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CAPACITIVELY COMPENSATED CONVERTER-FED ASYNCHRONOUS MOTOR

István SCHMIDT

Department of Electrical Machines
Technical University of Budapest
H–1521 Budapest, Hungary

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Abstract

The capacitively compensated asynchronous motor is able to operate at changing frequency with line commutated converters. Having a constant value capacitor only a narrow range of speed can be covered economically. The paper shows a simple approximative Park vector method for the calculation of the steady state condition.

Keywords: asynchronous motor, capacitor unit, line commutated converter, nonsinusoidal supply, harmonic conditions.

1. Introduction

In the rectifier and inverter mode of operation of the line commutated converters, the voltage of the a.c. network is commutating. The role of the 3 phase network can be substituted by the induced voltage of a properly overexcited synchronous motor as well. The solution shown in *Fig. 1* is called load commutated converter-fed synchronous motor (CFSM).

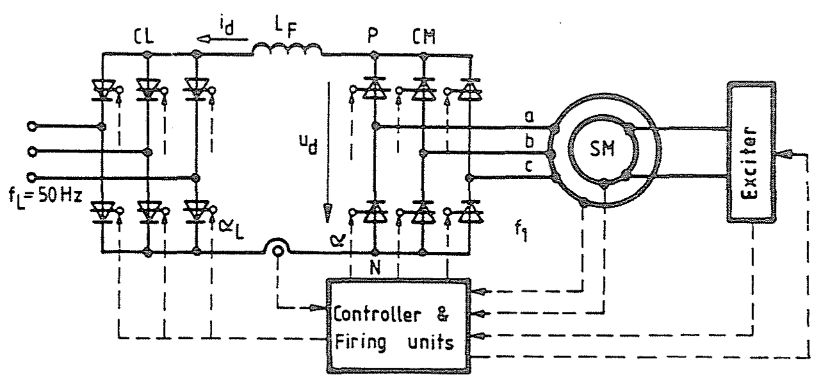

Fig. 1. Circuit diagram of the converter-fed synchronous motor

A squirrel cage asynchronous motor supplemented with series or parallel capacitors (made to become capacitive as resultant) is able to operate from a line commutated converter. As usually the parallel compensation is applied in practice, in the following it will be discussed in brief.

The basic circuit shown in *Fig. 2* is similar to *Fig. 1* of the CFSM. The solution shown in *Fig. 2* is called capacitively compensated converter-fed asynchronous motor (CFAM). In the CFSM and in the CFAM both the CL line and the CM motor side converters are operating with line commutation. The capacitive compensation is assured by the capacitance unit which consists of the capacitances denoted with C. L_K is the commutation inductance of the CM motor side converter. The line commutation of the thyristors of the CM is performed by the $\overline{u}$ terminal voltage of the capacitively compensated AM asynchronous motor. As L_K is very small from the point of view of the permissible di/dt necessary (about $L_K \approx 0.01$), thus the commutations in the CM can be regarded with good approximation as instantlike.

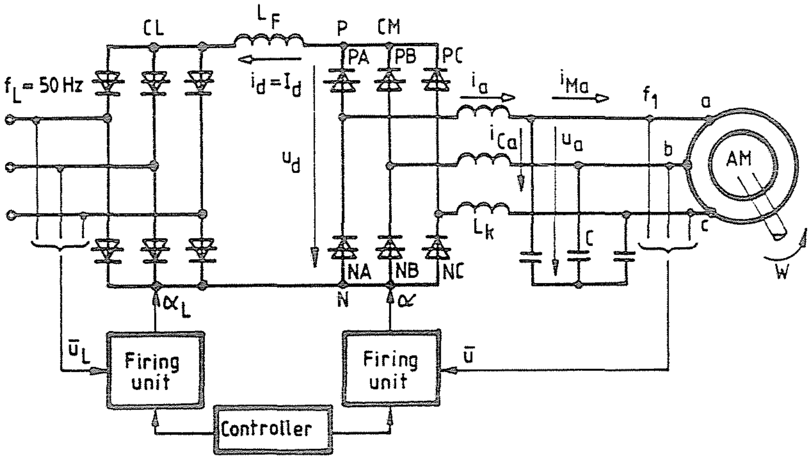

Fig. 2. Circuit diagram of the capacitively compensated converter-fed asynchronous motor

Analyzing the CFAM the following simplifications can be found:

a) the d.c. current is perfectly smooth ($L_F = \infty$), that is $i_d(t) = I_d =$ const.,

b) the overlap angle in the CM converter is zero ($\delta = 0°$).

2. Working Conditions with Fundamental Harmonics

The discussion based upon the fundamental harmonic can be carried out by using the simple equivalent circuit of *Fig. 3* (iron loss of the motor has been neglected). Here $\overline{Z}_1$ is the impedance of the asynchronous motor which depends on the $W_1 = 2\pi f_1$ and $W_r = W_1 - W = 2\pi f_r$ stator and rotor circuit angular frequency, respectively, at constant (R, R_r, L' and L_m) machine parameters. At f_1 = const. frequency and at $\overline{U}_1 = U_1$ = const. voltage, the $\overline{I}_{M1} = \overline{U}_1/\overline{Z}_1 = I_{M1}e^{j\varphi_M}$ fundamental harmonic current of the motor forms the well-known circle diagram as a function of W_r. This is the circle denoted by K in *Fig. 4*. In the figure it has been presumed that at $f_1 = f_{1n} = 1$ supply frequency and at $U_1 = U_n = 1$ supply voltage in the rated point marked with N, $I_{M1n} = I_n = 1$ and $\varphi_{Mn} = -30°(\cos\varphi_{Mn} \approx 0.867)$, and in the no load point $I_{M10} = I_n/3 = 1/3$ and $\varphi_{M0} = -90°$. The latter means the neglection of the stator resistance of the motor ($R \approx 0$), as a consequence of which the $\Psi_1 = U_1/W_1$ stator flux is independent of the load, and thus the motor and generator points are symmetrical to the imaginary axis.

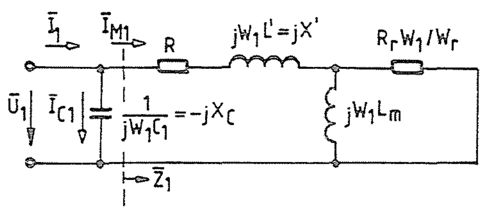

Fig. 3. Equivalent circuit for the fundamental harmonics

The fundamental harmonic current of the capacitor unit having C capacitance ($X_C = 1/(W_1C)$ reactance) is according to the following equation:

$$\overline{I}_{C1} = \frac{U_1}{-jX_C} = jU_1W_1C = I_{C1}e^{j90°} \tag{1}$$

constant at constant voltage and frequency. The a.c. current of the CM converter is the resultant of the load dependent $\overline{I}_{M1}$ and that of the constant $\overline{I}_{C1}$:

$$\overline{I}_1 = \overline{I}_{M1} + \overline{I}_{C1} = I_1e^{j\varphi}\,. \tag{2}$$

To enable the CM converter to operate with line commutation, an $\overline{I}_1$ resultant current having a capacitive φ phase angle is needed. It can be

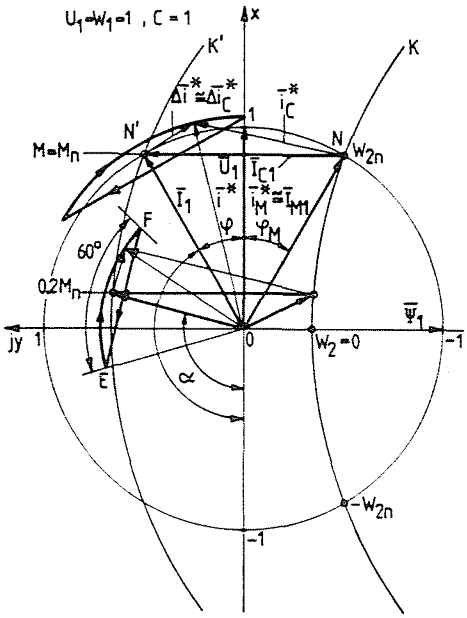

Fig. 4. Current vector diagram at rated frequency

seen from (1, 2) and from *Fig. 4* that, in spite of the inductive $\overline{I}_{M1}$ motor current, the resultant current can be turned into capacitive with $\overline{I}_{C1}$.

Measuring the α firing angle of the thyristors of the CM converter from the natural commutation instant of the $\overline{u}_1$ fundamental harmonic voltage of the capacitor unit, as a result of the instantlike commutation

$$\alpha = 180^\circ - \varphi \tag{3}$$

is the interrelation between the α firing angle and the φ phase angle. In a safe inverter mode of operation the largest firing angle is $\alpha_{\max} = 150 - 160^\circ$, to which a phase angle of $\varphi_{\min} = 20 - 30^\circ$ belongs. The minimum value of the capacitance to be inserted is determined by the

$$\varphi = \mathrm{arc}(\overline{I}_1) \geq \varphi_{\min} \tag{4}$$

inequality (angle condition). It can be seen from *Fig. 4* that, from this point of view, the point of the greatest load is critical.

The C capacitor in accordance with N rated point of rated frequency should be selected so big that, in the compensated rated point denoted with N', the phase angle of $\overline{I}_1$ current should be $\varphi = 30^\circ$. This compensation results according to (3) in a firing angle of $\alpha = 150°$. With this choice in the rated (N and N') points in accordance with the ONN' equal sided triangle, all currents are of rated value: $I_{M1} = I_{C1} = I_1 = I_n = 1$. This means based on (4) that the reactance of the inserted condenser unit per phase is

$$X_C = X_{Cn} = \frac{1}{W_{1n}C} = \frac{U_n}{I_{C1n}} = 1$$

at the $f_1 = f_{1n}$ frequency. This also means that in relative units $C = 1$ is the inserted capacitance. The fundamental harmonic rated power of the selected capacitor is at the rated frequency $Q_{C1n} = U_n I_{C1n} = U_n I_n = P_n = 1$, i.e. it corresponds to the rated seeming power of the motor.

At $U_1 = U_n$ and $W_1 = W_{1n}$ rated voltage and frequency, the $\overline{I}_{M1}$ current of the motor moves on the K circle. At the same time the $\overline{I}_1$ current of the CM converter is positioned on the capacitor current shifted K' circle with the constant $\overline{I}_{C1n} = jU_n W_{1n} C = jI_n = j1$. It can be seen that the φ phase angle of the $\overline{I}_1$ current (and therefore the α firing angle as well) strongly depends on the load of the motor.

If the frequency and the voltage are decreased proportionally and simultaneously related to the rated values (the rule of $f_1/f_{1n} = U_1/U_{1n} = v$ is kept), then for about $f > 10\,\mathrm{Hz}$, (until the R stator resistance is negligible) the flux of the motor is $\Psi_1 = \Psi_n = 1 = \text{const.}$, and thus the K circle diagram (the $\overline{I}_{M1}(W_r)$ current) remains unchanged. As both the reactance and the voltage of the capacitor unit, according to $X_C = X_{Cn}/v$ and $U_1 = vU_n$, respectively, depend on the frequency, thus in accordance with (1) the current of the capacitor changes with the square of the frequency:

$$\overline{I}_{C1} = jvU_n v W_{1n} C = v^2 \overline{I}_{C1n} \tag{5}$$

(in relative units $W_1 = f_1 = v$). Since at a given frequency $I_{C1} = \text{const.}$, thus in accordance with (2) for each $f_1 \leq f_{1n}$ frequency as current vector diagram individually a circle diagram is obtained. At changing frequency *Fig. 5* demonstrates the working conditions (the K' circles correspond to the K circle shifted by $\overline{I}_{C1}$). At the point P of motor operation, the currents $\overline{I}_{C1}$ and $\overline{I}_1$ valid at various frequencies have been marked. It can be seen that a small frequency change (between $1 \geq f_1 \geq 0.8$) can result in a significant change of $\overline{I}_{C1}$ and $\overline{I}_1$. Increasing the frequency in points P', P'' and P''', the φ phase angle of the $\overline{I}_1$ current of the CM converter will become greater, and the α firing angle will become smaller. Beyond a given current the curves K' cut the real axis (then $\varphi = 0°$, $\cos\varphi = 1$), which means the condition of being overexcited ceases. In the figure the $\varphi_{\min} = 20°$ straight line marked with H is the $\alpha_{\max} = 160°$ limit of the safe inverter mode of operation. The parts of the $\overline{I}_1$ current vector diagrams which are on the right side of this straight line cannot fall into the operational range. In rectifier (generator) mode of operation the $(-x)$ axis corresponding to $\alpha = 0°$ can be better approached as compared to the former one.

At a frequency $f_1 = vf_{1n}$ greater than the rated one $(v > 1)$, the voltage is maintained at $U = U_n =$ const. value, and because of this the flux decreases:

$$\Psi_1 = \frac{U_1}{W_1} = \frac{U_n}{vW_{1n}} = \frac{\Psi_n}{v}. \qquad (6)$$

The $\overline{I}_{M1}(W_r)$ current changes proportional to the Ψ_1 flux. In accordance with (1) the current of the capacitor unit is in this range proportional to the frequency:

$$\overline{I}_{C1} = jU_n vW_{1n}C = v\overline{I}_{C1n}. \qquad (7)$$

In *Fig. 5* the $W_1 = 1.1$ and 1.2 field weakened current vector diagrams are drawn as well.

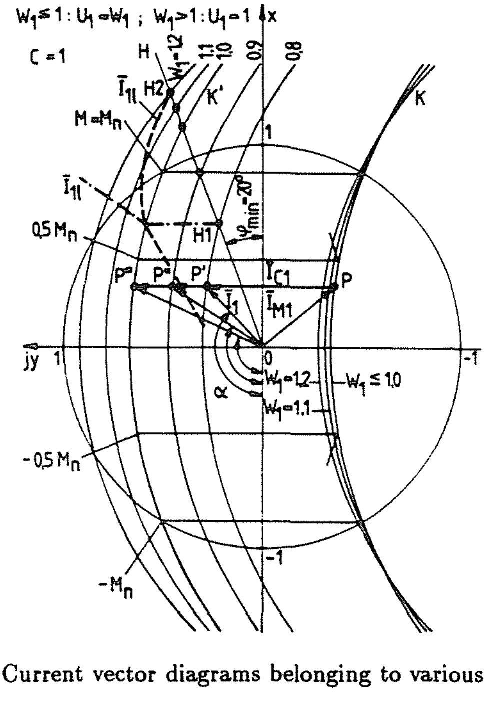

Fig. 5. Current vector diagrams belonging to various frequencies

The $\overline{I}_1$ current vector diagrams obtained at the individual frequencies are qualitatively similar to the current vector diagrams of a synchronous machine at constant excitation. The increase of the frequency has the same effect as if the excitation of the synchronous machine were increased.

From the $\overline{I}_1$ current vector diagrams the motor torque can be read in the following way. The torque of the CFAM can be calculated based on (2):

$$M = \overline{\Psi}_1 \times \overline{I}_{M1} = \overline{\Psi}_1 \times \overline{I}_1 = \Psi_1 I_1 \cos\varphi = -\Psi_1 I_1 \cos\alpha . \tag{8}$$

To a given torque in the $\Psi_1 = \Psi_n$ constant flux range a constant real power component (the horizontal straight line originating from the K circle diagram) belongs, in the $\Psi_1 = \Psi_n/v$ field weakened range a real power component increasing with v belongs to it. In *Fig. 5* in the K and the K' circle diagrams, the points of $M = \pm M_n = \pm 0.8$ and $M = \pm M_n/2$ are connected with a thin line. The vertical cuts of the straight lines denoted with H determine in the constant flux range maximum torques obtainable at the individual frequencies, but in the field weakened range they determine the obtainable maximum power values.

The angular velocity of the Park vectors, i.e. that of the synchronous angular velocity with $\delta = 0°$:

$$W_1 = \frac{\pi}{3\sqrt{3}} \frac{U_d}{\Psi_1 \cos\alpha} \tag{9}$$

(U_d is the d.c. voltage). According to $W = W_1 - W_r$ the angular velocity of the rotor deviates from it with the M torque dependent W_r.

After this, if the $M_l(W)$ characteristic of the load is known, at $C =$ const. the $\overline{I}_{1l}(W_1)$ working points pertaining to it can be defined. The $\overline{I}_{1l}$ characteristic drawn in *Fig. 5* with a dash-and-dot line corresponds to the $M_l \approx 0.7M_n =$ const. load in the $0.8 \leq W_1 \leq 1.2$ range. In the same range the $\overline{I}_{1l}$ characteristic drawn with dotted line shows an example for a load increasing with speed (e.g. for a pump).

This limit working point serves as a basis for the calculation of the capacitor C, which is getting onto the H straight line. For the constant load C has to be chosen for the minimum frequency value (point H_1). However, at loads having a pump character the situation is just the opposite; there from the point of view of the capacitor the point of the largest frequency (at the same time of the largest load) is authoritative (point H_2).

The control of the W angular velocity (indirectly that of the W_1 synchronous angular velocity) and that of the Ψ_1 flux is a fundamental control engineering task in the CFAM. There are two independent places in the CL line and the CM motor side converter where intervention is possible (with the α_L and the α firing angles), and thus these two control tasks can be solved. Obviously, the limits resulting from the line commutation of the CM have to be taken into consideration.

At the CFAM the optimum operation would correspond to the situation in *Fig. 5* where each motor working point would be on the H denoted straight line of the safety limit of the inverter operation. This could be assured only by the speed and load dependent continuous change of the capacitance. The changing C capacitor would play such a role at the CFSM as the changing excitation of the synchronous machine does. At a CFAM compensated with a continuously variable capacitance, the cast of roles could be similar to that of the CFSM: the Ψ_1 flux can be maintained with C, with the α_L firing angle of CL (with U_d) the W_1 synchronous angular velocity can be changed, however, the working point can be optimized with the α firing angle of the CM converter. Unfortunately, a continuously variable heavy current capacitor can be realized only by complicated and expensive electronic circuits, and so for the present task it cannot be applied economically enough.

In the case of a compensation with constant C, corresponding to *Fig. 5* acceptable conditions are obtained only in a relatively narrow frequency band. In the high frequency and low load points the working conditions are very bad already within the $0.8 \leq f_1 \leq 1.2$ frequency covering strip. In these cases the large (nearly the rated) reactive component of the $\overline{I}_1$ current unnecessarily loads the CM and CL converters and the $f_L = 50\,\text{Hz}$ network (bad $\cos\varphi$ and $\cos\varphi_L$).

3. Upper Harmonic Conditions

Fig. 6a shows the approximate substituting circuit of the capacitively compensated asynchronous motor valid for $f_1 > 10\,\text{Hz}$ nonsinusoidal supply (L' is the transient inductance of the stator).

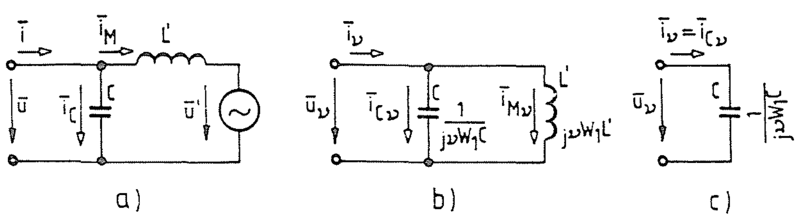

Fig. 6. Equivalent circuit for nonsinusoidal supply. *a*) Equivalent circuit for instantaneous values; *b*), *c*) Equivalent circuit for upper harmonics

The fundamental harmonics which depend on the working point determine the $\overline{u}' = \overline{U}' e^{jW_1 t}$ inserted voltage and the $\overline{i} = \overline{i}_1 + \Delta\overline{i}$ converter current. The latter is determined by the fundamental harmonic currents of (2) because of the $\delta = 0°$ instantlike commutation. In *Fig. 4*, the $\bar{i}^* = \bar{I}_1 + \Delta\bar{i}^*$ current in the synchronous rotating coordinate system of the converter has been drawn in two points, as $M = M_n$ and $0.2\,M_n$ as well. The $\bar{i}^*$ current moves along an α dependent 60° arc of a circle having a radius of $I_0 = (2/\sqrt{3})I_d = (\pi/3)I_1$ (I_d is the d.c. current). $\bar{I}_1$ points to the point of gravity of this arc. The firings occur in the point marked with F, and thus from this point because of the $\delta = 0°$ overlap angle $\bar{i}^*$ moves over to point E with a sudden jump. The $\overline{U}'$ amplitude of the inserted $\bar{u}'$ voltage can be determined from the $\overline{U}' = \overline{U}_1 - jW_1 L'\bar{I}_{1M} \approx I_{1M}(jW_1 L_m) * (R_r W_1/W_r)$ equation ($*$ is here the symbol of the replus). From *Fig. 3* at $R = 0$ approximation, the correctness of the second form can be realized. The $\overline{U}'$ complex amplitude is a function of the f_1 frequency (of W_1) and that of the load.

Based on the equivalent circuit of *Fig. 6a* for a given $\bar{u}'$ inserted voltage and for a given $\bar{i}$ converter current all current and voltage time functions can be calculated. However, by the application of a further approximation a simpler calculation is offered.

The resultant impedance acting against the upper harmonics can be calculated on the basis of the equivalent circuit of *Fig. 6b*:

$$\overline{Z}_\nu = \frac{\overline{U}_\nu}{\bar{I}_\nu} = \frac{-j\nu W_1 L'}{\nu^2 W_1^2 L'C - 1}\,. \tag{10a}$$

As the investigated circuit can be economically applied only in a narrow frequency band around f_{1n}, therefore the magnitude of the necessary capacitor is $C \approx 1$. The transient inductance is $L' \approx 0.2$. With this data it can be realized that for the upper harmonics of $\nu = 1 + 6k$ orders ($k = \pm 1$, $\pm 2, \ldots$) approximately the

$$\overline{Z}_\nu \approx \frac{-j}{\nu W_1 C} = -jX_{C\nu} \tag{10b}$$

impedance, i.e. the equivalent circuit of *Fig. 6c* can be used for calculation purposes. The above approximation corresponds to the fact that the flow of the $\Delta\bar{i}$ upper harmonic currents of the $\bar{i}$ converter current is circulated wholly through the capacitor unit, and thus the motor current consists purely of fundamental harmonic ($\bar{i}_M \approx \bar{i}_{M1}$, i.e. for $\nu \neq 1$ $\bar{i}_{M\nu} = \bar{0}$ and $\bar{i}_{C\nu} = \bar{i}_\nu$).

Based on *Fig. 6c* the amplitude of the $\bar{u}_\nu$ terminal voltage and that of the $\bar{i}_{C\nu} = \bar{i}_\nu$ capacitor (converter) current are in the following relation with each other:

$$\overline{U}_\nu = \frac{-j}{\nu W_1 C}\bar{I}_\nu\,, \qquad U_\nu = \frac{U_1}{\nu^2}\frac{I_1}{I_{C1}} \tag{11a,b}$$

($I_\nu = |I_1/\nu|$, $U_1 = I_{C1}/(W_1 C)$). From this the distortion factor characteristic of the pulsation of the terminal voltage can be calculated:

$$k_u = \frac{\Delta U_{\mathrm{eff}}^2}{U_1^2} = \sum_{\nu \neq 1} \frac{U_\nu^2}{U_1^2} = \left(\frac{I_1}{I_{C1}}\right)^2 \sum_{\nu \neq 1} \frac{1}{\nu^4} \approx \left(\frac{I_1}{I_{C1}}\right)^2 \; 0.00215 . \tag{12}$$

In the case of $I_1 = I_{C1}$ this corresponds numerically to the distortion factor of the flux of the simple voltage inverter supply. However, in general, the k_u factor is even smaller than that, since in a great part of the operational range $I_1/I_{C1} < 1$. Because of the approximation applied in the upper harmonic impedance, in the reality the voltage distortion factor is somewhat larger than usual.

The time function of the nonsinusoidal $\bar{i}_C = \bar{i}_{C1} + \Delta\bar{i}_C$ capacitor current and that of the $\bar{u} = \bar{u}_1 + \Delta\bar{u}$ have to be determined now. The nodal equation (2) is valid for instantaneous values as well:

$$\bar{i}_C = \bar{i} - \bar{i}_M \approx \bar{i} - \bar{i}_{M1} , \qquad \Delta\bar{i}_C = \Delta\bar{i} - \Delta\bar{i}_M \approx \Delta\bar{i} . \tag{13a,b}$$

Corresponding to the approximate equation established for the current deviations, the $\Delta\vec{i}^*$ converter current deviation drawn in *Fig. 4* indicates at the same time the $\Delta\vec{i}_C^*$ deviation of the capacitor current. The $\Delta I_{C\mathrm{eff}} \approx \Delta I_{\mathrm{eff}}$ resultant effective value of the upper harmonic currents of the capacitor is characteristic of the additional conducting losses. In knowledge of the ΔI_{eff} effective value of the current deviation of the converter, the current surplus which loads the capacitor unit can be calculated:

$$\Delta I_{C\mathrm{eff}} \approx \Delta I_{\mathrm{eff}} = k_i I_1^2 \approx k_{i0} I_1^2 \tag{13c}$$

($k_{i0} \approx 0.097$ is the distortion factor, I_1 is the fundamental harmonic of the converter current belonging to $\delta = 0°$ overlap angle).

Forder investigating the 1/6 period of an angle $\alpha < W_1 t < \alpha + \pi/3$ which begins with the firing of the NC thyristors of the CM converter we find that the fundamental harmonics and the current of the CM converter change in the following way:

$$\bar{u}_1 = U_1 e^{jW_1 t} , \qquad \bar{i}_{C1} = I_{C1} e^{j\left(\frac{\pi}{2} + W_1 t\right)} , \tag{14a,b}$$

$$\bar{i} = \bar{I}_0 = I_0 e^{\frac{j2\pi}{3}} , \qquad \bar{i}_1 = I_1 e^{j(\pi - \alpha + W_1 t)} , \tag{14c,d}$$

$$\bar{i}_{M1} = I_{M1} e^{j(-\varphi_M + W_1 t)} \tag{14e}$$

($I_0 = (2/\sqrt{3}) I_d$, $I_1 = (2\sqrt{3}/\pi) I_d$). Substituting the $\bar{i}$ and $\bar{i}_{M1}$ currents into (13a), the $\bar{i}_C(t)$ capacitor current will be obtained. *Fig. 7a,* which shows

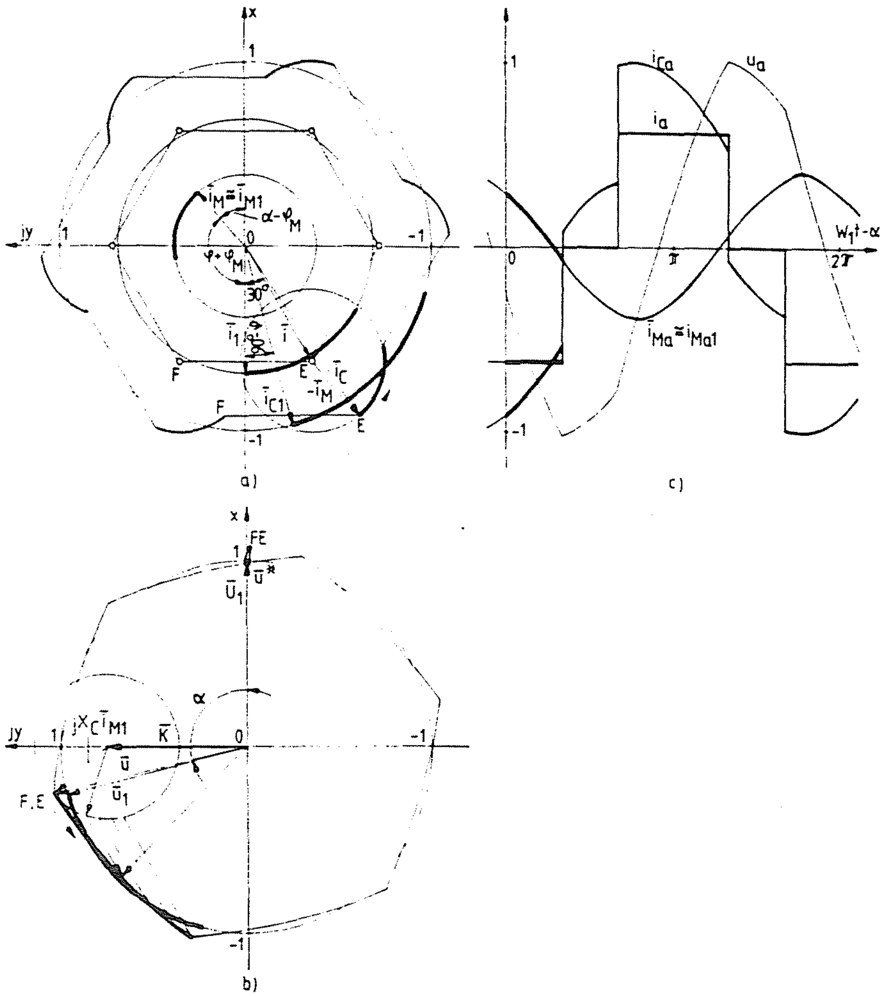

Fig. 7. Park vectors and time functions of the $M = 0.2M_n$ small load point. *a*) The current Park vectors; *b*) Park vector of the terminal voltage; *c*) Time functions

the Park vectors of the currents, has been drawn for the $M = 0.2M_n$ point of small load. The investigated tact has been made thicker, and the angles marked at the starting point of the tact can be understood on the basis of the equations (14).

From the $\bar{i}_C$ current corresponding to equation $\bar{i}_C = C\mathrm{d}\bar{u}/\mathrm{d}t$, the terminal voltage of the capacitor unit (the motor) can be determined. In the investigated 1/6 period:

$$\overline{u} = jX_C\overline{i}_{M1} + X_C\overline{I}_0\,(W_1t - \alpha) + \overline{K}\,. \tag{15}$$

Substituting from (14) the $\overline{i}_{M1}$ current of the motor and the $\overline{i} = \overline{I}_0$ current of the converter, the $\overline{u}(t) = u_x(t) + ju_y(t)$ complex time function of the voltage Park vector will be obtained. The $\overline{K}$ integration constant which has the dimension of the voltage according to the $\overline{u}(\alpha) = \overline{u}(\alpha+\pi/3)e^{-j(\pi/3)}60^\circ$ symmetry condition will get the $\overline{K} = j(\pi/3)X_CI_0$ value. *Fig. 7b* shows the $\overline{u}$ and $\overline{u}_1$ voltage vectors. In the figure, for the tact investigated, a construction resulting from the vectorial equation (15) is shown. From the small deviation of the leaf shaped curve of the $\overline{u}^*$ voltage in the synchronous rotating coordinate system and from the $\overline{u}^* = \overline{U}_1 = U_1$ fundamental harmonic it can be seen that $\overline{u}$ is very near to the $\overline{u}_1$ fundamental harmonic.

In *Fig. 7c* the time functions can be seen: i_a is the current of the converter, i_{Ca} that of the capacitor unit and $\overline{i}_{Ma}$ that of the a phase of the motor, while u_a is the voltage of phase a.

The Park vectors and the time functions are working point dependent. To demonstrate this, in *Fig. 8* the Park vectors and the time functions of the $M = M_n$ point of rated torque (point N' of *Fig. 4*) are given. In spite of the significant increase of the pulsation of the $\overline{i}_C$ current of the capacitor unit, the terminal voltage is still very near to the sinusoidal shape. The current in the squirrel cage asynchronous motor of the capacitive compensated drive is even in the reality of near sinusoidal shape, and thus the additional loss and the torque pulsation can be neglected. The capacitor unit, however, is loaded with a significant amount of upper harmonic currents.

4. Firing Control of the Machine-side Converter

Just as with the CFSM, it is advisable to fire the thyristors of the CM motor side converter of the CFAM with self control. The firing (the synchronization) can be performed from the terminal voltage considering the narrow frequency band and the near sinusoidal shape of the terminal voltage. In this simple case the firing control of the CM motor side and CL line side converters can become very similar. The firing of CM can be controlled by the $\overline{u}$ vector of the voltage or from the phase (or line) magnitudes of the terminal voltage. The firing control operated from the $\overline{u}$ terminal voltage equals with good approximation the firing control operated from the $\overline{u}_1$ fundamental harmonic voltage. From *Figs. 7b* and *8b* (from $\overline{u}^*$ and $\overline{U}_1$) it can be seen that the shift between vectors $\overline{u}$ and $\overline{u}_1$ is in space smaller than 5°.

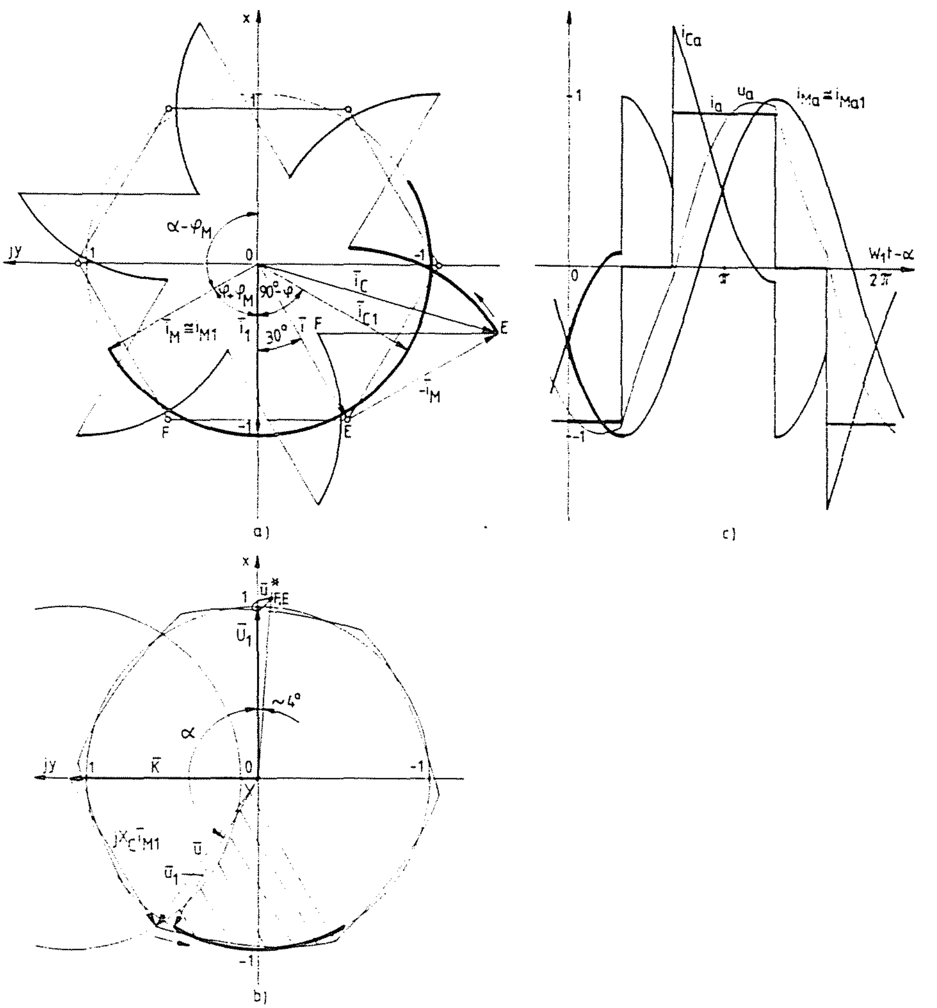

Fig. 8. Park vectors and time functions of the rated load point. *a*) The current Park vectors; *b*) Park vector of the terminal voltage; *c*) Time functions

Its simplicity is the advantage of the CFAM against the inverter-fed drives having forced commutation, and the application of the squirrel cage motor against the CFSM. However, besides these advantages it has some disadvantageous properties as well. Starting causes a serious problem. The simplest solution is starting from the $f_L = 50\,\mathrm{Hz}$ network. The asynchronous motor and the C capacitor unit can be switched to the network either simultaneously or one after the other. After the direct starting the transition from the constant frequency network operation to the variable frequency converter operation has to be solved. During the transition the excitation increase of the capacitively compensated motor might cause a problem. This problem may occur at the switch-off of the drive as well. It is the $1/\sqrt{LC} \leq W_1 \leq 1/\sqrt{L'C}$ angular frequency range ($L = L' + L_m$) where we have to count with the capacitive excitation increase. Having average parameters $1/\sqrt{3 \cdot 1} \approx 0.58 \leq W_1 \leq 1/\sqrt{0.2 \cdot 1} \approx 2.24$, i.e. $0.58 \cdot 50 = 29\,\text{Hz} \leq f_1 \leq 2.24 \cdot 50 = 112\,\text{Hz}$ is this range. After the excitation increase, the asynchronous machine brakes in generator mode of operation at the $W_r = -(R_r/R)W_1$ angular frequency of the rotor circuit.

Considering the advantages and disadvantages, the field of application of the CFAM is fairly narrow. It can be recommended especially for large power, small and medium voltage, asynchronous motor driving pumps.

At the CFAM drive applying in the CM converter extinguishable semiconductors (e.g. GTO), a PWM current inverter-fed asynchronous motor can be obtained. In this case with a much smaller capacitor the full frequency range could be covered. Nowadays this drive is going to be more employed.

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