and P_{AF0} equals the output power P_o.

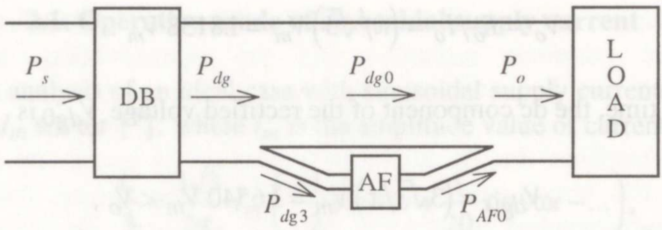

Fig. 3. Power flow diagram for the circuit in Fig. 1.

2.2. Operation mode of twelve-pulse supply current

The twelve-pulse operation mode enables one to eliminate lower supply-current harmonics up to the eleventh and ensure the power factor $PF = 0.99$. Therefore, in practice, such an operation mode is usually an acceptable approximation of the operation mode of the sinusoidal supply current [2,4]. Figure 4 shows the corresponding ideal normalized current waveforms in the circuit in Fig. 1. In this case, $I_m = 1$ equals the fundamental harmonic amplitude of the twelve-pulse supply currents i_A, i_B, and i_C. The characteristic current values can be expressed as follows:

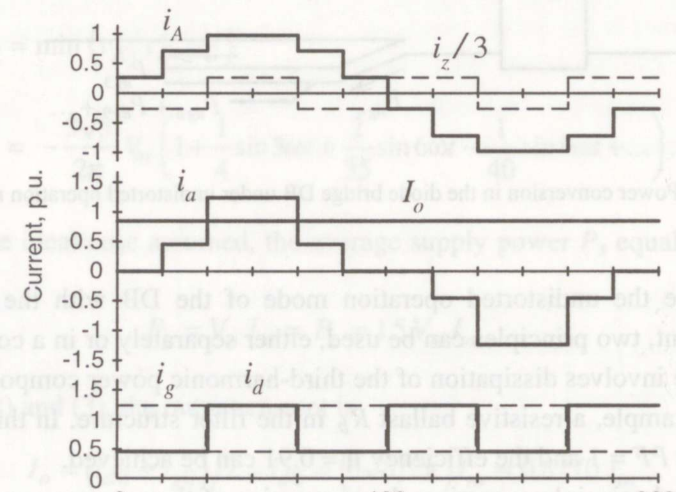

Fig. 4. Normalized current waveforms in the circuit in Fig. 1 for the twelve-pulse supply current mode.

The three-level supply current

$$i_A = I_m\left(\sin\omega t + \frac{1}{11}\sin 11\omega t + \frac{1}{13}\sin 13\omega t + \ldots\right) \tag{11}$$

has a minimum level

$$I_{A\min} = \frac{1}{3}\left|i_z\right| = \frac{\pi}{12} I_m = 0.2618\, I_m\ , \tag{12}$$

a medium level

$$I_{A\,\mathrm{med}} = \left(\frac{1}{3} + \frac{1}{\sqrt{3}}\right)\left|i_z\right| = \frac{\pi\left(1+\sqrt{3}\right)}{12} I_m = 0.7152\, I_m\ , \tag{13}$$

and a maximum level

$$I_{A\max} = \left(\frac{2}{3} + \frac{1}{\sqrt{3}}\right)\left|i_z\right| = \frac{\pi\left(2+\sqrt{3}\right)}{12} I_m = 0.9770\, I_m\ . \tag{14}$$

The two-level rectified currents i_d and i_g have a minimum level

$$i_{d\min} = i_{g\min} = \frac{\left|i_z\right|}{\sqrt{3}} = \frac{\pi}{4\sqrt{3}} I_m = 0.4534\, I_m \tag{15}$$

and a maximum level

$$i_{d\max} = i_{g\max} = \left(1 + \frac{1}{\sqrt{3}}\right)\left|i_z\right| = I_{A\max} + \frac{\left|i_z\right|}{3} = \frac{\pi\left(1+\sqrt{3}\right)}{4\sqrt{3}} I_m = 1.2388\, I_m\ . \tag{16}$$

The output current is

$$I_o = \frac{i_{d\max} + i_{d\min}}{2} = i_{d\max} - \frac{\left|i_z\right|}{2} = \frac{\pi\left(2+\sqrt{3}\right)}{8\sqrt{3}} I_m = 0.8461\, I_m\ . \tag{17}$$

The rectangular zero-sequence current is

$$i_z = -\frac{\pi}{4} I_m\, sign(\sin 3\omega t) = -0.7854\, I_m\, sign(\sin 3\omega t)$$

$$= -I_m\left(\sin 3\omega t + \frac{1}{3}\sin 9\omega t + \frac{1}{5}\sin 15\omega t + \ldots\right). \tag{18}$$

From (9) and (17), the average dc power of the DB is

$$P_{dg0} = V_{dg0}\, I_o = 1.3995\, V_m I_m = 0.9330\, P_s < P_s = P_o\ . \tag{19}$$

Consequently, like in the operation mode of the sinusoidal supply current, the DB consumes the power $P_s = 1.5\, I_m V_m$ at the supply frequency and converts it to the dc power P_{dg0} and to ac, typically to the third-harmonic power P_{dg3}. Likewise, the power P_{dg3} is consumed in the AF and converted into the additional portion of the dc power P_{AF0} that corresponds to the power flow diagram in Fig. 3.

3. PROPOSED IMPLEMENTATION OF THE TWELVE-PULSE OPERATION MODE

To implement the twelve-pulse operation mode, AF in the circuit in Fig. 1 has to ensure the required current relations (11)–(18). For that purpose, diode-switched inductor AFs can be used [6,8,9]. A simple diode-switched inductor AF ensures high efficiency, close-to-unity power factor, and high reliability. However, its main disadvantage is that relatively large stored energy and inductance values are needed to guarantee level current steps [6,7].

The inductor energy, inductance, and size can be reduced considerably replacing the time-variable diode-switched inductors by the diode-switched transformer T_3 and the relatively small coupled inductors L_d and L_g as proposed in Fig. 5. The role of the coupled inductors is to reduce the even harmonics of the rectified currents i_d and i_g. The appropriate modulation of the currents i_d and i_g is provided by the triple-frequency transformer T_3 with the primary turns $N_{nO} = N_{\rm I}$ and secondary turns $N_{fj} = N_{ef} = N_{lk} = N_{km} = N_{\rm II}$.

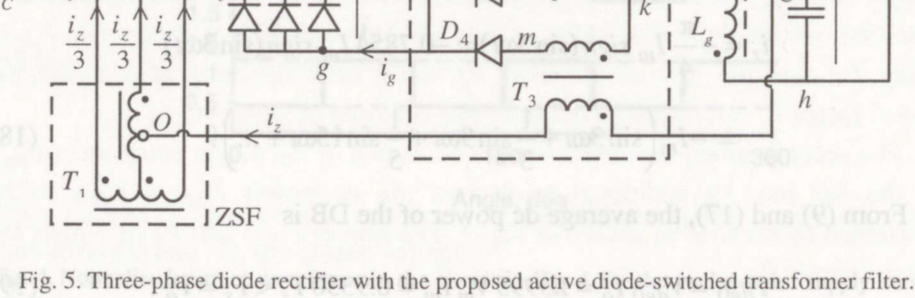

Fig. 5. Three-phase diode rectifier with the proposed active diode-switched transformer filter.

From (4) and (5), the primary voltage of the transformer is

$$v_{nO} = \frac{3\sqrt{3}}{8\pi} V_m \left(-\sin 3\omega t + \frac{1}{10} \sin 9\omega t - \dots \right). \tag{20}$$

Consequently, the secondary voltages of the transformer are

$$v_{fj} = v_{ef} = v_{lk} = v_{km} = v_{nO} \, N_{\mathrm{II}} / N_{\mathrm{I}} \, . \tag{21}$$

In the circuit in Fig. 5, the diodes $D_1, \dots, D_4$ are switched on and off alternatively due to the sign variation of the transformer voltages. During the interval $0 \le \omega t \le 30°$, we have $v_{nO} < 0$. Therefore, the diodes D_1 and D_3 are reverse-biased, while the diodes D_2 and D_4 are conducting. Assuming zero magnetization current, the total amper-turns of the transformer are

$$i_z N_{nO} + i_{d\min} N_{ef} + i_{g\max} N_{km} = 0 \tag{22}$$

and

$$i_z N_{\mathrm{I}} = -\left(i_{d\min} + i_{g\max} \right) N_{\mathrm{II}} \, . \tag{23}$$

From (15), (16), and (23), the transformer ratio required to ensure twelve-pulse operation mode equals

$$\frac{N_{\mathrm{II}}}{N_{\mathrm{I}}} = \frac{-i_z}{i_{d\min} + i_{g\max}} = \frac{3}{3 + 2\sqrt{3}} = 0.4641 \, . \tag{24}$$

During the interval $30° \le \omega t \le 60°$ with $v_{nO} > 0$, the diodes D_2 and D_4 are reverse-biased, while the diodes D_1 and D_3 are conducting. Consequently, the total amper-turns of the transformer are

$$i_z N_{nO} + i_{d\max} N_{jf} + i_{g\min} N_{lk} = 0 \, . \tag{25}$$

From (22) and (25), we obtain that rms current of the windings N_{ef} and N_{lk} is equal to $i_{d\min} / \sqrt{2}$, of the windings N_{km} and N_{jf} to $i_{d\max} / \sqrt{2}$, and of the winding N_{nO} to $|i_z|$.

Let us take into account the fact that flux linkage of the transformer depends mainly on the third-harmonic component of the winding voltages. As a result, the rated volt-amperes of a single winding are equal to the product of the winding rms current value and to the rms value of the third-harmonic component $V_{N\mathrm{I}3}$ or $V_{N\mathrm{II}3}$ of the corresponding winding voltage. From (20) and (24), we obtain

$$V_{N\mathrm{I}3} = \frac{3\sqrt{3}}{8\pi\sqrt{2}} V_m = 0.1462 \, V_m \; ; \; V_{N\mathrm{II}3} = \frac{N_{\mathrm{I}}}{N_{\mathrm{II}}} V_{N\mathrm{I}3} = 0.06785 \, V_m \, . \tag{26}$$

From (15), (16), (22), (25), and (26), the total volt-amperes of the transformer windings are

$$S_{T3} = 2\,V_{NII3}\,\frac{i_{d\min} + i_{d\max}}{\sqrt{2}} + V_{NI3}\left|i_z\right| = 0.2771\,V_m\,I_m = 0.1848\,P_s\,. \quad (27)$$

Hence, the equivalent volt-ampere rating of the tripled-frequency transformer T_3 becomes

$$VA_{T3} = \frac{1}{2}\,S_{T3} = 0.0924\,P_s\,, \quad (28)$$

i.e. 9.24 % of the rated supply power P_s.

In Fig. 5, the passive ZSF has been implemented by the autotransformer T_1 in the Scott connection to provide the zero-potential terminal O of the symmetrical sinusoidal supply voltages. Its equivalent volt-ampere rating VA_{T1}, calculated following the procedure used in the case of the transformer T_3, becomes

$$VA_{T1} = \frac{1}{2}\,S_{T1} = 0.2746\,P_s\,. \quad (29)$$

Analysis of the power transfer in the AF shown in Fig. 5 indicates that the conversion of the third-harmonic power P_{dg3} into the additional dc power P_{AF0} takes place in the diodes $D_1, \ldots, D_4$. In fact, these diodes, together with the transformer secondary windings, constitute two full-wave auxiliary rectifiers supplied by the third-harmonic voltage. It is essential for the rectified third-harmonic voltage to have a sixth-harmonic ripple-voltage component to compensate the undesired sixth-harmonic component of the voltage v_{dg}. As a result, the task of the inductors L_d and L_g is to smooth mainly the twelfth-harmonic ripple.

Analysis of the conversion processes indicates that the volt-amperes rating of the transformer T_3 can be reduced, replacing the two full-wave auxiliary rectifiers by two half-wave rectifiers as shown in Fig. 6. To ensure the twelve-pulse operation mode, the transformer turns have to be chosen as follows.

In the case of the four-winding configuration (Fig. 6*a*), $N_{ej} = N_{lm}$, $N_{jf} = N_{mk}$ and consequently

$$\frac{N_{ej} + N_{jf}}{N_{jf}} = \frac{i_{d\max}}{i_{d\min}} = 1 + \sqrt{3}\,,\ N_{ej} = \sqrt{3}\,N_{jf}\,. \quad (30)$$

In the case of the three-winding configuration (Fig. 6*b*), $N_{ef} = N_{lk}$ and we have

$$\frac{N_{ef}}{N_{nO}} = \frac{\left|i_z\right|}{i_{d\min}} = \sqrt{3}\,,\ N_{ef} = \sqrt{3}\,N_{nO}\,. \quad (31)$$

The volt-ampere ratings of the transformers shown in Figs. 6*a* and 6*b* are equal to $VA_{T3a} = 0.0777\,P_s$ and $VA_{T3b} = 0.0495\,P_s$, respectively.

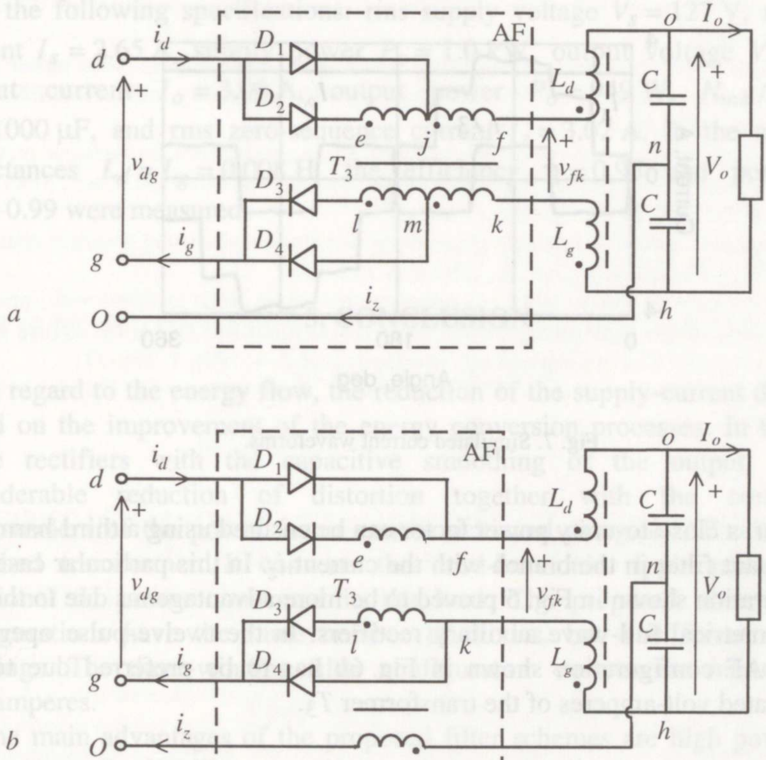

Fig. 6. Alternative active diode-switched transformer filters: *a,* four-winding configuration; *b,* three-winding configuration.

4. SIMULATION AND EXPERIMENTAL RESULTS

To verify the feasibility of the proposed AFs, the system illustrated in Figs. 5 and 6 has been tested using computer simulation. In addition, a laboratory rectifier corresponding to Fig. 5 has been built and tested.

The three-phase diode rectifier, using alternative AF configurations, was simulated with the following specifications: $v_A = \sqrt{2}\,127\sin 100\pi t$, $C = 2000\ \mu\text{F}$, and $R = 100\ \Omega$. The optimum transformation ratios were used. The simulation proved that the input and output characteristics of the rectifier do not depend on the AF configuration used. The simulated current waveforms i_A, $i_z/3$, and I_o, shown in Fig. 7, correspond to $L_d = L_g = 0.01$ H and to the power factor

$PF = 0.99$. A further increase of the inductances L_d and L_g results in more level steps of the twelve-pulse supply currents, but without essential increase in the power factor.

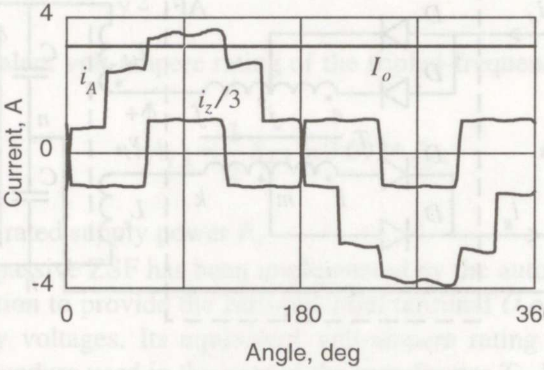

Fig. 7. Simulated current waveforms.

However, a close-to-unity power factor can be ensured using a third-harmonic series-resonant filter in the branch with the current i_z. In this particular case, the AF configuration shown in Fig. 5 proved to be more advantageous due to the use of the symmetrical full-wave auxiliary rectifiers. In the twelve-pulse operation mode, the AF configuration shown in Fig. 6*b* has to be preferred due to the minimum rated volt-amperes of the transformer T_3.

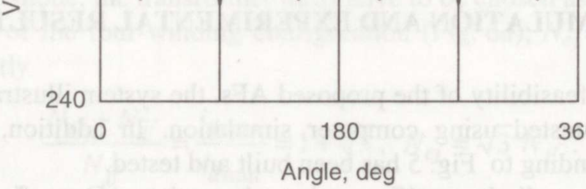

Fig. 8. Simulated voltage waveforms.

The simulated voltage waveforms v_{dg}, v_{fk}, V_o are shown in Fig. 8. Note that due to the third-harmonic power conversion into an additional portion of the dc power taking place in the diodes $D_1, \ldots, D_4$, $V_o > v_{dg0}$ and $v_{fk} > v_{dg}$. Note also that the fundamental ripple frequency of the voltage v_{fk} is increased twice as compared to that of the voltage v_{dg}. The experimental rectifier was implemented with the following specifications: rms supply voltage $V_s = 127$ V, rms supply current $I_s = 2.65$ A, supply power $P_s = 1.0$ kW, output voltage $V_o = 308$ V, output current $I_o = 3.08$ A, output power $P_o = 949$ W, $N_{mk}/N_{nO} = 0.5$, $C = 1000$ μF, and rms zero-sequence current $I_z = 3.02$ A. In the case of the inductances $L_d = L_g = 0.008$ H, the efficiency $\eta = 0.95$ and power factor $PF = 0.99$ were measured.

5. CONCLUSION

In regard to the energy flow, the reduction of the supply-current distortion is based on the improvement of the energy conversion processes. In three-phase diode rectifiers with the capacitive smoothing of the output voltage, a considerable reduction of distortion together with the corresponding improvement of the power factor can be achieved using an AF with a diode-switched transformer. It converts the third-harmonic ripple power into an additional dc power consumed in the load. The proposed alternative filter configurations have the same number of diodes, but a different number of windings. The three-winding filter configuration ensures the minimum rated volt-amperes.

The main advantages of the proposed filter schemes are high power factor, efficiency, and reliability as well as a relatively small volt-ampere rating of the filter transformer.

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UUED DIOODKOMMUTATSIOONIGA TRANSFORMATOORSED AKTIIVFILTRID MAHTUVUSLIKU SILUMISEGA KOLMEFAASILISTE DIOODALALDITE VÕIMSUSTEGURI PARANDAMISEKS

Tiiu SAKKOS ja Vello SARV

Nüüdisaegsete võimsuselektronmuundurite põhielementideks on pooljuhtlülitid, mis paratamatult genereerivad moonutusharmoonilisi. Oluliseks moonutuste allikaks on ka mahtuvusliku silumisega alaldid. Kolmefaasiliste dioodalaldite moonutuste märgatavaks vähendamiseks ning võimsusteguri suurendamiseks on esitatud dioodkommuteeritava trafoga aktiivfiltrite uued skeemid. Nende töö põhineb alaldi pulsatsioonivõimsuse muundamisel täiendavaks väljundvõimsuseks.

Kolme esitatud ja analüüsitud filtriskeemi iseloomustavad lisaks heale võimsustegurile väikesed kaod ning väike arvutuslik võimsus. 12-pulsiline töörežiim võimsusteguriga 0,99 on tagatav lihtsa kolmemähiselise filtritrafoga, mille arvutuslik võimsus on ainult 5% toitevõimsusest.