Nonlinear Integral Backstepping Control for Induction Motors

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***Abstract-* A novel structure of nonlinear integral backstepping control has been proposed for induction motors. We can see clearly that the structure of the controller generated by the classical version of backstepping is composed of a proportional action, which was added a derivative action on errors. Such a structure makes the system sensitive to measurement noise. The lack of integration means also the appearance of a constant static error, caused mainly by non-zero mean disturbances. The solution to this problem is to design a new version of the backstepping with integral action.**

By using the proposed integral backstepping controller, the system has a better load disturbance rejection capability. The effectiveness of this proposed control structure is verified by simulation as well as by experiment under critical disturbance conditions.

I. Introduction

Control of induction motors (IM) can become very complex depending on desired performance. This complexity is due mainly to the following reasons: the analytical model of induction machines is nonlinear, multi-variable and highly coupled, presence of parametric uncertainties and the need to take into account their variation in time.

The first control architectures of induction motors were based on traditional scalar control and can guarantee only modest performance. In many applications, it is necessary to use more sophisticated controls, consistent with the expected performance but more complex.

With technological advances in power electronics and digital electronics fields, it became possible to design the real implementation for control algorithms, whatever their degree of complexity and execution time.

Subsequently, several nonlinear control techniques have been studied to control the induction motors. They were developed to replace vector control, while providing both a separate control of flux and torque and a good robustness against the parametric variations. Among the techniques applied to induction motor control: input-output feedback linearization [1], passivity-based control [2], [3] and sliding mode control [4], [5].

Since last two decades, the nonlinear control called "backstepping" became one of the most popular control techniques for a wide range of nonlinear systems classes [6-10]. It is distinguished by its ability to easily guarantee the global stabilization of system, even in the presence of parametric uncertainties [11]. The design of the control law is based mainly on the construction of Lyapunov functions associated.

The efficient Backstepping control scheme for induction motor control is in association with indirect field oriented control. It can be designed in many versions and applied to many IM models. We can use the classical mathematical model [12, 13], or the multi-scalar one [14]. The second can be used in order to reduce the complexity of the first one. This simplifies the IM model and the adaptation of backstepping controller becomes easier.

In this paper, a new version structure of integral backstepping controller is used for speed, flux and current regulators in indirect field oriented control (IFOC). The integral action allows successfully rejection of disturbances. The overall aspect of the structure is maintained exactly as in the conventional IFOC like speed control loop, flux control loop and two current control loops. Flux and load torque estimators are also provided to fulfill the requirement of the backstepping controller.

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II. Model Description And Classical Control Design

Under the assumptions of linearity of the magnetic circuit and neglecting iron losses, the fifth-order nonlinear model of IM, after field oriented transformation is expressed in the rotate stator frame *(d,q)* as [15]:

$$\begin{aligned}
&\frac{d\omega}{dt}=\mu\lambda_{2d}i_{1q}-\frac{T_L}{J}\\
&\frac{di_{1q}}{dt}=-\eta_1 i_{1q}-\beta n_p\omega\lambda_{2d}-n_p\omega i_{1d}-R_2\left(\eta_2 I_{1q}+\alpha L_m\frac{i_{1q}\,i_{1d}}{\lambda_{2d}}\right)\\
&\qquad+\frac{1}{\sigma L_1}u_{1q}\\
&\frac{d\lambda_{2d}}{dt}=-\alpha R_2\lambda_{2d}+\alpha L_m R_2 i_{1d}\\
&\frac{di_{1d}}{dt}=-\eta_1 i_{1d}+n_p\omega i_{1q}+R_2\left(-\eta_2 I_{1d}+\alpha\beta\lambda_{2d}+\alpha L_m\frac{i_{1q}^2}{\lambda_{2d}}\right)\\
&\qquad+\frac{1}{\sigma L_1}u_{1d}\\
&\frac{d\rho}{dt}=n_p\omega+\alpha L_m R_2\frac{I_{1q}}{\lambda_{2d}}
\end{aligned}\tag{1}$$

With

$$\alpha=\frac{1}{L_2},\quad \beta=\frac{L_m}{\sigma L_1 L_2},\quad \eta_1=\frac{R_1}{\sigma L_1},\quad \eta_2=\frac{L_m^2}{\sigma L_1 L_2^2},\quad \mu=\frac{3n_pL_m}{2JL_2}$$

Where,

$$\sigma=1-\frac{L_m^2}{L_1L_2}$$

i_{1d}, i_{1q} the stator current in *(d,q)* frame

$\lambda_{2d}, \lambda_{2q}$ the rotor flux linkage in *(d,q)* frame

u_{1d}, u_{1q} the stator voltage input in *(d,q)* frame

ω the rotor angular speed

ρ the angle between the rotor flux linkages

n_p the number of pole pairs

L_m the mutual inductance

L_1, L_2 the stator and rotor inductances respectively

R_1, R_2 the stator and rotor resistances respectively

J the moment of inertia

T_L the applied load torque

The basic idea of the backstepping design is to chose recursively some appropriate functions of state variables as "virtual control" inputs for single-input-single-output subsystems of the overall system. Then, backstepping design is divided into various design steps. In each step, an extended Lyapunov function is associated to achieve the stability of the whole system [12].

Step 1

As the rotor angular speed and the rotor flux amplitude are the tracking objectives, we define first the tracking errors as

$$\begin{aligned}
&e_1=\omega_{ref}-\omega\\
&e_3=\lambda_{ref}-\lambda_{2d}
\end{aligned}\tag{2}$$

Then the error dynamical equations are

$$\begin{aligned}
&\dot e_1=\dot\omega_{ref}-\mu\lambda_{2d}i_{1q}+\frac{T_L}{J}\\
&\dot e_3=\dot\lambda_{ref}+\alpha R_2\lambda_{2d}-\alpha L_m R_2 i_{1d}
\end{aligned}\tag{3}$$

Since our objectives require that the two errors converge to zero, and require also that the current must be regulated and limited, we could satisfy this two objectives by viewing i_{1q} and i_{1d} as "virtual control" in the above equations and use them to control e_1, e_3. We use the Lyapunov function

$$V=\frac{1}{2}e_1^2+\frac{1}{2}e_3^2\tag{4}$$

To derive stabilizing feedback virtual controls i_{1q} and i_{1d} as follows. The derivatives of V along the error equations (3) is

$$\begin{aligned}
\dot V&=e_1\dot e_1+e_3\dot e_3\\
&=e_1\left[\dot\omega_{ref}-\mu\lambda_{2d}i_{1q}+\frac{T_L}{J}\right]\\
&\quad+e_3\left[\dot\lambda_{ref}+\alpha R_2\lambda_{2d}-\alpha L_m R_2 i_{1d}\right]\\
&=-k_1e_1^2-k_3e_3^2+e_1\left[k_1e_1+\dot\omega_{ref}-\mu\lambda_{2d}i_{1q}+\frac{T_L}{J}\right]\\
&\quad+e_3\left[k_3e_3+\left[\dot\lambda_{ref}+\alpha R_2\lambda_{2d}-\alpha L_m R_2 i_{1d}\right]\right]
\end{aligned}\tag{5}$$

Where k_1, k_3 are positive design constants that determine the closed-loop dynamics. If the stabilizing virtual controls are chosen as

$$\begin{aligned}
&(i_{1q})_{ref}=\frac{1}{\mu\lambda_{2d}}\left[k_1e_1+\dot\omega_{ref}+\frac{T_L}{J}\right]\\
&(i_{1d})_{ref}=\frac{1}{\alpha L_m R_2}\left[k_3e_3+\dot\lambda_{ref}+\alpha R_2\lambda_{2d}\right]
\end{aligned}\tag{6}$$

We obtain

$$\dot V=-k_1e_1^2-k_3e_3^2\le 0\tag{7}$$

Then, the virtual controls in (6) are chosen to satisfy the control objectives and also provide references for the next step of the backstepping design.

Step 2

Now, the new tracking objectives are the signals current i_{1q} and i_{1d}. So we define again error signals involving the desired variables in (6)

$$e_2 = (i_{1q})_{ref} - i_{1q} = \frac{1}{\mu\lambda_{2d}}\left[k_1 e_1 + \dot{\omega}_{ref} + \frac{T_L}{J}\right] - i_{1q}$$
$$e_4 = (i_{1d})_{ref} - i_{1d} = \frac{1}{\alpha L_m R_2}\left[k_3 e_3 + \dot{\lambda}_{ref} + \alpha R_2 \lambda_{2d}\right] - i_{1d} \quad (8)$$

Then the error equations (3) can be expressed as

$$\dot{e}_1 = -k_1 e_1 + \mu\lambda_{2d} e_2$$
$$\dot{e}_3 = -k_3 e_3 + \alpha L_m R_2 e_4 \quad (9)$$

Also, the dynamical equations for the error signals e_2, e_4 can be computed as

$$\dot{e}_2 = \phi_1 - \frac{1}{\sigma L_1} u_{1q}$$
$$\dot{e}_4 = \phi_2 - \frac{1}{\sigma L_1} u_{1d} \quad (10)$$

Where $\phi_i' s(i = 1,2)$ are known signals that can be used in the control, and their expressions are as follows

$$\phi_1 = \frac{k_1}{\mu\lambda_{2d}}(-k_1 e_1 + \mu\lambda_{2d} e_2) - \frac{\alpha R_2}{\mu\lambda_{2d}^2}(L_m i_{1d} - \lambda_{2d})$$
$$(k_1 e_1 + \dot{\omega}_{ref} + \frac{T_L}{J}) + \frac{1}{\mu\lambda_{2d}}\ddot{\omega}_{ref} + \eta_1 i_{1q} + \beta n_p \omega\lambda_{2d}$$
$$+ n_p \omega i_{1d} + R_2(\eta_2 i_{1q} + \alpha L_m \frac{i_{1q} i_{1d}}{\lambda_{2d}})$$
$$\phi_2 = \frac{k_3}{\alpha L_m R_2}(-k_3 e_3 + \alpha L_m e_4) + \frac{1}{\alpha L_m R_2}\ddot{\lambda}_{ref}$$
$$-\frac{\alpha R_2}{L_m}(\lambda_{2d} - L_m i_{1d}) + \eta_1 i_{1d} - n_p \omega i_{1q}$$
$$-R_2\left(-\eta_2 i_{1d} + \alpha\beta\lambda_{2d} + \alpha L_m \frac{i_{1q}^2}{\lambda_{2d}}\right) \quad (11)$$

Now we extend the Lyapunov function in (4) to include the state-variables e_2, e_4 as

$$V_e = \frac{1}{2}\left[e_1^2 + e_2^2 + e_3^2 + e_4^2\right] \quad (12)$$

We use V_e to derive both the control algorithm. To this end, we again compute the derivate of V_e along with the error equations (9), (10)

$$\dot{V}_e = e_1\dot{e}_1 + e_2\dot{e}_2 + e_3\dot{e}_3 + e_4\dot{e}_4$$
$$= e_1(-k_1 e_1 + \mu\lambda_{2d} e_2)$$
$$+ e_2\left(\phi_1 - \frac{1}{\sigma L_1} u_{1q}\right)$$
$$+ e_3(-k_3 e_3 + \alpha L_m R_2 e_4)$$
$$+ e_4\left(\phi_2 - \frac{1}{\sigma L_1} u_{1d}\right)$$
$$= -k_1 e_1^2 - k_2 e_2^2 - k_3 e_3^2 - k_4 e_4^2$$
$$+ e_2\left[\mu\lambda_{2d} e_1 + k_2 e_2 + \phi_1 - \frac{1}{\sigma L_1} u_{1q}\right]$$
$$+ e_4\left[\alpha L_m R_2 e_3 + k_4 e_4 + \phi_2 - \frac{1}{\sigma L_1} u_{1d}\right] \quad (13)$$

Where k_2, k_4 are positive design constants that determine the closed-loop dynamics.
From the above we can obtain the control laws as

$$u_{1q} = \sigma L_1[\mu\lambda_{2d} e_1 + k_2 e_2 + \phi_1]$$
$$u_{1d} = \sigma L_1[\alpha L_m R_2 e_3 + k_4 e_4 + \phi_2] \quad (14)$$

Which leads to

$$\dot{V}_e = -k_1 e_1^2 - k_2 e_2^2 - k_3 e_3^2 - k_4 e_4^2 \le 0 \quad (15)$$

III. Integral Backstepping Design

The introduction of integrators in the model will increase the model of two states. We start by drift once equations (2) and (4) of the model (1) and introducing new state variables i_q and i_d, we obtain the new augmented model [16, 17]:

$$\frac{d\omega}{dt} = \mu\phi_{rd} i_{sq} - \frac{T_L}{J}$$
$$\frac{d\phi_{rd}}{dt} = -\alpha R_r \phi_{rd} + \alpha L_m R_r i_{sd}$$
$$\frac{di_{sq}}{dt} = i_q$$
$$\frac{di_{sd}}{dt} = i_d$$
$$\frac{di_q}{dt} = \frac{d}{dt}(-\eta_1 i_{sq} - \beta n_p \omega\phi_{rd} - n_p \omega i_{sd} - R_r\left(\eta_2 i_{sq} + \alpha L_m \frac{i_{sq} i_{sd}}{\phi_{rd}}\right)) + w_q$$
$$\frac{di_d}{dt} = \frac{d}{dt}(-\eta_1 i_{sd} + n_p \omega i_{sq} + R_r\left(-\eta_2 i_{sd} + \alpha\beta\phi_{rd} + \alpha L_m \frac{i_{sq}^2}{\phi_{rd}}\right)) + w_d \quad (16)$$

The application of backstepping to this new model allows the calculation of interim orders w_q and w_d. They are given by

$$w_q = \frac{d}{dt}[\mu\phi_{rd} e_1 + k_2 e_2 + \psi_1] + k_5 e_5 + e_2$$
$$w_d = \frac{d}{dt}[\alpha L_m R_r e_3 + k_4 e_4 + \psi_2] + k_6 e_6 + e_4 \quad (17)$$

This allows a simple integration to return to u_{sq} and u_{sd}, so written :

$$u_{sq} = \sigma L_s \int w_q dt = \sigma L_s \left[\mu \phi_{rd} e_1 + k_2 e_2 + \psi_1 + k_5 \int e_5 dt + \int e_2 dt \right]$$
$$= u_{sq_0} + \sigma L_s \left[k_5 \int e_5 dt + \int e_2 dt \right]$$
$$u_{sd} = \sigma L_s \int w_d dt = \sigma L_s \left[\alpha L_m R_r e_3 + k_4 e_4 + \psi_2 + k_6 \int e_6 dt + \int e_4 dt \right]$$
$$= u_{sd_0} + \sigma L_s \left[k_6 \int e_6 dt + \int e_4 dt \right] \tag{18}$$

With

$$e_5 = (i_q)_{ref} - i_q$$
$$= \mu \phi_{rd} e_1 + k_2 e_2 + \psi_1$$
$$e_6 = (i_d)_{ref} - i_d \tag{19}$$
$$= \alpha L_m R_r e_3 + k_4 e_4 + \psi_2$$

Where k_5, k_6 are positive design constants.

In (18), one sees the components u_{sq}, u_{sd} which correspond to the classical version of backstepping, plus a term containing the integrator introduced by the modification.

IV. Adaptive Rotor Flux Estimator

The rotor flux has been estimated using the simplified IM model (1) obtained after the application of the field oriented transformation. This allows estimation of the rotor-flux vector in polar form is given by

$$\dot{\hat{\lambda}}_{2d} = -\alpha R_2 \hat{\lambda}_{2d} + \alpha L_m R_2 i_{1d} \tag{20}$$

V. Load Torque Observer

The load torque is needed for backstepping control. It is calculated by the load torque observer [18]:

$$\hat{t}_L = \frac{1}{\tau_0}\left(z - \frac{J}{n_p}\omega \right)$$
$$\frac{dz}{dt} = n_p \frac{2}{3} L_m \lambda_{2d} i_{1q} - \hat{t}_L \tag{21}$$

$\hat{t}_L$ is the observed load torque, τ_0 is the observer time constant and z is the observer state.

VI. Simulation And Experimental Study

The overall configuration of the control system for IM is shown in Fig. 1. The voltage-fed inverter is used for generating space vector pulse width modulated (SVPWM) three phase voltages for the induction motor.

The parameters of the induction motor used are given in Appendix. Parameters of the backstepping controller are: k_1= 900, k_2=900, k_3= 900, k_4= 900, k_5=0.01 and k_6=0.01.

A. Simulation Results

The effectiveness of the proposed controllers (classical and integral backstepping) has been verified by simulation in Matlab/Simulink environment. The results are shown in Fig. 2 and Fig. 3.

The speed reference is a step function followed by smoothening equal to 300 tr/mn. The reference flux is set to 0.5 Wb. A constant load torque of 6.1 N.m is applied from the instant t = 3 s for the first case (classical backstepping), and 9 N.m is applied from instant t=8.7 s for second case (integral backstepping).

B. Experimental Results

The block schematic of the experimental setup is shown in Fig. 4. Experimental setup consists of a squirrel cage induction motor of rating: Star/Delta 380V/220V AC, 8.9 A/ 15.5 A , 4 kW, p.f. 0.82, 1440 r.p.m, 50 Hz and is coupled with a powder brake. The rotor shaft of the induction motor is fitted with a optical position encoder with 1024 lines per rotation for measuring angular position and speed. The induction motor receives power from an SVPWM inverter of 1000 V, 30 A rating. The dSPACE interface generates the SVPWM pulses for the inverter and takes the signals of the measured currents for phase "a" and "b" through ADCs and angular position signals through encoder. It takes also the speed command from the dSPACE "controldesk" and generates the voltage command for the SVPWM inverter. The signal for angular positions are sampled for every 750 microsecond interval and the current signals are sampled for every 200 microsecond interval. The computation for control algorithm is done within a time step of 200 microsecond. The inverter switching frequency is kept at 10kHz using the slave DSP. The control and estimation algorithms need as inputs (1) the stator currents and (2) the encoder position. For that filters are used:

- Digital low pass filter is used for filtering encoder position signal.
- Digital synchronous resonating filter [19] is used for reducing high-frequency noise in the stator current signals arising out of SVPWM and electrical grid.

In experimental mode, load torque is applied at the instant t=3 s for classical backstepping and at t=8.7 s for integral backstepping.

The experimental results are shown in Fig. 5 and Fig. 6.

C. Discussion

As shown in the simulation (Fig. 2 and Fig. 3) and in the experimental (Fig. 5 and Fig. 6) results, performance of integral backstepping control is better than the classical one. Bandwidth in speed response is smaller and better capability of disturbance rejection (Fig. 6).

We find that, with the utilization of the proposed nonlinear integral backstepping control on an induction motor drive, the controlled speed and flux amplitude can quickly and accurately track the desired references, under presence of disturbance. The validity and effectiveness of the practical implementation of this control structure are thus demonstrated satisfactorily.

VII. CONCLUSION

A new and effective version of backstepping technique control with SVPWM inverter is presented. We have shown by simulation and by experiment that the proposed nonlinear control algorithm with integral action achieved very good tracking performance. This improvement is manifested for signal quality speed (bandwidth smaller) and for almost total rejection of the disturbance (load torque) even this disturbance is much higher as applied in the case of the classical backstepping. The other interesting feature of the proposed method is that it is simple and easy to implement in real time.

APPENDIX

INDUCTION MOTOR DATA

Stator resistance	1.34 ohms;
Rotor resistance	1.18 ohms;
Mutual inductance	0.17 H;
Rotor inductance	0.18 H;
Stator inductance	0.18 H;
Number of pole pairs	2
Motor load inertia	0.0153 kgm^2;

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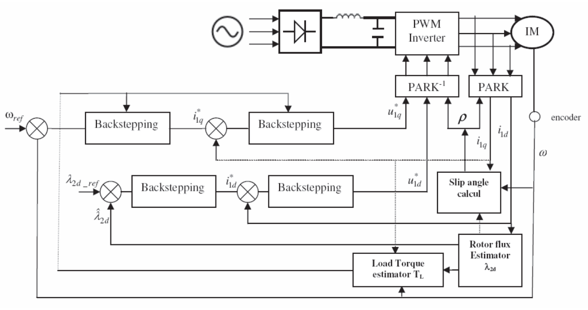

Fig. 1. Overall block diagram of the control scheme for IM

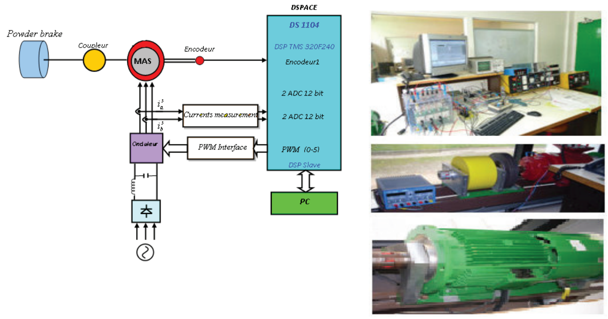

Fig. 4. The experimental setup

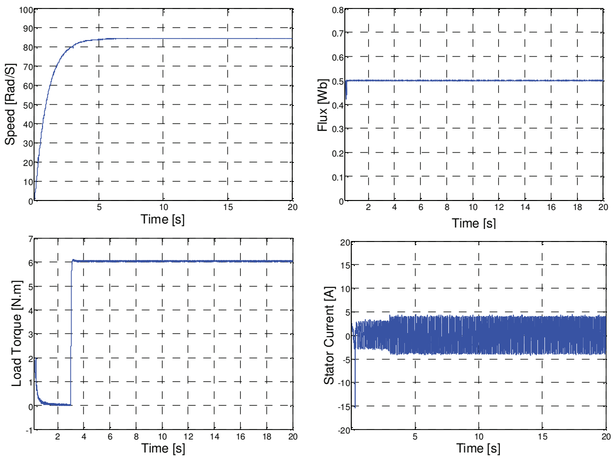

Fig. 2. Tracking performance (Simulation results) Classical backstepping

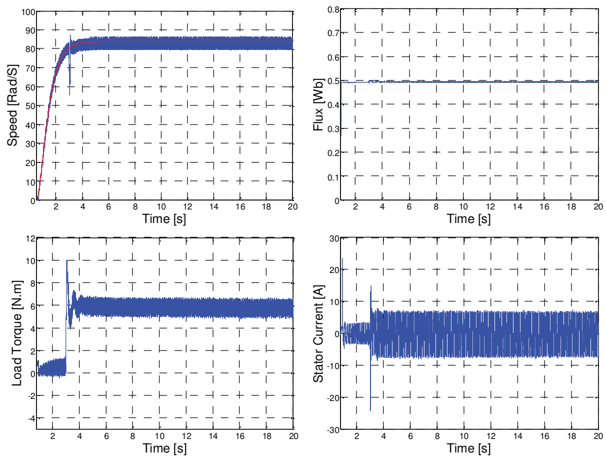

Fig.5. Tracking performance (experimental results) Classical backstepping

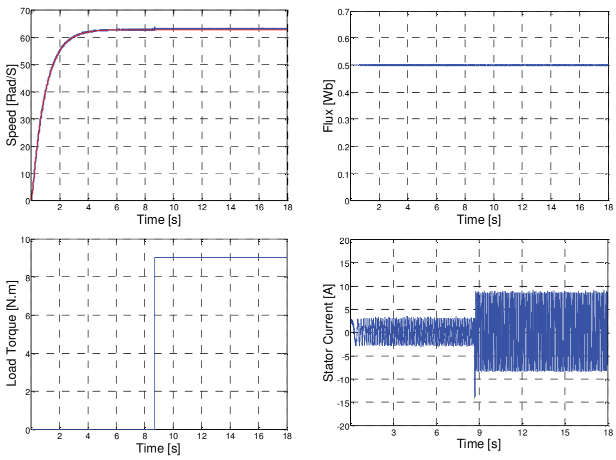

Fig. 3. Tracking performance (Simulation results) Integral backstepping

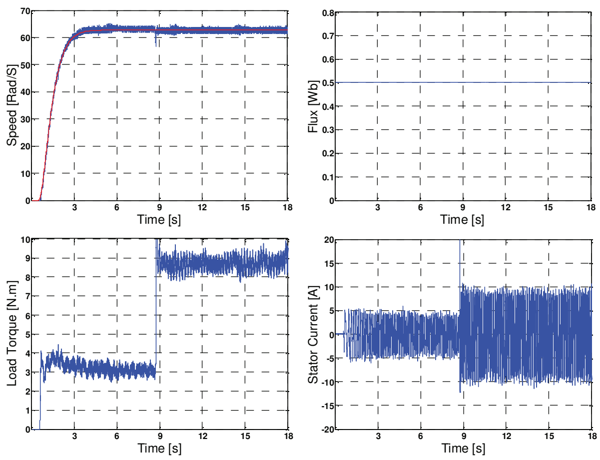

Fig.6. Tracking performance (experimental results) Integral backstepping