Simplified Speed-Sensorless Vector Control for Induction Motors and Stability Analysis

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This paper presents a new simplified speed-sensorless vector control method for induction motors. The output voltage of the d -axis PI current controller with a decoupling control is used to compute the flux angle and to control the speed in correspondence with its reference. System stability is discussed by the root loci computed from a linear model. The effectiveness of the proposed method is demonstrated by nonlinear simulations and experiments.

Keywords: speed-sensorless vector control, induction motor, root loci, linear model, stability analysis

1. Introduction

Speed-sensorless vector control of induction motors (IM) allows high performance control of torque and speed and the system is widely used. Many model reference adaptive system (MRAS) based methods are studied. Representative speed estimation schemes use flux simulator⁽¹⁾⁽²⁾, full-order observer^{(3)(6)–(8)} reduced-order observer⁽¹⁰⁾, or sliding-mode full-order observer⁽⁴⁾⁽⁵⁾⁽⁹⁾. Manipulated design of the observers can improve the stability of the sensorless system even at low speed regenerating operation. However, the configurations of these systems are relatively complicated. It is because the MRAS based methods need state observer and many PI controllers (d - q currents, speed and speed estimation). On the other hand, some simplified speed sensorless vector control methods are proposed. Simplifying the system configuration by removing the current regulators is proposed⁽¹¹⁾, and the stability is improved by adding a flux stabilizing controller using derivative of magnetic current⁽¹²⁾. However, these papers have no information about the stability of regenerating mode. A sensorless method using the induced d - and q -axis voltages obtained by a voltage model has been proposed⁽¹³⁾, and a similar method is applied to railway vehicle traction⁽¹⁴⁾. However, the stable region is not clear in these papers. Furthermore, a primary flux control method is proposed⁽¹⁵⁾ and the stability is improved at regenerating mode⁽¹⁶⁾.

In this paper, we propose a new simplified speed-sensorless vector control method of IM based on rotor flux⁽¹⁸⁾⁽²⁰⁾. By considering a flux vector obtained by a voltage model in which the derivative term is neglected and a flux vector of a current model, the flux angle is estimated so as to align these vectors. Since the output voltage of d -axis PI current controller is used for both the flux angle estimation and speed

control (q -axis voltage control), the system is simplified and stabilized at regenerating mode. This scheme is not reported in conventional simplified methods. A linear model of the proposed system is derived in state space equation. System stability is discussed by showing root loci of the linear model. By virtue of the stability analysis, we design the parameters of controller. Transient responses of the proposed system are demonstrated by nonlinear simulation and experiment.

2. Proposed Method

The d - q voltage model and current model of the induction motor which are used for the derivation of proposed system are described by (1)–(4) as follows:

voltage model:

$$e_{sd}^* = (R_s^* + \sigma L_s p) i_{sd}^* - \omega^* \sigma L_s i_{sq} + \frac{Mp}{L_r} \psi_{rd}^v - \frac{\omega^* M}{L_r} \psi_{rq}^v \quad \dots \dots \dots (1)$$

$$e_{sq}^* = \omega^* \sigma L_s i_{sd}^* + (R_s^* + \sigma L_s p) i_{sq} + \frac{\omega^* M}{L_r} \psi_{rd}^v + \frac{Mp}{L_r} \psi_{rq}^v \quad \dots \dots \dots (2)$$

current model:

$$0 = -\frac{M}{\tau_r^*} i_{sd}^* + \left(\frac{1}{\tau_r^*} + p \right) \psi_{rd}^* - (\omega^* - \hat{\omega}_r) \psi_{rq}^* \quad \dots \dots \dots (3)$$

$$0 = -\frac{M}{\tau_r^*} i_{sq} + (\omega^* - \hat{\omega}_r) \psi_{rd}^* + \left(\frac{1}{\tau_r^*} + p \right) \psi_{rq}^* \quad \dots \dots \dots (4)$$

where, $p = d/dt$, $\sigma = 1 - M^2/(L_s L_r)$, $\tau_r^* = L_r/R_r^*$

The $\hat{\omega}_r$ is the estimated speed. We define the d - q axis which rotates synchronously with the rotor flux of the current model so as to satisfy that $\psi_{rq}^* = 0$ as shown in Fig. 1. By assuming that the d -axis current command i_{sd}^* is constant, we obtain (5) and (6) from (3) and (4) respectively.

$$\psi_{rd}^* = M i_{sd}^* \quad \dots \dots \dots (5)$$

$$\omega^* = \hat{\omega}_r + \omega_e \quad \dots \dots \dots (6)$$

where, $\omega_e = \frac{i_{sq}}{\tau_r^* i_{sd}^*}$

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 Fig. 1. The d - q axis and rotor flux vectors

Fig. 2. Block diagram of proposed speed-sensorless system

By neglecting the derivative terms, the voltage model becomes:

$$e_{sd}^* = R_s^* i_{sd}^* - \omega^* \sigma L_s i_{sq} - \frac{\omega^* M}{L_r} \psi_{rq}^v \quad (7)$$

$$e_{sq}^* = \omega^* \sigma L_s i_{sd}^* + R_s^* i_{sq} + \frac{\omega^* M}{L_r} \psi_{rd}^v \quad (8)$$

In Fig. 1, the rotor flux space vector obtained by the voltage model is denoted by ψ_r^v and that of current model is ψ_r^i . When the rotor speed is unknown, the voltage model flux ψ_r^v can be used to estimate the flux direction. The flux angle θ^* is adjusted by changing ω^* such that the flux ψ_{rq}^v becomes zero. When $\psi_{rq}^v > 0$ and ψ_r^v is leading than d -axis as shown in (a), the controller must increase the value of ω^* . When $\psi_{rq}^v < 0$ and ψ_r^v is lagging than d -axis as shown in (b), the controller must decrease the value of ω^* . We propose a speed sensorless vector control system as shown in Fig. 2.

We define the induced voltage as

$$e_d^* = -\frac{\omega^* M}{L_r} \psi_{rq}^v \quad (9)$$

Then, from (7) we have

$$e_{sd}^* = R_s^* i_{sd}^* - \omega^* \sigma L_s i_{sq} + e_d^* \quad (10)$$

The e_d^* is obtained by the output of d -axis PI current controller with a decoupling control as shown in Fig. 2⁽¹⁸⁾. We estimate the ω^* by using (9) to converge the q -axis flux zero⁽¹⁷⁾ as follows:

$$\omega^* = \omega_r^* + \omega_e - \omega_d \quad (11)$$

where, $\omega_d = K_\omega e_d^*$, $K_\omega = \text{sign}(\omega^*) |K_\omega|$.

The $\text{sign}(\omega^*)$ is 1 when $\omega^* > 0$ and -1 when $\omega^* < 0$.

By assuming that ψ_{rd}^v is equal to ψ_{rd}^* , we compute e_{sq}^* as follows

$$e_{sq}^* = (\omega_r^* + \omega_e + \omega_c) L_s i_{sd}^* \quad (12)$$

$$\text{where, } \omega_c = \frac{K_\omega}{s T_\omega} e_d^* \quad (13)$$

The term ω_c compensates the q -axis stator resistance voltage drop in steady-state condition. By the integral control of (13), e_d^* becomes zero in steady-state, therefore, the actual speed ω_r is equal to the reference speed ω_r^* when motor constants are correct. From (6) and (11), we have

$$\omega_d = \omega_r^* - \hat{\omega}_r \quad (14)$$

Therefore, the speed control is accomplished by (13) even in transient state.

In proposed system of Fig. 2, any observer is not used and d -axis current controller (PI) and flux frequency controller (P) and speed controller (I) are used. Therefore, the proposed system is much simpler than conventional observer based systems.

3. Stability Analysis

To describe the IM shown in Fig. 2, we chose the d - q axis which rotates synchronously with θ^* . Then, well-known induction motor d - q model is obtained. In this paper, the following assumptions are made:

- 1) Voltage control is ideal, therefore
- 2) The machine constants of controller are equal to those of actual machine.

The state variable e_{cd} is defined as

$$p e_{cd} = K_i (i_{sd}^* - i_{sd}) \quad (16)$$

The equation of motion is described as

$$\frac{2}{P} J p \omega_r = \frac{P M}{2 L_r} (i_{sq} \psi_{rd} - i_{sd} \psi_{rq}) - T_L \quad (17)$$

where, P : number of poles, T_L : load torque.

By using (9)–(13), (15)–(17), and the d - q model of IM⁽¹⁷⁾, the following non-linear state equation is obtained.

$$p \mathbf{x} = \mathbf{f}(\mathbf{x}, i_{sd}^*, \omega_r^*, T_L) \quad (18)$$

where,

$$\mathbf{x} = [i_{sd} \ i_{sq} \ \psi_{rd} \ \psi_{rq} \ \omega_r \ e_{cd} \ \omega_c]^T \quad (19)$$

Transient responses are computed by solving (18).

Taking a small perturbation of state variables at steady-state operating point, a linear model of proposed system is derived from (18). A linear model of the proposed system is derived in form (Refer to the Appendix.)

$$p \Delta \mathbf{x} = \mathbf{A} \Delta \mathbf{x} + \mathbf{B} \Delta \omega_r^* + \mathbf{B}_L \Delta T_L \quad (20)$$

where, $\Delta \mathbf{x} = [\Delta i_{sd} \ \Delta i_{sq} \ \Delta \psi_{rd} \ \Delta \psi_{rq} \ \Delta \omega_r \ \Delta e_{cd} \ \Delta \omega_c]^T$. System stability is discussed by using root loci by computing the eigenvalues of the matrix $\mathbf{A}$. We confirmed that the transient responses of linear model are close to those of the nonlinear model during small displacements about a steady

state operating point⁽²⁰⁾. Therefore the derivation of linear model is reliable.

System is stable when all eigenvalues of the matrix A of linear model (20) lie inside the left half of s -plane.

The d -axis PI current controller is designed as $K_p = 14.7$, $K_i = 3395$ having cut off frequency 1500 rad/s . The control parameters K_ω and T_ω are studied by using the root loci. As the result explained below, we have selected $|K_\omega| = 5.0$ and $T_\omega = 0.05$ as designed parameters.

The constants of induction motor used for stability analysis are: number of poles $P = 4$, stator resistant $R_s = 1.54 \Omega$, rotor resistance $R_r = 0.787 \Omega$, stator and rotor inductance $L_s = L_r = 0.115 \text{ H}$, mutual inductance $M = 0.11 \text{ H}$, and moment of inertia including load $J = 0.0126 \text{ kg-m}^2$.

Figure 3 shows the trajectories of dominant poles at speed command 50 min^{-1} for the change of slip speed N_{sl} . Slip speed 80 min^{-1} almost corresponds to the rated torque. Although the system is unstable at plugging region in which the slip speed is less than -50 min^{-1} , the system is stable at regenerating region when $-50 \text{ min}^{-1} < N_{sl} < 0 \text{ min}^{-1}$.

When the speed commands are 100 min^{-1} , 1000 min^{-1} and 1500 min^{-1} , the trajectories are shown in Fig. 4, Fig. 5 and Fig. 6 respectively. It is observed that the system is stable both in regenerating and in motoring operations. The roots move left side of s -plane and the system becomes more stable as the speed command increases.

Figures 7 and 8 shows the root loci when the gain $|K_\omega| = 2.0$ at speed 100 min^{-1} and 1500 min^{-1} respectively. It is observed that the system becomes unstable (Fig. 7) or weakly damped (Fig. 8) in regenerating region when K_ω is small. It is noted that small value of $|K_\omega|$ can be relatively acceptable in motoring region.

Figure 9 shows the unstable region when $|K_\omega| = 5.0$ and $T_\omega = 0.05$ for the changes of speed command and slip speed. The shaded area means unstable region. The unstable region is observed in plugging region and very low speed motoring region. However, all regenerating region is stable.

Figure 10 shows the unstable region when $|K_\omega| = 5.0$ and $T_\omega = 0.2$. Although the large value of T_ω causes unstable region of regenerating operation, the stable region of motoring and plugging operations is expanded. So if we change the

Fig. 3. Trajectories of poles at speed $N_r^* = 50 \text{ min}^{-1}$ for the change of slip speed with $|K_\omega| = 5.0$ and $T_\omega = 0.05$

Fig. 4. Trajectories of poles at speed $N_r^* = 100 \text{ min}^{-1}$ for the change of slip speed with $|K_\omega| = 5.0$ and $T_\omega = 0.05$

Fig. 5. Trajectories of poles at speed $N_r^* = 1000 \text{ min}^{-1}$ for the change of slip speed with $|K_\omega| = 5.0$ and $T_\omega = 0.05$

Fig. 6. Trajectories of poles at speed $N_r^* = 1500 \text{ min}^{-1}$ for the change of slip speed with $|K_\omega| = 5.0$ and $T_\omega = 0.05$

Fig. 7. Trajectories of poles at speed $N_r^* = 100 \text{ min}^{-1}$ for the change of slip speed with $|K_\omega| = 2.0$ and $T_\omega = 0.05$

Fig. 8. Trajectories of poles at speed $N_r^* = 1500 \text{ min}^{-1}$ for the change of slip speed with $|K_\omega| = 2.0$ and $T_\omega = 0.05$

Fig. 9. Unstable region with parameters $|K_\omega| = 5.0$ and $T_\omega = 0.05$

Fig. 10. Unstable region with parameters $|K_\omega| = 5.0$ and $T_\omega = 0.2$

Fig. 11. Unstable transient responses with parameters $|K_\omega| = 5.0$, $T_\omega = 0.05$ and $T_L = -4.0 \text{ N}\cdot\text{m}$ for the step change of speed command from 20 min^{-1} to 22 min^{-1}

$|K_\omega|$ and T_ω , the unstable region can be reduced.

As the results of stability analysis, large value of $|K_\omega|$ stabilize the system. Too small value of T_ω increases the unstable region in motoring mode⁽²⁰⁾ and too large value of T_ω increases it in regenerating mode.

Figure 11 shows the transient responses when the system is operated at unstable region in Fig. 9 for the step change of speed command from 20 min^{-1} to 22 min^{-1} with parameters $|K_\omega| = 5.0$ and $T_\omega = 0.05$ and $T_L = -4.0 \text{ N}\cdot\text{m}$ ($N_{sl} = -35.2 \text{ min}^{-1}$ at $t = 0$). Both the non-linear model of (18) and the linear model of (20) have similar responses in short period after the step change of speed reference.

4. Experimental Results

The proposed control system is implemented by a DSP (TMS320C32)-based PWM inverter. Sampling period used in this system is $200 \mu\text{s}$. Because of the dead time and the non-ideal features of IGBT influence on the output voltage of the inverter, a compensating algorithm is developed for the experimental system⁽¹⁷⁾. The tested induction machine is 1.5 kW and the constants have been described in chapter 3.

Figure 12 shows the experimental results for the step

(a) 50→150→50 min⁻¹ (b) 500→600→500 min⁻¹ (c) 1000→1100→1000 min⁻¹

Fig. 12. Transient responses for the step change of speed command. (Motoring operation $|K_\omega| = 5.0$; $T_\omega = 0.05$; $T_L = 4.0$ N·m)

(a) 50→150→50 min⁻¹ (b) 500→600→500 min⁻¹ (c) 1000→1100→1000 min⁻¹

Fig. 13. Transient responses for the step change of speed command. (Motoring operation $|K_\omega| = 5.0$; $T_\omega = 0.05$; $T_L = 4.0$ N·m)

change of the speed command N_r^* in motoring operations. In case (a), N_r^* is stepped from 50 min⁻¹ to 150 min⁻¹ and then down to 50 min⁻¹. In cases (b) and (c), the speed commands are similarly stepped as 500→600→500 min⁻¹ and 1000→1100→1000 min⁻¹ respectively. N_r is actual motor speed (ω_r) and N^* is synchronous speed (ω^*). The control parameters are set as $|K_\omega| = 5.0$ and $T_\omega = 0.05$. The load

torque T_L is set to 4.0 N·m (half of rated torque) by a DC generator connected to IM. In any cases, quick responses of N_r are obtained and the changes of torque current i_{sq} are similar to those of vector control system with speed sensor. The non-linear simulation results computed by (18) are shown in Fig. 13. Figures 13(a), (b) and (c) corresponds to Figs. 12(a), (b) and (c) respectively. The experimental results agree well

(a) 50→150→50 min⁻¹ (b) 500→600→500 min⁻¹ (c) 1000→1100→1000 min⁻¹

Fig. 14. Transient responses for the step change of speed command (Regenerating operation $|K_\omega| = 5.0$; $T_\omega = 0.05$; $T_L = -4.0$ N·m)

with the computed results except for high frequency ripples. The high frequency ripples of N^* and e_d^* are caused by PWM voltage control in experimental system. However, since the flux angle θ^* is obtained by integrating ω^* , the actual rotor speed has little ripples.

Figures 14 and 15 show the experimental results and the non-linear simulation results in regenerating operations re-

(a) 50→150→50 min⁻¹ (b) 500→600→500 min⁻¹ (c) 1000→1100→1000 min⁻¹

Fig. 15. Transient responses for the step change of speed command (Regenerating operation $|K_\omega| = 5.0$; $T_\omega = 0.05$; $T_L = -4.0$ N·m)

spectively. Same change of speed commands are tested under the load torque $T_L = -4.0$ N·m. In any cases, quick responses of N_r are obtained. The experimental results agree well with the computed results too. The high frequency ripples of N^* in Fig. 14(c) are smaller than those in Fig. 12(c). The reason for this difference can be considered that the amplitude modulation ratio of regenerating operation is smaller if the voltage

control has allowance under limited DC-bus voltage; the distortion of stator currents is reduced. The larger we choose the gain $|K_\omega|$, the larger high frequency ripples of N^* are induced. From this point, we have selected the gain $|K_\omega|$ to 5.0.

5. Conclusions

The conclusions drawn from this study are summarized as follows:

- (1) We have proposed a new simplified speed-sensorless vector control method of IM.
- (2) The flux angle computation and the torque and speed control are realized by using the output voltage of d -axis PI current controller with a decoupling control.
- (3) A linear model is derived and the root loci obtained the model is used to design the control parameters.
- (4) It is demonstrated that the proposed system can realize stable operation in both motoring and regenerating modes by the result of root loci.
- (5) The experimental results agree with those of non-linear simulation and the usefulness of proposed method is verified.

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Appendix

$$A = \begin{bmatrix} -a_1 - \frac{K_P}{\sigma L_s} & 0 & \frac{M}{\sigma L_s L_r \tau_r} \\ -\omega^* - i_{sd} K_\omega K_P & -a_1 + \frac{1}{\sigma \tau_r^*} - \frac{1}{\tau_r^*} & -\frac{M \omega_r}{\sigma L_s L_r} \\ \frac{M}{\tau_r} + \psi_{rq} K_\omega K_P & \frac{\psi_{rq}}{\tau_r^* i_{sd}^*} & -\frac{1}{\tau_r} \\ -\psi_{rd} K_\omega K_P & \frac{M}{\tau_r} - \frac{\psi_{rd}}{\tau_r^* i_{sd}^*} & \omega_r - \omega^* \\ -a_2 \psi_{rq} & a_2 \psi_{rd} & a_2 i_{sq} \\ -K_i & 0 & 0 \\ -\frac{K_\omega K_P}{T_\omega} & 0 & 0 \end{bmatrix}^*$$

$$* = \begin{bmatrix} \frac{M \omega_r}{\sigma L_s L_r} & \frac{M \psi_{rq}}{\sigma L_s L_r} & \frac{1}{\sigma L_s} & 0 \\ \frac{M}{\sigma L_s L_r \tau_r} & -\frac{M \psi_{rd}}{\sigma L_s L_r} & i_{sd} K_\omega & \frac{i_{sd}^*}{\sigma} \\ \omega^* - \omega_r & -\psi_{rq} & -\psi_{rq} K_\omega & 0 \\ -\frac{1}{\tau_r} & \psi_{rd} & \psi_{rd} K_\omega & 0 \\ -a_2 i_{sd}^* & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & \frac{K_\omega}{T_\omega} & 0 \end{bmatrix}$$

$$a_1 = \frac{R_s}{\sigma L_s} + \frac{M^2}{\sigma L_s L_r \tau_r}, \quad a_2 = \frac{M P^2}{4 J L_r}$$

$$B = \begin{bmatrix} 0 \\ \frac{i_{sd}^*}{\sigma} - i_{sd} \\ \psi_{rq} \\ -\psi_{rd} \\ 0 \\ 0 \\ 0 \end{bmatrix}, \quad B_L = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ -\frac{P}{2J} \\ 0 \\ 0 \end{bmatrix}$$

Mineo Tsuji (Senior Member) was born in December 1953. He received the Ph.D. degree from Kyushu University in March 1981 in electrical engineering, became a lecturer at Nagasaki University in April 1981 and became a Professor in July 2001 in the Division of Electrical Engineering and Computer Science, Graduate School of Engineering, Nagasaki University. His current interests are control of AC machines and electrical power systems. Dr. Tsuji is a member of IEEE. He received a Prize Paper Award from the IEEJ in 1998.

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