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A SENSORLESS DIRECT TORQUE CONTROL INDUCTION MOTOR DRIVER WITH A FULL-ORDER FLUX OBSERVER BASED ON PARTICLE SWARM OPTIMIZATION PARAMETER ESTIMATORS

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Abstract. *A sensorless direct torque control induction motor drive based on full-order stator flux observer is presented. The identified rotor speed is acquired stably from the proposed full order stator flux observer, and the speed feedback gain parameters of the stator flux observer are designed by the particle swarm optimization (PSO) algorithm to improve that the speed feedback gain parameters are determined by try and error. The simulation and experiment of the proposed control scheme is established by toolbox Matlab/Simulink® and PC_based platform and the results confirmed the correctness of the proposed system.*

Keywords: Induction motor drive, Direct torque control (DTC), Speed sensorless, Full-order flux observer, Particle swarm optimization (PSO)

1. **Introduction.** The dynamic of induction motors is nonlinear and variant, so the control of the motors is more difficult. By comparison with the conventional induction motors control, the sensorless DTC technology just has two control loops, flux loop and torque loop, and is much simpler and can reduce the complexity of the control to a large application [1,2]. Latest literature shows that the field speed and slip speed are often estimated to calculate the rotor speed in DTC applications by the field oriented control (FOC) [3]. The full-order observer and the model reference adaptive system (MRAS) architecture are also used to determine the stability of the rotor speed estimator [4-8]. To determine feedback gains is always a key problem using MRAS architecture to identify rotor speed and mismatch feedback gain parameters always cause the instability of sensorless direct torque control induction motor drive. One of the latest relative research is designed feedback gain parameters with constant coefficient [6], so the speed estimator cannot correct work in all speed range. Others are usually designed feedback gain parameters by try and error [5], and it has not even been discussed [4,7,8]. However, the literature only focuses on the estimated strategy of observer gain. The PSO algorithm was presented by Kennedy and Eberhart in 1995, and is a Swarm Intelligence. It is suitable for dynamic parameter estimators with fast convergence and fewer setting parameters, and is widely used in optimization problems [9-11]. The PSO algorithm is presented to improve the correctness of rotor speed estimator and reduce the procedure of determining correct feedback gain parameters in this paper.

2. **Principle.**

2.1. **Mathematical dynamic equation of the induction motor.** Let the stator flux and the stator current be set as status variables $\mathbf{x}$ and then dynamic equations of the induction motor reference to a dual-axis (d-q) synchronous rotational coordinate system are expressed by Equation (1) and Equation (2) is denoted as output function.

$$\dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}\mathbf{u}, \quad \mathbf{x} = \begin{bmatrix} \mathbf{i}_s^e \\ \boldsymbol{\varphi}_s^e \end{bmatrix} \tag{1}$$

$$i_s^e = [1 \quad 0]\mathbf{x} = C\mathbf{x} \tag{2}$$

wherein $\mathbf{A} = \begin{bmatrix} -\left(\frac{R_s}{L_\sigma} + \frac{1}{\sigma\tau_r} - j\omega_r\right) & \frac{1}{L_\sigma}\left(\frac{1}{\tau_r} - j\omega_r\right) \\ -R_s & 0 \end{bmatrix}$, $\mathbf{B} = \begin{bmatrix} \frac{1}{L_\sigma} \\ 1 \end{bmatrix}$, ω_r represents the rotor electrical speed, R_s represents the stator resistance, L_s and L_r represent the stator inductance and rotor inductance, $L_\sigma = \sigma L_S$ represents the leakage inductance coefficient and $\sigma = L_m^2/L_S L_r$. Furthermore, let $\mathbf{i}_s^e = i_{ds}^e + j i_{qs}^e$, $\boldsymbol{\varphi}_s^e = \phi_{ds}^e + j\phi_{qs}^e$ represents the d-q components of status variables $\mathbf{x}$ in synchronous rotational coordinate; and after substituting them into Equation (1), a differential expression using $i_{ds}^e, i_{qs}^e, \phi_{ds}^e, \phi_{qs}^e$ as state variables can be expressed by Equation (3).

$$p\begin{bmatrix} i_{ds}^e \\ i_{qs}^e \\ \phi_{ds}^e \\ \phi_{qs}^e \end{bmatrix} = \begin{bmatrix} -\left(\frac{R_s}{L_\sigma} + \frac{1}{\sigma\tau_r}\right) & \omega_e - \omega_r & \frac{1}{L_\sigma\tau_r} & \frac{\omega_r}{L_\sigma} \\ -(\omega - \omega_r) & -\left(\frac{R_s}{L_\sigma} + \frac{1}{\sigma\tau_r}\right) & -\frac{\omega_r}{L_\sigma} & \frac{1}{L_\sigma\tau_r} \\ -R_s & 0 & 0 & \omega_e \\ 0 & -R_s & -\omega_e & 0 \end{bmatrix}\begin{bmatrix} i_{ds}^e \\ i_{qs}^e \\ \phi_{ds}^e \\ \phi_{qs}^e \end{bmatrix} + \begin{bmatrix} \frac{v_{ds}^e}{L_\sigma} \\ \frac{v_{qs}^e}{L_\sigma} \\ v_{ds}^e \\ v_{qs}^e \end{bmatrix} \tag{3}$$

Accordingly, the mechanical equation of the motor is expressed by Equation (4).

$$T_e = J_m p\omega_{rm} + B_m\omega_{rm} + T_L \tag{4}$$

In the equation, J_m represents the motor inertia, B_m represents the viscous friction coefficient, and T_L represents the load torque. The electromagnetic torque T_e is expressed by Equation (5), and P represents the polar pair.

$$T_e = \frac{3}{4}P\left(\phi_{ds}^e i_{qs}^e - \phi_{qs}^e i_{ds}^e\right) \tag{5}$$

2.2. **The principle of rotor speed estimator with full-order flux observer.** According to latest literature, the stator flux oriented vector control technique often uses an open loop to estimate the rotor speed and flux. To obtain the better controlling quality, this study uses a closed-loop full-order flux observer to solve the problems incurred by the open loop architecture and submits a method of the estimators. The study chooses the stator current and the stator flux as state variables to build the flux observer whereby the rotor speed can be estimated, as shown in Equation (6) redeemed by Equation (3) plus the full-order observer gain $G_1 \sim G_4$, thereby creating a closed loop architecture to tune the stability of the rotor speed estimator [5].

$$\begin{aligned} p\begin{bmatrix} \hat{i}_{ds}^e \\ \hat{i}_{qs}^e \\ \hat{\phi}_{ds}^e \\ \hat{\phi}_{qs}^e \end{bmatrix} &= \begin{bmatrix} -\left(\frac{R_s}{L_\sigma} + \frac{1}{\sigma\tau_r}\right) & (\omega_e - \hat{\omega}_r) & \frac{1}{L_\sigma\tau_r} & \frac{\hat{\omega}_r}{L_\sigma} \\ -(\omega - \hat{\omega}_r) & -\left(\frac{R_s}{L_\sigma} + \frac{1}{\sigma\tau_r}\right) & -\frac{\hat{\omega}_r}{L_\sigma} & \frac{1}{L_\sigma\tau_r} \\ -R_s & 0 & 0 & \omega_e \\ 0 & -R_s & -\omega_e & 0 \end{bmatrix}\begin{bmatrix} \hat{i}_{ds}^e \\ \hat{i}_{qs}^e \\ \hat{\phi}_{ds}^e \\ \hat{\phi}_{qs}^e \end{bmatrix} + \begin{bmatrix} \frac{1}{L_\sigma} & 0 \\ 0 & \frac{1}{L_\sigma} \\ 1 & 0 \\ 0 & 1 \end{bmatrix}\begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix} \\ &\quad + G\left(\mathbf{i}_s^e - \hat{\mathbf{i}}_s^e\right) \end{aligned} \tag{6}$$

In the equation, "ˆ" represents the estimated values of the state variable, $G = [G_1 \; G_2 \; G_3 \; G_4]^T$ represents the observer gain matrix of the observer. Equation (6) is further expressed by a spatial vector expression as follows:

$$p\hat{\mathbf{x}} = \mathbf{A}\hat{\mathbf{x}} + \mathbf{B}u + \mathbf{G}\left(\mathbf{i}_s^e - \hat{\mathbf{i}}_s^e\right) \tag{7}$$

Equation (7) is an estimation equation. If an output equation is defined as

$$\hat{\mathbf{i}}_s^e = [1 \quad 0]\hat{\mathbf{x}} = C\hat{\mathbf{x}} \tag{8}$$

If an error vector obtained is written with $\mathbf{e} = \mathbf{x} - \hat{\mathbf{x}}$, linearize Equations (1) and (7) to obtain

$$p\mathbf{e} = (\mathbf{A} - \mathbf{GC})\mathbf{e} + \left(\mathbf{A} - \hat{\mathbf{A}}\right)\hat{\mathbf{x}} \tag{9}$$

wherein $|\hat{\boldsymbol{\varphi}}_s^e| = \left|\hat{\Phi}_s^e\right|$. Based on the MRAS architecture, the rotor speed estimator is shown in Figure 1, and the rotor speed estimator is defined as

$$\hat{\omega}_r = -\left(K_p + K_i \int dt\right)\varepsilon \tag{10}$$

wherein

$$\varepsilon = \left(p\hat{\mathbf{i}}_s^e J\right)^T pe_i = p\hat{i}_{ds}^e pe_{iq}^e,$$

$$G_q(s) = \frac{s^3 + m_1 s^2 + \left(y + \omega_e^2\right)s + \omega_e(m + \omega_e m_1)}{\left[s^2 + m_1 s + \left(y - \omega_e w - \omega_e^2\right)\right]^2 + \left[(m_2 + 2\omega_e)s + (m + \omega_e m_1)\right]^2},$$

$$m = \frac{G_4}{L_\sigma \tau_r} - \omega_r(G_3 + R_s), \quad m_1 = G_1 + \frac{R_s}{L_s} + \frac{1}{\sigma\tau_r}, \quad w = G_2 - \omega_r, \quad J = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix},$$

$$y = \frac{1}{L_\sigma \tau_r}(G_3 + R_s) + \omega_r G_4,$$

K_p and K_i represent the feedback gain parameters and are determined by try and error in latest literature.

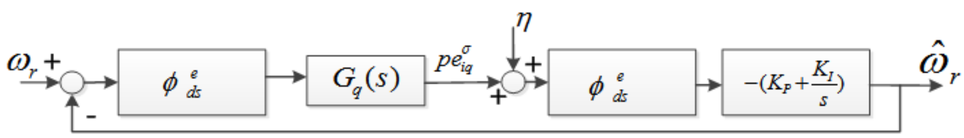


FIGURE 1. The architecture of rotor speed estimator

2.3. **Direct torque control.** The control principle of DTC is to calculate the three-phase voltage of the motor by the switching state of the inverter and the DC bus voltage V_{dc} firstly and the three-phase voltage of the motor is expressed by

$$V_{as} = \frac{v_{dc}}{3}(2S_a - S_b - S_c) \tag{11}$$

$$V_{bs} = \frac{v_{dc}}{3}(-S_a + 2S_b - S_c) \tag{12}$$

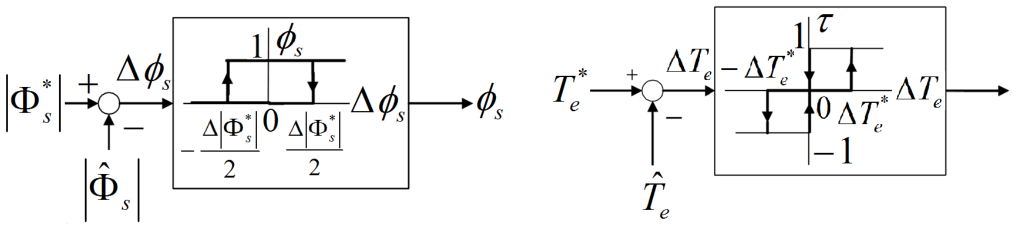
$$V_{cs} = \frac{v_{dc}}{3}(-S_a - S_b + S_c) \tag{13}$$

wherein the power switch, S_X, the value is 1 or 0 that represents the power switch, power on or cutoff state, when the switch is $x = a, b, c$. The three-phase current of the motor can be measured by a Hall-effect-based current sensor. After reference to a dual-axis (d-q) static rotational coordinate system, use

$$\hat{\boldsymbol{\varphi}}_s^s = \hat{\phi}_{ds}^s + \hat{\phi}_{qs}^s = \frac{\tau_c}{1 + p\tau_c}(\mathbf{v}_s^s - R_s\mathbf{i}_s^s) + \frac{\tau_c}{1 + p\tau_c}\boldsymbol{\varphi}_s^* \tag{14}$$

Estimating the stator flux, $\hat{\omega}_e = \frac{d}{dt}\tan^{-1}\left(\hat{\phi}_{qs}^s \middle/ \hat{\phi}_{ds}^s\right)$ represents the field electrical synchronous speed and τ_c represents the time constant. Compare these two values with the torque command and the stator flux command $|\Phi_s^*|$ respectively to obtain the torque error $\Delta T_e = T_e^* - \hat{T}_e$ and the stator flux error $\Delta\phi_s = |\Phi_s^*| - \left|\hat{\Phi}_s\right|$ and input them into a hysteresis controller of the inverter shown in Figure 2(a) and Figure 2(b) to decide the switching state of the power transistor of the inverter.

2.4. **PSO algorithm.** The PSO algorithm, the flow chart shown in Figure 3, has the merits of quick convergence and fewer parameters and is adapted to dynamic environments, so it is widely applied to optimization problems. This study uses the PSO algorithm to



(a) (b)

FIGURE 2. The block diagram of DTC control loop (a) stator flux Φ_s, (b) torque T_e

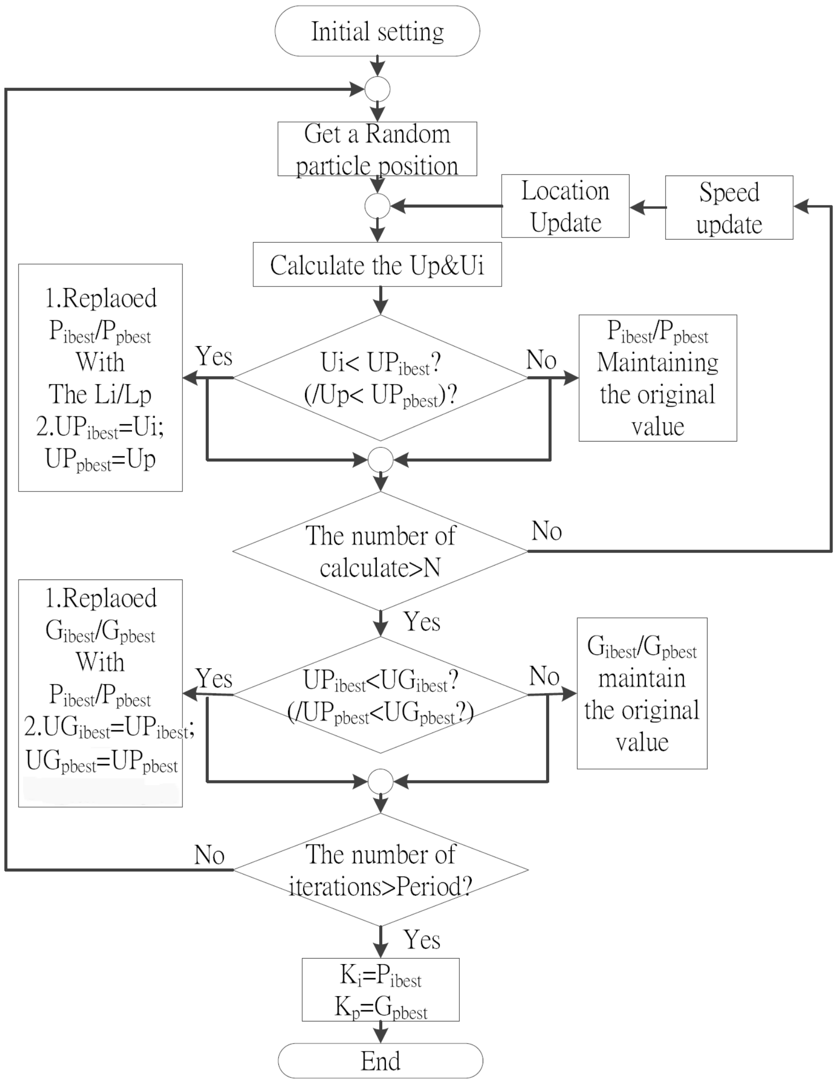


FIGURE 3. The flow chart of PSO

tune or adjust the feedback gain parameters of rotor speed estimator, and the algorithm of which is initially made by distributing particles randomly. Each of the particles firstly calculates a present fitness value and decides according to the personal best fitness value and the swarm's best fitness value. Then, update the speed and locations to calculate the particle fitness value, the algorithm of which is expressed by

$$V_i(k+1) = w \times V_i(k) + C_1 \times Rand \times (P_{best} - L_i) + C_2 \times Rand \times (G_{best} - L_i) \quad (15)$$

$$L_i(k+1) = L_i(k) + V_i(k+1) \quad (16)$$

wherein $V_i(k)$ represents the present particle speed and $V_i(k+1)$ is next particle speed, $L_i(k)$ represents the particle position now and $L_i(k+1)$ represents the next particle position, P_{best} represents the best location of the personal particle, G_{best} represents the best location of the global particle, C_1 represents the learning factor of the personal particle, C_2 represents the learning factor of the global particle, and Rand is an evenly distributed random constant between $[0.0, 1.0]$. To improve the convergence to the local solution efficiently, the study uses an inertia weight method [6] which is executed by adding a weight factor w into the conventional PSO and it will be reduced to 90% after iteration loop. The range of w setting this parameter is approximately set between $[0, 1.5]$ [7]. Otherwise, $Ui = Up = |\omega_r - \hat{\omega}_r|$, UP_{ibest} represents the smallest value of Ui, UP_{pbest} represents the smallest value of Up, UG_{ibest} represents the smallest value of UP_{ibest}, and UG_{pbest} represents the smallest value of UP_{pbest}. The learning factor in the PSO algorithm is divided into two parts, wherein the range of setting both of the learning factor of the personal particle C_1 and the learning factor of the global particle C_2 are approximately set between $[0, 4.0]$ [8].

3. **Results of Simulations and Experiments.** To verify the feasibility of the full-order flux observer, the PSO algorithm is used to tune the feedback gain parameters, and a full-order sensorless direct torque module built by MATLAB/Simulink®. The architecture shown in Figure 4 is used to simulate and analyze firstly. Then, the experiments platform, the architecture shown in Figure 4, drives a squirrel-cage induction motor (IM) drive that is 3-phase, 0.75kw, 220V and Δ-pin to different experiments with a PC-based adapter in the inverter and $K_p = G_{pbest}$ and $K_i = G_{ibest}$.

Finally, two speed commands are used, such as 300rpm and 1800rpm. Figures 5 and 6 show the responses of the study presented at T_e = 2N-m, N = 5, Period = 5, in

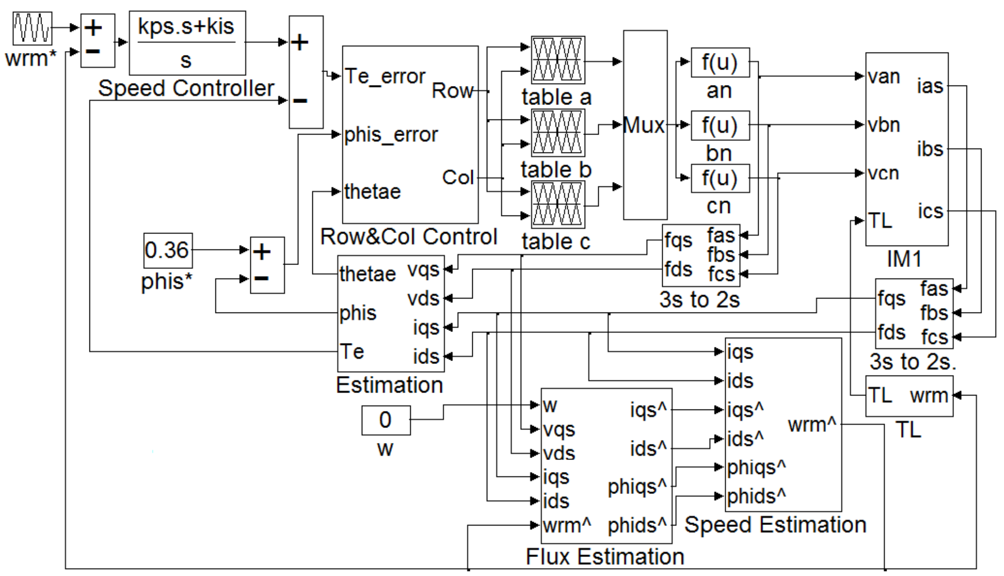


FIGURE 4. Architecture of the Simulink direct torque induction motor control

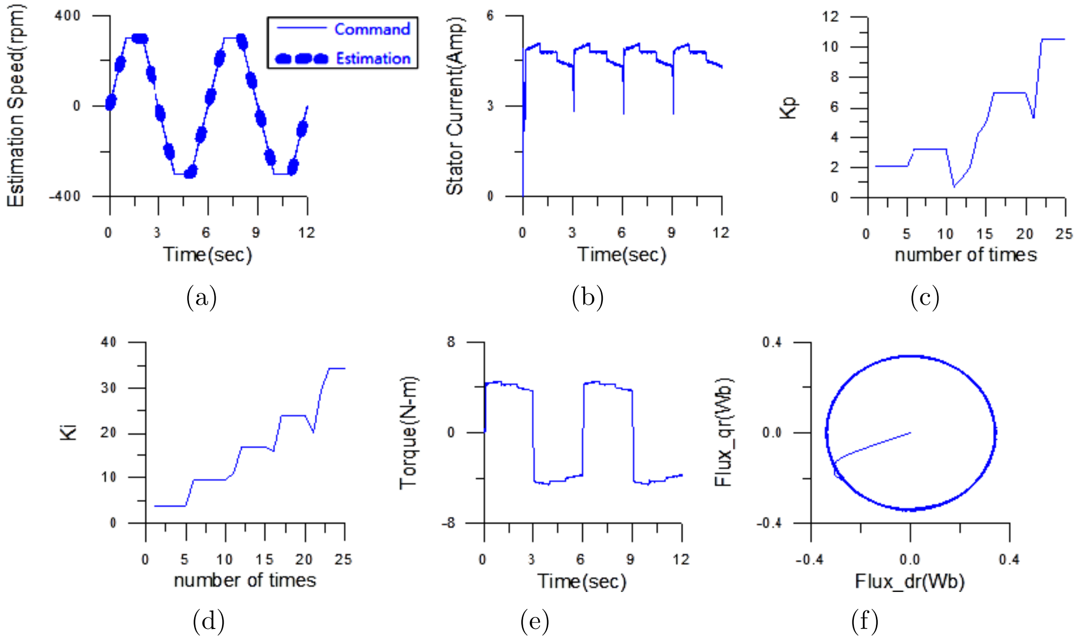


FIGURE 5. Simulation result at T_e = 2N-m, N = 5, Period = 5, ω_r = 300rpm: (a) rotor speed, (b) stator current, (c) K_p, (d) K_i, (e) estimated torque, (f) estimated stator fiux

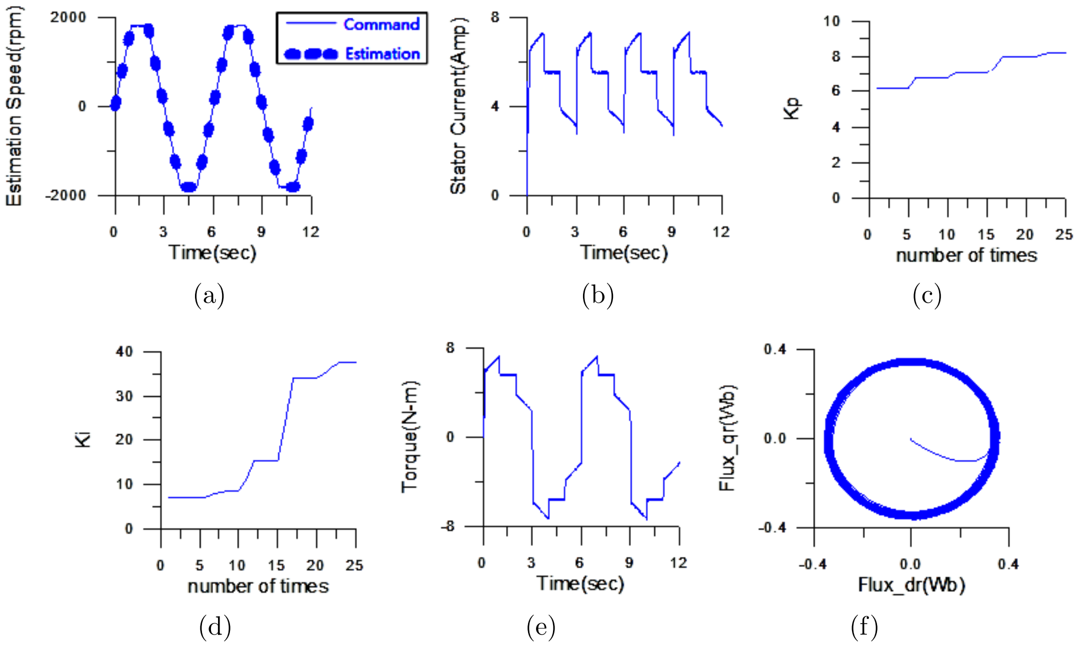


FIGURE 6. Simulation result at T_e = 2N-m, ω_r = 1800rpm, N = 5, Period = 5: (a) rotor speed, (b) stator current, (c) K_p, (d) K_i, (e) estimated torque, (f) estimated stator fiux

MATLAB/Simulink® and calculate $K_p = 10.5360$ and $K_i = 34.3965$ at 300rpm, $K_p = 8.1333$ and $K_i = 37.6765$ at 1800rpm, to the experiments. The experiments results are shown in Figures 7 and 8. Both verify the method the study presents can estimate the actual rotation speed in DTC and catch the commanded speed rapidly.

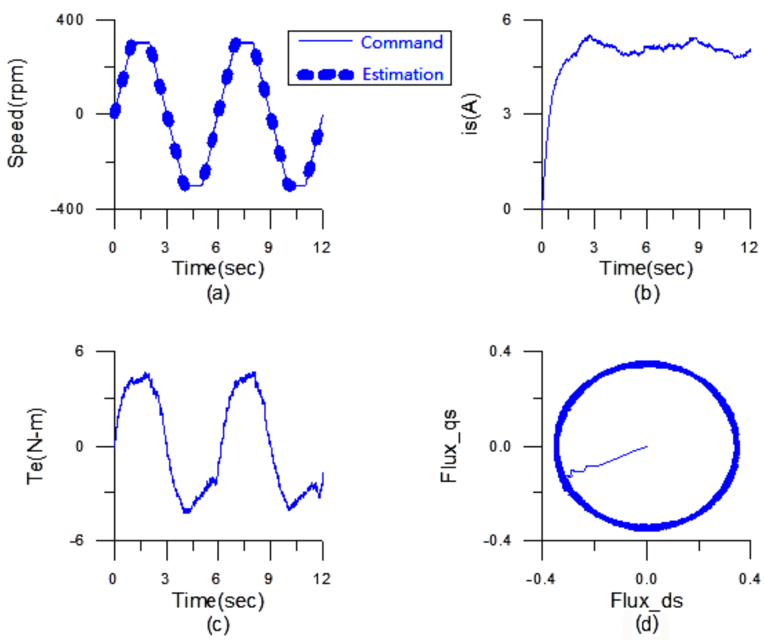


FIGURE 7. Experimental result at $T_e = 2$N-m, $\omega_r = 300$rpm, $K_p = 10.5360$, $K_i = 34.3965$: (a) rotor speed command and estimation, (b) stator current, (c) estimated torque, (d) estimated stator fiux

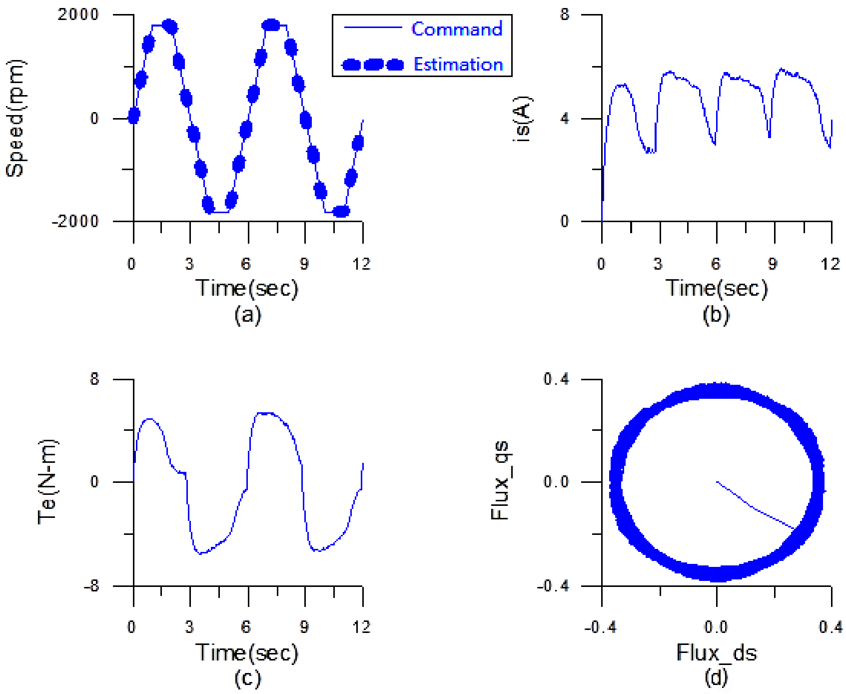


FIGURE 8. Experimental result at $T_e = 2$N-m, $\omega_r = 1800$rpm, $K_p = 8.1333$, $K_i = 37.6765$: (a) rotor speed command and estimation, (b) stator current, (c) estimated torque, (d) estimated stator fiux

4. **Conclusions.** This study applies a full-order stator flux observer to a sensorless direct torque control induction motor firstly. Then, the PSO algorithm is used to obtain proper feedback gain parameters and the results of simulation and experiment are confirmed that the PSO algorithm has feasibility, and can improve the performance of the system efficiently. Otherwise, these results prove that the PSO algorithm is a simple method to

determine the feedback gain parameters, K_p and K_i, directly better than those by try and error. It also obviates instability problem of the rotor speed estimator by setting incorrect feedback gain parameters. Finally, the result of this study can be applied to estimating parameters in nonlinear system, complex system, and further, to avoiding the problem caused by parameters mismatch.

REFERENCES

[1] A. B. Jidin, N. R. B. N. Idris, A. B. M. Yatim, M. E. Elbuluk and T. Sutikno, A wide-speed high torque capability utilizing overmodulation strategy in DTC of induction machines with constant switching frequency controller, *IEEE Trans. Power Electron*, vol.27, no.5, pp.2566-2575, 2012.

[2] C. Zhang, X. Nian, T. Wang and W. Gui, Adaptive rotor resistance estimation in the low-speed range of speed sensorless DTC controlled IM drives, *Proc. of IEEE Int. Conf. on Neural Networks*, Perth, Australia, vol.4, pp.1942-1948, 1995.

[3] C. Liu, *AC Motor Control: Vector Control and Direct Torque Control Principle*, 2nd Edition, The Group Bookstore, Taipei, 1992.

[4] W. Sun, Y. Yu, G. Wang, B. Li and D. Xu, Robustness improvement of speed estimation in speed-sensorless induction motor drives, *IEEE Trans. Ind. App.*, vol.52, no.3, pp.2525-2536, 2016.

[5] Y. B. Zbede, S. M. Gadoue and D. J. Atkinson, Model predictive MRAS estimator for sensorless induction motor drives, *IEEE Trans. Ind. Electron.*, vol.63, no.6, pp.3511-3521, 2016.

[6] W. C. Pu, Y. C. Luo and P. Y. Chu, Sensorless stator field-oriented controlled IM drive at low speed with R_r estimator, *Mathematical Problems in Engineering*, vol.2014, 2014.

[7] C. Schauder, Adaptive speed identification for vector control of induction motors without rotational transducers, *IEEE Trans. Ind. Appl.*, vol.28, no.5, pp.1054-1061, 1992.

[8] M. Rashed et al., A new stable MRAS-based speed and stator resistance estimators for sensorless vector control induction motor drive at low speeds, *Proc. of IEEE-IAS Annu. Meeting*, Salt Lake City, UT, pp.1181-1188, 2003.

[9] W. Li, Y. Huang and T. Chang, Adaptive inertia weight particle swarm algorithm improvement of research, *Journal of Information Science*, 1997.

[10] Y. Shi and R. C. Eberhart, A modified particle swarm optimizer, *Proc. of the IEEE International Conference on Evolutionary Computation*, pp.69-73, 1998.

[11] Masahiro, Q. Wu and L. Zhuang, Meta-PSO optimization parameters used for classification, *International Symposium on Information Technology*, China, 1997.