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ANALYSIS OF ZERO SEQUENCE CURRENT OF ASYNCHRONOUS MOTORS FOR SENSORLESS CONTROL

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Abstract. A sensorless vector control method with the identification of non-measurable state variables through the use of zero sequence current for induction motors with windings connected in delta has been considered. It is proved that this method allows to create a system with a wide control range of the motor speed.

Keywords: zero sequence current, asynchronous motors, sensorless control.

Introduction. Sensorless vector control [1] of induction motors implies absence of any sensor on the shaft and inside the machine using appropriate estimation of state variables based on a mathematical model. However, most of existing methods of sensorless estimation are based on the use of an idealized mathematical model of an induction motor, that leads to significant problems when running at a low speed. That is why for the synthesis of a wide range control it is necessary to apply estimation which uses the anisotropic properties of the engine.

In [2] a method for diagnosis of the engine faults based on the analysis of zero sequence current in the case when the motor windings are connected in delta is presented. Therefore, it is appropriate to analyze the use of zero sequence current for the analysis of the position of the induction motor anisotropy axis.

Materials and methods. Expansion of the range of rotational speed control of sensorless drives is only possible through improving the methods of flux and rotor speed estimation of the engine when running at a speed which is close to zero. An idealized model of the induction motor cannot satisfy these requirements. The methods of sensorless identification based on the anisotropic properties of the machine include high- voltage or input test vectors to the fundamental voltage that feeds the engine. Besides, the current response to the input of additional voltage is analysed, and the anisotropy axis position is defined. Since the engine windings are connected in delta, the presence of anisotropy leads to the zero sequence current, the possibility of using this signal for sensorless estimation of non-measurable state variables should be studied.

Results. The equation of induction motor stator circuit electric equilibrium can be written as:

$$v_A = L_{\sigma A}\frac{di_A}{dt} + i_A R_A + \frac{d\psi_A}{dt}; \quad (1)$$

$$v_B = L_{\sigma B}\frac{di_B}{dt} + i_B R_B + \frac{d\psi_B}{dt}; \quad (2)$$

$$v_C = L_{\sigma C}\frac{di_C}{dt} + i_C R_C + \frac{d\psi_C}{dt}, \quad (3)$$

where v_A, v_B, v_C – voltage of the motor windings; i_A, i_B, i_C – stator phase currents; $L_{\sigma A}$, $L_{\sigma B}$, $L_{\sigma C}$ – inductance of coil scattering; R_A, R_B, R_C – active windings resistance; $\frac{d\psi_A}{dt}$, $\frac{d\psi_B}{dt}$, $\frac{d\psi_C}{dt}$ – the back-EMF of the motor.

Heterogeneity of the asynchronous machine that results from steel saturation or the presence of discrete rotor bars leads to changes in leakage inductance depending on the position of the axis of the corresponding anisotropy. Assuming the assumption about the sinusoidal character of inductance modulation generated by the anisotropy, we can write:

$$L = \begin{bmatrix} l_a(t) & 0 & 0 \\ 0 & l_b(t) & 0 \\ 0 & 0 & l_c(t) \end{bmatrix}; \quad (4)$$

$$l_a(t) = l_0 + l_{aн}\cos 2\theta_{aн}; \quad (5)$$

$$l_b(t) = l_0 + l_{aн}\cos\left(2\theta_{aн} - \frac{2\pi}{3}\right); \quad (6)$$

$$l_c(t) = l_0 + l_{aн}\cos\left(2\theta_{aн} + \frac{2\pi}{3}\right); \quad (7)$$

where L – the matrix of the own inductances of the machine, l_0 – a constant component of the machine winding inductance, $l_{ан}$ – the winding inductance component that is modulated by the presence of the machine anisotropy, $\theta_{ан}$ – the position the axis of anisotropy relative to the A coil axis of the motor.

In order to determine the position of the axis of anisotropy using a high-frequency input signal to the voltage of the motor stator or using test vectors. The introduction of high-frequency signal is undesirable because of additional energy loss, acoustic noise and electromagnetic torque ripple precipitation. Therefore, a more rational solution is the introduction of two test vectors within the period of pulse-width modulation (PWM). As a standalone voltage inverter is used in most modern frequency-controlled drives, it is advisable to use as test vectors those ones that correspond to the base vectors of the scheme. Such vectors are shown in Fig. 1. Vectors that are opposite in direction should be used (eg, V_1 and V_4) during each PWM period. By applying them to the same length of time, you can avoid any effect on the output voltage of the inverter.

Finding the position of the axis of anisotropy causes no difficulties, provided that the measured values of leakega inductance observed effect of single anisotropic properties. But this condition is quite complex for real asynchronous machine, which has two large anisotropic properties: one that is associated with a change in inductance due to saturation of the steel under the influence of the main flux, and one that is associated with the presence of discrete rotor bars. To reduce the influence of bars on the rotor machine is skewed, but studies show [4-8], it cannot reduce it to the level of modulation, which can be neglected it.

The change of leakage inductance provided by simultaneous influence of anisotropy, associated with saturation and anisotropy of the rotor bars can be written as:

$$L_{\sigma A} = L_{\sigma c} + L_{\sigma н}\cos(2\omega_н t) + L_{\sigma pc}\cos(n\omega_{pc} t + \phi_0); \quad (8)$$

$$L_{\sigma B} = L_{\sigma c} + L_{\sigma н}\cos\left(2\omega_н t - \frac{2\pi}{3}\right) + L_{\sigma pc}\cos\left(n\omega_{pc} t - \frac{2\pi}{3} + \phi_0\right); \quad (9)$$

$$L_{\sigma C} = L_{\sigma c} + L_{\sigma н}\cos\left(2\omega_н t + \frac{2\pi}{3}\right) + L_{\sigma pc}\cos\left(n\omega_{pc} t + \frac{2\pi}{3} + \phi_0\right), \quad (10)$$

where $L_{\sigma c}$ – the average leakage inductance, $L_{\sigma н}$ – leakage inductance component modulated by the presence of anisotropy saturation machine, $L_{\sigma pc}$ – leakage inductance component that is modulated by the presence of machine rotor bars, $\omega_н$ – speed of rotational axis anisotropy caused by the presence of anisotropy that equal to rotation of the machine field, n – number of machine rotor bars, ω_{pc} – speed of induction motor shaft, ϕ_0 – initial position angle anisotropy due to the presence of rotor bars.

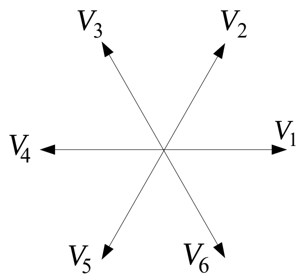
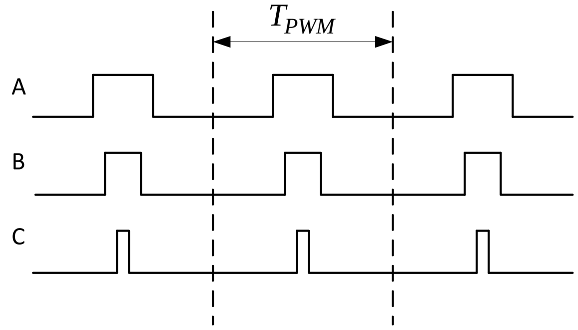
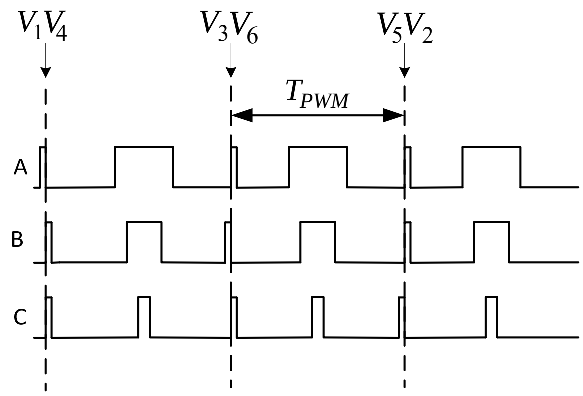


Figure 1. Test voltage vectors



a)



b)

Figure 2. Waveform of the inverter control without applying test vectors (a) and with applying test vectors (b)

Consider the equation of state of the induction motor in case of applying a test vector.

The equivalent circuit of the system is shown in Fig. 3. The system of equations can be written as

$$\begin{cases} V_{dc} = E_{ab} + L_{\sigma ab}\dfrac{di_{ab}}{dt}; \\ 0 = E_{bc} + L_{\sigma bc}\dfrac{di_{bc}}{dt}; \\ -V_{dc} = E_{ca} + L_{\sigma ca}\dfrac{di_{ca}}{dt}. \end{cases} \quad (11)$$

Zero sequence current for this case can be expressed with the system (11) as follows

$$\frac{di_0}{dt} = \frac{V_{dc}}{L_{\sigma ab}} - \frac{V_{dc}}{L_{\sigma ca}} - \frac{E_{ab}}{L_{\sigma ab}} - \frac{E_{bc}}{L_{\sigma bc}} - \frac{E_{ca}}{L_{\sigma ca}}. \quad (12)$$

When working in the area of low speeds the back-EMF of the motor is low compared to the voltage level of the inverter DC, so it can be neglected

$$\frac{di_0}{dt} = \frac{V_{dc}}{L_{\sigma ab}} - \frac{V_{dc}}{L_{\sigma ca}}. \quad (13)$$

Substituting (8)-(10) in (13) and after completing the simplification and neglecting components with small amplitudes

$$\frac{di_0}{dt} \approx V_{dc}\frac{-\sqrt{3}L_{\sigma H}\sin\left(2\omega_H t + \frac{\pi}{3}\right) - \sqrt{3}L_{\sigma pc}\sin\left(n\omega_{pc}t + \frac{\pi}{3} + \phi_0\right)}{L_{\sigma c}^{\ 2}} = k\left(L_{\sigma H}\sin\left(2\omega_H t + \frac{\pi}{3}\right) + L_{\sigma pc}\sin\left(n\omega_{pc}t + \frac{\pi}{3} + \phi_0\right)\right) \quad (14)$$

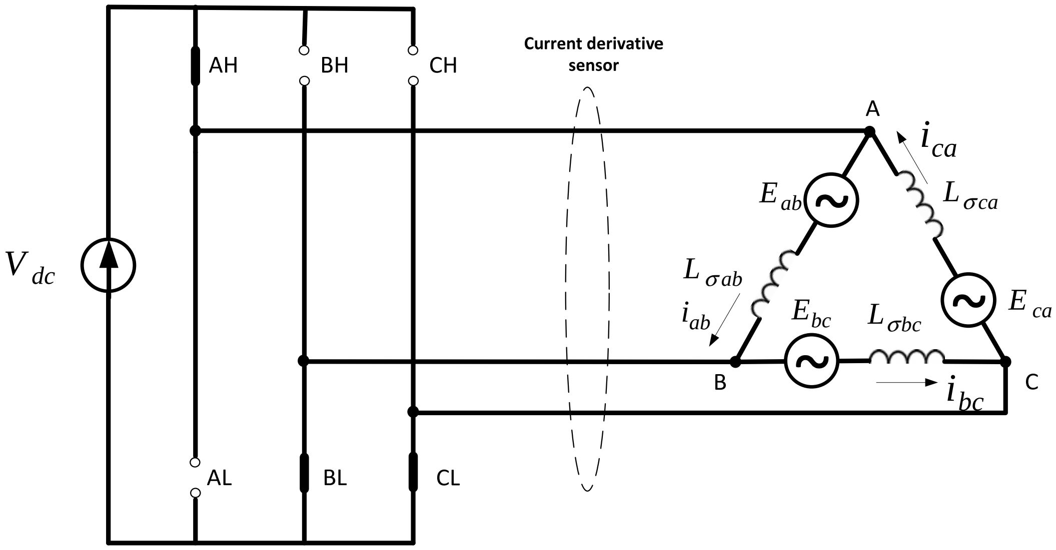


Figure 3. Replacement scheme of frequency-controlled drive in case of applying the test vector V_1

Thus, the equation (14) shows that the signal derivative zero sequence current is the sum of two components: the first is modulated anisotropy saturation $L_{\sigma H}\sin\left(2\omega_H t + \frac{\pi}{3}\right)$, and the second - the presence of discrete rotor bars $L_{\sigma pc}\sin\left(n\omega_{pc}t + \frac{\pi}{3} + \phi_0\right)$. In this case, the actual problem is the separation of these components to create opportunities for individual assessment of motor rotor position and direction of the main flux.

A discrete-field model has been made in Ansoft Maxwell 3D in order to carry out the analysis. The simulation results of the system are presented in Fig. 4-5. Of these, it is evident that when running at low speed error estimation of rpm studied way less than 3%, which is satisfactory for the use of this signal in the system sensorless vector control of induction motors [9-11].

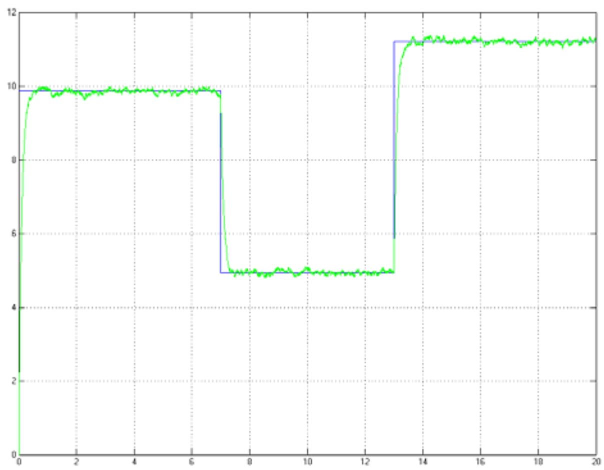


Figure 4. Diagram of the signal of initialized speed and of speed measured with the investigated method

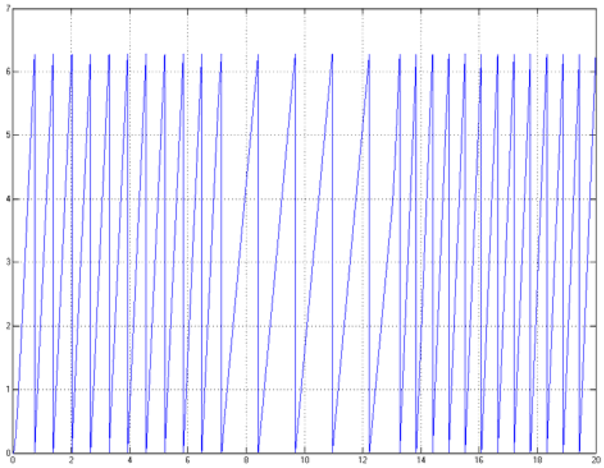


Figure 5. Diagram of rotor position angle measured with the investigated method

Conclusions. The study confirmed the possibility of using a zero sequence current signal for determination the position of the anisotropy axis and, consequently, the position of the vector of the main stream of the machine or the motor rotor depending on the nature of the anisotropy. Besides, to achieve a satisfactory performance for the accuracy characteristics of non-measurable state variables one should use a sensor of the current derivative. The advantage of this method over traditional ones is the possibility of using only one sensor instead of three.

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