Identification And Adaptive Control Strategy for Saturated Wound-Rotor Induction Motors

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Abstract. This paper presents an online identification method and an adaptive control of an induction machine operating under saturated conditions. The system parameters may vary with the operating point. The system model is formulated in the α/β reference frame. A parameter estimation algorithm is derived from the machine's dynamic and algebraic equations. It is used in the design of an indirect adaptive controller for speed and rotor flux using the so-called backstepping approach. The control law is developed using the estimated fluxes since the actual fluxes are not directly measurable and the estimated time-varying machine inductances converge almost instantaneously to their true values. MATLAB/Simulink simulation results show indeed the fast convergence of estimated parameters. They also demonstrate the effectiveness of the proposed adaptive control strategy.

INTRODUCTION

The induction machine remains one of the most widely used electrical machines in industrial applications because of its robustness and relatively low cost [1]. It plays an important role in variable-speed drives, where high requirements in terms of dynamic performance and robustness are increasingly sought [2]. However, achieving such performance generally relies on the availability of a sufficiently accurate machine model, often based on the assumption of constant parameters. In practice, this assumption is difficult to justify. Indeed, the machine resistances and inductances may vary significantly under the effect of operating conditions such as temperature rise and magnetic saturation [3]. The latter introduces significant nonlinearities into the electromagnetic behavior of the machine. It degrades the accuracy of the nominal model and, consequently, the performance of conventional control strategies. In this context, the development of methods capable of accounting for both saturation and parameter variability constitutes a major challenge for the advanced control of induction machines [4].

Several studies have been devoted to the identification of induction machine parameters [5] and the synthesis of adaptive control laws [6]. Nevertheless, in many approaches, the electrical parameters are assumed to be constant or slightly varying, and saturation effects are neglected or taken into account in a simplified manner. In addition, parameter identification and control for saturated induction motors are often treated separately. This limits the overall adaptability of the system in the presence of rapid changes in operating conditions. It therefore appears relevant to develop a unified framework in which fast online parameter estimation directly supports the control design. It is within this perspective that the present work is situated.

The paper considers an induction machine model formulated in the α/β reference frame. The state variables are the stator and rotor fluxes, together with the rotor speed. It does not rely on any restrictive assumption regarding the resistances and inductances. A time-varying parameter identification algorithm is developed using the machine dynamic equations and algebraic relations. This approach leads to an estimator capable of tracking online the unknown parameters of the machine under saturated conditions.

Next, the estimated parameters are used for the synthesis of an adaptive controller for speed and rotor flux. The proposed control strategy relies on an appropriate reformulation of the model, together with the backstepping design approach. Since the actual machine fluxes are not directly measurable, the design is based on their estimates obtained from the parameters identified online. This approach thus makes it possible to construct a coherent architecture in which parameter identification and adaptive control are tightly coupled.

The originality of the paper lies in the proposal of an integrated framework that combines the online estimation of the time-varying parameters and the synthesis of an adaptive controller for the induction machine. The simulation results illustrate the relevance of the approach. The convergence of the estimated parameters to the true parameters is almost instantaneous. The proposed control scheme achieves satisfactory dynamic performance.

The paper is organized as follows. Section I presents the induction machine model adopted in this study. It also introduces a change of variables that leads to a reformulation of the model in a new state space. The new model is well suited to the implementation of the backstepping design approach for the synthesis of the speed and rotor flux control laws. Section II presents in detail the methodology adopted to derive the identification algorithm for the time-varying parameters of the induction machine. Section III is dedicated to the synthesis of the adaptive controller based on the estimated parameters. Section IV presents the simulation results obtained to assess and validate the performance of both the identification algorithm and the proposed controller. Finally, the conclusion is presented in the last section.

I. INDUCTION MOTOR MODEL

The dynamic model adopted in this study uses stator and rotor fluxes and rotor speed as state variables. The algebraic relationships relating fluxes to currents are also included. No constant-parameter assumption is made for the machine inductances and resistances; therefore, these parameters may vary with operating conditions while the model remains valid.

Nevertheless, in order to lighten the analytical developments, it will be assumed that the machine operates under balanced conditions. Under this assumption, the representation in the α/β reference frame can be used to describe the machine's dynamics. The dynamic model of the induction machine is then written in the following form: [7]

$$\frac{d}{dt}\psi_{s\alpha} = -R_s i_{s\alpha} + u_{s\alpha} \qquad 1.1$$

$$\frac{d}{dt}\psi_{s\beta} = -R_s i_{s\beta} + u_{s\beta} \qquad 1.2$$

$$\frac{d}{dt}\psi_{r\alpha} = -R_r i_{r\alpha} - \omega\psi_{r\beta} \qquad 1.3$$

$$\frac{d}{dt}\psi_{r\beta} = -R_r i_{r\beta} + \omega\psi_{r\alpha} \qquad 1.4$$

$$\frac{d}{dt}\omega = \frac{M}{\Delta}(\psi_{s\beta}\psi_{r\alpha} - \psi_{s\alpha}\psi_{r\beta}) - \frac{T_L}{J} \qquad 1.5a$$

Equivalently,

$$\frac{d}{dt}\omega = M(i_{s\beta}i_{r\alpha} - i_{s\alpha}i_{r\beta}) - \frac{T_L}{J} \qquad 1.5b$$

The algebraic relationships between the fluxes and the currents have the following form. It should be specified that the inductances and resistances appearing in these expressions may change when the machine is operating.

$$\psi_{s\alpha} = L_s i_{s\alpha} + M i_{r\alpha} \qquad 2.1$$

$$\psi_{rb} = L_s i_{sb} + M i_{rb} \qquad 2.2$$

$$\psi_{r\alpha} = L_r i_{r\alpha} + M i_{s\alpha} \qquad 2.3$$

$$\psi_{rb} = L_r i_{rb} + M i_{sb} \qquad 2.4$$

Where $\Delta = L_s L_r - M^2$ and L_r, L_s and M are the rotor inductance, the stator inductance and the mutual inductance, respectively. R_r and R_s are the rotor resistance and the stator resistance, respectively. $i_{s\alpha}$, $i_{s\beta}$ $i_{r\alpha}$ and $i_{r\beta}$ are the stator and rotor current components in the (α/β) reference frame. ω is the rotational speed of the machine rotor in rad/s. T_L is the mechanical load torque in N-m. J is the machine inertia.

We introduce the following change of variables with a view to synthesizing the speed and flux control laws for the induction machine. The resulting transformation leads to a model that is well suited to the implementation of the backstepping-based control design approach.

$$\Sigma = (\psi_{s\beta}\psi_{r\alpha} - \psi_{s\alpha}\psi_{r\beta}) \qquad 3.1$$

$$\Phi = (\psi_{s\alpha}\psi_{r\alpha} + \psi_{s\beta}\psi_{r\beta}) \qquad 3.2$$

$$\Psi = (\psi_{r\alpha}\psi_{r\alpha} + \psi_{r\beta}\psi_{r\beta}) \qquad 3.3$$

$$\Gamma = (\psi_{s\alpha}\psi_{s\alpha} + \psi_{s\beta}\psi_{s\beta}) \qquad 3.4$$

It should be noted that the new state variables have a physical interpretation. The first variable is proportional to the electromagnetic torque of the machine. The second variable corresponds to the dot product of the stator and rotor flux vectors expressed in the α/β reference frame. It provides a measure of the degree of magnetic alignment between these two flux vectors. It therefore reflects the strength of their electromagnetic coupling. The third and fourth state variables correspond respectively to the squared magnitudes of the rotor and stator fluxes.

The dynamic model of the induction machine can be reformulated as follows:

$$\frac{d}{dt}\Psi = -\frac{2R_rL_s}{\Delta}\Psi + \frac{2R_rM}{\Delta}\Phi \qquad 4.1$$

$$\frac{d}{dt}\Phi = -\frac{(R_rM + R_sL_s)}{\Delta}\Phi + \frac{(R_rM - R_sL_s)}{\Delta}\Psi + \omega\Sigma + u_1 \qquad 4.2$$

$$\frac{d}{dt}\omega = \frac{M}{\Delta}\Sigma - \frac{T_L}{J} \qquad 4.3$$

$$\frac{d}{dt}\Sigma = -\frac{(R_rL_s + R_sL_r)}{\Delta}\Sigma - \omega\Phi + u_2 \qquad 4.4$$

$$\frac{d}{dt}\Gamma = -\frac{2R_sL_r}{\Delta}\Gamma + \frac{2R_sM}{\Delta}\Phi + \frac{2\Phi}{\Psi}u_1 - \frac{2\Sigma}{\Psi}u_2 \qquad 4.5$$

Where

$$u_1 = \psi_{r\alpha}u_{s\alpha} + \psi_{r\beta}u_{s\beta} \qquad 4.6$$

$$u_2 = \psi_{r\alpha}u_{sb} - \psi_{r\beta}u_{sa} \qquad 4.7$$

II. PARAMETER IDENTIFICATION ALGORITHM

This section presents the approach used to obtain the estimation algorithm for the potentially time-varying parameters of the induction machine. These parameters are the stator and rotor resistances, the stator and rotor inductances, and the mutual inductance. The algorithm is based on the dynamic equations and the algebraic equations describing the induction machine. In order to shorten the presentation, only the procedure associated with the α-axis component of the stator flux is detailed in the main body of the text. It will be shown how the dynamic equation and the algebraic equation relating the stator flux to the stator and rotor currents can be used to derive the first equation required for the parameter identification algorithm. The derivations for the four other equations are provided in the appendix.

Let us first consider the dynamics of the α-axis component of the stator flux and the algebraic equation relating the stator and rotor currents to the stator flux.

$$\frac{d}{dt}\psi_{s\alpha} = -R_s i_{s\alpha} + u_{s\alpha} \qquad 5.1$$

$$\psi_{s\alpha} = L_s i_{s\alpha} + M i_{r\alpha} \qquad 5.2$$

First, Equation (5.1) is filtered using the transfer function 1/(s+ α) where α is a positive real number. The expressions obtained after this filtering operation, together with the corresponding simplifications, are presented in the equations below.

$$\frac{1}{s+\alpha}\left(\dot{\psi}_{s\alpha}\right)+\frac{1}{s+\alpha}(R_s i_{s\alpha})=\frac{1}{s+\alpha}(u_{s\alpha}) \qquad 6$$

$$\psi_{s\alpha}-\frac{\alpha}{s+\alpha}(\psi_{s\alpha})+\frac{1}{s+\alpha}(R_s i_{s\alpha})=\frac{1}{s+\alpha}(u_{s\alpha}) \quad 7$$

Next, the flux equation (Equation 5.2) is substituted into the above expression to obtain the following equation.

$$L_s i_{s\alpha}+M i_{r\alpha}\ -\frac{\alpha}{s+\alpha}(L_s i_{s\alpha}+M i_{r\alpha}\ )+\frac{1}{s+\alpha}(R_s i_{s\alpha})=\frac{1}{s+\alpha}(u_{s\alpha}) \qquad 8$$

The swapping lemma [8] is then used to further decompose this equation. This lemma provides a rigorous way to handle the result of filtering the product of two time-varying signals. Its application ultimately leads to the following equation.

$$L_s i_{s\alpha}+M i_{r\alpha}\ -\alpha\left\{L_s\frac{1}{s+\alpha}(i_{s\alpha})-\frac{1}{s+\alpha}\left[\dot{L}_s\frac{1}{s+\alpha}(i_{s\alpha})\right]\right\}-\alpha\left\{M\frac{1}{s+\alpha}(i_{r\alpha})-\frac{1}{s+\alpha}\left[\dot{M}\frac{1}{s+\alpha}(i_{r\alpha})\right]\right\}+\left\{R_s\frac{1}{s+\alpha}(i_{s\alpha})-\frac{1}{s+\alpha}\left[\dot{R}_s\frac{1}{s+\alpha}(i_{s\alpha})\right]\right\}=\frac{1}{s+\alpha}(u_{s\alpha}) \qquad 9$$

We now define the equations below. They represent the realizations of the filters that appear in Equation (9). The first equation is the realization of the filter $\frac{1}{s+\alpha}(i_{s\alpha})$ applied to the stator current $i_{s\alpha}$. The second equation represents the realization of the filter applied to the rotor current $i_{r\alpha}$. The third equation is the realization of the filter applied to the stator voltage $u_{s\alpha}$. The fourth equation represents the realization of the filter applied to $\dot{L}_s\frac{1}{s+\alpha}(i_{s\alpha})+\dot{M}\frac{1}{s+\alpha}(i_{r\alpha})$-$\dot{R}_s\frac{1}{s+\alpha}(i_{s\alpha})$.

$$\dot{v}_{isa}=-\alpha_1 v_{isa}+i_{sa} \qquad 10.1$$

$$\dot{v}_{ira}=-\alpha_1 v_{ira}+i_{ra} \qquad 10.2$$

$$\dot{v}_{usa}=-\alpha_1 v_{usa}+u_{sa} \qquad 10.3$$

$$\dot{\omega}_1^{sa}=-\alpha\omega_1^{sa}+\alpha\dot{L}_s v_{isa}+\alpha\dot{M}v_{ira}-\dot{R}_s v_{isa} \qquad 10.4$$

Consequently, Equation (9) can be equivalently reformulated as follows:

$$L_s i_{s\alpha}+M i_{r\alpha}\ -\alpha L_s v_{isa}-\alpha M v_{ira}+R_s v_{isa}-\omega_1^{sa}=v_{usa} \qquad 11$$

This equation constitutes the first of the five relationships required for deriving the identification algorithm for the five unknown parameters of the induction machine.

The method used to derive Equation (11) is then applied to the other equations of the induction machine dynamics. This results in four additional equations needed to complete the derivation of the parameter identification algorithm. For the sake of conciseness, the details of these derivations are omitted. The resulting equations are as follows:

$$L_s i_{sb}+M i_{rb}\ -\beta L_s v_{isb}-\beta M v_{irb}+R_s v_{isb}-\omega_1^{sb}=v_{usb} \qquad 12.1$$

$$M i_{sa}+L_r i_{ra}\ -\gamma M v_{isa}-\gamma L_r v_{ira}+R_r v_{ira}-\omega_1^{ra}=0 \qquad 12.2$$

$$M i_{sb}+L_r i_{rb}\ -\delta M v_{isb}-\delta L_r v_{irb}+R_r v_{irb}-\omega_1^{sb}=0 \qquad 12.3$$

$$\omega-\sigma v_\omega-\frac{M}{J}v_{Te}+\omega_1^{\omega}=-v_{Tl} \qquad 12.4$$

We now introduce the vector of unknown parameters, defined as follows:

$$\theta=(R_s\quad R_r\quad L_s\quad L_r\quad M)^T$$

It can be shown that all the previously established equations can be rewritten in the following compact form:

$$\dot{W}=AW-\Xi\dot{\theta} \qquad 13.1$$

$$\Psi\theta=\mathrm{Y}+\mathrm{W} \qquad 13.2$$

where

$$A=-diag(\alpha\quad\beta\quad\gamma\quad\delta\quad\sigma)^T$$

$$W=(\omega_1^{sa}\quad\omega_1^{sb}\quad\omega_1^{ra}\quad\omega_1^{rb}\quad\omega_1^{\omega})^T$$

$$Y=(v_{usa}\quad v_{usb}\quad 0\quad 0\quad -\omega+\sigma v_\omega+v_{Tl})^T$$

$$\Psi=\begin{bmatrix} v_{isa} & 0 & i_{sa}-\alpha v_{isa} & 0 & i_{r\alpha}-\alpha v_{ira}\\ v_{isb} & 0 & i_{sb}-\beta v_{isb} & i_{ra}-\gamma v_{ira} & i_{rb}-\beta v_{irb}\\ 0 & v_{ira} & 0 & i_{rb}-\delta v_{irb} & i_{sa}-\gamma v_{isa}\\ 0 & v_{irb} & 0 & 0 & i_{sb}-\delta v_{isb}\\ 0 & 0 & 0 & 0 & -\eta v_{Te}\end{bmatrix}$$

$$\Xi=-\Psi+diag\left(i_{s\alpha}\quad i_{s\beta}\quad i_{r\alpha}\quad i_{r\beta}\quad 0\right)$$

Where η=1/J. This leads to the estimation algorithm for the variable parameters of the induction machine operating under saturated conditions. It can be formulated as follows [9]

$$\frac{d}{dt}\widehat{W}=A\widehat{W}-\rho\Psi^{-1}\Xi\left\{Y+\widehat{W}-\Psi\widehat{\theta}\right\} \qquad 14.1$$

$$\frac{d}{dt}\widehat{\theta}=\rho\Psi^{-1}(Y+\widehat{W}-\Psi\widehat{\theta}) \qquad 14.2$$

$\widehat{W}$ is an estimate for the vector W and $\widehat{\theta}$ is an estimate for θ.

It was shown in [9] that the estimated parameters converge almost instantaneously to the exact values of the unknown parameters. Based on this result, the next section focuses on using these estimates to synthesize an adaptive control law for the induction machine.

III. ADAPTIVE CONTROL DESIGN

In this section, we design an adaptive controller for the induction machine operating under saturated conditions. We assume that the machine parameters may change with the operating conditions. Estimates of these parameters are obtained using the estimation algorithm presented in the previous section. Since it was shown that the estimated parameters converge almost instantaneously to the true unknown values, the controller design is based on the estimated flux components rather than the actual fluxes. The actual fluxes are not directly measurable. The machine flux estimates are defined as follows.

$$\hat{\psi}_{s\alpha}=\hat{L}_s i_{s\alpha}+\widehat{M}i_{r\alpha} \qquad 15.1$$

$\hat{\psi}_{rb} = \hat{L}_s i_{sb} + \hat{M} i_{rb}$ 15.2

$\hat{\psi}_{r\alpha} = \hat{L}_r i_{r\alpha} + \hat{M} i_{s\alpha}$ 15.3

$\hat{\psi}_{rb} = \hat{L}_r i_{rb} + \hat{M} i_{sb}$ 15.4

The equivalent expressions of the state variables associated with the machine dynamics can then be written as follows:

$\hat{\Sigma} = (\hat{\psi}_{s\beta}\hat{\psi}_{r\alpha} - \hat{\psi}_{s\alpha}\hat{\psi}_{r\beta})$ 16.1

$\hat{\Phi} = (\hat{\psi}_{s\alpha}\hat{\psi}_{r\alpha} + \hat{\psi}_{s\beta}\hat{\psi}_{r\beta})$ 16.2

$\hat{\Psi} = (\hat{\psi}_{r\alpha}\hat{\psi}_{r\alpha} + \hat{\psi}_{r\beta}\hat{\psi}_{r\beta})$ 16.3

For readability and simplicity of notation, the actual machine signals are used in the derivation, although their estimated counterparts are used in implementation.

The following paragraph is devoted to the design of the flux controller. Its objective is to impose a prescribed flux level on the machine rotor. To derive the expression of this control law, we will rely on the dynamics given in Equations (4.1) and (4.2). The backstepping design approach will serve as the methodological framework for structuring the controller. From this perspective, Equation (4.1) can first be rewritten in the following form, after introducing the flux error variable as well as certain parameters defined later.

$\frac{d}{dt}\widetilde{\Psi} = -\theta_1\Psi + \theta_2\Phi - \dot{\Psi}_{ref}$ 17

We also introduce the following parameters. They appear in the system dynamics and depend directly on the machine parameters.

$\theta_1 = \frac{2R_r L_s}{\Delta}$ and $\theta_2 = \frac{2R_r M}{\Delta}$

The flux error is also defined as the difference between the flux and its reference.

$\widetilde{\Psi} = \Psi - \Psi_{ref}$ 18

Let $\hat{p}_2$ be an estimate of the inverse of parameter θ_2.

$\hat{p}_2 = 1/\hat{\theta}_2$ 19

Equation (14) is then used to determine the reference for the variable Φ. This reference is constructed so as to ensure the stabilization of the dynamics described by Equation (14), if it were considered as the control variable. It can be written as follows:

$\Phi^* = \hat{p}_2 v_1 - \frac{k}{4} v_1^2 \widetilde{\Psi}$ 20

Where

$v_1 = -k_1\widetilde{\Psi} + \hat{\theta}_1\Psi + \dot{\Psi}_{ref} - \frac{k}{4}\widetilde{\Psi}\Psi^2$

k_1 and k are two positive gains. It should be noted that this reference depends on the estimated parameters, or more precisely on quantities derived from the estimates of the machine parameters. We then introduce the following error signal, together with its dynamics. The latter will be used to derive the expression of the actual control input u_1.

$\widetilde{\Phi} = \Phi - \Phi^*$ 21.1

$\frac{d}{dt}\widetilde{\Phi} = -\theta_3\Phi + \theta_4\Psi + \omega\Sigma + u_1 - \dot{\Phi}^*$ 21.2

with

$\theta_3 = \frac{(R_r M + R_s L_s)}{\Delta}$ $\theta_4 = \frac{(R_r M - R_s L_s)}{\Delta}$

It can then be shown that, if u_1 is chosen according to the following expression, the dynamics (4.1) and (4.2) are stabilized simultaneously. The resulting expression for u_1 will later be used to determine the α and β components of the stator voltage.

$u_1 = -k_2\widetilde{\Phi} + \hat{\theta}_3\Phi - \hat{\theta}_4\Psi - \omega\Sigma + \dot{\Phi}^* + \hat{\theta}_2\widetilde{\Psi} - \frac{k}{4}\widetilde{\Phi}\left(\Psi^2 + \widetilde{\Psi}^2 + \Phi^2\right)$ 23

k_2 is a positive gain. The following paragraph presents the design of the speed controller for the induction machine. First, the speed error is introduced. Then, the associated dynamics is obtained in the following form:

$\dot{\widetilde{\omega}} = -\theta_5\Sigma - \theta_6 - \dot{\omega}_{ref}$ 24

where $\widetilde{\omega} = \omega - \omega_{ref}$

The parameter θ_5 together with the estimate of its inverse are defined below.

$\theta_5 = \frac{M}{\Delta}$ and $\hat{p}_5 = 1/\hat{\theta}_5$ 25

First, the reference for the variable Σ is obtained. This reference is obtained by treating Σ as a virtual control input intended to stabilize the speed error dynamics. The corresponding expression for the reference of Σ is given by:

$\Sigma^* = \hat{p}_5 v_3 - \frac{k}{4} v_3^2 \widetilde{\omega}$ 26

where $v_3 = \frac{T_L}{J} - k_3\widetilde{\omega} - \frac{k}{4}\widetilde{\omega} + \dot{\omega}_{ref}$

k_3 is a positive gain. The error signal associated with Σ is then introduced, and its dynamics is obtained as shown below.

$\dot{\widetilde{\Sigma}} = -\theta_6\Sigma - \omega\Phi + u_2 - \dot{\Sigma}^*$ 27

where $\widetilde{\Sigma} = \Sigma - \Sigma^*$ and $\theta_6 = \frac{(R_r L_s + R_s L_r)}{\Delta}$

Finally, this latter dynamic equation is used to derive the expression of u_2. It can be written in the following form:

$u_2 = -k_4\widetilde{\Sigma} + \hat{\theta}_6\Sigma + \omega\Phi + \dot{\Sigma}^* + \hat{\theta}_5\widetilde{\omega} - \frac{k}{4}\widetilde{\Sigma}(\Sigma^2 + \widetilde{\omega}^2)$ 28

k_4 is a positive gain. The signal u_2 will later be used to determine the components of the stator voltage to be applied to the machine inverter. The α and β components of the stator voltage are obtained from Equations (4.6) and (4.7). The corresponding expressions take the following form:

$u_{s\alpha} = \frac{\hat{\psi}_{r\alpha}u_1 - \hat{\psi}_{r\beta}u_2}{\hat{\psi}_{r\alpha}^2 + \hat{\psi}_{r\beta}^2}$ 29.1

$u_{s\beta} = \frac{\psi_{r\alpha}u_2 + \psi_{r\beta}u_1}{\hat{\psi}_{r\alpha}^2 + \hat{\psi}_{r\beta}^2}$ 29.2

It is important to recall that, in the expressions above, the actual variables have been systematically replaced by their respective estimates. Stability of the closed-loop system can be established using a Lyapunov-based argument. The proof is omitted for brevity.

Note that the derivatives of certain signals are required to implement u_{sa} and u_{sb}. It is not recommended to use numerical differentiation, as it could amplify measurement noise. Instead, we recommend obtaining the derivatives of these signals by using the filter given below.

Indeed, the derivative of a signal y denoted by $\mathrm{d}y$, is obtained as follows [10]:

$\dot{x}_1 = 5sign(y - x_1)$ 30.1

$\dot{x}_2 = 0.25(\dot{x}_1 - x_2)$ 30.2

$\mathrm{d}y = x_2$ 30.3

The next section presents simulation results used to assess the performance of the induction motor parameter estimation algorithm and the proposed adaptive controller.

IV. SIMULATION RESULTS

MATLAB/Simulink simulations are carried out to evaluate the performance of the proposed parameter estimation algorithm, as well as that of the adaptive controller. The complete dynamic model of the saturated induction machine is implemented. The overall simulation scheme includes the induction machine model, the parameter estimation block, and the adaptive control block. The characteristics of the wound-rotor induction machine used for the simulations are given in the table below. The controller gains are also provided in the same table.

Table 1: Machine and controller nominal parameters

R_s	R_r	L_s	L_r	M
5.3 Ω	3.3 Ω	0.365 H	0.375 H	0.34 H
k_1	k_2	k_3	k_4	k
100	60	100	60	1

The following scenario is used to evaluate the performance of the identification algorithm as well as that of the adaptive control scheme. The stator and rotor resistances R_s and R_r generally vary with temperature. We therefore assumed that their increase (63.77 % for R_s and 68.03% for R_r) occurs simultaneously with the increase in the load torque up to its nominal value. At startup, the load torque is 2.5 Nm, and at t= 2.5s, it increases to 5.8 Nm. In the machine overspeed operating mode, flux weakening, that is, a reduction in the magnitude of the rotor flux, is required in order to limit the voltage of the PWM inverter DC bus. The application of this technique leads to a decrease in the stator, rotor, and mutual inductances due to magnetic saturation. In our test, the increase in inductances is 13.7% for L_s, 13.3% for L_r and 14.7% for M. At t=5s, the speed is increased to 10% above its nominal value, and the rotor flux is reduced by 21.2%. In Figure 1, the reference speed ω_{ref} increases from 104.5 rad/s to 115 rad/s between t = 5 s and t = 6.5 s. It can be observed that the rotational speed tracks the reference speed despite the parameter variations. At t=2.5s when the resistances R_s and R_r, vary, a slight increase in ω_r relative to ω_{ref} is observed, but it quickly returns to ω_{ref}. The controller is therefore robust to variations in R_r and R_s.

Also, at t= 5s, the inductances L_s, L_r and M vary. The rotational speed deviates slightly and then it returns to the reference speed. The speed control is also robust to inductance variations caused by magnetic flux saturation. In Figure 2, Ψ_{ref} is reduced from its nominal value of 1.16 Wb to 0.98 Wb in order to limit the inverter bus voltage while maintaining the maximum torque. This flux reduction implies a reduction in the inductances L_s, L_r and M. It can be observed that Ψ tracks its reference Ψ_{ref} with a steady-state error close to zero, even though the parameters L_s, L_r and M vary. The controller is therefore robust to variations in these parameters. Figures 3 and 4 illustrate the evolutions of the rotor and stator resistances, and of the rotor, stator, and mutual inductances, together with their estimates generated by the identification algorithm.

It can be observed that the resistances start increasing at t = 2.5 seconds and reach their constant values at t = 4 seconds. It is also observed that the estimated values track the actual values perfectly and almost instantaneously. This confirms the performance predicted in [9]. Figure 5 illustrates the magnitude of the stator voltage to be applied to the machine inverter. It can be observed that it remains within acceptable limits while adjusting to the variations imposed by the changes in the operating point. These results show that the use of estimated parameters and estimated fluxes in the controller design does not degrade its performance. The adaptive nature of the approach allows the control law to continuously adjust to the actual operating condition of the machine.

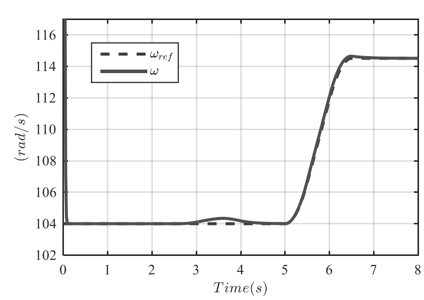

Figure 1: Reference and actual speeds

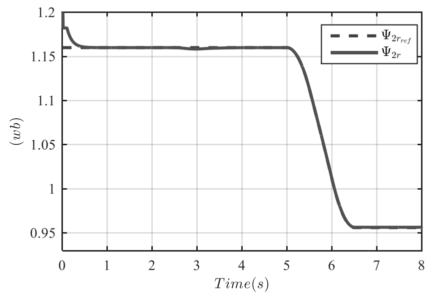

Figure 2: Reference and estimated rotor fluxes

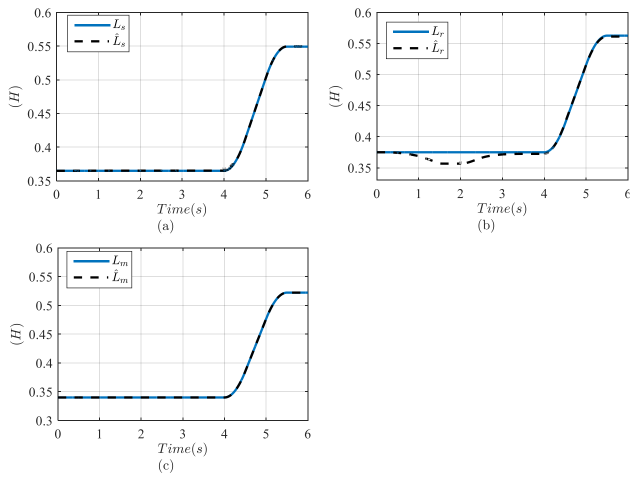

Figure 3: Actual and estimated values of Ls, Lr and M

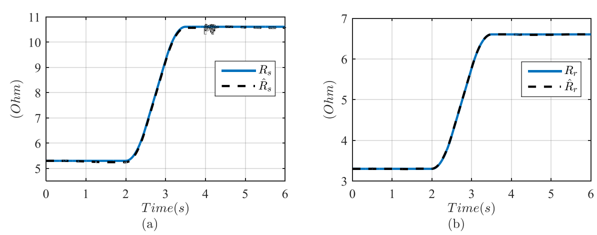

Figure 4 : Actual and estimated values of Rs and Rr

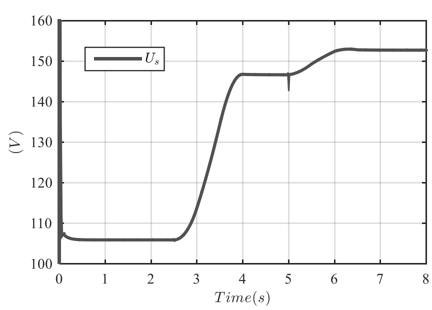

Figure 5: Magnitude of the stator voltage

V. Conclusion

This article has presented a unified approach for parameter identification and adaptive control of the induction machine operating under saturated conditions. Its parameters may vary significantly with the operating point. The first contribution is the derivation of an estimation algorithm directly based on the dynamic and algebraic equations of the machine. The second contribution is the synthesis of an adaptive controller that uses the estimated parameters and fluxes. A backstepping approach is used to derive the control laws for the flux and speed controllers. The originality of the proposed approach is the integration of online identification and adaptive control within a single framework suited to machine operation under magnetic saturation. The simulation results confirm the effectiveness of the method.

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