A New Method of Sensorless Vector Control for Induction Motors Using Instantaneous Reactive Power

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This paper proposes a new method of sensorless vector control for induction motors, in which the motor speed is estimated by both instantaneous reactive power and variation of magnetizing current. The proposed system is constructed in a synchronously rotating reference frame and is very simple. Another advantage is that the proposed system can work in regenerating mode without any special technique. A linear model is deduced to analyze system stability. According to the trajectories of poles and zeroes, the system gain is selected to be suited for operating in a wide speed range and in motoring or regenerating mode. The feasibility of the proposed system is verified by digital simulation.

Keywords: induction motor, sensorless vector control, instantaneous reactive power, speed estimation, stability analysis

1. Introduction

Induction motors have been predominantly the most widely used motors in industrial and domestic applications for reasons of simplicity, cost, robustness and reliability. Speed sensor used in the most effective vector-controlled schemes does present problems. It lowers the system reliability, especially in hostile environments, and requires careful cabling arrangement with special attention to electrical noise ⁽¹⁾. In addition, a speed sensor is difficult to be mounted on motor shaft for some special purpose. To overcome these problems, a large variety of different schemes without speed sensor has been reported ^{(2)~(4)}. Based on stator terminal measurements, the rotor speed can be estimated by an equivalent circuit model of the induction motor. The estimated speed accuracy and the system stability are greatly under the influence of the model parameters, especially stator and rotor resistance ⁽⁵⁾. In addition, speed sensorless control schemes increase software intricacy more or less and require high-performance processors. However, sensorless vector control is

necessary to reduce system complication and to increase reliability for applicability ⁽⁴⁾.

To overcome the above problems, an approach of indirect vector control to estimate motor speed by instantaneous reactive power is proposed ⁽⁶⁾. An advantage of this approach is that the speed estimation is robust against the stator resistance variation. Another advantage is achieving much wider-bandwidth speed control. However, a disadvantage is that this approach is unstable in regenerating mode ⁽⁷⁾. To solve this problem, an approach with the Extended Kalman Filter algorithm is proposed ⁽⁷⁾. However, this approach is complicated. Stability analysis of the whole system is difficult because the above two approaches are constructed in a stationary reference frame.

In this paper, we propose a new method and study the system stability ⁽⁸⁾. The motor speed is estimated by both instantaneous reactive power and variation of magnetizing current. This instantaneous reactive power is obtained from the vector control theory ⁽⁹⁾. Therefore, the new system is simpler than Peng's one. Another advantage is that this system can do its work even in

regenerating mode without any complicated technique. Because the new system is constructed in a synchronously rotating reference frame, a linear model of the whole system is easy to be deduced. Then, the influence of this system gain is fully taken into consideration in different operating conditions (at low or high speed, in motoring or regenerating mode). According to the analytical investigation, various kinds of control gains are selected to be suited for the system different operations. The feasibility of the proposed system is verified by digital simulation.

2. Description of Proposed System

An induction motor can be described by voltage and current models in a synchronously rotating reference frame as follows:

$$e_{sd} = (r_s + \sigma L_s p)i_{sd} - \omega^* \sigma L_s i_{sq} + e_{md} \quad (1)$$

$$e_{sq} = \omega^* \sigma L_s i_{sd} + (r_s + \sigma L_s p)i_{sq} + e_{mq} \quad (2)$$

$$0 = -\sigma_r M i_{sd} + \sigma_r \psi_{rd} + \omega_r \psi_{rq} + \frac{L_r}{M} e_{md} \quad (3)$$

$$0 = -\sigma_r M i_{sq} - \omega_r \psi_{rd} + \sigma_r \psi_{rq} + \frac{L_r}{M} e_{mq} \quad (4)$$

where, i_{sd} , i_{sq} : d - q axis stator current;

e_{sd} , e_{sq} : d - q axis stator voltage;

ψ_{rd} , ψ_{rq} : d - q axis flux linkage;

ω_r : rotor angular speed;

ω^* : electrical angular speed;

p : differential operator;

$\sigma_r = r_r / L_r$, $\sigma = 1 - M^2 / (L_s L_r)$.

And the other symbols used in the above equations are shown in Table 1. The counter EMF e_{md} and e_{mq} can be expressed as

$$e_{md} = \frac{M}{L_r} (p \psi_{rd} - \omega^* \psi_{rq}) \quad (5)$$

$$e_{mq} = \frac{M}{L_r} (p \psi_{rq} + \omega^* \psi_{rd}) \quad (6)$$

According to the vector control theory, a synchronously rotating reference frame is chosen for $\psi_{rq} = 0$. The instantaneous reactive power into the rotor is defined as

$$q_m = i_{sd} e_{mq} - i_{sq} e_{md} \quad (7)$$

The actual currents i_{sd} and i_{sq} can be replaced with i_{sd}^* and i_{sq}^* if the current control is ideal.

Table 1. Parameter of 3-phase IM

Parameter	Symbol	Value
Rated output power		2.2 kW
Rated torque		12.2 N·m
Rated Current		8.6 A
Rated rotational speed		1720 rpm
Moment of Inertia	J	0.0671 kg · m ²
Pole number	P	4
Stator resistance	r_s	0.0662 Ω
Rotor resistance	r_r	0.0645 Ω
Stator self inductance	L_s	0.086 H
Rotor self inductance	L_r	0.086 H
Magnetizing inductance	M	0.082 H

From eq.(1) and eq.(2), the instantaneous reactive power of the voltage model used in the controller can be written as

$$q_m = i_{sd}^* \{ e_{sq} - \omega^* \sigma L_s i_{sd}^* - (r_s^* + \sigma L_s p) i_{sq}^* \} - i_{sq}^* \{ e_{sd} - r_s^* i_{sd}^* + \omega^* \sigma L_s i_{sq}^* \} \quad (8)$$

where i_{sd}^* is constant. From eq.(3), the d -axis flux in the controller is constant and can be written as

$$\psi_{rd}^* = M i_{sd}^* \quad (9)$$

According to the above description, the counter EMF of the current model can be expressed as

$$\hat{e}_{md} = 0 \quad (10)$$

$$\hat{e}_{mq} = \frac{M^2}{L_r} \omega^* i_{sd}^* \quad (11)$$

Therefore, the instantaneous reactive power of the current model can be written as

$$\hat{q}_m = \frac{M^2}{L_r} \omega^* i_{sd}^{*2} \quad (12)$$

From eq.(4), the flux speed can be written as

$$\omega^* = \hat{\omega}_r + \sigma_r^* \frac{i_{sq}^*}{i_{sd}^*} \quad (13)$$

Here $\hat{\omega}_r$ is the estimated speed in a sensorless vector control system. We propose a new method to estimate rotor speed as follows:

$$q_m - \hat{q}_m = e_{sq} i_{sd}^* - e_{sd} i_{sq}^* - \omega^* L_s (i_{sd}^{*2} + \sigma i_{sq}^{*2}) - \sigma L_s i_{sd}^* p i_{sq}^* \quad (14)$$

$$\begin{aligned} \hat{\omega}_r = & K_{\omega p} \left(1 + \frac{1}{T_{\omega i p}} \right) (q_m - \hat{q}_m) \\ & + K_e \left(1 + \frac{1}{T_{e p}} \right) (i_{sd}^* - i_{sd}) \end{aligned} \quad (15)$$

As the speed command changes greatly, the motor operating state changes between motoring mode and regenerating mode temporarily. It

causes variations in the d -axis actual current and the d -axis rotor flux. To decrease the error between the actual speed and the estimated speed, the error of the d -axis current is added to estimate the rotor speed. In addition, the speed estimation is robust against stator resistance because eq.(14) does not contain stator resistance. In order to avoid the differentiation of the q -axis current in eq.(14), pi_{sq}^* is approximated by a new variable pq as

$$q \equiv \frac{1}{1+T_q p} i_{sq}^* \quad (16)$$

$$pi_{sq}^* \approx pq = \frac{i_{sq}^*}{T_q} - \frac{q}{T_q} \quad (17)$$

Fig.1 shows the proposed system. The speed PI control is expressed as

$$i_{sq}^* = K_s \left(1 + \frac{1}{T_s p}\right) (\omega_r^* - \hat{\omega}_r) \quad (18)$$

The current PI control is written as

$$e_{sd}^* = K_p \left(1 + \frac{1}{T_{ip}}\right) (i_{sd}^* - i_{sd}) - \omega^* \sigma L_s i_{sq}^* \quad (19)$$

$$e_{sq}^* = K_p \left(1 + \frac{1}{T_{ip}}\right) (i_{sq}^* - i_{sq}) + \omega^* L_s i_{sd}^* \quad (20)$$

3. Stability Analysis

The analysis of the proposed system shown in Fig.1 is presented. To simplify the stability analysis, the motor voltages are assumed to be controlled ideally, or more precisely,

$$e_{sa} = e_{sa}^*, \quad e_{sb} = e_{sb}^*, \quad e_{sc} = e_{sc}^* \quad (21)$$

To derive the equation for IM in the d - q axis revolving synchronously with ω^* , eq.(21) can be rewritten as

$$e_{sd} = e_{sd}^*, \quad e_{sq} = e_{sq}^* \quad (22)$$

3.1 Steady-State Solution

The steady-state solution of the proposed system can be obtained by setting $p=0$. The estimated rotor speed is set equal to the speed command. Load torque T_L , i_{sd}^* and ω_r^* can be given arbitrarily. For convenience of analysis, a given T_L is replaced by a given slip frequency. From eq.(3) and eq.(4), the fluxes of IM can be calculated as follows:

$$\psi_{rd} = \frac{\sigma_r M [\sigma_r i_{sd}^* + (\omega^* - \omega_r) i_{sq}^*]}{\sigma_r^2 + (\omega^* - \omega_r)^2} \quad (23)$$

$$\psi_{rq} = \frac{\sigma_r M [\sigma_r i_{sq}^* - (\omega^* - \omega_r) i_{sd}^*]}{\sigma_r^2 + (\omega^* - \omega_r)^2} \quad (24)$$

Fig.1. Block diagram of the proposed system

The stator voltages can be calculated from eq.(1) and eq.(2) as follows:

$$e_{sd} = r_s i_{sd}^* - \omega^* \sigma L_s i_{sq} - \frac{M}{L_r} \omega^* \psi_{rq} \quad (25)$$

$$e_{sq} = \omega^* \sigma L_s i_{sd}^* + r_s i_{sq} + \frac{M}{L_r} \omega^* \psi_{rd} \quad (26)$$

Therefore, i_{sq}^* can be determined by the following equation.

$$e_{sq} i_{sd}^* - e_{sd} i_{sq}^* - \omega^* L_s (i_{sd}^{*2} + \sigma i_{sq}^{*2}) = 0 \quad (27)$$

3.2 Linear Model

The system stability is analyzed by a linear model that is deduced from the above nonlinear equations at a steady-state operating point. The linear model of IM can be derived from eq.(1) to eq.(4) and the equation of motion.

$$p \Delta \mathbf{x}_s = \mathbf{A}_s \Delta \mathbf{x}_s + \mathbf{B}_s \Delta \mathbf{u}_s + \mathbf{B}_L \Delta T_L \quad (28)$$

where, $\Delta \mathbf{x}_s = [\Delta i_{sd} \quad \Delta i_{sq} \quad \Delta \psi_{rd} \quad \Delta \psi_{rq} \quad \Delta \omega_r]^T$

$$\Delta \mathbf{u}_s = [\Delta e_{sd} \quad \Delta e_{sq} \quad \Delta \omega^*]^T$$

The linear model of the controller is expressed as

$$p \Delta \mathbf{z} = \mathbf{A}_z \Delta \mathbf{x}_s + \mathbf{A}_z \Delta \mathbf{z} + \mathbf{B}_z \Delta \mathbf{u}_s + \mathbf{B}_r \Delta \omega_r^* \quad (29)$$

$$\Delta \mathbf{u}_s = \mathbf{F}_x \Delta \mathbf{x}_s + \mathbf{F}_z \Delta \mathbf{z} + \mathbf{F}_r \Delta \omega_r^* \quad (30)$$

where, $\Delta \mathbf{z} = [\Delta e_\omega \quad \Delta e_s \quad \Delta e_{vd} \quad \Delta e_{vq} \quad \Delta q \quad \Delta \hat{\omega}_r]^T$

The state variables Δe_ω , Δe_s , Δe_{vd} and Δe_{vq} are caused by PI controllers for the speed estimation, speed control, d - q axis current control, respectively. In order to deduce the state equation, a first order lag element is added to calculate $\hat{\omega}_r$ in eq.(15). The above matrixes are given in Appendix. From eq.(28), eq.(29) and eq.(30), the linear model of the whole system can be written as

$$p \Delta \mathbf{x} = \mathbf{A} \Delta \mathbf{x} + \mathbf{B} \Delta \omega_r^* + \mathbf{B}_L' \Delta T_L \quad (31)$$

where, $\Delta \mathbf{x} = [\Delta \mathbf{x}_s^T \quad \Delta \mathbf{z}^T]^T$

Fig. 2. Root trajectories of poles and zeros

$$A = \begin{bmatrix} A_s + B_s F_x & B_s F_z \\ A_x + B_z F_x & A_z + B_z F_z \end{bmatrix}$$

$$B = \begin{bmatrix} B_s F_r \\ B_z F_r + B_r \end{bmatrix}, \quad B_L = \begin{bmatrix} B_L \\ 0 \end{bmatrix}$$

The speed command is chosen as the input variable and the actual speed is chosen as the output variable as follows:

$$\Delta \omega_r = C \Delta x \quad (32)$$

So, the speed transfer function can be given as

$$G(s) = \frac{\Delta \omega_r}{\Delta \omega_r^*} = \frac{C \text{adj}(sI - A) B}{|sI - A|} \quad (33)$$

3.3 Results of the Analysis

In the following study, the parameters of 3-phase induction motor given in Table 1 are used.

Fig.2 (a) shows the root trajectories of poles and zeros on the s-plane calculated by eq.(33), in which $K_{\omega p}$ changes from 0.01 to 1. When $K_{\omega p}$ increases, the pole p_5 is located in an unstable region and the system becomes unstable, as No. 5 of p_5 in Fig.2 (a). A suitable value of $K_{\omega p}$ is selected to be 0.1. Fig.2 (b) shows the root trajectories in which K_e changes from -60 to 20. The pole p_1 moves into an unstable region when K_e is a positive. So, the value of K_e must be smaller than zero. Fig.2 (c) shows the root trajectories, in which motor speed is 1800 rpm and slip frequency is 80 rpm or -80 rpm (full load). The pole p_5 moves into a stable region when K_e changes from -20 to -10, as No. 1 and No. 2 of p_5 . If $K_{\omega p}$ is a negative, another pole moves into a stable region in regenerating mode, as No. 3 and No. 4 of p_3 . From this

Fig.3. Trajectories of poles and zeros with variation of stator or rotor resistance

figure, we know that a K_e of -10 is suited to operate at higher speed and a K_{wp} of -0.1 is suited to operate in regenerating mode.

Fig.2 (d) shows the root trajectories, in which motor speed is 80 rpm and slip frequency is -80 rpm or 80 rpm. From this figure, we know that the system is more stable if T_e changes from 0.01 to 0.001 and T_{wi} changes from 0.02 to 0.002. If the motor is operating in regenerating mode and a K_{wp} is -0.1, No. 3 of the pole p_2 moves into a stable region, as No. 4 of this pole. As a conclusion, a set of adaptive system gain shown in Table 2 is selected for operating in a wide speed range and in motoring or regenerating mode.

Fig.3 (a) shows the root trajectories, in which the actual value of stator resistance r_s changes from $0.6 r_{so}$ to $1.6 r_{so}$, r_{so} is the nominal value of stator resistance. The stator resistance r_s has no influence over the system stability because r_s is not used for speed estimation. Fig.3 (b) shows the root trajectories in which the actual value of rotor resistance r_r changes from $0.6 r_{ro}$ to $1.6 r_{ro}$, and r_{ro} is the nominal value of rotor resistance. The p_2 shown in this figure will move into an unstable region when r_r is smaller than $0.6 r_{ro}$. However, this case can be avoided in practice.

Fig.4 shows the operation area in which the system gain shown in Table 2 is used for calculation and rotor resistance changes from $0.8 r_{ro}$ to $1.5 r_{ro}$. Although there is an unstable area in lower-speed range when the slip frequency is a negative, the proposed system can work in a wide speed range, as shown in this figure.

Table 2. Adaptive PI gain for this system

$\hat{N}_r$ [rpm]	K_e	T_e	$T_{\omega i}$	$K_{\omega p}$	
				$i_{sq}^* \geq 0$	$i_{sq}^* < 0$
1800 }	-10.0	0.01	0.02	0.1	-0.1
800 }					
500 }	-20.0	0.001	0.002		
200 }					
0

Fig.4. Unstable area with r_r

4. Digital Simulation

To study the performance of the system during transients, the most common approach is to use the differential equations of IM model. For verifying the proposed system, the nonlinear model is used for simulation. Variables of the motor part are still in a continuous system and calculated by

the Runge-Kutta method. The control part is digitized by the backward Euler method and the sampling period T_c is 256 μ s. The dead time caused by the digitized controller is equal to T_c and considered in the following simulation. For example, eq.(15) can be rewritten as follows:

$$\begin{aligned}\hat{\omega}_r^k = & \hat{\omega}_r^{k-1} + K_{ap}(1 + \frac{T_c}{T_{ai}})(q_m^{k-1} - \hat{q}_m^{k-1}) \\ & - K_{ap}(q_m^{k-2} - \hat{q}_m^{k-2}) - K_e(i_{sd}^* - i_{sd}^{k-2}) \\ & + K_e(1 + \frac{T_c}{T_e})(i_{sd}^* - i_{sd}^{k-1})\end{aligned}\quad (34)$$

Fig.5 shows the transient responses for the step change of speed command. Fig.5 (a) shows the simulated result that K_e is zero and the step change of speed command is 200 rpm. When the speed command steps down to 500 rpm at 6 second, the actual speed can not converge to 500 rpm. Although the motor is still desired to work in motoring mode, the torque-producing current is a negative. This is because only the information of instantaneous reactive power is not enough to control an induction motor in regenerating mode. This result is similar to that reported by Kubota *et al*⁽⁷⁾. To overcome this problem, both of the information of instantaneous reactive power and the variation of magnetizing current are used to estimate motor speed. Fig.5 (b) shows the result that K_e is -20 and the slip frequency is 80 rpm. Fig.5 (c) shows the result that the slip frequency is -80 rpm. From these figures, it is confident that the proposed method can work in both motoring and regenerating mode.

As stator resistance increases with the inner temperature of IM, the actual value of r_s is greater than the nominal value when the motor is in operation. In the following simulations, the actual value of r_s is set to be 1.5 r_{s0} .

Fig.6 shows the result that the motor operates at lower speed and the step change of speed command is 40 rpm. Fig.6 (a) is the result that the slip frequency is 80 rpm. Fig.6 (b) is the result that the slip frequency is -30 rpm. Fig.7 shows the result that the motor operates at higher speed and the step change of speed command is 200 rpm. Fig.7 (a) is the result that the motor works in motoring mode. Fig.7 (b) is the result that the motor works in regenerating mode. From Fig.6 and Fig. 7, we can know that the proposed system can

Fig.5. Transient responses for step change of N_r

(a) Motoring mode

(b) Regenerating mode

Fig.6. Transient responses at lower speed

(a) Motoring mode

(b) Regenerating mode

Fig.7. Transient responses at higher speed

(a) Motoring mode

(b) Regenerating mode

Fig.8. Transient responses when $r_r = 0.8r_{r0}$

work in a wide speed range and is suited to operate in both motoring and regenerating mode.

Fig.8 shows the result that rotor resistance r_r is $0.8 r_{ro}$. Fig.8 (a) and Fig.8 (b) are the results that the motor operates in motoring mode or in regenerating mode, respectively. From these two figures, we know that the proposed system is stable even if r_r is smaller than r_{ro} . The speed accuracy of the system is under the influence of rotor resistance variation. The above simulation results confirm the validity of the proposed system and the inherent robustness to stator resistance. In addition, the system gain must be changed depending upon the motor speed, as shown in Table 2, but it is not very strict. In a practical control system, a hysteresis band of the motor speed can be used to ensure smooth transition between the adjacent gains. Another method that the gains as a smooth function of the motor speed are stored in a look-up table can be also used.

5. Conclusion

In this paper, a new method of speed sensorless vector control for induction motors is proposed. The motor speed is estimated by both instantaneous reactive power and variation of magnetizing current. A linear model of the whole system is deduced to analyze system stability. A set of system gain is selected to be suited for operating in the different conditions. Digital simulation shows that the proposed system can work in a much wide speed range and in motoring or regenerating mode without any special technique.

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Appendix

$$A_s = \begin{bmatrix} a_3 & \omega^* & a_1 \sigma_r M & a_1 \omega_r M & a_1 M \psi_{rq} \\ -\omega^* & a_3 & -a_1 \omega_r M & a_1 \sigma_r M & a_1 M \psi_{rd} \\ \sigma_r M & 0 & -\sigma_r & \omega^* - \omega_r & -\psi_{rq} \\ 0 & \sigma_r M & \omega_r - \omega^* & -\sigma_r & \psi_{rd} \\ -a_2 \psi_{rq} & a_2 \psi_{rd} & a_2 i_{sq} & -a_2 i_{sd} & -R_\omega / J \end{bmatrix}$$

$$a_1 = \frac{1}{\sigma L_s L_r}, \quad a_2 = \frac{P^2 M}{4 J L_r}, \quad a_3 = -a_1 (r_s L_r + \sigma_r M^2)$$

$$B_s = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 & i_{sq} \\ 0 & \frac{1}{\sigma L_s} & -i_{sd} \\ 0 & 0 & \psi_{rq} \\ 0 & 0 & -\psi_{rd} \\ 0 & 0 & 0 \end{bmatrix}, \quad B_L = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ -\frac{P}{2J} \end{bmatrix}$$

$$A_z = \begin{bmatrix} 0 & -\frac{a_4 K_{\omega p}}{T_{\omega i}} & 0 & 0 & \frac{\sigma L_s i_{sd}^* K_{\omega p}}{T_q T_{\omega i}} & \frac{a_4 K_{\omega p} K_s}{T_{\omega i}} \\ 0 & 0 & 0 & 0 & 0 & -\frac{K_s}{T_s} \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & \frac{K_p}{T_i} & 0 & 0 & 0 & -\frac{K_s K_p}{T_i} \\ 0 & \frac{1}{T_q} & 0 & 0 & -\frac{1}{T_q} & -\frac{K_s}{T_q} \\ \frac{1}{T_\omega} & -\frac{a_4 K_{\omega p}}{T_\omega} & \frac{K_e T_i}{K_p T_e T_\omega} & 0 & \frac{\sigma L_s i_{sd}^* K_{\omega p}}{T_q T_\omega} & a_5 \end{bmatrix}$$

$$a_4 = e_{sd}^* + 2\sigma L_s \omega^* i_{sq}^* + \sigma L_s i_{sd}^* / T_q$$

$$a_5 = -1/T_\omega + a_4 K_{\omega p} K_s / T_\omega$$

$$A_x = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ -\frac{K_p}{T_i} & 0 & 0 & 0 & 0 \\ 0 & -\frac{K_p}{T_i} & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ -\frac{K_e}{T_\omega} & 0 & 0 & 0 & 0 \end{bmatrix}$$

$$B_z = \begin{bmatrix} -\frac{K_{\omega p}}{T_{\omega i}} i_{sq}^* & \frac{K_{\omega p}}{T_{\omega i}} i_{sd}^* & -\frac{K_{\omega p}}{T_{\omega i}} a_6 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \\ -\frac{K_{\omega p}}{T_\omega} i_{sq}^* & \frac{K_{\omega p}}{T_\omega} i_{sd}^* & -\frac{K_{\omega p}}{T_\omega} a_6 \end{bmatrix}$$

$$a_6 = L_s i_{sd}^{*2} + \sigma L_s i_{sq}^{*2}$$

$$B_r = \begin{bmatrix} -\frac{a_4 K_{\omega p} K_s}{T_{\omega i}} & \frac{K_s}{T_s} & 0 & \frac{K_4 K_p}{T_i} & \frac{K_s}{T_q} & -\frac{a_4 K_{\omega p} K_s}{T_\omega} \end{bmatrix}^T$$

$$F_x = \begin{bmatrix} -K_p & 0 & 0 & 0 & 0 \\ 0 & -K_p & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \end{bmatrix}, \quad F_r = \begin{bmatrix} -a_7 K_s \\ K_s (K_p + \sigma_r^* L_s) \\ \sigma_r^* K_s / i_{sd}^* \end{bmatrix}$$

$$a_7 = \sigma L_s (\omega^* + \sigma_r^* i_{sq}^* / i_{sd}^*)$$

$$F_z = \begin{bmatrix} 0 & -a_7 & 1 & 0 & 0 & a_7 K_s - \sigma L_s i_{sq}^* \\ 0 & K_p + \sigma_r^* L_s & 0 & 1 & 0 & L_s i_{sd}^* - K_s (K_p + \sigma_r^* L_s) \\ 0 & \sigma_r^* / i_{sd}^* & 0 & 0 & 0 & 1 - \sigma_r^* K_s / i_{sd}^* \end{bmatrix}$$

$$C = [0 \ 0 \ 0 \ 0 \ 1 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$$

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