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# Nonlinear Integral Backstepping Control for Induction Motor drive with Adaptive Speed Observer using Super Twisting Strategy

(Full text in English)

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## Abstract

This paper presents a nonlinear Integral backstepping control approach applied to an induction motor fed by a power voltage source without speed sensor. We present the implementation of the integral backstepping control for induction motor drives based on the field oriented control technique. We also present the mechanism of adaptive super twisting speed and rotor flux observers with the only assumption that from stator voltages and currents are measurable. The objective is to improve the speed control, the rotor flux control under load torque disturbances and parameter variations. The overall stability is demonstrated using Lyapunov theory. Simulation results prove clearly high robustness against disturbances; the estimated flux and rotor speed converge to their actual values.

**Keywords:** field oriented control, induction motor drive, integral backstepping, lyapunov function, sensorless drive, super twisting

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## 1. Introduction

The induction motor (IM) plays an important role in industrial applications, due to simplicity of design and maintenance, robustness and low cost [1]. The model of IM is nonlinear, multivariable and greatly coupled have a fairly complex control. The progress in nonlinear control and the considerable technological advances, both in the field of power electronics and in microelectronics, made possible the implementation of high performance control of induction motor, making it a must in the areas of variable speed and fast torque control. This motor is a serious candidate for nonlinear control [2].

In this paper, we use a backstepping technique to synthesize a sensorless vector control of induction motor. The backstepping technique is a systematic and recursive method of synthesis of nonlinear control law based on Lyapunov functions that ensure the stabilization of each synthesis step [3]. At each step of the process, a virtual control is generated for ensuring the convergence of the system to its steady state [4]. This technique allows the synthesis of control laws taking into account disturbances or parameters mismatch [5]. The modification of Backstepping control by adding Integral action is an interesting

solution to improve the robustness of backstepping control and to eliminate residual errors [6].

The state variables used in the development of induction motor control (flux, speed ...) are often difficult to obtain for technical or economic reasons (difficulty or impossibility of measuring, fragility, size, cost and maintenance). Dedicated sensors represent the weak link in the chain, so we try to fulfil their role by estimation using electrical variables. Therefore, the mechanical speed sensor is completely replaced by software which are based on calculation algorithms [7].

During last years, various sensorless control schemes have been presented and are now used in many fields of electric drive. Estimators and observers can be classified into the following categories: Estimators in open loop, estimators based on harmonic analysis, MRAS techniques, nonlinear and adaptive observers. Estimators are sensitive to parameters variations. Using observers improves the robustness of the estimation in regard of parametric variations and measurement noises [8].

The proposed technique is based on the approach developed by H. Kubota and al [9] which uses an adaptive observer to estimate the motor speed. The contribution of this work is based on the replacement of the adaptive



mechanism from a conventional PI controller, to a super-twisting adaptive control algorithm. The aim is to boost the accuracy of the estimated speed and decrease the influence of the parameter uncertainties.

## 2. Mathematical Model of Induction Motor and Vector control

### 2.1. Model of IM in (d-q) Reference Frame

The mathematical model of the three-phase induction motor can be expressed in the (d-q) synchronous rotating frame by the following nonlinear equations [10] [11]:

$$\begin{aligned}\frac{di_{sd}}{dt} &= a_1 i_{sd} + \omega_s i_{sq} + a_2 \phi_{rd} \\ &\quad - a_3 \omega_r \phi_{rq} + b V_{sd} \\ \frac{di_{sq}}{dt} &= a_1 i_{sq} - \omega_s i_{sd} + a_3 \phi_{rd} \omega_r \\ &\quad + a_2 \phi_{rq} + b V_{sq} \\ \frac{d\phi_{rd}}{dt} &= a_4 i_{sd} - a_5 \phi_{rd} + (\omega_s - \omega_r) \phi_{rq} \\ \frac{d\phi_{rq}}{dt} &= a_4 i_{sq} - a_5 \phi_{rq} - (\omega_s - \omega_r) \phi_{rd} \\ \frac{d\omega_r}{dt} &= a_6 (i_{sq} \phi_{rd} - i_{sd} \phi_{rq}) - a_7 \omega_r - a_8 T_l\end{aligned}\quad (1)$$

with:

$$\begin{aligned}a_1 &= -b \left( R_s + \left( \frac{L_m}{L_r} \right)^2 R_r \right), \quad a_2 = \frac{L_m R_r}{(\sigma L_s L_r^2)} \\ a_3 &= -\frac{L_m}{(\sigma L_s L_r)}, \quad a_4 = \frac{L_m R_r}{L_r}, \quad a_5 = \frac{R_r}{L_r}, \quad a_6 = \frac{p^2 L_m}{J L_r} \\ a_7 &= \frac{f_c}{J}, \quad a_8 = \frac{p}{J}, \quad \sigma = 1 - \left( \frac{L_m^2}{L_s L_r} \right), \quad b = \frac{1}{\sigma L_s}\end{aligned}$$

### 2.2. Direct field Oriented Control for Induction Motor

The field oriented control strategy is very efficient to get high-performances for induction motor drive in terms of torque and speed. The principle of this technique is to make the behaviour of the induction motor similar to that of a DC motor, for which a natural decoupling exists between the control of the torque and the magnetic flux. This is done by correctly computing the dq reference frame to cancel the quadrature component of the flux. We must find the right Park's angle so that the rotor flux vector is entirely on the d-axis, the decoupling is obtained by imposing

the following condition [12] [13]:

$$\phi_r = \phi_{rd} = L_m \cdot i_{sd} \text{ and } \phi_{rq} = 0 \quad (2)$$

## 3. Integral Backstepping control

### 3.1. Application principle

The basic idea of backstepping control is to make closed loop systems equivalent to first-order subsystems in cascade that are Lyapunov stable. This gives ensures robustness and asymptotic global stability.

Our aim to control the flux and speed variables. To let them follow particular trajectories, we chose as intermediate variables, the stator currents. They must follow their reference values defined by the "virtual control". Finally, we compute the voltage controls (stator voltages) required to let the "virtual controls" converge to the desired values with regards to the stability of the associated Lyapunov function [14] [15] [16].

### 3.2. Application of conventional backstepping control to induction motor

In our work, we provide Integral backstepping control by using the model of the leakage oriented induction machine equation (1) [17].

Our objective is to synthesize the expression of the control variables  $V_{sd}$  and  $V_{sq}$  to let the state variables of the induction motor follow the desired references. By making a variables change, the new ones are the errors between the set points and the state variables. This control is presented in four steps [6], as we will show in this section.

The motor has to reach the desired speed under a rated magnetic flux  $\phi_{rd}$ . We must first compute speed and rotor flux errors.

#### a) First step "speed loop"

For the electrical speed  $\omega_r$ , we define the tracking error as:

$$e_1 = \omega_r^* - \omega_r \quad (3)$$

And their dynamics are given by:

$$\dot{e}_1 = \dot{\omega}_r^* - \dot{\omega}_r \quad (4)$$

Then the error dynamical equations are:

$$\dot{e}_1 = \omega_r^* - a_6 i_{qs} \phi_{rd} + a_7 \omega_m + a_8 T_l$$

The first Lyapunov function is chosen such that:



$$v_1(e_1) = \frac{1}{2} e_1^2 \quad (5)$$

Its derivative is:

$$\dot{v}_1(e_1) = e_1 \cdot \dot{e}_1 \quad (6)$$

$$\dot{v}_1(e_1) = e_1 (\dot{\omega}_r^* - a_6 \cdot i_{sq} \cdot \phi_{rd} + a_7 \cdot \omega_r + a_8 \cdot T_l)$$

The objectives of tracking are performed ( $\dot{v}_1(e_1) < 0$ ) choosing the quadrature current as virtual control which represents the stabilizing function as follows:

$$i_{sq}^* = \frac{1}{\phi_{rd} \cdot a_6} (k_1 e_1 + \omega_r^* + a_7 \cdot \omega_r + a_8 \cdot T_l) \quad (7)$$

with  $k_1$  is a positive constant.

The derivative of the Lyapunov function becomes:

$$\dot{v}_1(e_1) = -k_1 e_1^2 < 0$$

Since the current  $i_{sq}$  is not a control input and is only one variable of the system with its own dynamics. We will use it to introduce the integral action, so we choose the desired dynamic behaviour of the virtual control  $i_{sq}^*$  as following:

$$i_{sq}^* = \frac{1}{\phi_{rd} \cdot a_6} (k_1 e_1 + \omega_r^* + a_7 \cdot \omega_r + a_8 \cdot T_l) + \lambda_1 \chi_1 \quad (8)$$

With  $\lambda_1$  is positive constant and  $\chi_1 = \int_0^t e_1(\tau) d\tau$  is the integral action brought in accordance with the following error "e". By introducing this integral in the virtual control, we ensure the convergence of the tracking error to zero in steady state.

#### b) Second step " $i_{sq}$ current loop"

For this step, our goal is the replacement of the virtual current control by computing control voltages.

We define the error of the  $q$ -axis component of the stator current and its reference.

$$e_2 = i_{sq}^* + \lambda_1 \cdot \chi_1 - i_{sq} \quad (9)$$

The derivative is written as:

$$\dot{e}_2 = \dot{i}_{sq}^* + \lambda_1 \cdot e_1 - \dot{i}_{sq} \quad (10)$$

$$\dot{e}_2 = \dot{i}_{qs}^* - \delta_1 - \frac{1}{\sigma L_s} v_{sq}$$

with

$$\delta_1 = -a_1 i_{sq} - \omega_s i_{sd} - a_3 \phi_{rd} \omega_r$$

$$v_2(e_2) = \frac{1}{2} e_2^2 \quad (11)$$

Its derivative is:

$$\dot{v}_2(e_2) = e_2 \cdot \dot{e}_2 \quad (12)$$

$$\dot{v}_2(e_2) = -e_2 \cdot \left( \dot{i}_{qs}^* - \delta_1 - \frac{1}{\sigma L_s} v_{sq} \right)$$

For a negative error derivative  $e_2$ , we must choose  $v_{sq} = \sigma L_s (k_2 e_2 + \dot{i}_{sq}^* - \delta_1)$  as the first control voltage.

$$\dot{v}_2(e_2) = -k_2 e_2^2 < 0$$

with  $k_2$  is a positive constant.

#### c) Third step "flux loop"

We define the tracking error as:

$$e_3 = \phi_{rd}^* - \phi_{rd} \quad (13)$$

And its derivative with respect to time leads to:

$$\dot{e}_3 = \dot{\phi}_{rd}^* - \dot{\phi}_{rd} \quad (14)$$

$$\dot{e}_3 = \dot{\phi}_{rd}^* + a_5 \cdot \phi_{rd} - a_4 \cdot i_{sd}$$

The third Lyapunov function is chosen such that:

$$v_3(e_3) = \frac{1}{2} (e_3^2) \quad (15)$$

Its derivative is:

$$\dot{v}_3(e_3) = e_3 \cdot \dot{e}_3 \quad (16)$$

$$\dot{v}_3(e_3) = e_3 (\dot{\phi}_{rd}^* + a_5 \cdot \phi_{rd} - a_4 \cdot i_{sd})$$

The tracking objectives are achieved ( $\dot{v}_3(e_3) < 0$ ) by choosing the reference of the current component as follows:

$$i_{sd}^* = \frac{1}{a_4} (k_3 e_3 + \dot{\phi}_{re}^* + a_5 \phi_{rd}) \quad (17)$$

with  $k_3$  is constant positive.



The derivative of the Lyapunov function becomes:

$$\dot{v}_3(e_3) = -k_3 e_3^2 < 0$$

The desired dynamic behaviour of the virtual control with integral action  $i_{sd}^*$  is given by:

$$i_{sd}^* = \frac{1}{a_4} (k_3 e_3 + \dot{\phi}_{re}^* + a_5 \phi_{rd}) + \lambda_2 \chi_2 \quad (18)$$

with  $\lambda_2$  is positive constant,

and  $\chi_2 = \int_0^t e_3(\tau) d\tau$

d) Fourth step “isd current loop”

$$e_4 = i_{sd}^* + \lambda_2 \cdot \chi_2 - i_{sd} \quad (19)$$

$$e_4 = i_{sd}^* + \lambda_2 \cdot e_3 - i_{sd}$$

The derivative of the error variable is:

$$\dot{e}_4 = \dot{i}_{sd}^* - \dot{i}_{sd} \quad (20)$$

$$\dot{e}_4 = \dot{i}_{sd}^* - \delta_2 - \frac{1}{\sigma L_s} V_{sd}$$

with

$$\delta_2 = -a_1 i_{sd} + \omega_s i_{sq} + a_2 \phi_{rd}$$

$$\dot{v}_4(e_2) = -e_4 \cdot \left( \dot{i}_{sd}^* - \delta_2 - \frac{1}{\sigma L_s} V_{sd} \right) \quad (21)$$

$$\dot{v}_4(e_4) = e_4 \cdot \dot{e}_4$$

For the error derivative  $e_4$  to be negative we must choose  $v_{sd} = \sigma L_s (k_4 e_4 + \dot{i}_{sd}^* - \delta_2)$  as the second control voltage.

$$\dot{v}_4(e_4) = -k_4 e_4^2 < 0$$

with  $k_4$  is a positive constant.

This leads to stabilize the system by the increased Lyapunov function of energy:

$$v_e = \frac{1}{2} [e_1^2 + e_2^2 + e_3^2 + e_4^2] \quad (22)$$

Its derivative is:

$$\dot{v}_e = -k_1 e_1^2 - k_2 e_2^2 - k_3 e_3^2 - k_4 e_4^2 \quad (23)$$

which leads to:

$$-k_1 e_1^2 - k_2 e_2^2 - k_3 e_3^2 - k_4 e_4^2 < 0 \quad (24)$$

For the proposed sensorless vector control, the rotor flux and position are determined by the following relationships:

$$\hat{\theta}_s = \arctg \left( \frac{\hat{\phi}_{r\beta}}{\hat{\phi}_{r\alpha}} \right) \quad (25)$$

$$\hat{\phi}_r = \sqrt{\hat{\phi}_{r\alpha}^2 + \hat{\phi}_{r\beta}^2} \quad (26)$$

The Integral backstepping control for induction motor with the proposed observer can be presented by the block diagram shown in Figure 1.

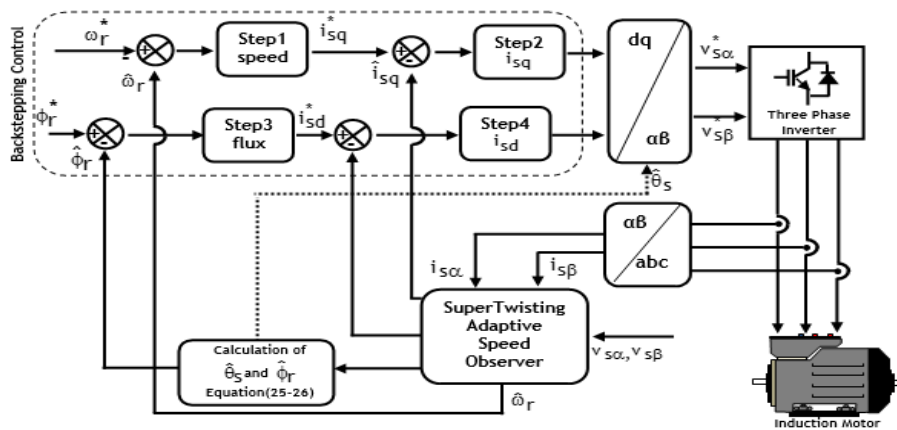


Figure 1. Block diagram of backstepping control scheme based induction motor drive

#### 4. Speed and flux estimation by the Adaptive Observer

In this section, the conventional adaptive observer is presented. Its goal is the

reconstruction of the rotor flux and speed without a mechanical sensor.

The rotational speed of the induction motor is not measured and is considered as an unknown parameter in the observer equation system.



#### 4.1. Model of observer

The mathematical model of induction motor, expressed in the  $(\alpha - \beta)$  stationary reference frame allows having a good compromise between stability and simplicity of observer, can be described by the following state equation [18]:

$$\begin{cases} \dot{X} = A \cdot X + B \cdot U \\ Y = C \cdot X \end{cases} \quad (27)$$

where

$$X = [i_{s\alpha} \ i_{s\beta} \ \phi_{r\alpha} \ \phi_{r\beta}]^T$$

$$U = v_s = [v_{s\alpha} \ v_{s\beta}]^T, \quad Y = i_s = [i_{s\alpha} \ i_{s\beta}]^T$$

The state equations can be rewritten as:

$$A = \begin{bmatrix} a_1 & 0 & a_2 & a_3 \cdot \omega_r \\ 0 & a_1 & -a_3 \cdot \omega_r & a_2 \\ a_4 & 0 & a_5 & -\omega_r \\ 0 & a_4 & \omega_r & a_5 \end{bmatrix} \quad (28)$$

$$B = \begin{bmatrix} a_6 & 0 \\ 0 & a_6 \\ 0 & 0 \\ 0 & 0 \end{bmatrix}, \quad C = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \quad (29)$$

The coefficients  $a_1, a_2, a_3, a_4, a_5$  and  $a_6$  have the same expressions as those defined in equation (1).

The state observer, which estimates the stator current and the rotor flux is written as follows [9]:

$$\hat{\dot{X}} = \hat{A} \cdot \hat{X} + B \cdot U + G(\hat{i}_s - i_s) \quad (30)$$

with

$$i_s = [i_{s\alpha} \ i_{s\beta}]^T \text{ and } \hat{i}_s = [\hat{i}_{s\alpha} \ \hat{i}_{s\beta}]^T$$

Are respectively the vectors of the measured and the estimated stator currents.  $G$  is the observer gain matrix which imposes its dynamics. It is calculated as follows [9-19]:

$$G = \begin{bmatrix} g_1 & g_2 & g_3 & g_4 \\ -g_2 & g_1 & -g_4 & g_3 \end{bmatrix} \quad (31)$$

By naming  $e = X - \hat{X}$ , the dynamics of the estimation error of the adaptive observer is obtained by:

$$\begin{aligned} \frac{d}{dt}e &= \frac{d}{dt}(X - \hat{X}) = (A + G \cdot C) \cdot (X - \hat{X}) + (A - \hat{A})\hat{X} \\ &= (A + G \cdot C) \cdot e - \Delta A \cdot \hat{X} \end{aligned} \quad (32)$$

with

$$\Delta A = \hat{A} - A$$

$$\Delta A = \hat{A} - A = \begin{bmatrix} 0 & 0 & 0 & c \cdot \Delta \omega_r \\ 0 & 0 & -c \cdot \Delta \omega_r & 0 \\ 0 & 0 & 0 & -\Delta \omega_r \\ 0 & 0 & \Delta \omega_r & 0 \end{bmatrix}$$

$$c = L_m / \sigma \cdot L_s \cdot L_r$$

$$\Delta \omega_r = \hat{\omega}_r - \omega_r : \text{Error on speed.}$$

The speed adaptation law is given by:

$$\frac{1}{\lambda} \cdot \frac{d}{dt} \Delta \omega_r = e_{is\alpha} \cdot \hat{\phi}_{r\beta} - e_{is\beta} \cdot \hat{\phi}_{r\alpha} / c \quad (31)$$

$$\hat{\omega}_r = \int \frac{\lambda}{c} (e_{is\alpha} \cdot \hat{\phi}_{r\beta} - e_{is\beta} \cdot \hat{\phi}_{r\alpha}) dt \quad (33)$$

$$\hat{\omega}_r = \int K_w (e_{is\alpha} \cdot \hat{\phi}_{r\beta} - e_{is\beta} \cdot \hat{\phi}_{r\alpha}) dt \quad (34)$$

with

$$K_w = \lambda \cdot \frac{1 - \sigma}{\sigma \cdot L_m}$$

This adaptation law has been established for almost constant speed, to improve the dynamic of speed observation, we replaced the pure integral by a PI controller as follows [20]:

$$\begin{aligned} \hat{\omega}_r &= K_p \cdot (e_{is\alpha} \cdot \hat{\phi}_{r\beta} - e_{is\beta} \cdot \hat{\phi}_{r\alpha}) \cdot dt \\ &\quad + K_i \int (e_{is\alpha} \cdot \hat{\phi}_{r\beta} - e_{is\beta} \cdot \hat{\phi}_{r\alpha}) \cdot dt \end{aligned} \quad (35)$$

with

$$e_{is\alpha} = i_{s\alpha} - \hat{i}_{s\alpha} \text{ And } e_{s\beta} = i_{s\beta} - \hat{i}_{s\beta}$$

$K_p$  And  $K_i$  are the proportional and integral gains respectively.

The matrix of the adjustable gain  $G$  of the observer will ensure the stability and the dynamics of desired observation.

The coefficient  $k$  is chosen so that the observer will responds faster than the system.

$$g_1 = (1 - k) \cdot \left( \frac{R_s}{\sigma \cdot L_s} + \frac{1 - \sigma}{\sigma \cdot T_r} + c \cdot \frac{1}{T_r} \right)$$

$$g_2 = (k - 1) \cdot \hat{\omega}_r$$



$$g_3 = \frac{(k^2 - 1)}{c} \left( \frac{R_s}{\sigma L_s} + \frac{1 - \sigma}{\sigma T_r} + c \cdot \frac{L_m}{T_r} \right) + \frac{(1 - k)}{c} \left( \frac{R_s}{\sigma L_s} + \frac{1 - \sigma}{\sigma T_r} + c \cdot \frac{1}{T_r} \right)$$

$$g_4 = \left( \frac{k - 1}{c} \right) \cdot \hat{\omega}_r$$

The structure of the adaptive observer is shown in Figure 2.

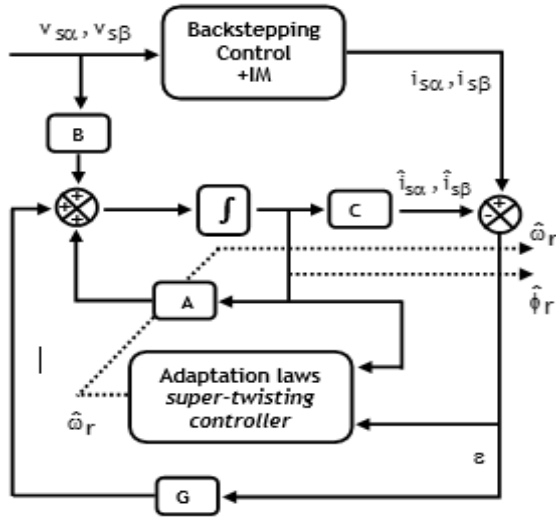
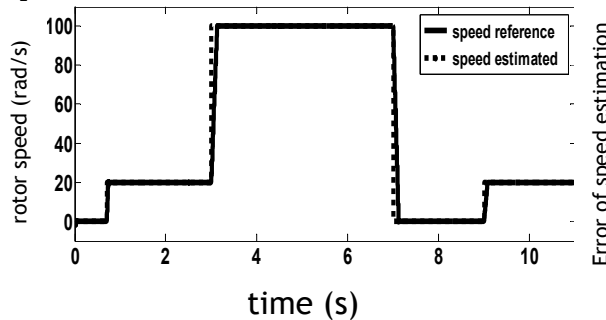


Figure 2. Block diagram of adaptive observer

## 5. Adaptive Super-Twisting Observer Design

The enhancement provided in this section consists in the replacement of the conventional PI controller by a super twisting control algorithm. The super twisting sliding mode algorithms are excellent controllers to solve problems of speed estimation and have the ability to compensate the uncertainties of the mathematical model [21] [22].

The super twisting algorithms have been developed in the context of higher order sliding mode control theory. Now, consider the super twisting control algorithm, which is given by [23]:



$$u = -\lambda_p |s|^r \operatorname{sgn}(s) + v$$

$$\dot{v} = -\lambda_i \operatorname{sgn}(s)$$
(36)

Where  $u$  represents the control signal,  $\lambda_i$  and  $\lambda_p$  are the control gains,  $s$  is a sliding variable,  $r$  is a positive constant,  $0 < r \leq 0.5$ .

The adaptation laws based the super twisting Control Algorithm are obtained for the estimation of speed can be designed as (34).

$$\hat{\omega}_r = \lambda_p |s|^r \operatorname{sgn}(s) + v$$

$$\dot{v} = -\lambda_i \operatorname{sgn}(s)$$
(37)

where the sliding variable is  $s = e_{is\alpha} \cdot \hat{\phi}_{r\beta} - e_{is\beta} \cdot \hat{\phi}_{r\alpha}$ , and gains  $\lambda_i$  and  $\lambda_p$  are positive constants.

A block diagram of the speed estimation control with super twisting sliding mode is shown in Figure 3.

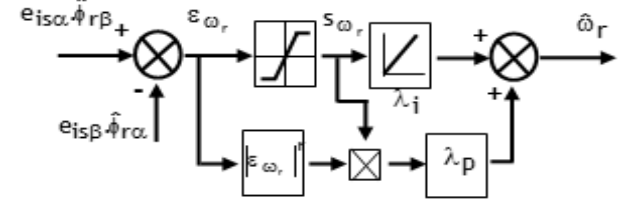
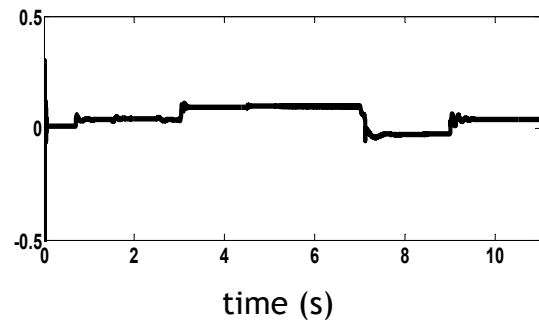
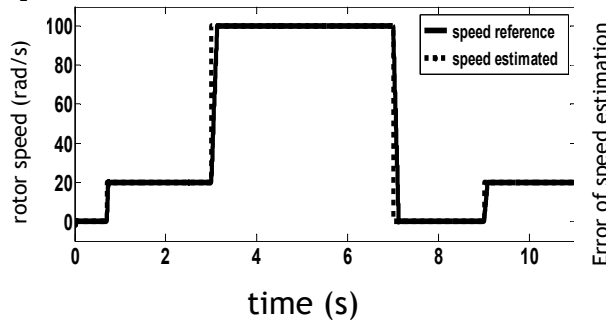


Figure 3. Block diagram of control loop of speed estimation with STSM regulator

## 5. Simulation results

The results exposed below were obtained by numerical simulations, performed for the nonlinear integral backstepping control without Speed Sensor associated with the adaptive observer applied to the induction motor drive for the values of the following coefficients:  $K_1 = K_2 = 500$ ,  $K_3 = K_4 = 1800$ ,  $\lambda_i = 10000$ ,  $\lambda_p = 10200$ ,  $k=0.96$ . These coefficients are obtained by try-error procedure.





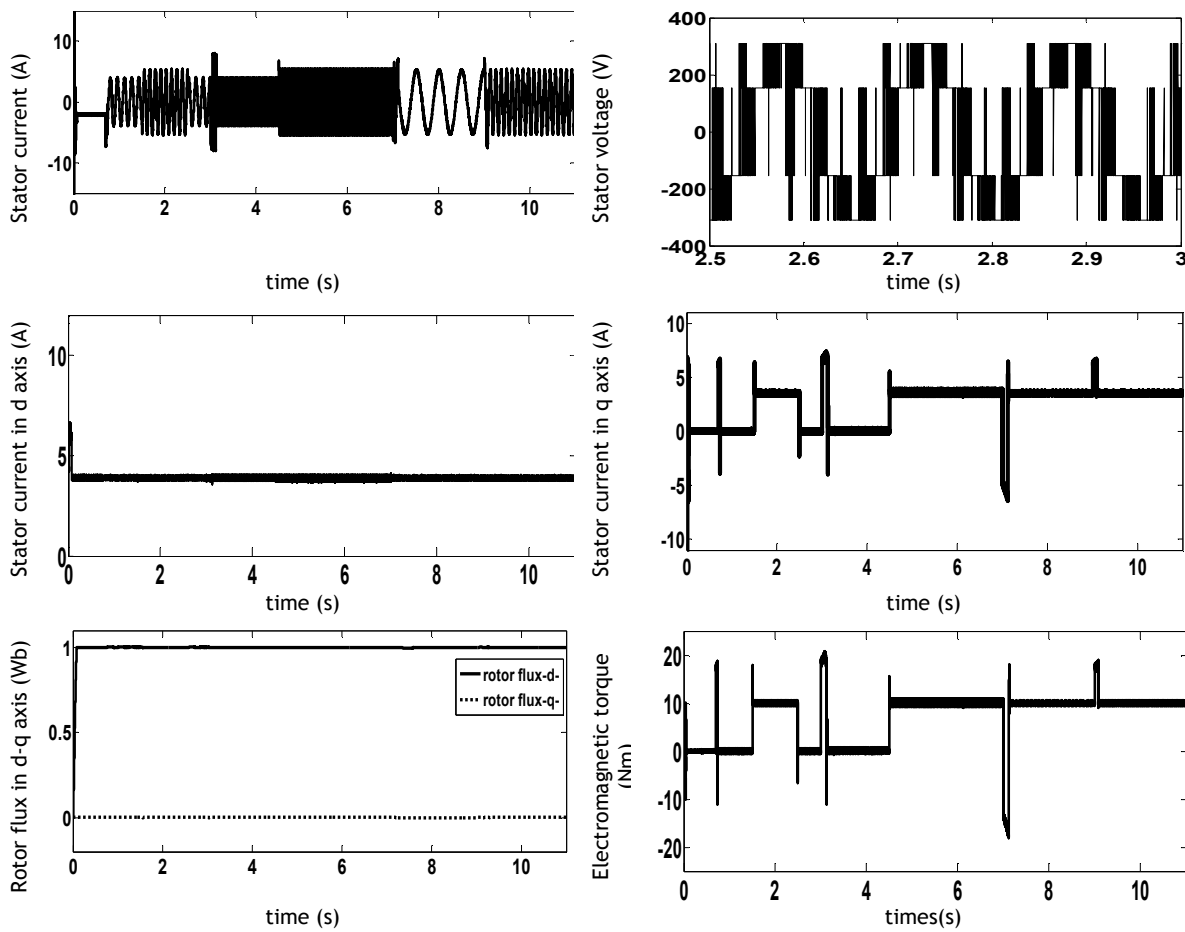


Figure 4. Performance of the proposed observer at high and low speed region with two load variations

These figures describe the high performance of the closed loop system (observer and controller) in terms of trajectory tracking and disturbance rejection. The speed of the machine correctly follows its reference even under conditions where the machine is unobservable (between 7 and 9 sec) [24], The disturbance rejection is very satisfactory as low speed or high speed. Nevertheless, there is a small difference at application times of load torque steps. The results also show a good flux control.

## 6. Conclusions

This paper presents the application of nonlinear integral backstepping technique for sensorless vector control of induction motor. The proposed method is to use an adaptive observer based on a super twisting control strategy for the estimation of the speed and the rotor flux by using Lyapunov theory to show the system stability. The speed and rotor flux are estimated only with sensing electrical variables.

Analysis of the results confirms the superiority of the super twisting approach that allows a sensorless control with a high level of performance not only in terms of precision and robustness of the estimation, but also in terms of reference tracking and disturbance rejection.

## 7. Nomenclature

$T_{em}, T_L$  : Electromagnetic torque, load torque.

$i_{sd}, i_{sq}$  : Stator currents.

$\phi_{rd}, \phi_{rq}$  : Rotor flux.

$\omega_m$  : Electrical angular.

$\sigma$  : Total leakage factor.

$R_s$  and  $R_r$  : Stator and rotor resistances.

$L_s$  and  $L_r$  : Stator and rotor inductances.

$L_m$  : Mutual inductance.

$p$  : Number of pole pairs.

$J$  : Moment of inertia.

$f_c$  : Friction coefficient.



## 8. Appendix

Table 1. Nominal parameters of the induction motor used:

Rated power	1.5kW
Rated voltage	220V
Rated speed	1428rpm
Nominal frequency	50Hz
Rated stator current	3.64
Stator resistance	4.85Ω
Rotor resistance	3.805Ω
Cyclic stator inductance	0.274H
Cyclic rotor inductance	0.274H
Mutual inductance	0.258H
Number of pole pairs	2
Moment of inertia	0.031kg/m <sup>2</sup>
Friction coefficient	0.00114Nm.s/rd

## 9. Acknowledgment

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