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SLIDING-MODE DIRECT FIELD-ORIENTED CONTROL OF SIX-PHASE INDUCTION MOTOR

ŚLIZGOWE BEZPOŚREDNIE STEROWANIE POLOWO-ZORIENTOWANE SZEŚCIOFAZOWYM SILNIKIEM INDUKCYJNYM

Abstract

The paper presents the combined control of a six-phase squirrel-cage induction motor with the application of Sliding-Mode Control and Direct Field-Oriented Control (SMC-DFOC). The mathematical model of the six-phase squirrel-cage induction motor has been presented. The design procedure of the control structure of SMC-DFOC has been described. Simulation studies on the considered control system have been carried out and the results of the simulation studies have been presented and discussed.

Keywords: six-phase induction motor, Direct Field-Oriented Control, Sliding-Mode Control, simulation studies

Streszczenie

W artykule przedstawiono metodę sterowania sześciofazowym silnikiem indukcyjnym klatkowym z zastosowaniem sterowania ślizgowego i bezpośredniego sterowania polowo-zorientowanego. Omówiono model matematyczny sześciofazowego silnika indukcyjnego. Opisano proces projektowania układu sterowania ślizgowego i bezpośredniego sterowania polowo-zorientowanego. Przeprowadzono badania symulacyjne rozpatrywanej struktury sterowania, przedstawiono i omówiono wybrane wyniki badań symulacyjnych.

Słowa kluczowe: sześciofazowy silnik indukcyjny, bezpośrednie sterowanie polowo-zorientowane, sterowanie ślizgowe, badania symulacyjne

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1. Introduction

In recent years, there has been a great interest in using multi-phase squirrel-cage induction motors (motors with the number of stator phases greater than three). It is caused by the development of power electronics frequency converters that can be operated as multi-phase systems of energy, powered by the industrial three-phase grid. The main advantages of multi-phase induction motors are: the reduction of torque pulsations and the reduction of losses caused by higher harmonics of motor currents [5, 6, 8, 9]. Multi-phase motors have great reliability because they can be conditionally operated at failure of one or more phases of the stator winding. The rated stator phase currents of multi-phase induction motors are smaller than the rated stator phase currents of three-phase induction motors of the same power [5, 6, 8, 9].

In the literature, there are some papers about the constructions and control methods of the multi-phase induction motors [1–6, 8, 9]. Most studies are based on the considerations of Field-Oriented Control or Direct Torque Control [5, 9]. However, a small number of publications constitutes the problems of using the Sliding-Mode Control for the multi-phase induction motors.

In this paper, the novel combined control method of a six-phase squirrel-cage induction motor with application of Sliding-Mode Control and Direct Field-Oriented Control has been presented. It is expected that the consideration of the sliding-mode control technique for the control systems of a multi-phase induction motor allows to achieve robust control of the motor state variables, fast dynamic responses and easy hardware implementation [1, 4, 6, 7, 10, 11]. It is known that the chattering of controlled variables is the main disadvantage of sliding-mode control. In the literature, there are some articles about the methods of elimination of these negative effects [1, 7, 10, 11].

The paper is divided into six sections. Section 1 provides an introduction. Section 2 is dedicated to the mathematical model of the six-phase squirrel-cage induction motor. In Section 3, the introduction to the Sliding-Mode Control is presented. Section 4 deals with the design method of Sliding-Mode Direct Field-Oriented Control. Simulation studies are presented and discussed in Section 5. Section 6 is dedicated to the concluding remarks.

2. Mathematical model of six-phase squirrel-cage induction motor

A mathematical model of six-phase squirrel-cage induction motor has been formulated on the basis of commonly used simplifying assumptions [5, 6, 8, 9]: stator and rotor phase windings are considered as concentrated; the symmetry of the phase motor windings and the linearity of motor magnetic circuits are assumed; the eddy currents and iron losses are neglected; the electromagnetic variables and parameters of rotor winding have been transformed to the stator winding side. In this paper, the six-phase squirrel-cage induction motor with the axes of stator windings shifted electrically by sixty degrees is considered. The star connection of six-phase stator winding with single isolated neutral point and short-circuited circuits of six-phase rotor winding are assumed.

The initial mathematical model of the six-phase squirrel-cage induction motor is formulated for motor state variables considered in the stator and rotor phase coordinate system. This model has the form of a set of differential equations with variable coefficients, which are changing as a function of the rotor angle position. By the use of appropriate transformation matrices, which are presented in [5, 9], the stator and rotor state variables are transformed to the state variables presented in the common rectangular coordinate system and the state variables presented in some other additional coordinate systems. This way, the simpler form of mathematical model of six-phase induction motor, described by the differential equations with constant coefficients, is obtained.

The final equations of the transformed mathematical model of the six-phase induction motor take the following form [5, 9]:

– the stator and rotor voltage equations, expressed in the general rectangular coordinate system *x-y*, which rotates relative to the stator at arbitrary angular electrical speed ω_k:

$$u_{sx} = R_s i_{sx} - \omega_k \psi_{sy} + \frac{d}{dt}\psi_{sx}, \qquad u_{sy} = R_s i_{sy} + \omega_k \psi_{sx} + \frac{d}{dt}\psi_{sy} \tag{1}$$

$$0 = R_r\, i_{rx} - \left(\omega_k - \omega_e\right)\psi_{ry} + \frac{d}{dt}\psi_{rx}, \qquad 0 = R_r\, i_{ry} + \left(\omega_k - \omega_e\right)\psi_{rx} + \frac{d}{dt}\psi_{ry} \tag{2}$$

– the stator voltage equations expressed in the additional coordinate system *z*1-*z*2:

$$u_{sz1} = R_s\, i_{sz1} + \frac{d}{dt}\psi_{sz1}, \qquad u_{sz2} = R_s\, i_{sz2} + \frac{d}{dt}\psi_{sz2} \tag{3}$$

– the equations for stator and rotor flux linkages:

$$\psi_{sx} = L_s i_{sx} + L_m i_{rx}, \qquad \psi_{sy} = L_s i_{sy} + L_m i_{ry} \tag{4}$$

$$\psi_{rx} = L_r i_{rx} + L_m i_{sx}, \qquad \psi_{ry} = L_r i_{ry} + L_m i_{sy} \tag{5}$$

– the equation of the motor electromagnetic torque:

$$T_e = \frac{n}{2} p_b \left(\psi_{ry} i_{sx} - \psi_{rx} i_{sy}\right) = \frac{6}{2} p_b \left(\psi_{ry} i_{sx} - \psi_{rx} i_{sy}\right) \tag{6}$$

– the mechanical motion equation:

$$T_e - T_m = \frac{J_m}{p_b} \cdot \frac{d}{dt}\omega_e \tag{7}$$

where:

$u_{sx}, u_{sy}, i_{sx}, i_{sy}, i_{rx}, i_{ry}$ – components of the stator voltage vector, stator current vector and rotor current vector in the *x-y* coordinate system, respectively,

$u_{sz1}, u_{sz2}, i_{sz1}, i_{sz2}$ – components of the stator voltage vector and stator current vector defined in the additional *z*1-*z*2 coordinate system, respectively,

$\psi_{sx}, \psi_{sy}, \psi_{rx}, \psi_{ry}$ – components of the stator and rotor flux linkage vector in the *x-y* coordinate system,
ψ_{sz1}, ψ_{sz2} – components of the stator flux linkage vector in the additional *z*1-*z*2 coordinate system,
ω_k – arbitrary angular speed of the coordinate system relative to the stator,
ω_e – the electrical angular speed of the motor,
T_e – the motor electromagnetic torque,
T_m – the load torque,
J_m – the inertia of the drive system,
R_s, R_r – stator and rotor phase resistance, respectively,
L_{ls}, L_{lr} – the leakage inductance of stator and rotor, respectively,
L_m – the motor magnetizing inductance,
$L_s = L_{ls} + L_m, L_r = L_{lr} + L_m$ – the total inductance of stator and rotor, respectively,
p_b – the number of motor pole pairs,
n – the number of stator phases.

In the mathematical model of the six-phase induction motor presented above, the equations for stator and rotor zero components and the equations for rotor components considered in the additional coordinate system have been omitted, because under the aforementioned assumptions, these state variables are always equal to zero.

On the basis of the torque equation (6), it can be stated that, in the multi-phase induction motor, only the state variables determined in the general coordinate system *x-y* are involved in the generation of the electromagnetic torque. The other state variables represented in the additional systems (labeled with the index *z*) do not contribute to the torque production. However, these variables should be considered in the exact analysis because the values of these variables may be different from zero and can influence upon the resultant stator phase currents and losses.

3. Introduction to the Sliding-Mode Control

Sliding mode control is a nonlinear control technique, featuring remarkable properties of accuracy, robustness, and easy tuning and implementation [10]. Nowadays, this type of control is widely used for the control of electromechanical systems and power electronic converters [7, 10, 11]. On this basis, the first studies have been taken on the use of the sliding mode control to control the multi-phase induction motors [1, 4, 6]. In this paper, the improved sliding mode control with the application of equivalent control has been studied and discussed.

Sliding mode control systems are designed to drive the system states onto a particular surface named the sliding surface. Once the sliding surface is reached, sliding mode control keeps the states on the sliding surface. Hence, the sliding mode control is a two-part controller design. The first part involves the design of a sliding surface; the second one is concerned with the selection of a control law that will make the switching surface attractive to the system state [10].

In the article, the following form of switching functions $s(t)$ has been used [1, 6, 7, 10]:

$$s_i(t) = x_i^*(t) - x_i(t), \tag{8}$$

where:

x_i^* – the reference value of controlled variable,
x_i – the real value of controlled variable,
$I = 1,\ldots, m$ – the number of the controlled variables.

The control laws applied in the controllers must meet the requirements for the existence and reachability of the sliding mode. In the sliding-mode control, the following condition has been included [1, 7, 10, 11]:

$$s(t) \cdot \frac{d}{dt} s(t) < 0 \tag{9}$$

In the applied method of the equivalent sliding-mode control, the control signal u^* is considered as the sum of two components: continuous component u_{eq} and discontinuous component u_n [1, 7, 11]:

$$u^* = u_{eq} + u_n \tag{10}$$

The continuous component u_{eq} is considered as the signal of the equivalent sliding-mode control. The continuous component u_{eq} is determined on the basis of the mathematical model of the control system assuming that all uncertainties are equal to zero. The nonlinear control system of six-phase induction motor can be described by the following general equation system [1, 11]:

$$\frac{dx}{dt} = f(t) + g(t) \cdot u(t) \tag{11}$$

where:

x – the vector, which represents the state variables of the system,
$f(t), g(t)$ – vector functions,
$u(t)$ – the vector of control signals.

The continuous component u_{eq} of the control signal of the sliding-mode controller can be determined as the solution of the following equation [1, 11]:

$$\frac{\partial s}{\partial t}\left[f(t) + g(t) \cdot u(t) \right] = 0 \tag{12}$$

And hence:

$$u_{eq} = -\left[\frac{\partial s}{\partial t} g(t) \right]^{-1} \frac{\partial s}{\partial t} f(t) \tag{13}$$

In order to account for real nonzero uncertainties, the discontinuous control signal is introduced. The discontinuous part u_n of the signal of sliding-mode control has been designed as the term defined by the following equation [1, 10]:

$$u_n = k_n \cdot \operatorname{sgn} s(t) \tag{14}$$

where $k_n > 0$ is the element of vector of the gain coefficients of the discontinuous part of the sliding-mode controller [1, 10].

4. Sliding-Mode Direct Field-Oriented Control

In this paper, the method of Direct Field-Oriented Control (DFOC) with equivalent sliding-mode controllers has been applied for the control of the six-phase induction motor. The main goal of a field-oriented control is obtaining decupled control of the electromagnetic torque and rotor flux of the induction motor.

The rotor field-oriented control is based on the mathematical model of the six-phase induction motor formulated in the rectangular coordinate system *x-y*, which rotates at angular speed $\omega_{\psi r}$ equal to the angular speed of the rotor flux vector:

$$\omega_k = \omega_{\psi r} = \frac{L_m i_{sy}}{\tau_r \psi_r} + \omega_e, \quad \tau_r = \frac{L_r}{R_r} \tag{15}$$

From the conditions for the rotor field-oriented control, it is required that the component ψ_{rx} of the rotor flux vector is equal to the magnitude of the rotor flux vector and the component ψ_{ry} of the rotor flux vector is equal to zero [1, 8]:

$$\psi_{rx} = \psi_r, \quad \psi_{ry} = 0 \tag{16}$$

In the conventional Direct Field-Oriented Control (DFOC) methods, the PI controllers have been generally used. In the improved control structure presented in this paper, the PI controllers have been replaced by the sliding-mode controllers. The developed scheme of DFOC with sliding-mode controllers is presented in Fig. 1. The estimator of the instantaneous magnitude of the rotor flux vector and the instantaneous angle of the rotor flux vector must be used in this method. The rotor-flux model based on the measured stator currents and motor angular speed is used in the estimation block. In the Sliding-Mode DFOC system, two outer control loops with Sliding-Mode Controllers (SMC) have been applied: the control loop for the motor angular speed ω_e and the control loop for the magnitude of the rotor flux vector ψ_r. The Sliding-Mode Controller of the motor angular speed determines the reference component i_{sy}^* of stator current vector, which is responsible for the control of motor electromagnetic torque. The Sliding-Mode Controller of the magnitude of the rotor flux vector determines the reference component i_{sx}^* of the stator current vector, which is responsible for the rotor flux control. In the presented control system, two inner control loops have been also applied: the control loop for the component i_{sx} of the stator current vector and the control loop for the component i_{sy} of the stator current vector. The output signals from the Sliding-Mode Controllers of the components i_{sx} and i_{sy} of the stator current vector are the reference values of the components u_{sx}^* and u_{sy}^* of the stator voltage vector. These reference values are then transformed to the stator phase coordinate system and are sent to PWM modulator, which determines the switching states of the six-phase Voltage Source Inverter. Detailed descriptions of the design and synthesis of the SMC controllers used in the DFOC structure are presented in the next sections.

4.1. Equivalent sliding-mode controller of motor angular speed

The equation of the motor electromagnetic torque (6) after taking into account the condition for the rotor flux orientation [1, 8] can be represented in the following form:

$$T_e = 3 p_b i_{sy} \Psi_r \tag{17}$$

The mechanical motion equation (7) can be written in the following form:

$$\frac{d\omega_e}{dt} = \frac{p_b\left(T_e - T_m\right)}{J_m} = \frac{p_b\left(3 p_b i_{sy} \Psi_r - T_m\right)}{J_m} \tag{18}$$

The switching function and its time derivative for SMC controller of motor angular speed can be represented in the following forms:

$$s_\omega = \omega_e^* - \omega_e \tag{19}$$

$$\frac{d}{dt} s_\omega = \frac{d}{dt}\omega_e^* - \frac{d}{dt}\omega_e \tag{20}$$

The equation (20) after substitution of the equation (18) takes the following form:

$$\frac{d}{dt} s_\omega = \frac{d}{dt}\omega_e^* - \frac{3 p_b^2 i_{sy} \Psi_r - p_b T_m}{J_m} \tag{21}$$

On the basis of the equation (21), the control signal $i_{sy}{}^*$ can be determined as the sum of two appropriate components: continuous component $i_{syeq}{}^*$ and discontinuous component $i_{syn}{}^*$ which are established by the separate control blocks. The control law of sliding-mode controller of motor angular speed can be presented in the form:

$$i_{sy}^* = \underbrace{\frac{J_m \frac{d}{dt}\omega_e^* + p_b T_m}{3 p_b^2 \Psi_r}}_{i_{syeq}^*} + \underbrace{k_{syn} \cdot \operatorname{sgn}\left(s_\omega\right)}_{i_{syn}^*} \tag{22}$$

where:

$k_{syn} > 0$ – the gain of the discontinuous part of the sliding-mode speed controller.

The sliding-mode controller of motor angular speed has been included in Fig. 1 as the block marked as “Speed controller”.

4.2. Equivalent sliding-mode controller of the magnitude of the rotor flux vector

After appropriate manipulations of the motor model equations, the magnitude of the rotor flux vector ψ_r can be presented in the form:

$$\frac{d}{dt}\psi_r = -R_r i_{rx} \tag{23}$$

The component i_{rx} of the rotor current vector is determined as:

$$i_{rx} = \frac{\psi_r - L_m i_{sx}}{L_r} \tag{24}$$

The equation (23) after substitution of the equation (24) takes the form:

$$\frac{d}{dt}\psi_r = \frac{L_m i_{sx} - \psi_r}{\tau_r} \tag{25}$$

The switching function and its time derivative for sliding-mode controller of the magnitude of the rotor flux vector can be represented in the following forms:

$$s_\psi = \psi_r^* - \psi_r \tag{26}$$

$$\frac{d}{dt}s_\psi = \frac{d}{dt}\psi_r^* - \frac{d}{dt}\psi_r \tag{27}$$

The equation (27) after substitution of the equation (25) has the form:

$$\frac{d}{dt}s_\psi = \frac{d}{dt}\psi_r^* + \frac{\psi_r - L_m i_{sx}}{\tau_r} \tag{28}$$

On the basis of the equation (28), the control signal i_{sx}^* can be determined as the sum of two components: continuous component i_{sxeq}^* and discontinuous component i_{sxn}^* which are established by the separate control blocks. After some manipulations, the following form is obtained:

$$i_{sx}^* = \underbrace{\frac{\tau_r}{L_m}\cdot\frac{d}{dt}\psi_r^* + \frac{\psi_r}{L_m}}_{i_{sxeq}^*} + \underbrace{k_{sxn}\cdot\operatorname{sgn}\left(s_\psi\right)}_{i_{sxn}^*} \tag{29}$$

where: $k_{sxn} > 0$ – the gain of the discontinuous part of the controller of magnitude of the rotor flux vector.

The sliding-mode controller of the magnitude of the rotor flux vector is included in the Fig. 1 as the block marked as "Flux controller".

4.3. Equivalent sliding-mode controller of component i_{sx} of stator current vector

With taking into account that for the rotor field-oriented control the $\psi_{ry} = 0$, the i_{ry} component of the rotor current vector can be expressed in the form:

$$i_{ry} = \frac{-L_m i_{sy}}{L_r} \tag{30}$$

The equations of the components of stator flux vector after some manipulations can be presented in the following forms:

$$\psi_{sx} = L_{ls} i_{sx} + \psi_r \tag{31}$$

$$\psi_{sy} = L_{ls} i_{sy} + \frac{L_{lr}\left(\omega_{\psi r} - \omega_e\right)\psi_r}{R_r}. \tag{32}$$

The switching function and its time derivative for the controller of the component i_{sx} of stator current vector are defined as:

$$s_{ix} = i_{sx}^* - i_{sx} \tag{33}$$

$$\frac{d}{dt} s_{ix} = \frac{d}{dt} i_{sx}^* - \frac{d}{dt} i_{sx} \tag{34}$$

The equation (1) of the component u_{sx} of the stator voltage vector after substitution of the equations (31) and (32) has the form:

$$u_{sx} = R_s i_{sx} - \omega_{\psi r}\left[L_{ls} i_{sy} + \frac{L_{lr}\left(\omega_{\psi r} - \omega_e\right)\psi_r}{R_r}\right] + L_{ls}\frac{d}{dt} i_{sx} \tag{35}$$

Equation (35) after some algebraic operations can be presented in the form:

$$\frac{d}{dt} i_{sx} = \frac{u_{sx} - R_s i_{sx} + \omega_{\psi r}\left[L_{ls} i_{sy} + \frac{L_{lr}\left(\omega_{\psi r} - \omega_e\right)\psi_r}{R_r}\right]}{L_{ls}} \tag{36}$$

Equation (34) after substitution of the equation (36) takes the following form:

$$\frac{d}{dt} s_{ix} = \frac{d}{dt} i_{sx}^* - \frac{u_{sx} - R_s i_{sx} + \omega_{\psi r}\left[L_{ls} i_{sy} + \frac{L_{lr}\left(\omega_{\psi r} - \omega_e\right)\psi_r}{R_r}\right]}{L_{ls}} \tag{37}$$

The output signal from the sliding-mode controller of the component i_{sx} of stator current vector can be considered as the sum of two components: continuous component $u_{sxeq}{}^*$ and discontinuous component $u_{sxn}{}^*$, which are determined by the separate control blocks:

$$u_{sx}^* = \underbrace{L_{ls}\frac{d}{dt}i_{sx}^* + R_s i_{sx} - \omega_{\psi r}\left[L_{ls}i_{sy} + \frac{L_{lr}\left(\omega_{\psi r} - \omega_e\right)\psi_r}{R_r}\right]}_{u_{sxeq}^*} + \underbrace{k_{suxn}\,\mathrm{sgn}\left(s_{ix}\right)}_{u_{sxn}^*} \tag{38}$$

where:

$k_{suxn} > 0$ – the gain of the discontinuous part of the controller of component i_{sx} of stator current vector.

The control part responsible for equivalent sliding-mode control of the i_{sx} component of the stator current vector is included in the Figure 1 as "i_{sx} controller".

4.4. Equivalent sliding-mode controller of i_{sy} component of stator current vector

The switching function and its time derivative for controller of the component i_{sy} of stator current vector are defined as:

$$s_{iy} = i_{sy}^* - i_{sy} \tag{39}$$

$$\frac{d}{dt}s_{iy} = \frac{d}{dt}i_{sy}^* - \frac{d}{dt}i_{sy} \tag{40}$$

The equation of the component u_{sy} of stator voltage vector (1) after substitution of the equations (31) and (32) and with using of the conditions required for the field-oriented control has the form:

$$u_{sy} = R_s i_{sy} + \omega_{\psi r}\left(L_{ls}i_{sx} + \psi_r\right) + \frac{d}{dt}L_{ls}i_{sy} \tag{41}$$

Equation (41) after some algebraic operations can be represented in the following form:

$$\frac{d}{dt}i_{sy} = \frac{u_{sy} - R_s i_{sy} - \omega_{\psi r}\left(L_{ls}i_{sx} + \psi_r\right)}{L_{ls}} \tag{42}$$

Equation (40) after substitution of the equation (42) has the following form:

$$\frac{d}{dt}s_{iy} = \frac{d}{dt}i_{sy}^* - \frac{u_{sy} - R_s i_{sy} - \omega_{\psi r}\left(L_{ls}i_{sx} + \psi_r\right)}{L_{ls}} \tag{43}$$

The output signal from the sliding-mode controller of the component i_{sy} of stator current vector can be considered as the sum of two components: continuous component $u_{syeq}{}^*$ and discontinuous component $u_{syn}{}^*$ which are determined by the separate control blocks:

$$u_{sy}^{*} = \underbrace{L_{ls}\frac{d}{dt}i_{sy}^{*} + R_s i_{sy} + \omega_{\psi r}\left(L_{ls}i_{sx} + \psi_r\right)}_{u_{syeq}^{*}} + \underbrace{k_{suyn}\,\mathrm{sgn}\left(s_{iy}\right)}_{u_{syn}^{*}} \tag{44}$$

where: $k_{suyn} > 0$ – the gain of the discontinuous part of controller of i_{sy} component of stator current vector.

The part responsible for equivalent sliding-mode control of the i_{sy} component of the stator current vector is included in the Fig. 1 as "i_{sy} controller".

4.5. The scheme of the Sliding-Mode Direct Field-Oriented Control

The system of the Sliding-Mode Direct Field-Oriented Control of six-phase squirrel-cage induction motor with application of the synthesis procedures presented above has been shown in Fig. 1.

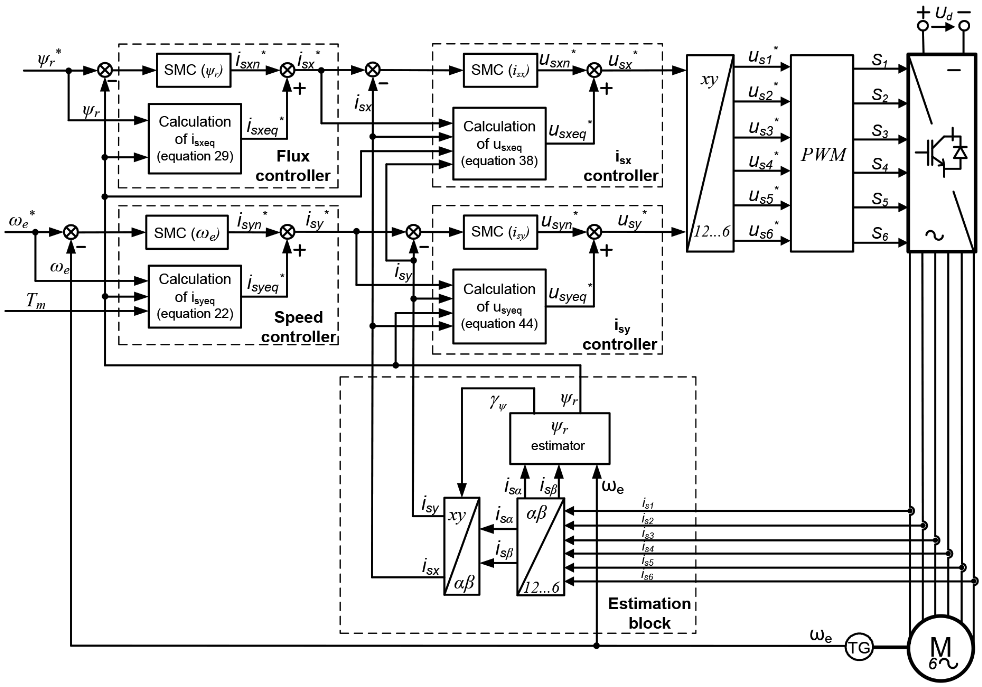

Fig. 1. Sliding-Mode DFOC of six-phase squirrel-cage induction motor

The appropriate transformation blocks of the motor stator currents have been applied in the Sliding – Mode Direct Field-Oriented Control system. The values of the stator currents transformed from the phase coordinate system to the *x-y* and α-β coordinate systems are used in the control system and in the estimator of the rotor flux vector. Six-phase induction motor is supplied by the six-phase two level Voltage Source Inverter.

5. Simulation results

Simulation studies have been carried out in order to verify the presented method of Sliding-Mode Control. The simulation model of the Sliding-Mode Direct Field-Oriented Control of six-phase induction motor has been implemented in Matlab-Simulink® Software. The simulation studies were carried out for six-phase squirrel-cage induction motor with the parameters: $P_N = 3$ kW, $U_{fN} = 230$ V, $f_N = 50$ Hz, $n_N = 2950$ rpm, $p_b = 2$, $R_s = 1.9\ \Omega$, $R_r = 2.1\ \Omega$, $L_{1s} = L_{1r} = 0.013$ H, $L_m = 0.6$ H. The nominal values were marked by the index N.

Figure 2 presents the waveforms of reference and calculated motor speed. It is assumed that, at first, the reference speed is rising to the half of the nominal speed, then for a certain period of time is constant, and then is rising again to the nominal speed. During the final trajectory, the reference speed is decreasing with the reversion of the nominal speed. During the period of operation with the nominal speed, the step change of the load torque has been forced. It can be stated that the calculated speed follows the reference speed with great accuracy.

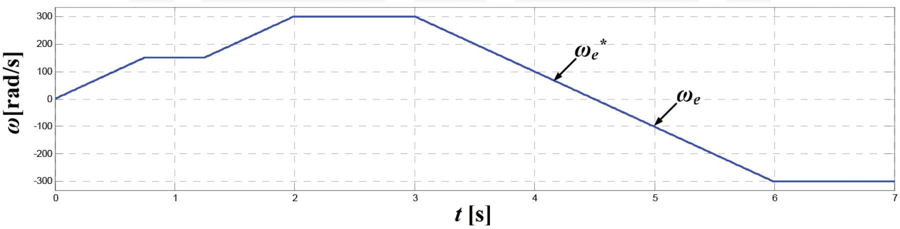

Fig. 2. The waveforms of speeds of six-phase induction motor for Sliding-Mode DFOC

Figure 3 presents the waveforms of the electromagnetic torque of the six-phase induction motor and the load torque. The obtained simulation results confirm the control of the electromagnetic torque with great accuracy and quick reaction of the Sliding-Mode DFOC for the changes of the load torque.

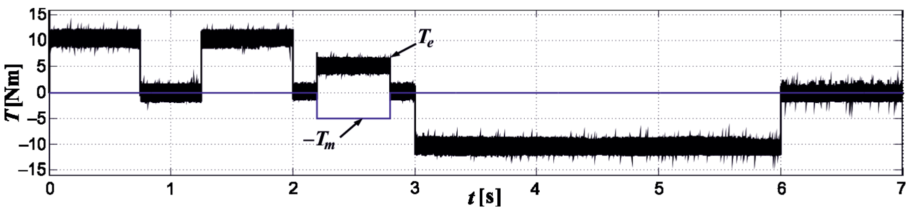

Fig. 3. The waveforms of the electromagnetic torque of six-phase induction motor and the load torque for Sliding-Mode DFOC

Figure 4 presents the stator phase current of the six-phase induction motor for the Sliding-Mode DFOC. The values of the stator current amplitude are great in dynamic states and are fixed at the steady-state operation. The simulation studies showed that for the considered control of the components of the stator current in the additional coordinate system $z1$-$z2$ have small amplitudes and they have no negative impact on the value of the phase currents.

Figure 5a presents the output signal i_{sy}^* from Sliding-Mode Controller of motor angular speed and Fig. 5b presents the output signal i_{sx}^* from Sliding-Mode Controller of magnitude of the rotor flux vector. In the presented examples of control signals, the continuous components and discontinuous control components are visible.

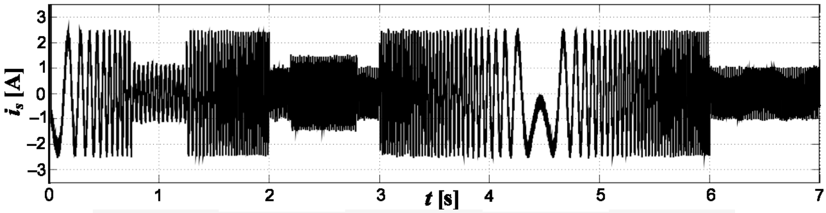

Fig. 4. The stator phase current of six-phase induction motor for Sliding-Mode DFOC

a) b)

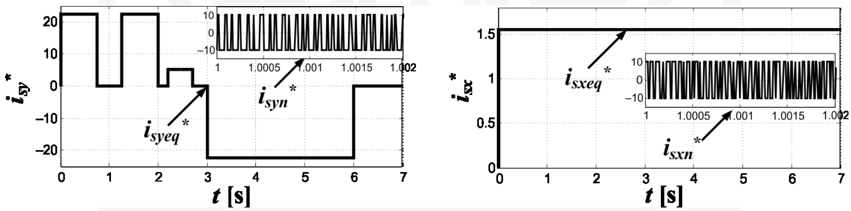

Fig. 5. Simulation results: a) the output signal i_{sy}^* from Sliding-Mode Controller of motor angular speed and b) the output signal i_{sx}^* from Sliding-Mode Controller of magnitude of the rotor flux vector

Figure 6 presents the trajectory of the rotor flux vector of the six-phase induction motor for Sliding-Mode DFOC. It can be stated that the magnitude of the rotor flux vector is regulated at the reference value without overshoot and have rated value at all operational times.

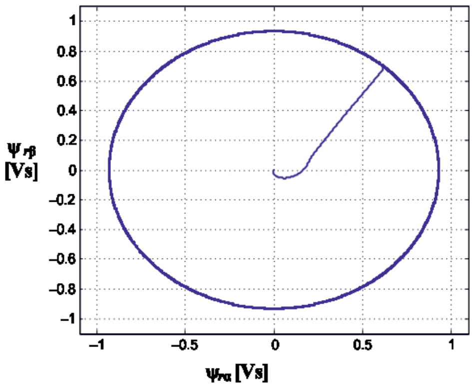

Fig. 6. The trajectory of the rotor flux vector of six-phase induction motor for Sliding-Mode DFOC

6. Conclusions

In this paper, the analysis and the design procedure of the novel structure of Sliding-Mode Direct Field-Oriented Control of six-phase induction motor have been presented.

The conducted simulation studies of the drive system with six-phase induction motor allow to draw a conclusion that the considered SMC-DFOC allows to obtain the control of the motor electromagnetic variables with great accuracy.

Based on the analysis of the simulation results, it can be stated that the real motor speed follows the reference motor speed with great accuracy. The values of motor electromagnetic torque and the amplitudes of the stator currents depend on the condition of the drive system. The magnitude of the rotor flux vector is regulated at the reference nominal value. During all tested states of the drive system, the overshoots and oscillations do not occur in the waveforms of the controlled electromagnetic variables.

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