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A Dynamic Converter of Monophase into Triphase Voltage. Part I: Structure and Components[1]

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1. Introduction

Triphase energy is the basic energy used to obtain rotational mechanic energy, thanks to the ease of creating oriented rotating magnetic field. This energy is supplied to the main type of motors in manufacturing – asynchronous motors with a cage rotor. However, such energy is not available everywhere. This concerns mainly some remote country or wild regions, where conducting of triphase energy is too expensive. In some less industrialized countries monophase energy is provided with one wire, using the earth as a second one. The same refers to a number of countries, where the houses are supplied with monophase or biphase (anti-phase voltage) and the application of various techniques requires triphase energy supply as well.

The devices, converting monophase energy into triphase with different degree of symmetry, deserve particular attention. The most direct approach to such type of conversion is rectification of monophase voltage and successive inversion to symmetric triphase voltage. This approach enables the obtaining of high degree of symmetry, the building in of different protections, high efficiency at different level of loading, work with a power factor close to 1. However, such solution is too expensive, not reliable enough and requires qualified exploitation.

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The present paper, consisting of two parts, discusses dynamic converters of monophase into triphase voltage, which use a standard triphase asynchronous motor with a cage rotor as a main component. According to such a concept, the converter is functionally simple, constructed by highly reliable components, that have proved their fitness during quite long, and in the case of the motor – century-old development. They enable the connection of several consumers, including monophase ones. With the help of some not very complicated means, the start moment can be definitely raised, which allows switching under nominal load or rapid gaining of nominal revolutions.

The current work, presented in two parts, investigates in details dynamic phase converters, so it might give the knowledge and relations required for the design of similar devices. The first part studies the structure and the nodes, composing the converter. The second part discusses the start and control in the achievement of a given degree of symmetry.

The first part of the paper considers dynamic converters of monophase into triphase voltage that include a standard triphase asynchronous motor with a cage rotor as a main component. Its structure is investigated and the components of a dynamic phase converter, designed according the scheme, shown on Fig. 1, are dimensioned.

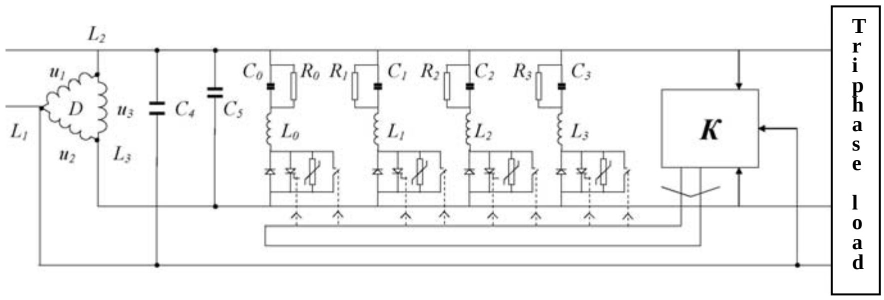

Fig. 1

The letter D denotes a triphase asynchronous motor with a cage rotor, which together with a set of capacitors, comprises a monophase-triphase converter. The constantly connected capacitors C_4 and C_5 compensate the reactive energy, consumed by D for the formation of magnetic field and symmetrize the output triphase voltage for unloaded or minimally loaded converter. The capacitor C_0 is switched on for a short time at big slip with the purpose to create big motor moment. It is mainly used at start and seldom – at considerable decrease of the revolutions of D as a result of incidental short overloading. The capacitors in battery C_1–C_3 are selectively turned on in order to support the symmetry within acceptable limits. The controller K monitors the symmetry of the output voltage and when the limits set are broken, it alters the capacitors, switched on in operating mode by the battery. The capacities are chosen after the dyadic law $C_1 = 2C_2 = 4C_3$ thus enabling linear scale of alteration. The total capacity must ensure optimal symmetry at nominal load. Diode-thyristor switching on is accepted, that enables activation of the capacitors at null current (and peak voltage) and lack of transient processes. The inductances L_0–L_3 limit the currents during converter passing to functioning or random non-sanctionary switching during converter's normal work. The varistors connected protect the valves from overvoltages, appearing in the line and particularly from overvoltages during converter switching on or non-sanctionary turning on of any thyristor. A possibility is built in for the controller to regulate the relays, which by their contacts shunt the switched on thyristors, in order to diminish the heat losses. The control is selected so that every commutation (switching on or off) of a capacitor is realized at open contacts and the commutation of the relays happens always at null voltages. A specific requirement towards these relays is that the open contacts must stand the doubled peak supply voltage of 1500 V. The resistors R_0–R_3 are connected for capacitors discharge after the converter is switched off.

2. Converter components

2.1. Capacitors

The capacitor C_4 is connected to the input clamps of the converter to compensate the magnetizing current of the asynchronous motor. Its capacity depends on the type and power of the motor and its design, and can be defined by its $\cos\varphi$ at nominal load. Its value is usually 0.95.

The following relations follow directly from motor equivalent scheme (Fig. 2):

$$jX = j\varpi L\,; \quad R = \frac{U^2}{P_n}\,, \quad \mathrm{tg}\varphi = \frac{X}{R} = \frac{XP_n}{U^2}.$$

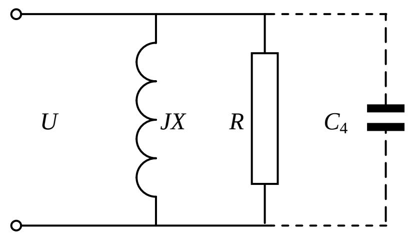

Fig. 2

For complete compensation with the help of capacitor C_4

$$\varpi L = \frac{1}{\varpi C_5} \text{ and } \frac{C_5}{P_n} = \frac{\mathrm{tg}\varphi}{\varpi U^2}$$

and for $\varpi = 2\pi 50 \quad \cos\varphi = 0.95$ and $U = 400$ V

$$\frac{C_5}{P_п} = 6\ \mu\text{F/kW}.$$

This value is just tentative. In the case of a real converter the capacitor is selected in relation to the minimal current consumed by the converter at idle running. For more powerful motors the requirements for greater value of $\cos\varphi$ are stricter and the relation above given must be decreased. Practically the measurements show that this relation decreases from 6 for $P_п = 4$ kW down to 2 for $P_п = 64$ kW.

The measurements necessary for the selection of an exact value of C_4 are quite simple – an experimentally chosen capacitor, which minimizes the current, consumed by the free of load converter.

The remaining 4 capacitors are selected in a way that ensures maximal symmetry of the voltages by module, at alteration of the load from zero up to nominal. The comparatively small step of capacity alteration (eight degrees), linearly changable, allows good compatibility with the real load. The total symmetrizing capacity C is defined for nominal load. The separate values are distributed according to a dyadic scheme. The value of C depends on the motor considered, but with sufficient approximation it may be determined by the condition that the reactive power of the capacitors must be equal to the nominal active power.

$$P = U^2 \varpi C,$$

$$C = \frac{P}{\varpi U^2}$$

For $U = 400\,\mathrm{V}$ and $f = 50\,\mathrm{Hz}$

$$C = 19.89P\ \mu\mathrm{F}.$$

In the expression P is given in kW.

Table 1 shows the calculated and rounded values of the total capacity for different power of the converter.

Table 1

P, kW	4	8	12	16	24	32	48	64
C, µF	80	160	240	320	480	640	960	1280

The capacitor C_4 remains constantly connected and its value may be assumed equal to the doubled step of alteration. In this way a given degree of symmetry is supported within the interval of idle running – active symmetrizing.

With an accepted constant step of alteration of the symmetrizing capacity, the base capacity (of capacitor C_3) is defined by the expression

$$C_3 + 2C_3 + 2C_3 + 4C_3 = C; \quad 9C_3 = C.$$

The capacity of a constantly turned on symmetrizing capacitor C_4 can be additionally corrected separately for the motor used. In idle running mode its value is chosen so that the voltage on it surpasses the input voltage by 3-4%. The altered value of C_4 with respect to the determined one does not cause alteration of the other capacitors, since they depend on the power consumed and only to a small extent – on motor parameters.

The start capacitor C_0 is connected with the purpose to reach big starting torque, for rapid transition to stationary mode and turning on under load. Its capacity is selected as a compromise between a high starting torque and acceptable starting current, consumed from the network. It remains connected up to obtaining of nominal revolutions – 0.3-2 s, depending on the motor power. The value $C_0 = (2 \div 4)C$ may be considered acceptable. In order to decrease its capacity, the controlled capacitors are connected with it.

Taking in mind the considerations above given, Table 2 shows the capacities for different in power converters. The capacitors denotations are as follows:

C_0 – a start capacitor,

C_1-C_3 – commutating capacitors,

C_4 – a constantly connected symmetrizing capacitor,

C_5 – a capacitor, compensating the magnetizing current of the motor.

The table contains also the effective value of the current passing through the capacitor. It must be taken into consideration when choosing the diode-thyristor pairs, the connecting wires, the cross section of the protection coils wires.

The table is composed for input voltage of 400 V and operation frequency of 50 Hz.

Due to the large values of the capacitors, the specific capacity, given in $\mu\text{F}/\text{cm}^3$ is an important factor in the selection of the type of the capacitors. The capacitors with polypropylene metalized dielectric are the most appropriate among the nowadays available.

Table 2

C, μF I, A	P, kW							
	4	8	12	16	24	32	48	64
C_0 I_0	200 25	400 50	500 63	600 75	1000 126	1200 151	2000 251	2400 302
C_3 I_3	10 1.3	20 2.5	30 3.8	40 5.0	60 7.5	80 10	100 13	150 19
C_2 I_2	20 2.5	40 5	60 7.5	80 10	120 15	160 20	200 25	300 38
C_1 I_1	40 5	80 10	120 15	160 20	240 30	320 40	400 50	600 75
C_4 I_4	20 2.5	20 5	60 7.5	80 10	100 13	120 15	150 19	180 23
C_5 I_5	20 2.5	40 2.5	60 7.5	80 10	100 15	120 20	150 25	180 38

For the commutating capacitors the discharge of energy during converter outage must be ensured. The capacitors used are with big capacity, with operation mode similar to the ones, compensating reactive energy in a power line. The time constant of the discharge $\tau = RC = 10\,\text{s}$ is considered acceptable. After switching off, most of the capacitors remain charged up to the peak value of the supply voltage. The energy, stored in each capacitor, is $W_C = \frac{1}{2}C2U_{\text{ef}}^2 = CU_{\text{ef}}^2$.

The power dissipated in the discharging resistors is $P_R = 2\frac{W_C}{\tau}$.

Table 3 shows the energy, stored in the capacitor, the value of the discharging resistor and the power dissipated in it for different values of the capacitors.

The choice of big values of the discharging resistors and long time of discharge is rather connected with the difficulties of its dissipation, than with energy saving. For large values of the capacitor it might be appropriate to increase the time constant τ.

Table 3. Energy, stored in the capacitor ($U_{ef} = 400$ V, $\tau = 10$ s)

C, μF	10	50	100	500	1000	2000	3200
W_C, J	1.6	8	16	80	160	320	512
R, kΩ	1000	200	100	20	10	5	3.1
P_R, W	0.32	1.6	3.2	16	32	64	102.4

2.2. Protection coils

For lossless turning on of the capacitors and mainly for protection against over-currents, a coil is connected in series to the circuit of each one. Its inductance must be selected so that the maximal value of the current passing through the valves must not overpass their limit values and also the single peak absorbed energy, represented by the factor $I^2t = \int i^2 dt$.

Every coil, combined with its respective capacitor, forms a circuit of second order with a small decrement of attenuation, which leads to the appearance of oscillatory processes after every initialization (new turning on) of the converter and to the possibility for occurrence during every period of the network.

The oscillatory system may be represented by the wave resistance $R = \sqrt{\frac{L}{C}}$, the resonance frequency $f = \frac{1}{2\pi\sqrt{LC}}$, and respectively the oscillation period $T = 2\pi\sqrt{LC}$, and null attenuation.

The peak loading during switching on depends on the phase of the supply voltage, when it is realized. The process will be analyzed for the most unfavorable phase.

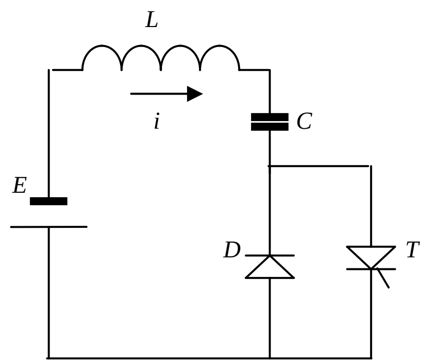

Fig. 3

The resonance frequency f is multiple higher than the network one f_0. This makes possible to regard the transient process as running at constant input voltage E. It is also assumed that the supply line is infinitely powerful. At switching on all the capacitors, through the respective diodes, are charged with current, altering according to the relation $i = \frac{E}{R} \sin 2\pi ft$ for $0 \le t \le \frac{T}{2}$.

At switching on, the motor is immobile and the voltage E is a result of the transformer transfer of energy from the supply phase and it is lower, (though with the same or opposite phase) than the supply voltage, hence $E < \sqrt{2}U$.

At $E = \sqrt{2}U$, for $t = \frac{T}{4}$ the current reaches the maximal value $I_p = \frac{\sqrt{2}U}{R}$.

The nominal effective value of the current running through the capacitor is $I_n = U\varpi_0 C$.

Here ϖ_0 denotes the angular frequency of the power line voltage.

For known I_n and feasible peak value of a semi-sinusoid with a duration $\frac{T}{2} = \pi\sqrt{LC}$, the necessary inductance of the coil is

(1) $$L \ge 2U \frac{I_n}{I_p} . \frac{1}{w_0 I_p} = 2U . \left(\frac{I_n}{I_p}\right)^2 . \frac{1}{w_0 I_n} .$$

For inductance, determined by this indicator, the resonance frequency is obtained as $f = f_0 \frac{I_p}{\sqrt{2} I_n}$.

The relation $\frac{I_p}{I_n}$ depends on the type of the valve and it usually accepts values within the limits 10-20, but for a single peak loading with a semi-sinusoid from the supply line 50 or 60 Hz. In the case considered the current pulse is also a semi-sinusoid, but with multiple higher frequency and thus with larger feasible value of the relation $\frac{I_p}{I_n}$.

Table 4 indicates the coils inductances for the different capacitors from Table 2, at $U_{ef} = 400\,\mathrm{V}$, $\frac{I_p}{I_n} = 15$, $f_0 = 50$ Hz.

Table 4

C, μF	10	20	30	40	60	80	120	160	240	320	480	640	1000	1500	2000	3000
I, nA	1.3	2.5	3.8	5	7.5	10	15	20	30	40	60	80	130	190	250	380
L, μH	8700	4500	3000	2260	1500	1130	750	570	377	283	188	141	87	60	45	30

The resonance frequency does not depend on the voltage and the capacity of the capacitor, but it is determined only by the network frequency and the relation $\frac{I_p}{I_n}$. From the relations

$$L = 2U\left(\frac{I_{\mathrm{n}}}{I_{\mathrm{p}}}\right)^2 . \frac{1}{\varpi_0 I_{\mathrm{p}}} \text{ and } \varpi^2 = \frac{1}{\sqrt{LC}}$$

the following is directly obtained:

(2) $$f = f_0 \frac{I_{\mathrm{p}}}{I_{\mathrm{n}}} \frac{1}{\sqrt{2}}.$$

For the given case

$$f = 50\frac{15}{\sqrt{2}} = 530 \text{ Hz},$$

at that, for $\frac{I_{\mathrm{p}}}{I_{\mathrm{n}}} = 10$, $f = 353\,\text{Hz}$.

The inductances, defined in this way, have too big values. For capacitors up to 100-150 μF it is more appropriate to use thyristors with nominal current, bigger than the actually running, and hence with larger peak current, thus decreasing the necessary inductance. On the contrary, for large capacitors – above 1000 μF, the use of bigger inductances may prove optimal for the decrease of thyristors pulse loading.

The thyristors must be selected by the indicator I^2t as well. It reflects more adequately the pulse single over-loading:

$$I^2t = \int_0^{T/2} \left(\frac{E}{R}\right)^2 \sin^2 \varpi t dt = \frac{1}{4} E^2 \frac{C}{L} T,$$

$$L = \left(\frac{\pi E^2}{2I^2t}\right)^2 C^3,$$

$$I_{ef} = \frac{E}{\sqrt{2}R},$$

$$T = \sqrt{LC}\, 2\pi = \frac{1}{f}, \quad \frac{E}{R} = \sqrt{2} I_{\mathrm{ef}},$$

$$I^2t = \left(\frac{1}{2}\frac{E}{R}\right)^2 \frac{1}{f},$$

(3) $$I^2t = \frac{1}{2f} I_{\mathrm{ef}}^2.$$

In order not to surpass the feasible value of I^2t, the actual inductance must be bigger than the one thus defined. For most of the thyristors this constraint is less exacting than the one for the peak current.

In order to maintain the thyristors parameters, the rate of current increase after triggering must not surpass their catalog set value. For most of the up-to-date thyristors it is above 50 A/μs and seldom – above 20 A/μs.

This speed in the converter is limited by the coil only and it has the value $\frac{di}{df} = \frac{E}{L}$.

The check must be done for the highest possible voltage in converter operation.

When the converter is switched on, the transient process ends up with capacitors charging up to the doubled input voltage. They remain charged with this voltage until their first activation – turning on in symmetrizing mode.

Thyristors activation takes place at minimal anode-cathode voltage U_{ak} that corresponds to the negative maximum of the voltage. At cyclic operation this voltage is close to zero. During the first activation after converter's switching on, the capacitor is charged up to the doubled value of this voltage and the thyristor is turned on at $U_{\text{ak}} = 2E - E = E$. A process begins, similar to the one at converter switching on, and it will not finish after time $t = \frac{T}{2}$, but will continue to develop, because the pair thyristor-diode is bi-directionally conductable. The process development after $t = T$ depends on the scheme accepted for thyristor control. For continuous activation, the oscillatory process continues until the complete dissipation of the over-energy, accumulated in the capacitor. At single activation of the thyristor for the network period, it remains turned on for the period $\frac{T}{2}$ only. At moment $t = \frac{T}{2}$, the current direction is reversed, it passes through the diode and the thyristor is deactivated. Under such control the dissipation of the over-energy is realized for several network periods and the valves are subjected to continuous stress loading.

A radical solution of the problem discussed is the restriction of the accumulated over-energy or its dissipation before the first activation of the respective thyristor. Fig. 4 shows one possible solution.

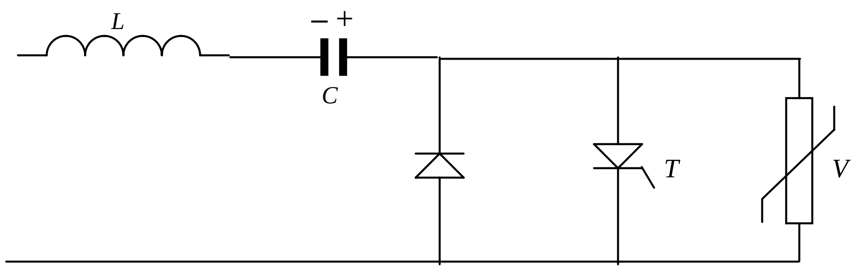

Fig. 4

The varistor V restricts the voltage on the valves to $E_V = 2\sqrt{2}U_{\max}$, where $U_{\max}$ is the maximal effective value of the input voltage and under the assumption that the converter is turned on at the moment of the maximum of the voltage negative semi-wave, as a result of the transient process it is charged up to $2\sqrt{2}U_{\mathrm{ef}}$.

During the next positive semi-wave the varistor is activated and it limits the voltage on C to $\sqrt{2}U_{\mathrm{ef}}$. The equivalent scheme, presenting the process, is shown on Fig. 5.

The initial conditions are $t=0$, $i=0$, $u_C=0$ and $\dfrac{di}{dt}=0$

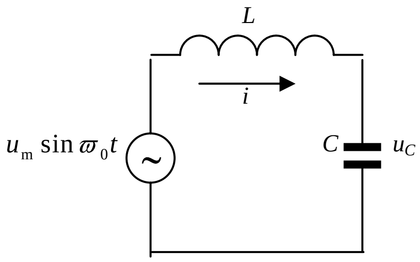

Fig. 5

The current stationary component is

$$i_{\mathrm{st}}=\frac{u_{\mathrm{m}}}{X}\sin\left(\varpi_0 t+\frac{\pi}{2}\right)=\frac{u_{\mathrm{m}}}{X}\cos\varpi_0 t\ ,$$

and the one of the free oscillations

$$i_{\mathrm{fr}}=\left(A+B\right)\cos\varpi t+j\left(A-B\right)\sin\varpi t,$$

$$X=\frac{1}{\varpi_0 c}-\varpi_0 L\ ,\quad \varpi=\frac{1}{\sqrt{LC}}\ ,$$

$$i=i_{\mathrm{st}}+i_{\mathrm{fr}}\ .$$

For $t=0$, $i=0$, $A+B=-\dfrac{u_{\mathrm{m}}}{X}$,

$$i=\frac{u_{\mathrm{m}}}{X}\left(\cos\varpi_0 t-\cos\varpi t\right),$$

(4) $$u_C=\frac{1}{C}\int i dt=\frac{1}{1-\left(\dfrac{\varpi_0}{\varpi}\right)^2}u_m\left(\sin\varpi_0 t-\frac{\varpi_0}{\varpi}\sin\varpi t\right).$$

The inequality $\varpi_0 \ll \varpi$ is always kept in the actual scheme.

The running current is composed of two components with equal amplitude. Due to the integrating action of the capacitor, the influence of the high frequency components of the current on the capacitor voltage u_C is quite weak.

The process will continue until the increase of the voltage u_C up to u_{m}. The capacitor is discharged by this value from the initial charge down to $2u_{\mathrm{m}}$ and the varistor is no longer conductable. Thanks to the energy, stored in the coil, the process can continue until the first zero of the current, but this cannot lead to considerable difference from u_{m} of the remainder voltage.

At converter switching on, some processes take place with big current and voltage over-loading and in order to limit them, the building in of additional components is implied, which are absolutely unnecessary for stationary operation of the converter.

The processes thus presented will not occur at switching on during the positive semi-wave.

The real processes are less intensive than the extremal processes herein considered. They do not account the losses and moreover – the lowered voltage at immobile motor. The influence of the losses is not considerable, because thanks to the protection means included the transient processes end for one period of the resonance oscillation.

3. Constructive dimensioning of the coils

It is assumed that the coils are made aerial, circular and with a square cross section of the wires forming the coil.

The inductance of a similar coil is determined by the expression

$$L = 5.10^{-8}\alpha d w^2 = A_L w^2 ,$$

$$A_L = 5.10^{-2}\alpha d \ \mu\text{H} / \text{T}^2 .$$

The coefficient α is determined according to the expression

$$\alpha = 2\pi\left[\left(1+\frac{\lambda}{6}\right)^2 \ln\frac{8}{\lambda^2} - 1.6967 + 0.408\lambda^2\right].$$

In the expression $\lambda = \dfrac{a}{d}$, d and a are respectively the coil length and side, as seen in Fig. 6.

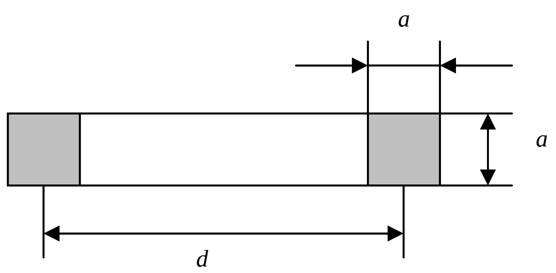

Fig. 6

For the coils considered the diameter d accepts values 0.15-0.35 m, a – 15-25 mm, at that λ varies within the limits 0.05-0.15.

Table 5 gives the relation $\alpha = \varphi(\lambda)$.

Table 5

λ	0.05	0.06	0.07	0.0714	0.08	0.0833	0.09	0.096	0.1	0.11	0.12	0.13	0.14	0.15
α	40.08	37.8	35.87	35.57	34.21	33.66	32.74	31.83	31.436	30.256	29.81	28.2	27.286	26.44

The values of α for different values of λ and diameter *d*, are given in Table 6.

Table 6

D, mm	200	200	250	300	300	350	300	250	350	350
λ	0.1	0.07	0.06	0.06	0.07	0.06	0.0833	0.1	0.096	0.0714
A_L, mH/T^2	0.314	0.36	0.473	0.473	0.538	0.66	0.505	0.393	0.55	0.623

The coils must be dimensioned with respect to inductance and power as well. For the coils in the start circuits large current loading must be accepted, because they turn on for a short time – up to 2 s and the frequency of turning on is not higher than 3 turnings per min. They must be designed with a stable mechanical shape in order to support the inductance expected. The coils positioning in the converter must be such that minimizes the electromagnetic interference among them.

The peak voltages on the coils can accept values twice higher than the power voltage, the voltage to common earth is increased by the phase one.

Four examples of coils dimensioning are given.

Example 1. d = 300 mm, a = 25, w = 16 turnings per min, λ = 0.0833; from Table 6 α = 33.66

$$L = 5\times10^{-2}\times33.66\times0.3\times16^2 = 129.3\ \mu\text{H}.$$

A coil like this can be made from wire with a cross section of 12 mm^2, looped in 4 layers, each layer with 4 turns. The resistance of such a conductor is 1.5 mΩ/m, the length of the wire for the whole coil is 15 m and the resistance – 0.0225 Ω.

For converters 48 and 64 kW the power dissipated in them is respectively

$$250^2\times0.0225 = 1.406\ \text{kW and } 300^2\times0.0225=2.025\ \text{kW}.$$

Such powers may be considered acceptable due to the short time of switching on.

However, it is more appropriate, without increasing the nomenclature of the coils to use two coils connected in parallel, made by wire 11.5×1 mm with strengthened isolation and average diameter of 300 mm. The dimensions for such a coil are preserved similar to the case above considered and hence the inductance will remain unaltered.

Example 2. d = 250 mm, w = 16, a = 25 mm,

$$L = 5\times10^{-2}\times0.25\times31.436 = 100.5\ \mu\text{H}.$$

If wire with cross section of 12 mm^2 is used, the resistance is

$$R = \pi\,0.25\times16\times1.5 = 18.8\ \text{m}\Omega.$$

Depending on the passing current, the coil can be made from wire with a different cross section, preserving the geometry.

Example 3. d = 200 mm, w = 16, a = 20, λ = 0.1, α = 31.436,

$$L = 5\times10^{-2}\times0.2\times31.436\times256 = 80.4\ \mu\text{H}.$$

With such geometry and selection of the wire cross section with respect to the passing current, all the coils with inductance 100 μH can be designed.

Example 4. d = 150 mm, w = 16, a = 15, λ = 0.1, α = 31.436,

$$L = 5\times10^{-2}\times0.15\times31.436\times256 = 60.2\ \mu\text{H}.$$

Coils with inductance of 60 μH can be made with such geometry.

The aerial coils are with very large dimensions, the magnetic field they create occupies a large volume and their placing creates problems due to mutual interference. The use of ferromagnetic circuits is attractive because of energy localization in a small volume with high density. The energy density is limited by the saturation induction B. For easy accessible materials it varies within the limits 1.1-1.8 T. For B = 1.5 T the density achieved is

$$\rho = \frac{1}{2}BH = \frac{1}{2}\frac{B^2}{\mu_0} = \frac{1}{2}.\frac{2.25}{4\pi10^{-7}} = 0.895\ \text{J/cm}^3.$$

For the converters considered the accumulated magnetic energy varies within the limits 1-450 J, which means volume of the clear air gap of 1.2-500 cm^3. The formation of a concentrated air gap with volume above 2 cm^3 requires too big magnetic circuits.

The magnetic circuits with distributed air gap in the form of toroid magnet dielectrics enable the design of the smallest inductances in low-power converters. For such magnetic circuits the achieved energy, set to the complete volume, ranges within the limits 15-39 mJ/cm^3 (Karson Electronics catalog). The volumes of the single toroids suggested reach 50 cm^3. If three toroids are used in parallel, the maximal feasible energy is 5.8 J.

Some of the coils can be made from magnet-electric circuits, but not all in one converter. Energy of 27 J is stored in the start inductance for a converter 4 kW. The necessary active volume of the magnetic circuit is $\frac{27}{39.10^{-3}} = 690\ \text{cm}^3$, and the whole volume occupied will be 1 l. It is obvious, that for the converters discussed, the use of aerial coils without ferromagnetic circuits, is the only acceptable solution.

4. Conclusion

Triphase energy is the basic one in obtaining rotational mechanics, thanks to the particular ease of creating oriented rotating magnetic field. This energy is supplied to the main type of motors in manufacturing – asynchronous motor with a cage rotor. However, this energy is not available everywhere. This concerns mainly some rural and desert districts, where providing triphase energy is too expensive. This refers also to the cases, when the households are supplied with single phase or bi-phase energy and the application of different techniques requires triphase energy supply.

The work above presented discusses dynamic converters of monophase into triphase voltage, using a standard triphase asynchronous motor with a cage rotor as a main component. According to the concept presented, the converter shows the following advantages:

- it is functionally simple;
- built up from highly reliable components, that have proved their fitness in very long exploitation;
- a motor with century-old qualities is used;
- it enables the connection of several consumers, including monophase ones;
- with the help of some not complex means the start moment can be raised, that allows switching on under nominal load or rapid acquiring of nominal revolutions;
- its price is not high in comparison with devices, converting monophase into tri-phase energy with different degree of symmetry, which is still quite expensive, not sufficiently reliable and requires qualified exploitation..

We believe that the converters presented will find wide application in the near future.

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Динамический конвертор монофазного напряжения в трифазное напряжение. Часть I: Структура и компоненты

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(Р е з ю м е)

Настоящая работа, представленна в двух частях этого издания, исследует подробно динамические фазовые конверторы, что помагает получить необходимые знания и завысимости для построения таких изделий. В первой части обсуждаются структура и компоненты конвертора. Во второй части исследуются старт и управление, которое приводит к заданной степени симметрии.