International Journal of Power Electronics and Drive Systems (IJPEDS)
Vol. 15, No. 1, March 2024, pp. 53~63
ISSN: 2088-8694, DOI: 10.11591/ijpeds.v15.i1.pp53-63

DSP in the loop implementation of an improved sliding mode control for induction motor drive

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Article Info

Article history:

Received Apr 7, 2023
Revised Jul 9, 2023
Accepted Jul 29, 2023

Keywords:

Chattering reduction
Induction motor
Processor in the loop
Sliding mode control
Space vector modulation

ABSTRACT

This paper deals with the design and processor in the loop implementation of an improved sliding mode control of three-phase induction motor. Studied control strategy uses a state-space model expressed in the rotor flux frame. Proposed control strategy is designed in two steps based in cascaded method. This controller is based on new smooth functions replacing sign functions to reduce chattering phenomenon. This work integrates also a calculation of controller parameters to ensure set response times. It focuses on the study of these proposed techniques and their effect on motor waveforms. For code generation of the control algorithm, implementation and validation by processor in the loop technique we have created a platform based on MATLAB/Simulink with code composer studio as integrated development environment and the development board LAUNCHXL-F28069M from Texas instruments. Obtained results show that the proposed strategy reduces chattering using a simple algorithm.

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1. INTRODUCTION

Induction machine is a multi-variable nonlinear system, thus conventional control strategies using PI controllers are not simple to use; because they need supplementary operations such as decoupling, linearization, hysteresis controllers, commutation table [1]–[3]. So, classical control strategies for asynchronous motor require complex algorithms and do not ensure always desired behavior [4]. Otherwise, it is possible to proceed to direct design of control law based on advanced techniques [5]–[7].

Sliding mode control is a robust control strategy for dynamic and nonlinear systems. It is based on the famous stability theory of Lyapunov [8]–[10]. This control strategy is especially used for nonlinear systems like induction motors and ensure disturbance rejection when controller parameters are sufficiently high. Proposed control strategy is designed in two steps based in cascaded method separating slow variables which are speed and flux and fast variables which are direct current and quadratic current.

This work proposes a simplified state-space induction machine equation expressed in the rotor flux reference frame. This is suitable for a methodical and easy design of the sliding mode control technique. The control law integrating equivalent control and discontinuous control is then determined and smooth functions are added to sliding mode control scheme for more chattering reduction. The control law requires also the rotor flux, so this paper proposes a rotor flux estimator design [1], [2].

Proposed control strategy, is tested first by simulation on MATLAB/Simulink environment and then validated by PIL technique [11]. Processor in the loop, also called DSP in the loop, is a validation technique used to test control algorithms executed in an actual DSP while the controlled system is implemented in a simulation environment like Simulink. Communication between both sides uses serial link. In this work codes related to the control algorithms are generated using MATLAB/Simulink with code composer studio.

2. INDUCTION MOTOR MODELLING

In two-phase reference frame, electromagnetic state-space representation of the three-phase induction machine is a four variables model. In this paper we choose stator currents i_{sd}, i_{sq} ; and two virtual currents $i_{\varphi d}, i_{\varphi q}$ that replace rotor flux components $\varphi_{rd}, \varphi_{rq}$ [1], [2]. Electromagnetic state-space model is expressed in (1), juxtaposed with the mechanical (2). Where Input variables are stator voltage components v_{sd}, v_{sq}.

$$\frac{d}{dt}\begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{\varphi d} \\ i_{\varphi q} \end{bmatrix} = \begin{bmatrix} -\frac{1}{T_c} & \frac{d\rho s}{dt} & \frac{1-\sigma}{\sigma Tr} & \frac{1-\sigma}{\sigma} p\Omega \\ -\frac{d\rho s}{dt} & -\frac{1}{T_c} & -\frac{1-\sigma}{\sigma} p\Omega & \frac{1-\sigma}{\sigma Tr} \\ \frac{1}{Tr} & 0 & -\frac{1}{Tr} & \frac{d\rho r}{dt} \\ 0 & \frac{1}{Tr} & -\frac{d\rho r}{dt} & -\frac{1}{Tr} \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{\varphi d} \\ i_{\varphi q} \end{bmatrix} + \frac{1}{\sigma Ls} \begin{bmatrix} 1 & 0 \\ 0 & 1 \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} v_{sd} \\ v_{sq} \end{bmatrix} \tag{1}$$

$$\frac{d\Omega}{dt} = -\frac{f}{J}\Omega + \frac{T_M}{J} - \frac{T_L}{J} \tag{2}$$

ρs and ρr are frame angles relative respectively to the stator and rotor. T_M and T_L are electromagnetic torque and load torque. J, f and Ω are respectively inertia, coefficient of friction and rotor velocity, with:

$$T_M = \frac{3}{2} p(1-\sigma) Ls\left(i_{\varphi d} i_{sq} - i_{\varphi q} i_{sd}\right) \tag{3}$$

$$i_{\varphi d} = \frac{\varphi_{rd}}{Msr}; i_{\varphi q} = \frac{\varphi_{rq}}{Msr} \tag{4}$$

$$\frac{1}{T_c} = \left(\frac{1}{\sigma Ts} + \frac{1-\sigma}{\sigma Tr}\right); \sigma = 1 - \frac{Msr^2}{Ls.Lr} \tag{5}$$

Ts and Tr are stator and rotor time constants; Ls, Lr and Msr are stator, rotor, and mutual inductances. Finally, σ is the dispersion coefficient and the number of pole pairs is p. The same frame is used for all variables, so:

$$\frac{d\rho s}{dt} = p\Omega + \frac{d\rho r}{dt} \tag{6}$$

2.1. Induction machine model in rotor flux frame

Sliding mode strategy can be designed for induction machine in fixed frame, however state-space equations and resulting control law are long and complex in this case. This paper proposes to design sliding mode control in the rotor flux reference frame where rotor flux is defined by only one state variable rather than two variables [12], [13]. In this new reference frame, electromagnetic state-space is represented in (7), juxtaposed with the mechanical (8),

$$\frac{d}{dt}\begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{\varphi d} \end{bmatrix} = \begin{bmatrix} -\frac{1}{T_c} & \frac{d\rho s}{dt} & \frac{1-\sigma}{\sigma Tr} \\ -\frac{d\rho s}{dt} & -\frac{1}{T_c} & -\frac{1-\sigma}{\sigma} p\Omega \\ \frac{1}{Tr} & 0 & -\frac{1}{Tr} \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{\varphi d} \end{bmatrix} + \frac{1}{\sigma Ls} \begin{bmatrix} 1 & 0 \\ 0 & 1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} v_{sd} \\ v_{sq} \end{bmatrix} \tag{7}$$

$$\frac{d\Omega}{dt} = -\frac{f}{J}\Omega + \frac{T_M}{J} - \frac{T_L}{J} \tag{8}$$

with:

$$T_M = \frac{3}{2} p(1-\sigma) Ls\, i_{\varphi d} i_{sq} \tag{9}$$

The resulting induction machine model contains only four variables rather than five, so it is simpler. In addition, the electromagnetic torque has a compact expression containing only one term.

2.2. Variables notation

Three-phase induction machine is now a four state-space variables system. Main controlled variables are rotor flux and motor speed while stator currents are secondary variables [14].

$$\frac{d}{dt}\Omega = -\frac{f}{J}\Omega - \frac{T_L}{J} + \frac{3p}{2J}(1-\sigma)Ls i_{\varphi d} i_{sq} \quad (10)$$

$$\frac{d}{dt}i_{\varphi d} = -\frac{1}{Tr}i_{\varphi d} + \frac{1}{Tr}i_{sd} \quad (11)$$

$$\frac{d}{dt}i_{sq} = -\frac{1}{Tc}i_{sq} - \omega_s i_{sd} - \frac{1-\sigma}{\sigma}p\Omega i_{\varphi d} + \frac{1}{\sigma Ls}v_{sq} \quad (12)$$

$$\frac{d}{dt}i_{sq} = -\frac{1}{Tc}i_{sq} - \omega_s i_{sd} - \frac{1-\sigma}{\sigma}p\Omega i_{\varphi d} + \frac{1}{\sigma Ls}v_{sq} \quad (13)$$

We note that frame velocity ω_s depends on the other variables, it is not a separate one.

$$\omega_s = p\Omega + \frac{1}{Tr}\frac{i_{sq}}{i_{\varphi d}} \quad (14)$$

We propose notations change of (15) for variables.

$$\begin{bmatrix} x_1 \\ x_3 \end{bmatrix} = \begin{bmatrix} \Omega \\ i_{\varphi d} \end{bmatrix}; \begin{bmatrix} x_2 \\ x_4 \end{bmatrix} = \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} \quad (15)$$

Thus, induction machine model is given (16), (17), (18), and (19).

$$\dot{x}_1 = -\frac{f}{J}x_1 - \frac{T_L}{J} + \frac{3p}{2J}(1-\sigma)Ls x_3 x_2 \quad (16)$$

$$\dot{x}_2 = -\frac{1}{Tc}x_2 + \omega_s x_4 + \frac{1-\sigma}{\sigma Tr}x_3 + \frac{1}{\sigma Ls}v_{sd} \quad (17)$$

$$\dot{x}_3 = -\frac{1}{Tr}x_3 + \frac{1}{Tr}x_2 \quad (18)$$

$$\dot{x}_4 = -\frac{1}{Tc}x_4 - \omega_s x_2 - \frac{1-\sigma}{\sigma}p x_1 x_3 + \frac{1}{\sigma Ls}v_{sq} \quad (19)$$

2.3. Vector model of the induction machine

To simplify controllers design by sliding mode strategy, this work proposes the use of two-dimensional vector for input and state-space variables [15]. So, vector state-space representation is expressed in (20) and (21). Thus, two variables are controlled in one step.

$$\begin{bmatrix} \dot{x}_1 \\ \dot{x}_3 \end{bmatrix} = -\begin{bmatrix} \frac{f}{J}x_1 + \frac{T_L}{J} \\ \frac{1}{Tr}x_3 \end{bmatrix} + \begin{bmatrix} 0 & \frac{3p}{2J}(1-\sigma)Ls x_3 \\ \frac{1}{Tr} & 0 \end{bmatrix}\begin{bmatrix} x_2 \\ x_4 \end{bmatrix} \quad (20)$$

$$\begin{bmatrix} \dot{x}_2 \\ \dot{x}_4 \end{bmatrix} = \begin{bmatrix} -\frac{1}{T_c}x_2 + \omega_s x_4 + \frac{1-\sigma}{\sigma Tr}x_3 \\ -\omega_s x_2 - \frac{1}{T_c}x_4 - \frac{1-\sigma}{\sigma}p x_1 x_3 \end{bmatrix} + \frac{1}{\sigma Ls}\begin{bmatrix} v_{sd} \\ v_{sq} \end{bmatrix} \quad (21)$$

3. SLIDING MODE CONTROL DESIGN OF THE ASYNCHRONOUS MOTOR

This paper proposes a cascaded sliding mode control designed in two steps using smooth functions, and a rotor flux estimator based on stator currents as input variables [16], [17]. Figure 1 gives the control scheme representing a main block for speed and flux control by generating secondary variables reference signals, then an auxiliary block for currents control generating the sliding mode control law signals. A flux estimator and a SVM modulator [18], [19]. In this scheme, slow loop controller ensures the control law given in (33), fast loop controller calculates the control law related to (40), and the estimator is based on (42) and (43).

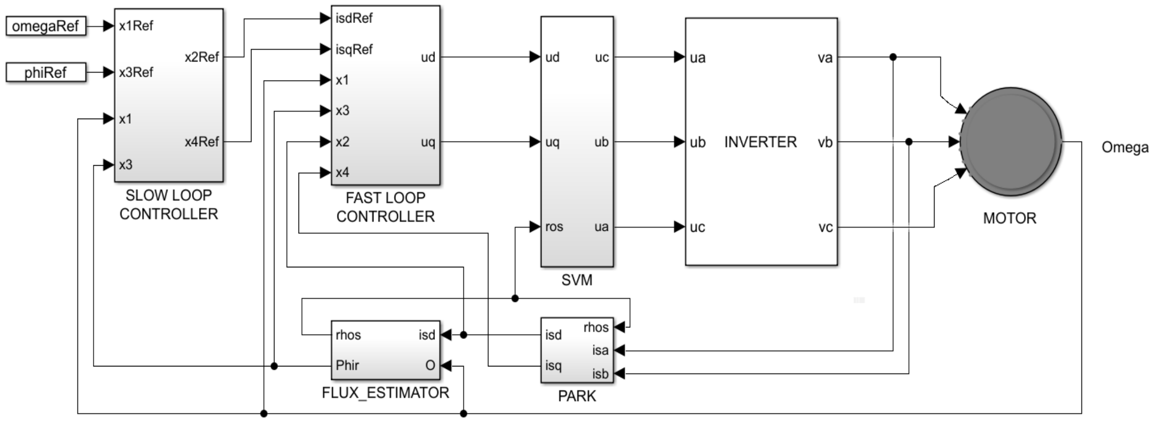

Figure 1. Synoptic diagram of proposed sliding mode control for induction motor

Finally, three-phase asynchronous machine model is represented in two compact (22) and (23) using matrix parameters and two-dimensional vectors for input and state-space variables [14]:

$$\dot{X}_1 = F_1(X_1) + G_1(X_1)X_2 \tag{22}$$

$$\dot{X}_2 = F_2(X_1, X_2) + G_2U \tag{23}$$

with:

$$X_1 = \begin{bmatrix} x_1 \\ x_3 \end{bmatrix};\ X_2 = \begin{bmatrix} x_2 \\ x_4 \end{bmatrix}; U = \begin{bmatrix} v_{sd} \\ v_{sq} \end{bmatrix} \tag{24}$$

and:

$$F_1(X_1) = -\begin{bmatrix} \frac{1}{Tm}x_1 + \frac{T_L}{J} \\ \frac{1}{Tr}x_3 \end{bmatrix} \tag{25}$$

$$F_2(X_1, X_2) = \begin{bmatrix} -\frac{1}{T_c}x_2 + \omega_s x_4 + \frac{1-\sigma}{\sigma Tr}x_3 \\ -\omega_s x_2 - \frac{1}{T_c}x_4 - \frac{1-\sigma}{\sigma}px_1x_3 \end{bmatrix} \tag{26}$$

$$G_1(X_1) = \begin{bmatrix} 0 & \frac{3p}{2J}(1-\sigma)Lsx_3 \\ \frac{1}{Tr} & 0 \end{bmatrix}; G_2 = \frac{1}{\sigma Ls} \tag{27}$$

It is a second-order vector system which can be easily controlled by double sliding mode controller, using cascaded control design method. In the proposed work, control design requires just two steps where control law integrates equivalent control and discontinuous control [20]–[22]. In order to simplify the control law, this work proposes to separate slow dynamic related to motor speed and rotor flux from fast dynamic related to stator currents; this offers also more robustness to the control behavior.

3.1. Slow dynamic stage controller design

In this stage, it is a question of controlling the main variable X_1 representing rotor flux and speed, through the auxiliary variable X_2. the objective is to express the secondary control law X_{2ref} taking:

$$X_2 = X_{2ref} \tag{28}$$

First, we define rotor speed and flux vector error (29) also considered as a sliding variable [23]; then we introduce the discontinuous function in the derivative of this sliding manifold (30).

$$\Sigma_1 = E_1 = X_1 - X_{1ref} \tag{29}$$

$$\dot{\Sigma}_1 = \dot{E}_1 = \dot{X}_1 - \dot{X}_{1ref} = F_1 + G_1 X_2 - \dot{X}_{1ref} = -K_1 sign(\Sigma_1) \tag{30}$$

$$K_1 = \begin{bmatrix} k_1 & 0 \\ 0 & k_3 \end{bmatrix}; sign(\Sigma_1) = sign(E_1) = \begin{bmatrix} sign(e_1) \\ sign(e_3) \end{bmatrix} \tag{31}$$

Candidate function of (32), called Lyapunov function, is a positive function. In addition, positive parameters in the diagonal matrix K_1 ensure that derivative of Lyapunov function is negative; this guarantee the stability in this stage [24].

$$V_1 = \frac{1}{2} E_1^T E_1 > 0 \rightarrow \dot{V}_1 = -E_1^T K_1 sign(E_1) < 0 \tag{32}$$

Thus, the intermediate control law is (33).

$$X_{2ref} = G_1^{-1}\left(-K_1 sign(\Sigma_1) - F_1 + \dot{X}_{1ref}\right) = G_1^{-1}(-K_1 sign(\Sigma_1) - F_1) \tag{33}$$

Calculation of the controller parameters can be done from the speed response time that we have set in this work to the value $trv = 200\ ms$ and flux response time that we have set to $trf = 20\ ms$. Indeed, since (30), the response times can be expressed by (34).

$$\begin{bmatrix} trv \\ trf \end{bmatrix} = \begin{bmatrix} \frac{|e_1(0)|}{k_1} \\ \frac{|e_3(0)|}{k_3} \end{bmatrix} \tag{34}$$

Thus, for an initial speed error of $|e_1(0)| = 100\ rads^{-1}$, and initial flux error $|e_3(0)| = 6\ A$ control parameters value is given by (35).

$$K_1 = \begin{bmatrix} 500 & 0 \\ 0 & 300 \end{bmatrix} \tag{35}$$

3.2. Fast dynamic stage controller design

In this step, we define the currents vector error (36) and we take it as a sliding variable, then we calculate its derivative by introducing the discontinuous term (37). Thus, we can determine the main control law (38). This control law generates driving signals for the SVM modulator.

$$\Sigma_2 = E_2 = X_2 - X_{2ref} = \begin{bmatrix} i_d \\ i_q \end{bmatrix} - \begin{bmatrix} i_{dref} \\ i_{qref} \end{bmatrix} \tag{36}$$

$$\dot{\Sigma}_2 = \dot{E}_2 = \dot{X}_2 - \dot{X}_{2ref} = F_2 + G_2 U - \dot{X}_{2ref} = -K_2 sign(\Sigma_2) \tag{37}$$

$$K_2 = \begin{bmatrix} k_2 & 0 \\ 0 & k_4 \end{bmatrix}; sign(E_2) = \begin{bmatrix} sign(i_d - i_{dref}) \\ sign(i_q - i_{qref}) \end{bmatrix} \tag{38}$$

Lyapunov function is a positive function and the diagonal matrix K_2 parameters are positives which ensures that candidate function derivative is negative (39); so, stability is guaranteed.

$$V_2 = \frac{1}{2} E_2^T E_2 > 0 \rightarrow \dot{V}_2 = -E_2^T K_2 sign(E_2) < 0 \tag{39}$$

In (40) gives the main control law that generates inverter control signals.

$$U = G_2^{-1}(-K_2 sign(\Sigma_2) - F_2 + \dot{X}_{2ref}) \approx G_2^{-1}(-K_2 sign(\Sigma_2) - F_2) \tag{40}$$

We calculate the parameters of this control law by setting a response time much shorter than that of the speed and flux control $trc = 2\ ms$. Direct and quadratic currents are limited to a value of 25 A. Thus:

$$trc = \frac{|e_2(0)|}{k_2} = \frac{|e_4(0)|}{k_4} \rightarrow k_2 = k_4 = \frac{25}{0.002} = 12500 \tag{41}$$

Finally, the control law is given by (33) and (40). Since rotor flux and speed references are constant, their derivative $\dot{X}_{1ref}$ is zero. In addition, the second step dynamic is very fast in relation to the second step dynamic thus the derivative $\dot{X}_{2ref}$ is neglected in the main control law.

3.3. Rotor flux estimation

Motor speed, stator currents and voltages are variables that can be measured easily. In this paper, rotor flux is assumed to be inaccessible, so it is estimated because it is necessary for control law elaboration. This work proposes the design of a simple rotor flux estimator generating flux magnitude and angle. In fact, only the direct component of the rotor flux has to be estimated because the quadrature one is zero in the rotor flux frame [25].

In (42) is obtained, in the rotor flux frame, from the three-phase asynchronous machine model. It is used to estimate the rotor flux magnitude.

$$\frac{d}{dt}i_{\varphi d} = -\frac{1}{Tr}i_{\varphi d} + \frac{1}{Tr}i_{sd} \quad (42)$$

Finally, rotor flux angle which is the reference frame position is estimated from (43) using estimated flux magnitude and by calculating the integral since the equation gives velocity.

$$\frac{d\rho s}{dt} = p\Omega + \frac{1}{Tr}\frac{i_{sq}}{i_{\varphi d}} \quad (43)$$

3.4. Sign function approximation

The sign function $sign(x)$ in sliding mode control causes chattering and degradation of the quality of control inputs absorbed currents and controlled variables [26], [27]. The replacement of the sign function by a smooth function like inverse tangent or hyperbolic tangent functions allows to overcome this problem [25], [28], [9]. We propose $\frac{2}{\pi}atan(\frac{x}{\delta})$ function that can be adjusted by the coefficient δ.

4. PROCESSOR IN THE LOOP VALIDATION

First of all, proposed control strategy is validated on Simulink. Algorithm code related to the studied control strategy is obtained using code generation technique from MATLAB/Simulink environment and then uploaded in the DSP that communicates with Simulink blocks. Experimentation is performed by processor in the loop: the SVM inverter and the induction machine are modeled on Simulink and controlled by the algorithm code running on the Texas Instruments $DSP\ TMS320F28069M$. Serial link between DSP and PC is ensured via USB port, this experimental platform is illustrated in Figure 2. Sampling frequency and SVM carrier frequency have the same value:

$$f_{PWM} = f_{SMP} = 10\ kHz \rightarrow T_{SMP} = 100\ \mu s$$

Controlled variables references are set to the values: rated rotor flux $0.35\ Wb$ and rotor speed $100\ rad/s$. They are applied at the starting time whereas rated load torque $20\ Nm$ is applied at time $t = 0.5\ s$. Maximum fundamental voltage that can be generated, by the SVM inverter from $539\ V$ DC bus, is $220\ V\ RMS$. The carrier amplitude is set to $311\ V$ so the SVM inverter average model is just unitary. Also, rated values and specification of the induction motor are given in Table 1.

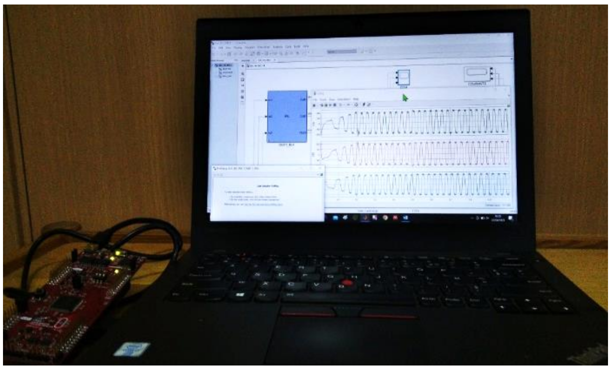

Figure 2. Validation platform using PIL technique for sliding mode control of induction motor

Table 1. System numerical specifications

Name	Value	Name	Value
Rated motor output power	: $3\ kW$	Moment of rotor inertia	: $J = 0.05\ kg.m^2$
Rated motor load torque	: $20\ Nm$	Coefficient of viscose friction	: $f = 0.005\ Nm/rad.s^{-1}$
Rated rotor speed	: $1455\ rpm; p = 2$	Stator and rotor time constants	: $Ts = 0.188\ s; Tr = 0.144\ s$
Rated supply voltage and frequency	: $220\vert 380V; 50\ Hz$	Current time constant	: $\tau = 7.36\ ms$
Stator and rotor resistances	: $Rs = 0.85\ \Omega; Rr = 0.16\ \Omega$	Machine dispersion coefficient	: $\sigma = 0.08587$
Stator and rotor cyclic inductances	: $Ls = 0.16\ H; Lr = 0.023\ H$	DC bus voltage	: $E = 220\sqrt{2}\sqrt{3} = 539\ V$
Cyclic mutual Inductance	: $Msr = 0.058\ H$		

Obtained results are given in Figures 3, 4, and 5. To prove the effect of smooth functions on chattering reduction we propose to test the same control strategy by PIL technique with sign function in the first step. Results of this test are given in Figures 6, 7, and 8. Responses represented in Figure 3 are rotor flux, electromagnetic torque, and motor speed. References for both rotor flux magnitude and motor speed, as controlled variables, are reached ensuring almost zero static error even when load torque is applied. Moreover, obtained dynamics are in accordance with response time requirements.

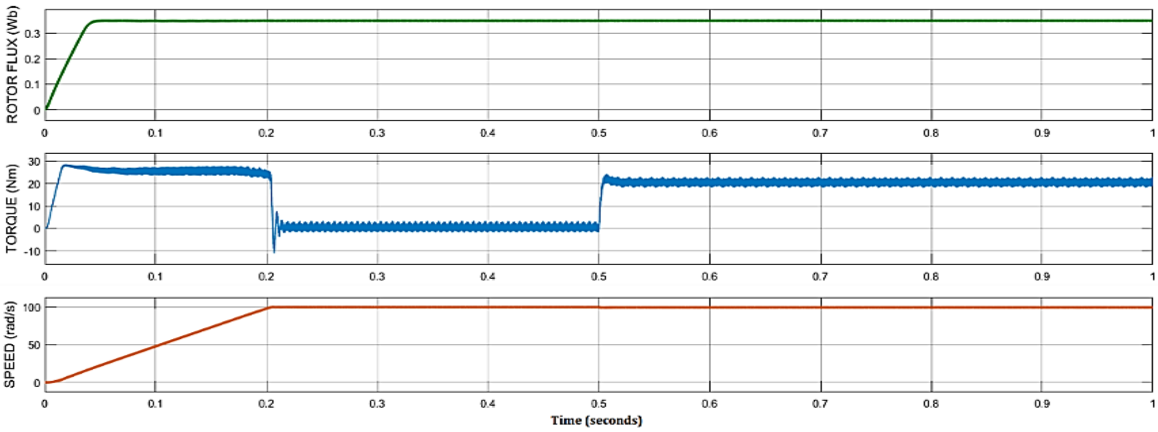

Figure 3. PIL responses for rotor flux magnitude, electromagnetic torque and speed

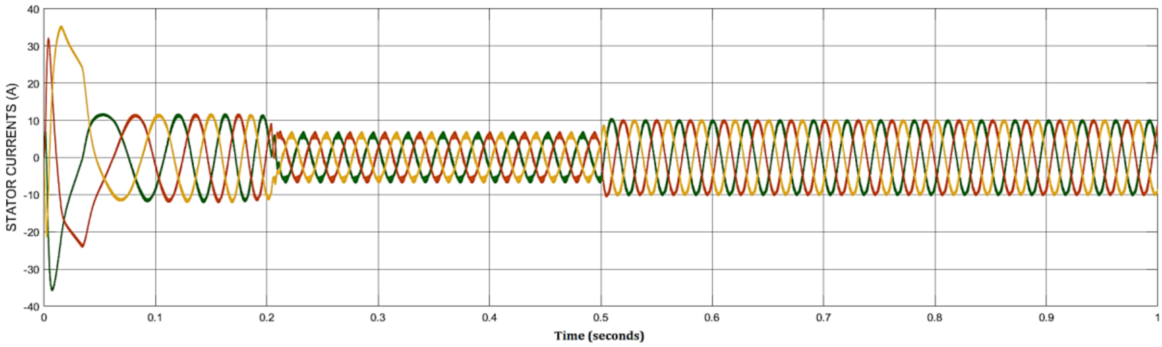

Figure 4. Stator currents responses obtained by PIL technique

Rejection of load torque disturbance is ensured as long as $\eta - reachability$ condition, that leads to condition (32), is guaranteed. Thus, for disturbance rejection, this condition requires sufficiently high controller parameters that can compensates load torque effect on the control law. PIL technique responses given in Figure 4 represents stator currents during steady-state on no load and under as well as their transient behavior. At start-up, stator currents show an increase which is limited to an acceptable value. This is the benefit of choosing a required response time for induction motor in accordance with its natural start. Obviously, an increase of stator currents is observed as soon as the load torque is introduced.

Figure 5 represents the inverter driving signals generated by the SVM block. This reference signals ensure that the induction motor is correctly powered by the three-phase voltage-source inverter. Obviously, obtained waveforms during the proposed experimentation are typical for SVM generator and we can note that the chattring effect is invisible. In Figure 6, rotor flux and torque have more ripple with sign function, this is caused by chattering phenomenon which is reduced by smooth function in last results. Visually, in Figure 7 stator currents have an excessive distortion compared to currents response in the first test.

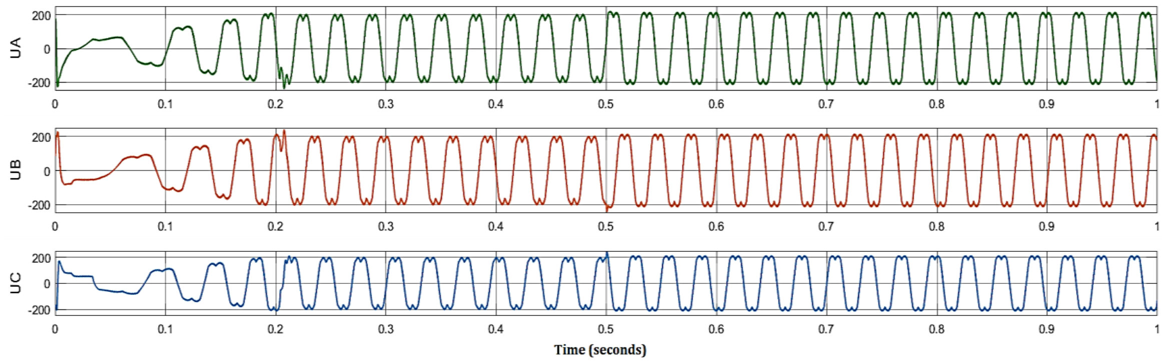

Figure 5. PIL responses for SVM signals driving the three-phase inverter

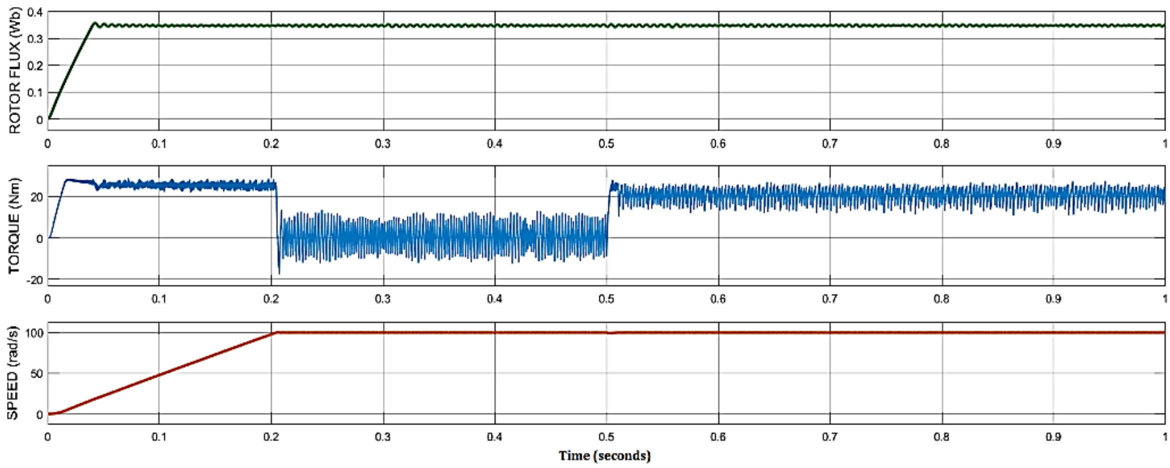

Figure 6. Rotor flux, torque and speed with sign function

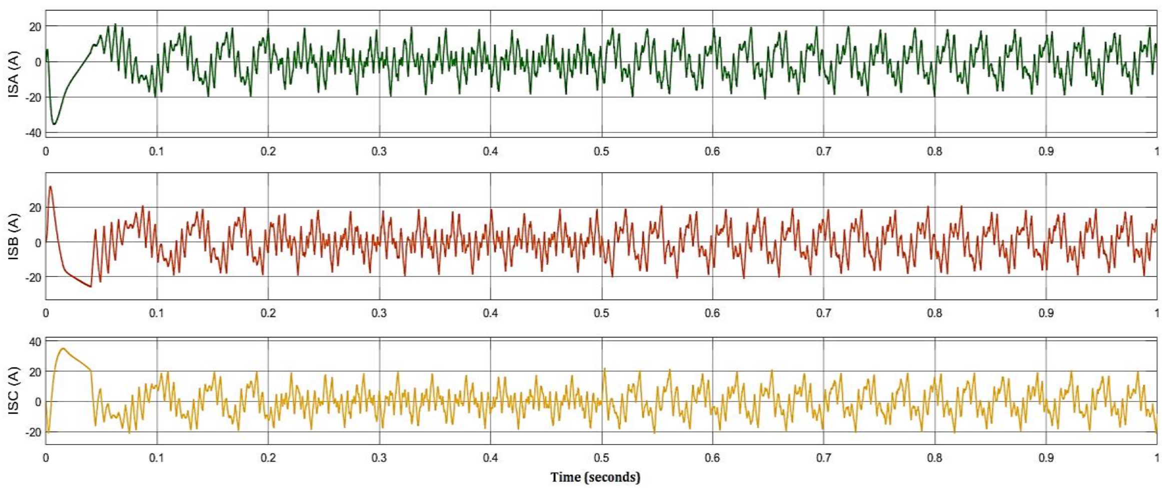

Figure 7. Stator currents with sign function

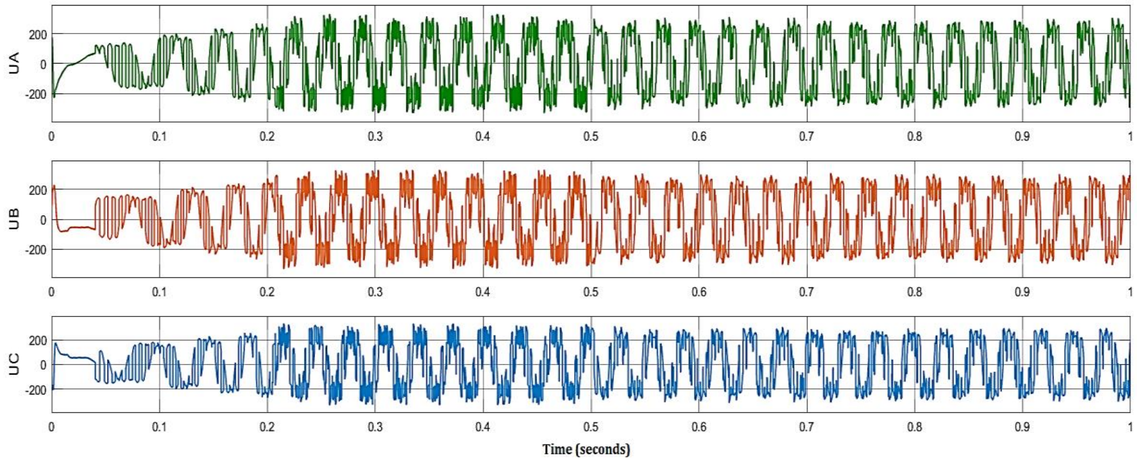

Figure 8. SVM input signals with sign function

5. CONCLUSION

In this paper, a reference frame aligned with rotor flux is used to simplify equations for the three-phase induction machine model. For more simplicity during controller design, obtained equations are represented using two-dimensional vector form. Thus, rotor flux and machine speed are processed in one step as well as direct and quadratic components of stator currents. After that, a novel sliding mode controller is proposed. This controller is designed in two steps using cascaded method to separate slow variables from fast variables.

In sliding mode control, sign functions bring chattering effect that affect almost all waveforms. To solve this problem, smooth functions are introduced. Also, since rotor flux is needed to perform this controller design an estimator is added for flux calculation. This paper is especially interested in parameters calculation for the proposed sliding mode controller. In fact, a calculating method is proposed depending on the required dynamic behavior in order to obtain precisely the mentioned response time.

This work proposes especially a method for calculating of the sliding mode controller parameters in order to satisfy precisely the dynamic-response requirements of the drive, namely response times. Before code generation, control algorithm is tested on Simulink using discrete mode for the sliding mode controller and continuous mode for the inverter and machine. Generated code is then transferred to an actual DSP in order to perform PIL validation using the development board LAUNCHXL-F28069M connected to the USB port of the PC that integrates MATLAB/Simulink and CCS. During PIL experimentation, the SVM inverter and the three-phase machine are implemented on Simulink and controlled by the algorithm running on the development board via serial link.

Experimental responses demonstrate that the expected behavior of the proposed sliding mode controller is obtained in term of required response time and precision. Moreover, the speed and flux responses have no overshot, the chattering effect is correctly reduced and the waveforms of stator currents and SVM signals responses are acceptable.

Also, we notice that this control strategy is capable to reject disturbance which is the load torque so no load torque observer is needed. Proposed sliding mode controller for induction machine is easily designed since sliding mode strategy was developed for linear and nonlinear systems. Moreover, proposed design uses cascaded method with separate dynamic and then gives simple control law and ensure expected responses.

Counter to several works, proposed paper is distinguished by a simple controller design and rigorous parameter calculation that guarantee the required behavior. In addition, proposed design uses smooth function, dynamic separation and it adopts a compact state-space equation based on vectors.

Proposed sliding mode controller design is easy to apply to induction machine and ensure robustness against load torque disturbance. Code generation allows easy implementation and PIL validation of the control algorithm. In future papers we intend to achieve a full experimental test-bench.

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