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A SENSORLESS ADAPTIVE BACKSTEPPING SLIDING MODE CONTROL OF DOUBLE-STAR INDUCTION MOTOR

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Keywords: Double-star induction motor; Adaptive backstepping sliding mode control; Parameter identification; Lyapunov stability.

A robust adaptive backstepping sliding-mode controller ABSMC is proposed for speed regulation of a double-star induction machine DSIM. This controller handles unknown or rapidly changing parameters while ensuring Lyapunov-based stability. First, a motor drive system model with lumped uncertainty is developed. Then, a nonlinear robust speed controller using the ABSMC theory is presented. In this technique, we employ it to guarantee speed tracking and parameter perturbation suppression. The simulation results indicate that the proposed control scheme effectively compensates for parameter perturbations while maintaining speed-tracking precision.

1. INTRODUCTION

At present, modern variable speed drive technology is based on the increasing use of the dual stator asynchronous machine due to its robustness, electromechanical reliability, and low cost [1]. Indeed, it is important to control this type of machine, considering parameter variations, as these can compromise control.

Several intelligent control techniques have been developed for the control of the double-stator induction motor DSIM. These techniques aim to achieve performance comparable to that of the direct current (DC) motor. The proportional-integral (PI) controller is one of the most widely used of these approaches. However, under complex operating conditions, the PI controller shows limited performance. As a result, it cannot meet the requirements of high-performance applications.

With the development of control theory and microcomputer technology, nonlinear control strategies have been applied, including adaptive control, backstepping control, sliding mode control, and fuzzy logic control [2–6].

Backstepping control is a systematic and recursive approach to nonlinear systems based on Lyapunov stability theory [7]. However, Backstepping control is a strict model-based control method. Thus, a single backstepping controller cannot ensure the under-complex working condition [2].

As is known, sliding mode control is a nonlinear control technique derived from variable structure control system theory, originally developed by Vladimir Utkin [8]. It is a popular approach among control engineers due to its several advantages. Particularly in motor drive applications, such as simple implementation, robustness, and good dynamic response. Moreover, the sliding mode dynamics can be easily stabilized through an appropriate choice of the sliding surface. However, its main drawback is the chattering phenomenon.

To reduce this phenomenon, the high-order sliding mode concept has been proposed as a preliminary solution. However, it didn't last long because the actual disadvantage is the stability proofs. Those are based on geometrical methods since the Lyapunov function proposes results in a difficult task. Now, the super twisting sliding mode controller provides one of the most attractive solutions in many industry domains [9].

Adaptive technology comes as a solution to systems where the variables (or several) vary over time, which affect the control process. It is developed for a wide variety of academic, military, and industrial applications. Therefore, the word adaptive, in the sense of process control, means that it changes its output in response to a change in the error. However, during development, based on known parameters, changes over time due to control problems. The engineers create the "self-tuning" or "auto-tuning" techniques. These have been developed since then to automate the tuning procedure [3,4,7]. Therefore, we note that there are controls called hybrid appearing as a competition of adaptive control. But they don't give any attractive results in the absence of the link, like fuzzy sliding mode control [10].

Recently, in industrial fields, engineers are using robust methods to update certain system parameters as an aid to diagnosing (or troubleshooting) failures. These estimators also reduce the reliance on physical sensors by replacing them in the event of failure or by minimizing their number. For example, an automotive scanner system and control algorithms for modern control in industrial production.

All sensors used to replace the information given by mechanical sensors are sometimes called software sensors. In the same context, several observation approaches for the asynchronous double-star machine without a mechanical sensor are developed. In the literature, we can find those based on artificial intelligence, like neural networks [11,12]. Another approach from the automatic side is based on a machine behavior model. Several observer-based approaches have been developed, including extended Kalman filters [13,14] and nonlinear observers such as sliding mode observers [15], as well as the extended Luenberger observer [16,17]. However, these approaches make it difficult to guarantee local convergence and stability. In addition, no proof of global convergence of the closed-loop system combining control and observer has been provided. Even though stability is often claimed. Moreover, the implementation of these methods leads to an increase in computational cost.

Historically, all control systems estimate certain parameters by adding a separate estimator, which increases costs. The problem is to implement only a single algorithm guaranteeing the same objective.

In this paper, to regulate the speed, flux, and currents of the DSIM and to estimate its parameters (R_r,R_s, L_s) using only an algorithm. The paper is organized as follows: Section two provides a reminder of the field-oriented control of the double-stator induction motor DSIM with relevant abbreviations. Section three reviews the ABSMC control problem and stability algorithm, using the closed-loop output error to re-estimate the DSIM parameters. Section four

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presents the numerical simulation results. Finally, a conclusion ends the paper.

2. PROBLEM FORMULATION AND MODELS

2.1 REMINDER OF INDIRECT FIELD ORIENTED CONTROL OF DSIM

Vector control is based on the decoupling of flux and torque. The principle of vector control, also called flux-oriented control, is obtained by regulating torque through one current component and flux through another component [3–6]. Vector control provides high industrial performance for asynchronous drives. In fact, if the rotor flux is aligned with the d-axis of the reference frame linked to the rotating field, rotor flux orientation is achieved as follows:

$$\Phi_{dr} = \Phi_r = 1, \phi_{qr} = 0. \tag{1}$$

The main objective of vector control is then to generate reference voltages for the static voltage converters supplying the DSIM. The notation X_{ref} denotes reference quantities, such as torque, flux, voltages, and currents. Applying rotor flux orientation to the machine equations leads to the expressions presented in [3–6]; ω_{sr_ref} denotes the slip angular frequency, given by:

$$\begin{cases} T_e = p\frac{L_m}{L_m+L_r}\left(I_{qs1\_ref} + I_{qs1\_ref}\right)\Phi_{r_{ref}}, \\ \omega_{sr\_ref} = \frac{R_r L_m}{(L_m+L_r)\Phi_{r\_ref}}\left(I_{qs1\_ref} + I_{qs2\_ref}\right), \\ \Phi_{\mathrm{s\_ref}} = \omega_{sr\_ref} + \omega_r\ , \\ \Phi_{r\_ref} = L_m\left(I_{ds1\_ref} + I_{ds2\_ref}\right), \end{cases} \tag{2}$$

$$\begin{cases} \frac{d}{dt}I_{ds1} = \frac{1}{L_s}\left[V_{ds1\_ref} - R_s I_{ds1} + \eta_1\right], \\ \frac{d}{dt}I_{ds2} = \frac{1}{L_s}\left[V_{ds1\_ref} - R_s I_{ds2} + \eta_2\right], \\ \frac{d}{dt}I_{qs1} = \frac{1}{L_s}\left[V_{qs1\_ref} - R_s I_{qs1} + \eta_3\right], \\ \frac{d}{dt}I_{qs2} = \frac{1}{L_s}\left[V_{qs2\_ref} - R_s I_{qs2} + \eta_4\right], \end{cases} \tag{3}$$

with:

$$\begin{cases} \eta_1 = +\omega_{s\_ref}\left(L_s I_{qs1} + T_r\Phi_{r_{ref}}\omega_{sr_{ref}}\right), \\ \eta_2 = -\omega_{s\_ref}\left(L_s I_{qs2} + T_r\Phi_{r_{ref}}\omega_{sr_{ref}}\right), \\ \eta_3 = +\omega_{s\_ref}\left(L_s I_{ds1} + \Phi_{r\_ref}\right), \\ \eta_4 = +\omega_{s\_ref}\left(L_s I_{ds2} + \Phi_{r\_ref}\right). \end{cases}$$

Equations (3) constituted the reference equation voltage system. Consequently, the speed and flux values equations for the system are estimated in [16] by: (DSIM parameters are given in Table 1).

$$\begin{cases} \frac{d\hat{\omega}}{dt} = \frac{1}{j}\left[p\frac{L_m}{L_m+L_r}\left(I_{qs1} + I_{qs2}\right)\Phi_{r_{ref}} - T_l - \beta_\omega\hat{\omega}\right], \\ \frac{d\hat{\Phi}_r}{dt} = -\frac{R_r}{L_m+L_r}\hat{\Phi}_r + \frac{R_r L_m}{L_m+L_r}\left(I_{ds1} + I_{ds2}\right). \end{cases} \tag{4}$$

2.2 PROBLEM EXPLANATION

Due to the various parameter perturbations in the complicated environments of the DSIM motor, one redefines the state equation in the following matrix form to realize online real-time estimation of system uncertainty by:

$$\frac{dx}{dt} = f_N(x) + \Delta f(x) + g(x)u, \tag{5}$$

where: $x = \left[I_{ds1}\ I_{ds2}\ I_{qs2}\ \hat{\omega}\ \hat{\Phi}_r\right]$; $u = \left[V_{ds1}\ V_{ds2}\ V_{qs1}\ V_{qs2}\right]$; $d = \frac{T_l}{j}$; $g(x) = a_{1N}$,

$$f_N(x) = \begin{bmatrix} -a_{2N}I_{ds1} + \omega_{s_{ref}}I_{qs1} + \omega_{s\_ref}\Phi_{r\_ref}\omega_{sr\_ref}\frac{T_r}{L_s} \\ -a_{2N}I_{qs1} - \omega_{s_{ref}}I_{ds1} - \omega_{s\_ref}\Phi_{r\_ref} \\ -a_{2N}I_{ds2} + \omega_{s\_ref}I_{qs2} + \omega_{s\_ref}\Phi_{r\_ref}\omega_{sr\_ref}\frac{T_r}{L_s} \\ -a_{2N}I_{qs2} - \omega_{s\_ref}I_{ds2} - \omega_{s\_ref}\Phi_{r\_ref} \\ a_{3N}\left(I_{qs1} + I_{qs2}\right) - a_{4N}\omega \\ -a_{5N}\Phi_r + a_{6N}\left(I_{ds1} + I_{ds2}\right) \end{bmatrix}, \tag{6}$$

$$\Delta f(x) = \begin{bmatrix} \Delta\alpha_{1N}V_{ds1} - \Delta a_{2N}I_{ds1} \\ \Delta\alpha_{1N}V_{qs1} - \Delta a_{2N}I_{qs1} \\ \Delta\alpha_{1N}V_{ds2} - \Delta a_{2N}I_{ds2} \\ \Delta\alpha_{1N}V_{qs2} - \Delta a_{2N}I_{qs2} \\ \Delta a_{3N}\left(I_{qs1} + I_{qs2}\right) - \Delta a_{4N}\hat{\omega} - d \\ -\Delta a_{5N}\hat{\Phi}_r + \Delta a_{6N}\left(I_{ds1} + I_{ds2}\right) \end{bmatrix}, \tag{7}$$

$a_1 = \frac{1}{L_s}$; $a_2 = \frac{R_s}{L_s}$; $a_3 = P\frac{L_m}{L_m+L_r}\frac{\Phi_{r\_ref}}{j}$; $a_4 = \frac{B_\omega}{j}$; $a_5 = \frac{R_r}{L_m+L_r}$; $a_6 = R_r\frac{L_m}{L_m+L_r}$, $T_r = \frac{L_r}{R_r}$ and: $\alpha_i = \alpha_{iN} + \Delta\alpha_{iN}$.

In the above, the subscript "*N*" stands for the nominal value of the parameters, and the symbol "Δ" represents the perturbation factor in the parameters. The variable ϑ is introduced as:

$$\vartheta = \left[\Delta\alpha_{1N}\quad \Delta\alpha_{2N}\quad \Delta\alpha_{3N}\quad \Delta\alpha_{4N}\quad \Delta\alpha_{5N}\quad \Delta\alpha_{6N}\right], \tag{8}$$

By substituting the expressions for $f_N(x)$, and $\Delta f(x)$ into eq (3), (4) the mathematical model of DSIM system, suitable for the robust ABSM controller design, is obtained as follows:

$$\begin{cases} \frac{d}{dt}I_{ds1} = f_{N1} + \alpha_{1N}V_{ds1} + \beta_1\vartheta^T, \\ \frac{d}{dt}I_{qs1} = f_{N2} + \alpha_{1N}V_{qs1} + \beta_2\vartheta^T, \\ \frac{d}{dt}I_{ds2} = f_{N3} + \alpha_{1N}V_{ds2} + \beta_3\vartheta^T, \\ \frac{d}{dt}I_{qs2} = f_{N4} + \alpha_{1N}V_{qs2} + \beta_4\vartheta^T, \\ \frac{d}{dt}\omega = f_{N5} + \beta_5\vartheta^T + d, \\ \frac{d}{dt}\Phi_r = f_{N6} + \beta_6\vartheta^T, \end{cases} \tag{9}$$

where:

$$\beta = \begin{bmatrix} V_{ds1} & -I_{ds1} & 0 & 0 & 0 & 0 \\ V_{qs1} & -I_{qs1} & 0 & 0 & 0 & 0 \\ V_{ds2} & -I_{ds2} & 0 & 0 & 0 & 0 \\ V_{qs2} & -I_{qs2} & 0 & 0 & 0 & 0 \\ 0 & 0 & I_{qs1} + I_{qs2} & -\omega & 0 & 0 \\ 0 & 0 & 0 & 0 & -\phi_r & I_{ds1} + I_{ds2} \end{bmatrix} \tag{10}$$

Therefore, the objective is to develop an updated law ϑ to estimate the system parameters. At the same time, a control strategy is designed to ensure accurate tracking of the desired speed and flux. The proposed controller relies exclusively on the estimated states and parameters.

3. DESIGN OF THE ABSMC CONTROLLER

The control objective of the DSIM speed regulation system is the designed controller. This controller can guarantee high performance in speed trajectory tracking and lumped uncertainty suppression, and it attempts to perform a parametric update. The proposed nonlinear control based on the ABSMC algorithm is presented and described step by step as follows:

STEP 1: SPEED REGULATORS

To design the speed controller, the speed tracking error and its time derivative along the solution trajectory of speed are defined as:

$$\begin{cases} e_\omega = \widehat{\omega} - \omega_{ref}, \\ \frac{de_\omega}{dt} = f_{N1} + \vartheta^T\beta_5 - d - \frac{d\omega_{ref}}{dt} \end{cases} \quad (11)$$

The first Lyapunov stability function is chosen as $V_1 = \frac{1}{2}e_\omega^2$. Define the energy function, $H_1 = \frac{dV_1}{dt}$, which yields

$$H_1 = e_\omega\left[f_{N1} + \vartheta^T\beta_5 - d - \frac{d\omega_{ref}}{dt}\right].$$

By applying the Lyapunov stability theorem as in the first step, we obtain:

$$H_1 = -k_\omega e_\omega^2 + e_\omega\left[\alpha_{3N}\left(I_{qs1} + I_{qs2}\right) - \alpha_{4N}\widehat{\omega} + \vartheta^T\beta_5 - \frac{d\omega_{ref}}{dt} - d + k_\omega e_\omega\right],$$

where k_ω is a positive constant. The chosen low attractive stratifying the necessary condition of stability ($H_1 < 0$) is the reference current I_{qs_ref} is determined by

$$I_{qs\_ref} = \frac{1}{2a_{3N}}\left[\alpha_{4N}\widehat{\omega} - \widehat{\vartheta}^T\beta_5 + \frac{d\omega_{ref}}{dt} + d - k_\omega e_\omega\right]. \quad (12)$$

STEP 2: FLUX REGULATORS

In this step, the second error variable e_Φ. To ensure the operation of the machine in the linear regime (out of saturation), a flux control is also carried out, following an imposed trajectory Φ_{r_ref}. To achieve this, we pose

$$\begin{cases} e_\Phi = \widehat{\Phi}_r - \Phi_{r_{ref}}, \\ \frac{de_\Phi}{dt} = f_{N6} + \vartheta^T\beta_6 - \Phi_{r_{ref}}. \end{cases} \quad (13)$$

We define the Lyapunov function $V_2 = \frac{1}{2}e_\Phi^2$, which constitutes the error flux, and we define the energy function, $H_1 = \frac{dV_2}{dt} = -k_\Phi e_\Phi^2 + e_\Phi\left[-a_{5N}\widehat{\Phi}_r + \alpha_{6N}(I_{s1} + I_{ds2}) - \vartheta^T\beta_6 + k_\Phi e_\Phi\right]$, where k_Φ is a positive constant, the necessary condition verified Lyapunov stability, H_2 is the reference current I_{ds_ref}, which is determined by

$$I_{ds\_ref} = \frac{1}{2\alpha_{6N}}\left[\alpha_N\widehat{\Phi}_r - \widehat{\vartheta}^T\beta_6 + \frac{d\Phi_{r_{ref}}}{dt} - k_\Phi e_\Phi\right], \quad (14)$$

where $\widehat{\vartheta}$ refers to the estimated value of the uncertain parameter ϑ. Note that the estimation error, $\tilde{\vartheta} = \widehat{\vartheta} - \vartheta$, and from its value we redefine the H_1, H_2, by

$$\begin{cases} H_1 = -k_\Omega e_\omega^2 + e_\Omega\widetilde{\Theta}^T B_5, \\ H_2 = -k_\Phi e_\Phi^2 + e_\Phi\widetilde{\Theta}^T B_6. \end{cases} \quad (17)$$

STEP 2: CURRENTS REGULATORS

With the implementation of field-oriented control, the ideal control inputs and tracking errors of stator currents on the d-q axis are adopted as:

$$\begin{cases} e_{d1} = I_{ds\_ref} - I_{ds1}, \\ e_{q1} = I_{qs\_ref} - I_{qs1}, \\ e_{d2} = I_{ds\_ref} - I_{ds2}, \\ e_{q2} = I_{qs\_ref} - I_{qs2}. \end{cases} \quad (18)$$

To ensure the speed tracking performance and uncertain disturbance rejection ability, the augmented Lyapunov function V is selected as:

$$\Psi = V_1 + V_2 + \frac{1}{2}\left[S_{d1}^2 + S_{d1}^2 + S_{d1}^2 + S_{d1}^2 + \tilde{\vartheta}^T\Gamma\tilde{\vartheta}\right],$$

where Γ denotes a positive definite symmetric matrix, S_{q1}, S_{d2} and S_{q2} are the integral sliding surfaces corresponding to e_{d1}, e_{d2}, e_{q1} and e_{q2} respectively, which are defined as follows

$$\begin{cases} S_{d1} = e_{d1} + \lambda_d\int_0^t e_{d1}dt, \\ S_{q1} = e_{q1} + \lambda_q\int_0^t e_{q1}dt, \\ S_{d2} = e_{d2} + \lambda_d\int_0^t e_{d2}dt, \\ S_{q2} = e_{q2} + \lambda_q\int_0^t e_{q2}dt, \end{cases} \quad (14)$$

where $\lambda_{d,q}$, the sliding surface gains are positive, define the function $H = \frac{dV_2}{dt}$ for the closed-loop system. Replacing the derivative of the current from eq. (3) in function yields

$$\begin{aligned} H = &-k_\omega e_\omega^2 - k_\Phi e_\Phi^2 - k_d S_{d1}^2 - k_q S_{q1}^2 - k_d S_{d2}^2 - k_q S_{q2}^2 - k_\vartheta\tilde{\vartheta}^T\tilde{\vartheta} + \\ &+\tilde{\vartheta}^T\left(-e_\omega\beta_5 + e_\Phi\beta_6 + \beta_1 S_{d1} + \beta_2 S_{q1} + \beta_3 S_{d2} + \beta_4 S_{q2} + \Gamma\frac{d\tilde{\vartheta}}{dt} + k_\vartheta\tilde{\vartheta}\right) \\ &+S_{d1}\left[f_{N1} + \alpha_{1N}V_{ds1} + \beta_1\tilde{\vartheta}^T - \frac{dI_{ds_{ref}}}{dt} + \lambda_d e_{d1} + k_d S_{d1}\right] \\ &+S_{q1}\left[f_{N2} + \alpha_{1N}V_{qs1} + \beta_2\tilde{\vartheta}^T - \frac{dI_{qs_{ref}}}{dt} + \lambda_q e_{q1} + k_q S_{q1}\right] \\ &+S_{d2}\left[f_{N3} + \alpha_{1N}V_{ds2} + \beta_3\tilde{\vartheta}^T - \frac{dI_{ds_{ref}}}{dt} + \lambda_d e_{d2} + k_d S_{d2}\right] \\ &+S_{q2}\left[f_{N4} + \alpha_{1N}V_{qs2} + \beta_4\tilde{\vartheta}^T - \frac{dI_{qs\_ref}}{dt} + \lambda_q e_{q2} + k_q S_{q2}\right] \end{aligned} \quad (15)$$

where k_d, k_q and k_ϑ refer to positive constants remaining to be designed. Thus, to guarantee the practical stability operation of the DSIM closed-loop system, the nonlinear robust control ABSMC law is designed as follows,

$$\begin{cases} V_{ds1\_ref} = \frac{1}{\alpha_{1N}}\left(-f_{N1} - \beta_1\widehat{\vartheta}^T + \frac{dI_{ds\_ref}}{dt} - \lambda_d e_{d1} - k_d S_{d1}\right), \\ V_{qs1\_ref} = \frac{1}{\alpha_{1N}}\left(-f_{N2} - \beta_2\widehat{\vartheta}^T + \frac{dI_{qs\_ref}}{dt} - \lambda_d e_{q1} - k_q S_{q1}\right), \\ V_{ds2\_ref} = \frac{1}{\alpha_{1N}}\left(-f_{N3} - \beta_3\widehat{\vartheta}^T + \frac{dI_{ds\_ref}}{dt} - \lambda_d e_{d2} - k_d S_{d2}\right), \\ V_{qs2\_ref} = \frac{1}{\alpha_{1N}}\left(-f_{N4} - \beta_4\widehat{\vartheta}^T + \frac{dI_{qs\_ref}}{dt} - \lambda_q e_{q2} - k_q S_{q2}\right). \end{cases} \quad (19)$$

These equations constituted the reference equation voltage system.

3.3 UPDATE ϑ WITH GLOBAL STABILITY

Then, the update law for the uncertain parameters can be estimated as:

$$\frac{d\widehat{\vartheta}}{dt} = \Gamma^{-1}\begin{bmatrix} \beta_1 S_{d1} + \beta_2 S_{q1} + \beta_3 S_{d2} + \beta_4 S_{q2} + e_\omega\beta_5 - \\ -e_\Phi\beta_6 - k_\vartheta\tilde{\vartheta} \end{bmatrix}. \quad (20)$$

Substituting eq. (20) into (18), yields,

$$\begin{aligned} H = &-k_\omega e_\omega^2 - k_\Phi e_\Phi^2 - k_d S_{d1}^2 - k_q S_{q1}^2 - \\ &-k_d S_{d2}^2 - k_q S_{q2}^2 - k_\vartheta\tilde{\vartheta}^T\tilde{\vartheta}. \end{aligned}$$

Redefine the Lyapunov function V, and it follows that $\frac{dV}{dt} = H < 0$. In view of the Lyapunov theorem, we can conclude that the closed-loop system is asymptotically stable. Therefore, with the tracking and estimation errors close to zero $t \to \infty$, it can realize the targets of speed flux tracking and disturbance rejection based on the above analysis.

Finally, the terms $\Delta\alpha_{iN}$ correspond to the elements of the estimated parameter vector $\widehat{\vartheta}$. It can be calculated using eq. (20). If the unknown parameters, for example, R_r, R_s and L_s, their estimated values can be deduced from the coefficients α_i using eq. (7) as follows:

$$\widehat{R}_r = \widehat{\alpha}_5(L_m + L_r), \widehat{R}_s = \widehat{\alpha}_2 L_{sn}, \widehat{L}_s = \frac{1}{\widehat{\alpha}_1}. \quad (21)$$

4. SIMULATION AND INTERPRETATION OF RESULTS

The verification and testing of the proposed control is synthesized for the asynchronous double stator motor using a simulation series by MATLAB, as described in Fig. 1. The used parameters are indicated in Table 1. The motor is fed by two identical inverters (INV1 and INV2).

The results presented in this paper were achieved with the following synthesis parameters: $k_\omega = k_\Phi = 200$, $P = 1000 I_6$, $k_\vartheta = 1000$, $k_d = 0.2$, $k_q = 0.5$.

Based on the updating of vector θ in eq (20), it is easy to identify or estimate machine parameters without knowing their nominal values. So, the estimator part in the control) adaptive update is based on the error and error integral of the values of measured quantities. Either measured or given by eq. (21), taking into consideration that the initial value of the vector $\hat{\vartheta}$ is 0.

The first objective is to control the rotational speed. The reference speed is rapidly increased to 250 rad/s, maintained at this value for 1.5 s, and then reversed to (−250 rad/s). In addition, the desired flux is set to 1 Wb. A positive load torque, representing an external disturbance, $T_l = 14$ Nm is applied at $t = 2$ s.

We can see that at start-up and during the transitory regime, speed increases linearly as a function of time, attaining its reference value at $t = 0.4$ s without overshooting. The electromagnetic torque stable to steady state (at $t = 0.4$ s). The quadrature currents initially reach 18 A and then evolve identically to the electromagnetic torque. The rotor current is approximately 30 A for 0.6 s, then decreases to stabilize at $t = 0.52$ s. The rotor flux along (d, q) shows peaks at start-up for a fraction of a second, then stabilizes and continues along its path according to its reference.

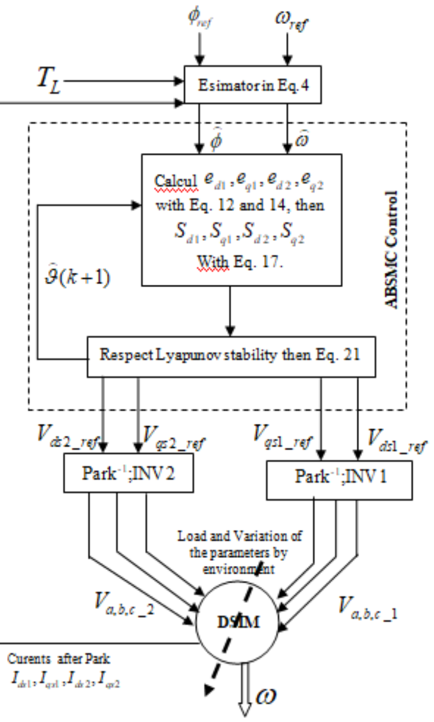

Fig. 1 – The control structure of the DSIM speed drive system with the proposed method.

Figure 2 shows the speed evolution of DSIM characteristics with ABSMC control.

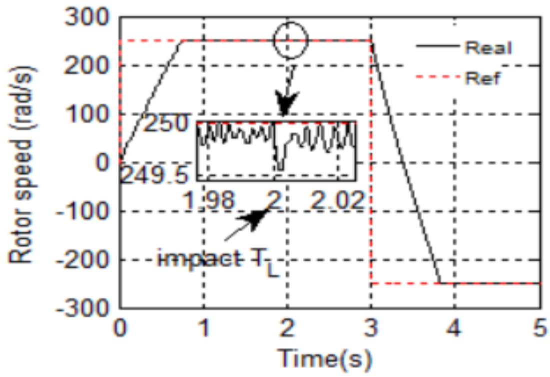

Fig. 2 – Dynamic response of the speed machine with application of the ABSMC control, when reversing the direction of speed rotation at time $t = 3$ s.

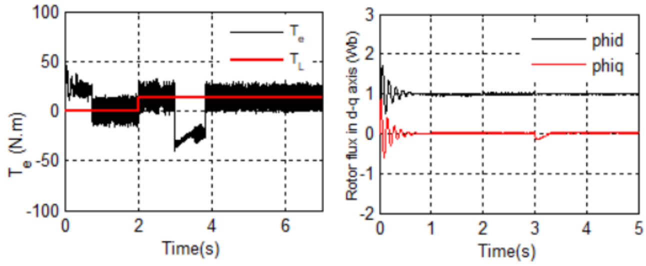

Fig. 3 – Dynamic response of the machine with application of the ABSMC control, when inverting the direction of speed at time $t = 3$ s.

According to Fig. 4, the estimation of the machine parameters DSIM, like rotor resistance, stator resistance, and inductance, is efficient, as its estimated value converges very rapidly to their nominal values. The results obtained show the good performance of the ABSMC control strategy.

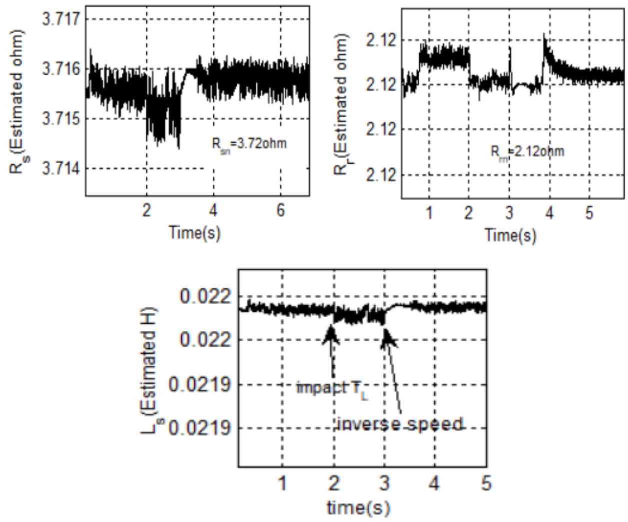

Fig. 4 – Evolution of rotor and stator resistance and inductance, during speed inversion.

The performance of the ABSMC control system in rotor resistance variation is shown in Fig. 5 and Fig. 6, where the rotor resistance is increased with some delay by 50 % of its nominal value from $t = 1.5$ s, with a load (a resistive torque of 14 Nm) introduced at $t = 2$ s. The speed value is varied in a different range, 250-100 rad/s. It can be clearly seen that there is no influence of the variation of the stator resistance on the load, while the speed stabilizes at the reference speed value of 100 rad/s.

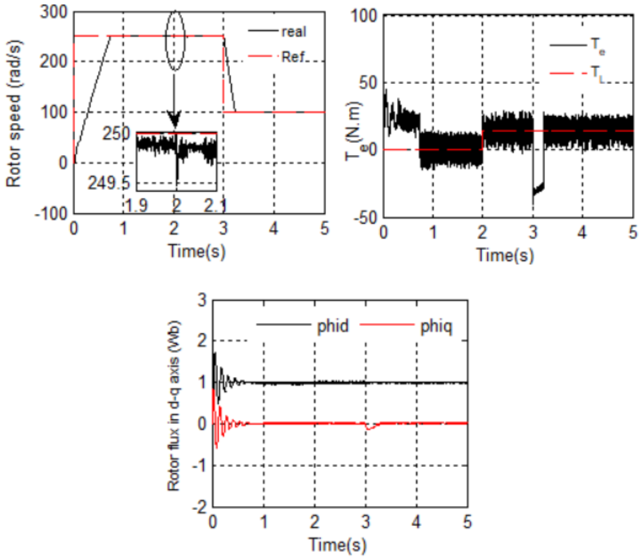

Fig. 5 – Dynamic response of the machine with application of the ABSMC control, when robustness test (in rotor resistance increased at $t = 1$ s).

On start-up. The electromagnetic torque has a damped sinusoidal form because of the effect estimator, eq. (4), then compensates for the applied load torque, and it should be noted that the stator resistance converges completely to its nominal value.

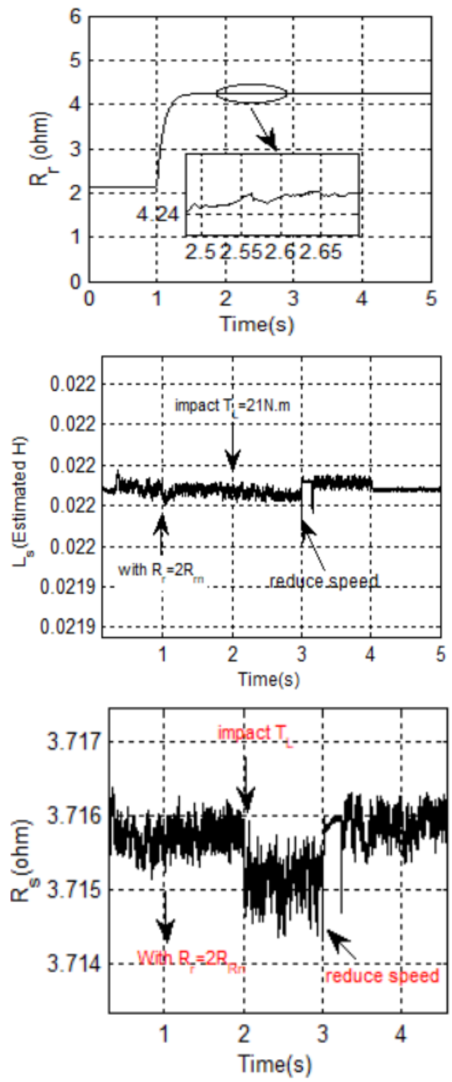

Fig. 6 – Parameter estimation with the application of the ABSMC control during the robustness test (with increased rotor resistance).

From the above, the results prove that the proposed ABSMC method effectively improves the performance of the system. It has the advantages of fast response, strong robustness against parameter variations, and external disturbances.

5. CONCLUSION

This paper proposes an adaptive Backstepping sliding mode controller ABSMC for a double-stator induction motor DSIM. This controller ensures asymptotic speed tracking, reduces chattering, and provides robust performance under parameter uncertainties via Lyapunov-based online adaptation.

The simulation results confirm the effectiveness of the proposed controller and demonstrate satisfactory dynamic performance. The controller ensures robustness and global stability. It provides a fast response without overshoot and maintains robustness under parameter variations and external disturbances.

Finally, in future work, we advise that we must estimate the speed, flux, and load torque using a Kalman filter before the experimental implementation of the proposed control.

CREDIT AUTHORSHIP CONTRIBUTION STATEMENT

CHAABANE HADJI: conceptualization, methodology, software, writing, original draft, validation.

DJALAL EDDINE KHODJA: Supervision, review & editing, results curation, investigation.

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Table 1

Parameters of the motor DSIM.

Parameters	Symbols & values
Stators resistances	$R_{s1} = R_{s2} = 3.72\ \Omega$
Rotor resistance	$R_r = 3.72\ \Omega$
Stator self-inductances	$L_{s1} = L_{s2} = 0.022$H
Rotor self-inductance	$L_r = 0.006$ H
Mutual inductance	$L_m = 0.3672$ H
Moment of inertia	j =0.0662 $\mathrm{kg.m^2}$
Viscous friction coefficient	β_ω= 0.001 N.m.s/rad
Nominal frequency	f = 50 Hz.

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