Research Article

Low-Speed Stability Optimization of Full-Order Observer for Induction Motor

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In terms of the instability of the full-order observer for the induction motor in the low-speed regenerative mode, the low-speed unstable region which leads to the extension of the commissioning cycle cannot be eliminated by the traditional adaptive law which aims at good system performance. It is proposed that the feedback gain matrix can control both the unstable region and the system performance both. To make a trade-off between the stability and performance by designing the feedback gain matrix is still an open problem. To solve this problem, first we analyze the cause of instability and derive constraints to ensure system stability by establishing a transfer function of the adaptive observing system for the speed. Then, with the derived constraints as the design criteria for the feedback gain matrix, a control strategy combining the weighted adaptive law with the improved feedback gain matrix is proposed to improve the stability at low speed. Finally, by comparing the traditional control strategy with the proposed control strategy through simulations and experiments, we show that the proposed control strategy achieves better performance with higher stability.

1. Introduction

The speed-sensorless vector control system of the induction motor abandons the photoelectric encoder and other traditional motor speed measurement devices, which reduces the cost of the system and enhances the reliability of system operation. At present, among speed identification methods for the speed-sensorless induction motor, the direct calculation method [1] directly uses the mathematical model of the induction motor for speed open-loop estimation. Although the structure is simple, this method features poor anti-interference ability and low-speed identification accuracy. The model reference adaptive control method [2, 3] takes the voltage model as an adjustable one that has a simple principle. However, the pure integrator in the voltage model causes DC bias and error in integral initial value, which leads to poor performance at low speed. The high frequency signal injection method [4] eliminates the problem of poor low-speed performance of the model reference adaptive control

method by taking advantage of the salient pole rotors. However, it depends heavily on the structural design of the motor and is not practical enough. In the adaptive full-order observer method [5], a state equation of the rotor-flux linkage and the stator current is established to predict the state of the motor in real time for the induction motor. The difference between the estimated value and the measured value of the stator current state is corrected and input by the gain matrix, and the estimated state is corrected in real time by feedback correction, thus forming a closed-loop state estimation to improve the performance of the speed identification system.

As a widely used speed identification tool, the adaptive full-order observer is unstable in the low-speed regenerative mode. To address this problem, there are many works on improving the speed identification system. In References [5, 6], the rotor-flux linkage error is ignored in the process of deriving the speed adaptive law using Popov's hyperstability theory. Although the immeasurability of the rotor-flux

linkage is considered, when the motor runs at low speed, the rotor-flux linkage error increases significantly, which results in inaccurate speed identification. In Reference [7], the rotor-flux linkage error is compensated in the adaptive law, which improves the accuracy and dynamic performance of the speed identification system. However, in the design of the weight coefficient of the rotor-flux linkage error in the scheme, filtering processing is required, which leads to an increase in system complexity. Since the poles of the motor model are in the left half plane of the s-plane, the model itself is stable [8]. In Reference [9], it is proposed that the poles of the full-order observer should be set on the left side of the motor pole. The scheme can improve the convergence speed of the full-order observer to a certain extent by setting a reasonable feedback gain matrix. However, the stability of the low-speed regenerative mode is still not effectively solved. In Reference [10], the transfer function of the open-loop full-order observer is analyzed, and the unstable region under the low-speed regenerative mode is given. In Reference [11], the regenerative instability problem is solved by improving the feedback gain matrix, but the pole position of the full-order observer is moved to the position close to the origin, which reduces the convergence speed of the system. References [12–17] provide a new idea for speed-sensorless performance optimization at low speed, but its algorithm is not practical due to its complexity.

In view of the shortcomings of the improved adaptive law [5–7] and the feedback gain matrix [8–11] of the full-order observer, an improved method combining the adaptive law with the feedback gain matrix is proposed to improve the dynamic performance and low-speed stability of the system, by introducing an adaptive law compensation method with adjustable weight coefficient and simplifying the feedback gain matrix with low-speed stability as the design criteria. The feasibility and effectiveness of this control strategy are supported by theoretical analyses and simulations.

2. Mathematical Model of Full-Order Observer for Induction Motor

With stator current and rotor-flux linkage of the induction motor as state variables, the state equation of the induction motor in the static coordinate system is given by

$$\begin{cases} \frac{d}{dt}x = Ax + Bu_s, \\ y = Cx. \end{cases} \quad (1)$$

By formula (1), the state equation of the full-order observer is obtained as follows:

$$\begin{cases} \frac{d}{dt}\hat{x} = \hat{A}\hat{x} + Bu_s + G(\hat{y} - y), \\ \hat{y} = C\hat{x}, \end{cases} \quad (2)$$

where $C = [I \ 0]$ is the output matrix, and the feedback gain matrix is as follows:

$$\begin{aligned} G &= [G_1 \ G_2]^T = [g_1I + g_2J \ g_3I + g_4J]^T, \\ A &= \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} = \begin{bmatrix} a_{11}I & a_{12}I + a_{12}'J \\ a_{21}I & a_{22}I + a_{22}'J \end{bmatrix} \\ &= \begin{bmatrix} \left(\frac{\delta-1}{\delta T_r} - \frac{R_s}{\delta L_s}\right)I & \frac{L_m}{\delta L_s L_r T_r}I - \frac{L_m \omega_r}{\delta L_s L_r}J \\ \frac{L_m}{T_r}I & -\frac{1}{T_r}I + \omega_r J \end{bmatrix}, \end{aligned} \quad (3)$$

in which $B = \begin{bmatrix} B_1 \\ 0 \end{bmatrix}$, $B_1 = (1/\delta L_s)I$, $I = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$, $J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$, $\delta = 1 - (L_m^2/L_s L_r)$, $T_r = (L_r/R_r)$, and $x = [i_s \ \psi_r]^T$ are the state variables, $y = i_s$ is the output variable, $i_s = [i_{s\alpha} \ i_{s\beta}]^T$ is the stator current, $\psi_r = [\psi_{r\alpha} \ \psi_{r\beta}]^T$ is the rotor-flux linkage, $u_s = [u_{s\alpha} \ u_{s\beta}]^T$ is the stator voltage, ω_r is the rotor speed, R_r and R_s are the rotor resistance and stator resistance, L_r , L_s , and L_m are the rotor inductance, stator inductance, and mutual inductance. The superscript “ $\wedge$ ” indicates the observed value.

An error equation is obtained by subtracting the state equation (1) of the induction motor from the state equation (2) of the full-order observer as follows:

$$\frac{d}{dt} \begin{bmatrix} e_i \\ e_\psi \end{bmatrix} = (A + GC) \begin{bmatrix} e_i \\ e_\psi \end{bmatrix} + \Delta\omega_r \begin{bmatrix} 0 & -\frac{J}{\varepsilon} \\ 0 & J \end{bmatrix} \begin{bmatrix} \hat{i}_s \\ \hat{\psi}_r \end{bmatrix}, \quad (4)$$

where $e_i = i_s - \hat{i}_s$, $e_\psi = \psi_r - \hat{\psi}_r$, and $\varepsilon = \delta L_s L_r / L_m$.

The speed adaptive law [18] can be obtained from the state error equation (4) by using Lyapunov stability theorem:

$$\begin{cases} \dot{\hat{\omega}}_r = K_p(\varepsilon_1 - \varepsilon_2) + K_i \int (\varepsilon_1 - \varepsilon_2)dt, \\ \varepsilon_1 = (e_{i_{s\alpha}} \hat{\psi}_{r\beta} - e_{i_{s\beta}} \hat{\psi}_{r\alpha}), \\ \varepsilon_2 = (e_{\psi_{s\alpha}} \hat{\psi}_{r\beta} - e_{\psi_{s\beta}} \hat{\psi}_{r\alpha}). \end{cases} \quad (5)$$

Note that it is impossible to obtain actual rotor-flux linkage, if it is assumed that the estimated flux linkage is equal to the actual flux linkage, $\varepsilon_2 = 0$, and the traditional speed adaptive law is obtained:

$$\dot{\hat{\omega}}_r = K_p \varepsilon_1 + K_i \int \varepsilon_1 dt. \quad (6)$$

When the motor operates at the medium-high speed, the flux linkage error term is small, which has little impact on the estimation of flux linkage when it is ignored. However,

when the motor operates at low speed, the rotor-flux linkage error will increase significantly, which leads to inaccurate observation.

3. Design of Speed Adaptive Law

3.1. Observer Based on Traditional Adaptive Law. By applying Laplace transform in the state error equation (4), we obtain

$$s \begin{bmatrix} e_i \\ e_\psi \end{bmatrix} = (A + GC) \begin{bmatrix} e_i \\ e_\psi \end{bmatrix} + \Delta\omega_r J \hat{\psi}_r \begin{bmatrix} -J \\ \varepsilon \\ J \end{bmatrix}, \quad (7)$$

where s is the differential divisor.

A closed-loop system composed of the error equation and speed adaptive link can be established by formulas (6) and (7). The system structure of this system is shown in Figure 1.

As shown in Figure 1, the input of the transfer function of the linear time-invariant forward path is $\Delta\omega_r J \hat{\psi}_r$. The output is the stator current error e_i , and the formula below is obtained:

$$G(s) = \frac{e_i}{\Delta\omega_r J \hat{\psi}_r}. \quad (8)$$

To facilitate the analysis of the stability of the full-order observer, the state error formula (7) is transformed into the rotor-flux linkage-oriented synchronously rotating coordinate system:

$$s \begin{bmatrix} e_i \\ e_\psi \end{bmatrix} = (A' + GC) \begin{bmatrix} e_i \\ e_\psi \end{bmatrix} + \Delta A' \begin{bmatrix} \hat{i}_s \\ \hat{\psi}_r \end{bmatrix}, \quad (11)$$

where $A' = \begin{bmatrix} A_{11} - \omega_1 J & A_{12} \\ A_{21} & A_{22} - \omega_1 J \end{bmatrix}$ and $\Delta A' = A' - \hat{A}'$. The state variables are the components under synchronously rotating coordinate systems m and t .

If the transfer function of the forward path is expressed by $G'(s)$ in coordinate systems m and t , formula (8) can be transformed into the following [10]:

$$\begin{bmatrix} e_{ism} \\ e_{ist} \end{bmatrix} = \begin{pmatrix} G'_{11}(s) & G'_{12}(s) \\ G'_{21}(s) & G'_{22}(s) \end{pmatrix} \begin{bmatrix} 0 \\ \hat{\psi}_r \end{bmatrix} \Delta\omega_r. \quad (12)$$

The elements of the transfer function $G'(s)$ matrix can be obtained by error equations under synchronously rotating coordinate systems [11].

The transfer function from m -axis component of stator current error to speed difference is expressed by $G'_m(s)$. The

FIGURE 1: Traditional observer structure in the stationary coordinate system.

By expanding formula (7) in s domain, the following formula is obtained:

$$\begin{cases} sIe_i = (A_{11} + G_1)e_i + A_{12}e_\psi - \frac{\Delta\omega_r}{\varepsilon} J \hat{\psi}_r, \\ sIe_\psi = (A_{11} + G_1)e_i + A_{12}e_\psi - \Delta\omega_r J \hat{\psi}_r. \end{cases} \quad (9)$$

Specific expression of transfer function of the linear time-invariant forward path is obtained by eliminating e_ψ in the simultaneous equations (9):

$$G(s) = \frac{s}{\varepsilon} \left[s^2 I - s(A_{11} + G_1 + A_{22}) + A_{22} \left(A_{11} + G_1 + \frac{A_{21} + G_2}{\varepsilon} \right) \right]^{-1}. \quad (10)$$

transfer function from t -axis component of stator current error to speed difference is expressed by $G'_t(s)$:

$$\begin{cases} G'_m(s) = \frac{e_{ism}}{\Delta\omega_r} = G'_{12}(s) \hat{\psi}_r, \\ G'_t(s) = \frac{e_{ist}}{\Delta\omega_r} = G'_{22}(s) \hat{\psi}_r. \end{cases} \quad (13)$$

The adaptive law equation is obtained by transforming the traditional adaptive law into coordinate systems m and t by coordinate transformation, as shown in the following equation:

$$\hat{\omega}_r = - \left(K_p + K_i \int dt \right) (i_{st} - \hat{i}_{st}) \psi_{rm}. \quad (14)$$

The structure diagram of the traditional full-order observer in the synchronously rotating coordinate system can be obtained by synthesizing equations (13) and (14), as shown in Figure 2.

3.2. Design of Improved Speed Adaptive Law. It can be seen from Figure 2 that the traditional adaptive full-order observer is a closed-loop system with single input and single output. In the closed-loop system, only the torque current

FIGURE 2: System structure diagram of the traditional full-order observer.

error component is involved in speed identification, and excitation current error component is not a part of the speed identification system.

By introducing the excitation current error component into the traditional speed identification system, equation (14) can be modified as follows:

$$\hat{\omega}_r = -\left(K_p + K_i \int dt\right) \left[(i_{st} - \hat{i}_{st})\psi_{rm} + M(i_{sm} - \hat{i}_{sm}) \right]. \quad (15)$$

If $M = L_r \psi_{st}$ and the introduced compensation term $M(i_{sm} - \hat{i}_{sm})$ is transformed into the static coordinate system, the compensation term is approximately equal to ε_2 [18]. So, it is the negligence of the flux linkage error term in the adaptive law of the traditional speed identification system that leads to the lack of excitation current error component in the synchronously rotating coordinate system, resulting in the inaccurate low-speed observation.

Considering that the actual value of rotor-flux linkage cannot be measured in actual application, the rotor-flux linkage error term ε_2 in the static coordinate system is transformed into detectable stator current:

$$\begin{aligned} \varepsilon_2 &= (e_{\psi_{sa}} \hat{\psi}_{r\beta} - e_{\psi_{s\beta}} \hat{\psi}_{r\alpha}) = \psi_{r\alpha} \hat{\psi}_{r\beta} - \hat{\psi}_{r\alpha} \psi_{r\beta} \\ &= \frac{\psi_{r\alpha} \hat{\psi}_{r\beta} - \hat{\psi}_{r\alpha} \psi_{r\beta}}{|\psi_r| \cdot |\hat{\psi}_r|} |\psi_r| |\hat{\psi}_r| \\ &= (\cos \theta \sin \hat{\theta} - \sin \theta \cos \hat{\theta}) |\psi_r| |\hat{\psi}_r| = \sin \Delta \theta |\psi_r| |\hat{\psi}_r|, \end{aligned} \quad (16)$$

where $|\psi_r|$ is the rotor-flux linkage vector module value and $\Delta \theta$ is the difference between the observed rotor-flux linkage vector angle $\hat{\theta}$ and the actual rotor-flux linkage vector angle θ .

The rotor-flux linkage vector angle difference can be replaced by the stator current vector angle difference [19]:

$$\sin \Delta \theta = \frac{i_{sa} \hat{i}_{s\beta} - \hat{i}_{sa} i_{s\beta}}{|\hat{i}_s| |\hat{i}_s|}. \quad (17)$$

By introducing equation (17) into equation (16), the following equation is obtained:

$$\varepsilon_2 = \frac{i_{sa} \hat{i}_{s\beta} - \hat{i}_{sa} i_{s\beta}}{|\hat{i}_s| |\hat{i}_s|} |\psi_r| |\hat{\psi}_r| = H(i_{sa} \hat{i}_{s\beta} - \hat{i}_{sa} i_{s\beta}), \quad (18)$$

where H is the weight coefficient. The accuracy and dynamic performance of the observer can be improved by adjusting the H value [20]. The typical value of parameter h can be designed as shown in the following equation:

$$\begin{cases} 0 < H < 0.05, & (\omega_1 < 0), \\ H = 0, & (\omega_1 \geq 0). \end{cases} \quad (19)$$

4. Stability Analysis and Improvement of Observer

4.1. Analysis of the Unstable Range for Full-Order Observer.

In theory, the stability of the full-order observer can be improved by weighting and compensating the adaptive law. However, the commissioning cycle will be extended, and there is a great blindness if the weight coefficient is adjusted in real time based on open-loop observation ($G = 0$). In addition, to improve the convergence speed of full-order observer speed identification, the open-loop gain is usually set to a large value. Considering that the root locus of the closed-loop transfer function starts from the open-loop pole and eventually tends to the open-loop zero point, an unreasonable weight coefficient will lead to a positive real part of the open-loop zero point of the observer, which causes instability as the closed-loop root locus of the full-order observer tends to open-loop zero point due to the large open-loop gain.

To analyze the unstable region of the open-loop observer and reasonably configure the feedback gain matrix to form a closed-loop full-order observer to eliminate the low-speed unstable region, the transfer function (10) of the linear time-invariant forward path can be simplified as follows:

$$G(s) = \frac{s}{\varepsilon} [s^2 I - s(aI + bI) + cI + dI]^{-1}, \quad (20)$$

where

$$\begin{aligned} a &= -g_1 + \frac{R_r}{\delta L_r} + \frac{R_s}{\delta L_s}, \\ b &= -g_2 - \omega_r, \\ c &= -\frac{R_r}{L_r} \left(-\frac{R_s}{\delta L_s} + g_1 + \frac{g_3}{\varepsilon} \right) - \omega_r \left(g_2 + \frac{g_4}{\varepsilon} \right), \\ d &= \omega_r \left(-\frac{R_s}{\delta L_s} + g_1 + \frac{g_3}{\varepsilon} \right) - \frac{R_r}{L_r} \left(g_2 + \frac{g_4}{\varepsilon} \right). \end{aligned} \quad (21)$$

According to Popov's hyperstability theorem, to ensure the asymptotic stability of the speed identification system, the transfer function of the linear time-invariant forward path should be a strictly positive real function:

$$G(j\omega) + G^*(j\omega) > 0, \quad \forall \omega > 0. \quad (22)$$

By introducing equation (20) and $s = j\omega_1$ into equation (20), a simplified equation is obtained:

$$\begin{cases} a > 0, \\ \omega_1^2 > \left(-\frac{d}{a}\right)^2, \end{cases} \quad (23)$$

where ω_1 is the synchronous angular frequency, so $\omega_c = -d/a$ is the critical angular frequency.

Formula (23) is the stability condition of the speed identification system, and the constraint condition $a > 0$ is naturally satisfied under open-loop observation ($G = 0$). If the motor operates in the forward rotation state and the synchronous frequency is positive, the unstable region of the open-loop observation speed identification system is as follows:

$$0 < \omega_1 < \omega_c = \frac{(R_s/\delta L_s)}{(R_s/\delta L_s) + (R_r/\delta L_r)} \omega_r < \omega_r. \quad (24)$$

The relationship between the electromagnetic torque and the speed of the induction motor is presented as follows:

$$T_e = n_p \frac{\psi_r^2}{R_r} (\omega_1 - \omega_r). \quad (25)$$

By introducing the boundary condition of the unstable region into equation (25), it is obtained that

$$\begin{cases} T_e = -n_p \frac{\psi_r^2}{R_r} \omega_r, \\ T_e = -n_p \frac{R_r/\delta L_r}{R_r/\delta L_r + R_s/\delta L_s} \frac{\psi_r^2}{R_r} \omega_r. \end{cases} \quad (26)$$

The graph of the unstable region is plotted with electromagnetic torque and speed, as shown in Figure 3(a). The shaded part in the figure is the unstable region, and the expression of the boundary line is shown in expression (26). In this case, the actual speed is greater than the synchronous speed and the slip frequency is negative, which means that the motor is in the dynamic braking state (unstable state).

4.2. Stability Improvement of Full-Order Observer. From the stability constraint expression (23), the stability of the full-order observer is subjected to the design of the feedback gain matrix. The stability of the observer can be improved by configuring a feedback gain matrix. To meet the low-speed stability requirements of the motor operation, the critical angular frequency ω_c is set to zero. At this point, the two boundary lines in Figure 3(a) coincide and the unstable region disappears, as shown in Figure 3(b). The stability constraint can be simplified as follows:

$$\begin{cases} g_1 < \frac{R_s}{\delta L_s} + \frac{R_r}{\delta L_r}, \\ \frac{R_r}{L_r} \left(g_2 + \frac{g_4}{\varepsilon}\right) = \omega_r \left(-\frac{R_s}{\delta L_s} + g_1 + \frac{g_3}{\varepsilon}\right). \end{cases} \quad (27)$$

According to this principle, the elements of the feedback gain matrix can be configured as follows [11]:

$$\begin{cases} g_1 = \frac{R_s L_r^2 + R_r L_m^2}{\delta L_s L_r^2} - k \frac{R_r}{L_r}, \\ g_2 = -k \omega_r, \\ g_3 = -\frac{L_m R_r}{L_r}, \\ g_4 = 0. \end{cases} \quad (28)$$

where k is the ratio of the observer pole to motor pole.

According to this design scheme, although global stability is achieved, the observer pole position is moved to the position close to the origin, which reduces the convergence speed of the system.

It can be seen from expression (28) that since the feedback gain matrix itself is time-varying and constantly updated, complicated element design will inevitably reduce its convergence performance. Therefore, in this paper, the feedback gain matrix is simplified.

$$\begin{cases} g_1 = k \frac{R_s L_r^2 + R_r L_m^2}{\delta L_s L_r^2}, \\ g_2 = 0, \\ g_3 = -\frac{L_m R_r}{L_r}, \\ g_4 = 0. \end{cases} \quad (29)$$

The final design scheme of the adaptive full-order observer can be obtained by synthesizing expressions (5), (18), and (29), as shown in Figure 4. The design scheme not only solves the problem of low-speed instability by reasonably designing the gain matrix but also improves the dynamic performance of the system by combining with the improved weighted adaptive law.

5. System Simulations and Experiments

5.1. System Simulations. In this paper, simulation of the decoupling vector control system of the full-order observer-based induction motor is carried out, and the simulation model of the control algorithm is constructed using MATLAB/SIMULINK, as shown in Figure 5.

In the simulation model, basic parameters of the induction motor are set as follows: $u_N = 380$ V, $P_N = 3*746$ W, $f = 50$ Hz, $R_s = 0.435$ Ω , $R_r = 0.816$ Ω , $L_m = 0.069$ H, $L_{lr} = L_{ls} = 0.002$ H, $J = 0.01$ kg·m², and $p = 2$.

Figures 6(a) and 6(b) are the speed waveforms of the traditional full-order observer control strategy and the

FIGURE 3: Diagram of the asynchronous motor $\omega_r - T_e$ based on (a) open-loop speed observer and (b) closed-loop speed observer.

FIGURE 4: The improved system structure diagram of the adaptive full-order observer.

FIGURE 5: Control block diagram of the system.

FIGURE 6: Speed waveform diagram of the control system at high speed. (a) Speed waveform of the traditional observer. (b) Improved full-order observer speed waveform.

improved full-order observer control strategy at high speed. From the speed graphs of two control strategies, in the high-speed and no-load state, the motor speed rises steadily to 1500 r/min in 0.25 s, and the overshoot of the improved full-order observer is lower than that of the traditional observer. At this point, the actual speed curve and the estimated speed curve of the two control strategies basically coincide, and both speed identification systems can accurately track the real speed.

In the low-speed regenerative braking mode, the given speed is set to 100 r/min and the given flux linkage to 0.9 Wb. From formula (25) we know that the critical value of power-generating load is -27 N·m. As a result, the load applied to the motor is set to -30 N·m.

Figure 7 is the speed waveform of the control system in the low-speed regenerative mode. To verify the stability of the control system under the regenerative state, the power-generating load is used for simulation experiment. As shown in the figure, the motor starts with no load, and then the speed is maintained at 100 r/min. At 0.5 s, the power-generating load of -30 N·m is suddenly applied to the motor. As the load applied exceeds the critical value, the traditional observer enters the unstable region. The observed speed becomes divergent and no longer converges to the actual speed, while the improved full-order observer converges to the actual speed stably. This is consistent with previous theoretical analysis, proving that the improved full-order observer control system has good low-speed stability.

Figure 8 shows the component diagram of rotor-flux linkage of the control system in the low-speed regenerative mode. When the power-generating load is suddenly applied at 0.5 s, the flux linkage of the traditional observer diverges, while the flux linkage of the improved full-order observer has accurate estimation without DC bias and error in integral initial value of open-loop estimation.

Figures 9 and 10 are the speed waveforms and their partial enlarged drawings of the improved full-order observer when load is added or reduced at low speed. In the low-speed state, the motor starts at no load and then steadily rises to 100 r/min at low speed. At 0.4 s, the load torque of

the motor steps from 0 to -30 N·m; at 0.6 s, the load torque steps from -30 N·m to 30 N·m. In this process, the estimated speed still tracks the actual speed in real time, showing that the control system has good dynamic performance when the load is added or reduced.

Figure 11(a) is the speed switch waveform at low speed. At 0.4 s, the given speed of the control system is stepped from 50 r/min to 30 r/min and from 30 r/min to 10 r/min at 0.6 s. In the figure, the control system not only can operate stably at extremely low speed but also has fast speed and small overshoot in the switching process. As shown in Figure 11(b), after the flux linkage is stabilized, the influence speed change is neglectable. It can be seen that the improved control scheme not only improves the dynamic performance but also has good low-speed stability.

5.2. System Experiments. The improved control algorithm is tested on a 5 kW induction motor doubly-fed platform, as shown in Figure 12. Motor 1 is the test motor, and Motor 2 the load motor. Some parameters of the motors in the experiment are as follows: $u_N = 380$ V, $P_N = 5$ kW, $f = 50$ Hz, $I_N = 11.1$ A, $p = 2$, $n_N = 1440$ r/min. In the test, Motor 1 works in the speed identification state and uses the speed obtained from speed identification to conduct closed-loop vector control. The stability of the control system at low speed is verified by observing the actual speed and estimated speed of Motor 1.

Figure 13 shows the three-phase stator current waveform of the induction motor at a low speed of 100 r/min. The three-phase stator current waveform at low speed is symmetrical and basically stable.

Figure 14 shows the waveform of the actual speed. When the given speed is switched from 600 r/min to 200 r/min and 100 r/min, respectively, the dynamic performance of the system is good during the whole process, and the motor operation is still stable when switched to the low-speed mode, which proves the effectiveness of the improved control strategy.

FIGURE 7: Speed waveform diagram of the control system in the low-speed regeneration mode. (a) Speed waveform of the traditional observer. (b) Improved full-order observer speed waveform.

FIGURE 8: Rotor-flux component diagram under the low-speed regeneration mode of the control system. (a) Rotor-flux linkage diagram of the traditional observer. (b) Improved rotor-flux linkage diagram of the full-order observer.

FIGURE 9: Waveform diagram of load speed at low speed after Improvement.

FIGURE 10: Local enlargement diagram of load addition and subtraction at improved low speed.

FIGURE 11: The improved waveform of speed switch and flux at low speed. (a) Speed switching waveform at low speed. (b) Flux linkage diagram under speed switching.

FIGURE 12: Asynchronous motor test platform.

FIGURE 13: Three-phase stator current waveform of the asynchronous motor at low speed.

FIGURE 14: Waveform of actual speed and speed.

6. Conclusion

In this paper, a control strategy for low-speed stability optimization of the induction motor based on the full-order observer is proposed. The low-speed instability of the full-order observer in the speed identification system is analyzed. The feedback gain matrix is designed to eliminate the unstable region of the control system, and the feedback gain matrix is simplified to improve the convergence speed. Combined with the weighted adaptive law, the good dynamic and static performance of the control system is achieved. The simulation results show that the control strategy can improve the stability at low speed and increase the accuracy of speed identification.

Data Availability

The raw/processed data required to reproduce these findings cannot be shared at this time as the data also form part of an ongoing study.

Conflicts of Interest

The authors declare that they have no conflicts of interest.

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