High performance backstepping control of induction motor with adaptive sliding mode observer

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It is well known that modern control of induction motor relies on a good dynamic model of the motor. Extensive research and activity have been devoted to the problem of induction motor control over the last decade. In this paper we introduce backstepping control with amelioration of performance to guarantee stability of the system. Accurate knowledge of the rotor speed and flux position is the key factor in obtaining a high-performance and high-efficiency induction-motor drive. Thus a sliding mode observer design is presented. Simulation results are included to illustrate good performance of backstepping control of sensorless induction motors with flux observer.

Key words: induction motor, backstepping control, sliding mode, flux observer design, estimation, sensorless control

1. Introduction

Induction motor serves as the example of difficult object to control thus new control and model identification methods need to be used. The motor constitutes nonlinear models with time varying parameters due to temperature variations and change in magnetic saturation level [1]. Nonlinear control problems can often be solved if full state information is available. In the last two decades, many modified nonlinear state feedback schemes such as input-output feedback linearization [2], passivity-based control [3, 4] and sliding-mode (SM) control [5] have been applied to the induction motor. Specially in the last few years, in the field of adaptive and robust control, a special control scheme known as 'backstepping' reached great attention [6, 7, 8, 9, 10]. This approach bases on a systematic procedure for the design of feedback control strategy. The strategy is suitable for the design of a large class of feedback linearizable systems exhibiting constant uncertainty. The strategy guarantees also global regulation and tracking for the class of nonlinear systems transformable into the parametric-strict feedback form. The idea of backstepping design is to select recursively some appropriate functions of state

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variables as pseudo-control inputs for lower dimension subsystems of the overall system. Each backstepping stage results in a new pseudo-control design, expressed in terms of the pseudo-control designs following from the preceding design stages. When the procedure terminates, a feedback design for the true control input results. This feedback achieves the original design objective by virtue of a final Lyapunov function, which is formed by summing up the Lyapunov functions associated with each individual design stage [11].

To control the induction motor drive the observer is used which constructs an estimates the unmeasured flux states. Several techniques in the literature have been used for flux, speed and (or) parameter estimation for the induction motor. In this paper we proposed an extended Kalman filter to estimate the rotor flux (or rotor currents) together with the rotor speed and the rotor speed time constant (or rotor resistance). This technique, however, is not robust against external disturbances (for example load torque). In other reference we used the induction motor equations to estimate the flux. Using independent subsystems for the rotor flux calculation, an estimation of the rotor speed (considered constant) was given using the Model Reference Adaptive System (MRAS) technique. Under load at low frequency this method gives poor results. A linear observer was proposed to estimate the rotor flux when the speed is constant. This approach is not robust against motor parameter variations and requires an adaptation mechanism for parameter identification. Thus we propose a sliding mode rotor flux observer to minimize the resistive parameters effects [12].

The control scheme involves the use of rotor sensorless speed measurements which lead to high costs and unreliability in the system. In the literature, some simple open loop methods can be used to determine the estimated speed in a fast way, however, they might be sensitive to improper parameters. On the other hand, some closed loop methods using speed observer are robust to mismatched parameters [13].

The paper is organized as follows. Modeling of induction motor is reviewed and details of backstepping control is presented in section 2. The sliding mode observer is discussed and the speed observer is described in section 3. Simulation results are delivered for illustration in section 4. Final section concludes the paper.

2. Backstepping control design of induction motor

To independently control the electromagnetic force and the flux we use (d-q) rotating reference frame synchronously with the rotor flux space vector. We applied this condition for $\phi_{rd} = \phi_r$, $\phi_{rq} = 0$ (symbols used throughout the paper are defined in a separate list in the end of the paper). The induction motor model of the fourth order, under assumption of linear magnetic circuits, is given by the following equations

$$\frac{di_{sd}}{dt} = -\left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma}\right)i_{sd} + \omega_s i_{sq} + \frac{MR_r}{\sigma L_s L_r^2}\phi_r + \frac{1}{\sigma L_s}V_{sd}$$

$$\frac{di_{sq}}{dt} = -\left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma}\right)i_{sq} - \omega_s i_{sd} - \frac{M}{\sigma L_s L_r}\omega_r \phi_r + \frac{1}{\sigma L_s}V_{sq}$$

$$\frac{d\phi_{rd}}{dt} = \frac{MR_r}{L_r}i_{sd} - \frac{R_r}{L_r}\phi_{rd} \tag{1}$$

$$\frac{d\Omega}{dt} = \frac{pM}{JL_r}(\phi_{rd}i_{sq}) - \frac{C_r}{J}$$

The controller is designed using the backstepping technique. This is done in three steps: first, the above machine model is reformulated in terms of appropriate tracking and control errors. The performance-oriented model obtained in such a way suggests a Lyapunov function. In the second step, a stabilizing control law is obtained. In the final step control law turns out.

2.1. Step 1

Let us introduce the tracking error of the rotor speed as

$$e_1 = \Omega - \Omega_{ref}. \tag{2}$$

where Ω_{ref} denotes the corresponding reference signal. In view of (1) time-derivation of (2) gives:

$$\dot{e}_1 = \frac{1}{J}\left(\frac{pM}{L_r}\phi_{rd}i_{sq} - C_r\right) - \dot{\Omega}_{ref} \tag{3}$$

where the last term is considered as a virtual control input. This motivates the following control error definition

$$e_{eff1} = \frac{pM}{JL_r}\phi_{rd}i_{sq} - \alpha_1 \tag{4}$$

where α_1 is a stabilizing function to be defined later. Substituting (4) in (3) yields:

$$\dot{e}_1 = e_{eff1} + \alpha_1 - \frac{Cr}{J} - \dot{\Omega}_{ref}. \tag{5}$$

We introduce now the Lyapunov function as

$$V_1 = \frac{1}{2}e_1^2 \tag{6}$$

and derivative error as

$$\dot{e}_1 = -k_1 e_1. \tag{7}$$

We obtain

$$\dot{V}_1 = -k_1 e_1^2 < 0 \quad k_1 > 0. \tag{8}$$

If the virtual control $\frac{p}{L_r}\phi_{rd}i_{sq}$ is effective (i.e. $e_{eff1} = 0$) then the stabilizing function:

$$\alpha_1 = \frac{C_r}{J} + \dot{\Omega}_{ref} - k_1 e_1.$$ (9)

Unfortunately, $\frac{p}{L_r}\phi_{rd}i_{sq}$ cannot be the effective control because i_{sq} is a state variable. The $e_{eff1} \neq 0$ and, consequently, the stabilizing function (9), together with (5), gives only

$$\dot{e}_1 = -k_1 e_1 + e_{eff1}.$$ (10)

2.2.   Step 2

Now, let us focus on the flux tracking error with his derivative:

$$e_2 = \phi_{rd} - \phi_{ref}$$ (11)

where ϕ_{ref} denotes the corresponding reference signal. In view of (1) we have:

$$\dot{e}_2 = \frac{MR_r}{L_r}i_{sd} - \frac{R_r}{L_r}\phi_{rd} - \dot{\phi}_{ref}.$$ (12)

Similarly, we introduce the control error:

$$e_{eff2} = \frac{MR_r}{L_r}i_{sd} - \alpha_2$$ (13)

where α_2 is a stabilizing function to be defined later. Substituting (13) in (12) gives:

$$\dot{e}_2 = -\frac{R_r}{L_r}\phi_{rd} - \dot{\phi}_{ref} + \alpha_2 + e_{eff2}.$$ (14)

Lyapunov function will be used as:

$$V_2 = \frac{1}{2}e_2^2.$$ (15)

We suggest derivative error as:

$$\dot{e}_2 = -k_2 e_2, \quad k_2 > 0.$$ (16)

As previously, if $\frac{MR_r}{L_r}i_{sd}$ is an effective control (i.e. $e_{eff2} = 0$) then the stabilizing function:

$$\alpha_2 = -k_2 e_2 + \frac{R_r}{L_r}\phi_{rd} + \dot{\phi}_{ref}.$$ (17)

As $\frac{MR_r}{L_r}i_{sd}$ cannot be an effective control (which means that $e_{eff2} \neq 0$), the stabilizing control, (17) together with (14), yields only

$$\dot{e}_2 = -k_2 e_2 + e_{eff2}.$$ (18)

2.3. Step 3

In the Step 3, effective control is presented and its dynamic errors. This step presents the main novelty of our approach.

$$
\begin{aligned}
e_{eff1} &= \frac{pM}{JL_r}\phi_{rd}i_{sq} - \alpha_1 \\
e_{eff2} &= \frac{MR_r}{L_r}i_{sd} - \alpha_2 \\
\dot{e}_{eff1} &= f_{11} + k_1 e_{eff1} - k_1^2 e_1 + \frac{pM}{JL_r}\phi_{rd}\frac{1}{\sigma L_s}V_{sq} \\
\dot{e}_{eff2} &= f_{22} + k_2 e_{eff2} - k_2^2 e_2 + \frac{MR_r}{L_r}\frac{1}{\sigma L_s}V_{sd}.
\end{aligned}
\tag{19}
$$

To this end, we again compute the derivative of Lyapunov function along the error equations (10), (18), and (19):

$$
V = \frac{1}{2}e_1^2 + \frac{1}{2}e_{eff1}^2 + \frac{1}{2}e_2^2 + \frac{1}{2}e_{eff2}^2. \tag{20}
$$

In such circumstances the control law of stator voltage turns out to be the following:

$$
\begin{aligned}
V_{sd} &= \frac{\sigma L_s L_r}{MR_r}\left(-f_{22} - e_2 - k_2 e_{eff2} + k_2^2 e_2 - k_4 e_{eff2}\right) \\
V_{sq} &= \frac{JL_r \sigma L_s}{pM\phi_r}\left(-f_{11} - e_1 - k_1 e_{eff1} + k_1^2 e_1 - k_3 e_{eff1}\right)
\end{aligned}
\tag{21}
$$

with k_3, $k_4 > 0$. We can express derivatives errors as follow:

$$
\begin{aligned}
\dot{e}_1 &= -k_1 e_1 + e_{eff1} \\
\dot{e}_2 &= -k_2 e_2 + e_{eff2} \\
\dot{e}_{eff1} &= f_{11} + k_1 e_{eff1} - k_1^2 e_1 + \frac{pM}{JL_r}\phi_{rd}\frac{1}{\sigma L_s}V_{sq} \\
\dot{e}_{eff2} &= f_{22} + k_2 e_{eff2} - k_2^2 e_2 + \frac{MR_r}{L_r}\frac{1}{\sigma L_s}V_{sd}.
\end{aligned}
\tag{22}
$$

Equation (22) can be simplified and represented as:

$$
\dot{e} = Ae + D \tag{23}
$$

with

$$
A = \begin{bmatrix} -k_1 & 0 & 1 & 0 \\ 0 & -k_2 & 0 & 1 \\ -k_1^2 & 0 & k_1 & 0 \\ 0 & -k_2^2 & 0 & k_2 \end{bmatrix}, \quad
D = \begin{bmatrix} 0 \\ 0 \\ f_{11} + \dfrac{pM}{JL_r}\phi_{rd}\dfrac{1}{\sigma L_s}V_{sq} \\ f_{22} + \dfrac{MR_r}{L_r}\dfrac{1}{\sigma L_s}V_{sd} \end{bmatrix}, \quad
e = \begin{bmatrix} e_1 \\ e_2 \\ e_{eff1} \\ e_{eff2} \end{bmatrix}
$$

Note that matrix *A* has only constant components to guarantee the stability of the control system and for improvement of simulation results.
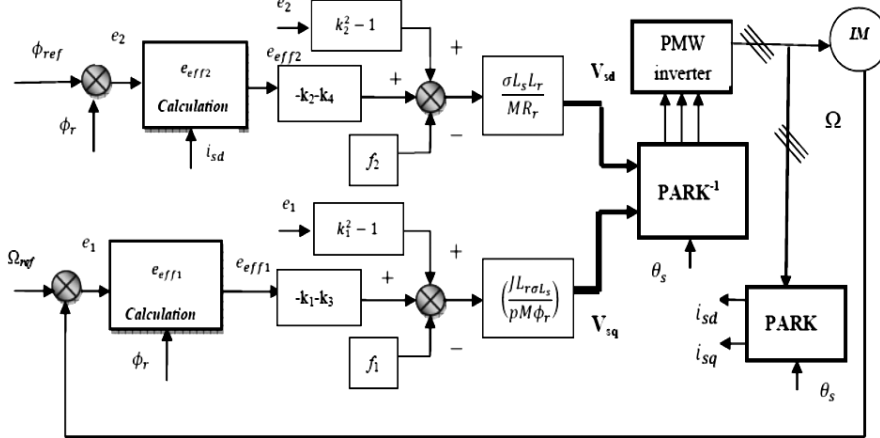

Figure 1. Block diagram of backstepping control.

Fig. 1 shows the different steps to turn out the control law of stator voltage. Accurate knowledge of the rotor speed and rotor flux are the keys factors in obtaining a high-performance and high-efficiency induction-motor drive, thus sliding mode observer design is presented in the next section.

3. Sliding mode observer design

Consider only the first four equations of the induction motor model given by (1). In the following, the speed will be considered as a varying parameter. The aim of the proposed observer is to estimate firstly the rotor flux. The observer is given by the following system:

$$\frac{d}{dt}\begin{bmatrix} \hat{i}_s \\ \hat{\phi}_r \end{bmatrix} =$$

$$\begin{bmatrix} -\dfrac{R_t}{\sigma L_s}I & \dfrac{M}{\sigma L_s L_r T_r}I - j\dfrac{\omega}{\sigma L_s} \\ \dfrac{M^2}{L_r T_r}I & -\dfrac{M}{L_r T_r}I + j\omega \end{bmatrix}\begin{bmatrix} i_s \\ \phi_r \end{bmatrix} + \begin{bmatrix} \dfrac{1}{\sigma L_s} \\ 0 \end{bmatrix}U_s + \begin{bmatrix} K_s \\ K_r \end{bmatrix}\begin{bmatrix} \mathrm{sgn}(S_1) \\ \mathrm{sgn}(S_2) \end{bmatrix} \quad (24)$$

where:

$$R_t = R_s + \frac{M^2}{L_r T_r}$$

is the total resistance restored to the stator,

$$I = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \qquad j = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$$

$$K = \begin{bmatrix} K_s \\ K_r \end{bmatrix}$$

is matrix gain and

$$S = \begin{bmatrix} S_1 \\ S_2 \end{bmatrix} = m \begin{bmatrix} i_{s\alpha} - \hat{i}_{s\alpha} \\ i_{s\beta} - \hat{i}_{s\beta} \end{bmatrix}$$

is the sliding surface which represent the error between the measured current components and those estimated where

$$m = \begin{bmatrix} 1 & \alpha \\ 0 & 1 \end{bmatrix}.$$

The coefficient α comprises between 0 and 1 and represents the freedom degree to be used to make dynamic comportment of the observer regular.

Thus, we use the K_r gain for action to fix the dynamics of convergence of the evaluation error flux (reduced system is equivalent) [13]. One considers then, the state variables of error estimation as follows:

$$\begin{bmatrix} \dot{\varepsilon}_{is} \\ \dot{\varepsilon}_{\phi r} \end{bmatrix} = \begin{bmatrix} 0 & \dfrac{M}{\sigma L_s L_r T_r} I - j \dfrac{\omega}{\sigma L_s} \\ 0 & -\dfrac{M}{L_r T_r} I + j\omega \end{bmatrix} \begin{bmatrix} \varepsilon_{is} \\ \varepsilon_{\phi r} \end{bmatrix} - \begin{bmatrix} K_s \\ K_r \end{bmatrix} \begin{bmatrix} \mathrm{sgn}(S_1) \\ \mathrm{sgn}(S_2) \end{bmatrix} \qquad (25)$$

3.1. Correction matrix gains K_s

To assure the asymptotic convergence of S to zero, one searches for the necessary conditions of stability, bounded to the value of the K_s gain. Let choose the following Lyapunov function:

$$V = \frac{1}{2} S^T S. \qquad (26)$$

In order to assure S convergent to zero, we must verify if derivative of V is strictly negative.

$$\dot{V} = S^T \dot{S} < 0 \Rightarrow \dot{V} = S^T m \dot{\varepsilon}_s < 0$$

$$\dot{V} = [S_1\ S_2] \begin{bmatrix} 1 & \alpha \\ 0 & 1 \end{bmatrix} \left(\frac{1}{\sigma L_s} \left(\frac{M}{L_r T_r} I - j\omega \right) \varepsilon_{\phi r} - K_s \begin{bmatrix} \mathrm{sgn}(S_1) \\ \mathrm{sgn}(S_2) \end{bmatrix} \right) < 0. \qquad (27)$$

If one puts

$$K_s = m^{-1} \begin{bmatrix} G_1 & 0 \\ 0 & G_2 \end{bmatrix} \tag{28}$$

where G_1 and G_2 are positive constants, then the derivative of the Lyapunov function becomes negative:

$$\dot{V} = S_1(f_1 + \alpha f_2 - G_1 \text{sgn}(S_1)) + S_2(f_2 - G_2 \text{sgn}(S_2)) < 0. \tag{29}$$

To assure the convergence of S to zero one must verify the following conditions:

If $S_1 > 0$ then $G_1 > f_1 + \alpha f_2$; elseif $S_1 \leqslant 0$ then $-G_1 < f_1 + \alpha f_2$

If $S_2 > 0$ then $G_2 > f_2$; elseif $S_2 \leqslant 0$ then $-G_2 < f_2$

So $G_1 > |f_1 + \alpha f_2|$, and $G_2 > |f_2|$. It only remains to choose values of G_1 and sufficiently large G_2 to verify the convergence of S toward zero, which means $\hat{i}_s \to i_s$.

3.2.   Correction matrix gains K_r

In order to calculate the gain correction K_r, we consider sliding on the surface S ($S = 0$, $\dot{S} = 0$). This hypothesis is verified, when the currents error is $\varepsilon_{is} = 0$, $\dot{\varepsilon}_{is} = 0$. We have then:

$$\dot{\varepsilon}_{is} = \frac{1}{\sigma L_s} \left[\frac{M}{L_r T_r} I - j\omega \right] \varepsilon_{\phi r} - K_s \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} = 0. \tag{30}$$

Then we have sliding surface given by as follows

$$\begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} = \frac{1}{\sigma L_s} K_S^{-1} \left[\frac{M}{L_r T_r} I - j\omega \right] \varepsilon_{\phi r}. \tag{31}$$

Using (31) we can express the equation for the error of the rotor flux as follows:

$$\dot{\varepsilon}_{\phi r} = \left(-\frac{M}{L_r T_r} I + j\omega - K_r K_s^{-1} \frac{1}{\sigma L_s} \left[\frac{M}{L_r T_r} I - j\omega \right] \right) \varepsilon_{\phi r}. \tag{32}$$

We can calculate the gain K_r correction by identification of an equivalent system with dynamics that assures the desire behavior

$$\dot{\varepsilon}_{\phi r} = -Q \varepsilon_{\phi r} \tag{33}$$

with Q – positive definite matrix:

$$Q = \begin{bmatrix} q_1 & 0 \\ 0 & q_2 \end{bmatrix}, \quad q_1, q_2 > 0.$$

We can recover the expression of K_r by:

$$K_r = \frac{\sigma L_s}{\beta} \begin{bmatrix} G_1 \left(\frac{M}{L_r T_r} q_1 - \beta \right) & -G_2 \left(q_1 \omega + \alpha \left(\frac{M}{L_r T_r} q_1 - \beta \right) \right) \\ G_1 \omega q_2 & G_2 \left(\frac{M}{L_r T_r} q_2 - \beta - \alpha \omega q_2 \right) \end{bmatrix}$$

where:

$$\beta = \left(\frac{M}{L_r T_r} \right)^2 + \omega^2 .$$

3.3.   Speed observer design

We consider the perfect motor model (supposing that resistances and inductances are perfectly known) and also the speed supposed as a parameter when uncertainty exists and is presented by:

$$\hat{\omega} = \omega + \Delta \omega \tag{34}$$

$$\begin{bmatrix} \dot{\varepsilon}_{is} \\ \dot{\varepsilon}_{\phi r} \end{bmatrix} =$$

$$\begin{bmatrix} 0 & \left(\frac{M}{\sigma L_s L_r T_r} \right) I - j\omega \\ 0 & -\left(\frac{M}{L_r T_r} \right) I + j\omega \end{bmatrix} \begin{bmatrix} \varepsilon_{is} \\ \varepsilon_{\phi r} \end{bmatrix} - \begin{bmatrix} K_s \\ K_r \end{bmatrix} \begin{bmatrix} \text{sgn}(S_1) \\ \text{sgn}(S_2) \end{bmatrix} - \Delta A_\omega^t \begin{bmatrix} i_s \\ \hat{\phi}_r \end{bmatrix} \tag{35}$$

where

$$\Delta A_\omega^t = (\omega - \hat{\omega}) \begin{bmatrix} 0 & \frac{-J}{\sigma L_s} \\ 0 & J \end{bmatrix} .$$

In this condition we can develop the derivation of V as:

$$\dot{V} = S_1 (f_1 + \alpha f_2 - G_1 \text{sgn}(S_1)) + S_2 (f_2 - G_2 \text{sgn}(S_2)) - S^T m (\omega - \hat{\omega}) \frac{J}{\sigma L_s} \hat{\phi}_r . \tag{36}$$

To compensate the term $S^T m (\omega - \hat{\omega}) \frac{J}{\sigma L_s} \hat{\phi}_r$, which sign is unknown, the following Lyapunov equation is considered

$$V_2 = V + \frac{(\omega - \hat{\omega})^2}{2\lambda_3} \tag{37}$$

where $\lambda_3 > 0$. Derivation of V_2 gives

$$\dot{V}_2 =$$

$$S_1 (f_1 + \alpha f_2 - G_1 \text{sgn}(S_1)) + S_2 (f_2 - G_2 \text{sgn}(S_2)) - S^T m (\omega - \hat{\omega}) \frac{J}{\sigma L_s} \hat{\phi}_r - \frac{\omega - \hat{\omega}}{\lambda_3} \dot{\hat{\omega}} . \tag{38}$$

To assure $\dot{V}_2 < 0$ for the development of adaptation low of speed we need to propose:

$$\dot{\hat{\omega}} = -\lambda_3 \frac{S^T m J \hat{\phi}_r}{\sigma L_s} \tag{39}$$

4.   Simulation results

The effectiveness of the proposed controller combined with the rotor flux and rotor speed estimation has been verified by simulations. The simulation results have been obtained under a constant load of 5 Nm applied at 0.5 s.

Table 10. Motor parameters

R_s	10 Ω	L_s	0.4642 H
R_r	6.3 Ω	L_r	0.4612 H
M	0.4212 H	J	0.02 kgm^2

Another parameters are as follows: output power 7.5kW, stator voltage 220/380 V, stator frequency 50 Hz. The speed tracking controller operated in a critical condition (rapid changes: 157,-157, 0, 5 rad/s). Fig. 1 shows the satisfactory performance of the speed tracking. One can see that the actual speed follows the speed command and estimated speed. Thus, the simulation results confirm that the proposed observer gives good results justified by rotor speed error converging to zero quickly.

We applied as well variation of 100% of the nominal rotor resistance between t = 0.3s and t=0.7s. Fig. 2. presents the obtained results. Note good observation of speed and flux. The influence of rotor resistance variation on the speed is negligible. Error speed converges to zero. Obtained results illustrate good convergence to desired value of speed and rotor flux.

5.   Conclusion

In this work we have presented a backstepping technique as a design tool to accommodate uncertainties and nonlinearities which are the inherent feature of control design for induction motor. This study has demonstrated that the design using backstepping technique is successful in the area of induction motor control systems. In this paper we present also the concept of rotor flux observer for sensorless speed control, which provides good precision as well. It is shown by simulation with the different tests that the proposed approach is robust.
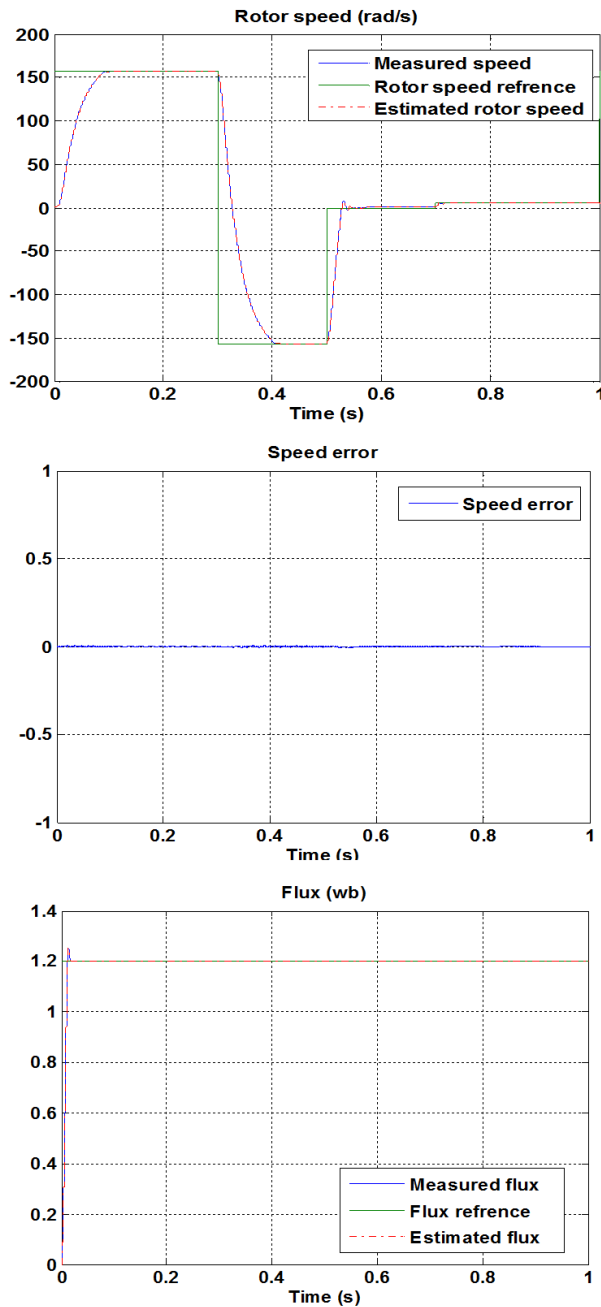
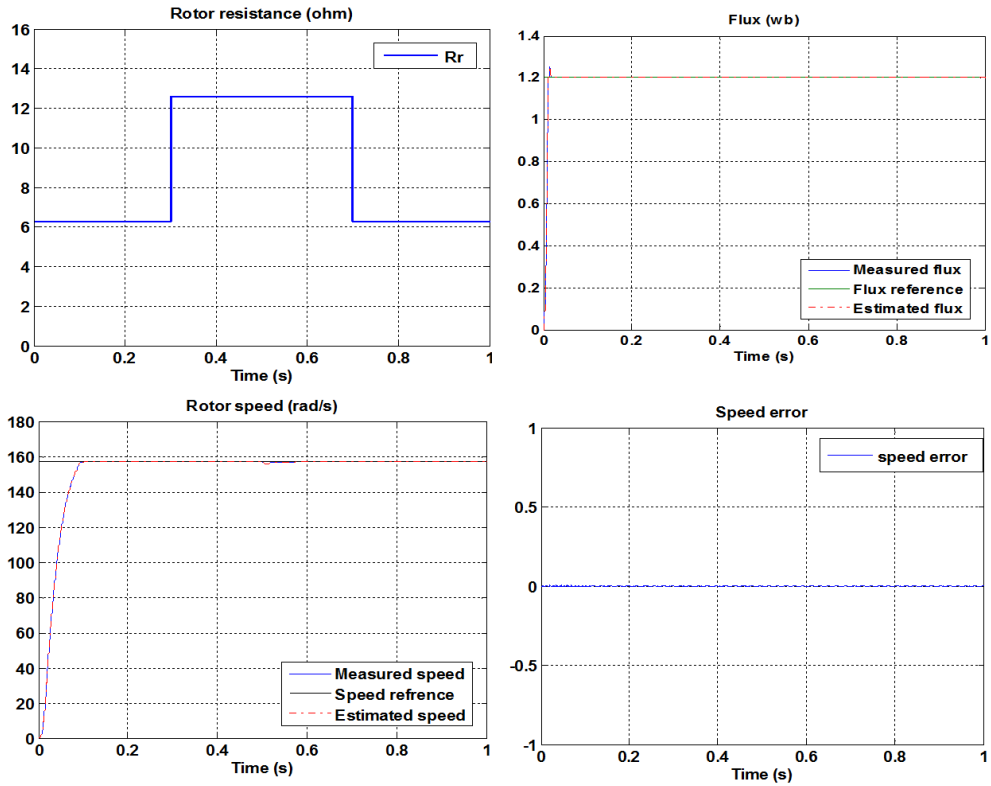Figure 2. Backstepping control for speed variation.

Figure 3. Backstepping control for rotor resistance variation.

List of symbols

s, r	indices for stator and rotor
R_s, R_r	stator and rotor resistances
L_s, L_r	self inductance of stator and rotor
M	mutual inductance
J	inertia moment of the moving element
f	viscous friction and iron-loss coefficient
T_r, T_s	rotor and stator time constant
σ	coefficient of dispersion
(d,q)	axes for direct and quadrate park subscripts
(α, β)	axes for stationary reference frame subscripts
$V_{sd}\ V_{sq}$	stator voltage in direct and quadrate park subscripts
i_s	stator currents
ϕ_s, ϕ_r	stator and rotor flux

g	slip
C_{em}	electromagnetic torque
C_r	load torque
ω_s, ω_r	stator and rotor angular frequency
p	number of pole pairs
ϕ_{ref}	rotor flux reference
Ω_{ref}	rotor speed reference
$V(x)$	Lyapunov function
$S(x)$	sliding surface
j	imaginary matrix
I	identity matrix
$\hat{x}$	estimated signal x

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