

A Sensorless Predictive Torque Control Strategy for Induction Motor Drives

Van-Quang-Binh Ngo
Faculty of Physics
University of Education, Hue University
Hue 530000, Vietnam
nvqbinh@hueuni.edu.vn

Kim-Anh Nguyen
Faculty of Electrical Engineering
University of Science and Technology, The University of Danang
Da Nang 550000, Vietnam
nkanh@dut.udn.vn

Khanh-Quang Nguyen
Faculty of Electrical Engineering
University of Science and Technology, The University of Danang
Da Nang 550000, Vietnam
nkquang@dut.udn.vn

Viet-Tuan Pham
Faculty of Physics
University of Education, Hue University
Hue 530000, Vietnam
pvtuan@hueuni.edu.vn

Abstract—This paper proposes a sensorless predictive torque control scheme for induction motor (IM) drives. The approach employs the IM dynamic model in the stationary reference frame to predict future stator currents for all possible inverter switching states. The optimal switching state is determined by minimizing a cost function that quantifies the error between the predicted torque, its reference, and the corresponding stator flux. To eliminate the speed sensor and reduce system complexity and cost, an adaptive Luenberger observer is adopted for rotor speed estimation. Simulation results in Matlab/Simulink demonstrate that the proposed method achieves steady-state and transient performance comparable to conventional FCS-MPC, in terms of speed settling time, torque ripple, and speed ripple.

Index Terms—Model predictive control, Direct torque control, Sensorless speed, Extended Luenberger observer, Induction motor.

I. INTRODUCTION

Induction motors are widely preferred in challenging industrial applications because of their simple structure, robust design, low cost, and capability to operate over a broad speed range. Their high efficiency and versatility make them essential across numerous sectors [1], [2]. In IM drive systems, field-oriented control (FOC) and direct torque control (DTC) are two commonly employed control strategies [1], [3]. FOC achieves fast dynamic performance and satisfactory steady-state behavior due to its cascaded structure employing PI controllers for both current and speed loops. Nevertheless, its performance is highly sensitive to the accuracy of motor parameters, which may vary due to noise, temperature changes, and friction effects. In contrast, DTC directly determines inverter switching states using a lookup table without requiring modulation techniques, enabling a rapid dynamic response. However, it typically suffers from significant torque ripple and a high switching frequency of the power devices.

In recent years, finite control set model predictive control (FCS-MPC) for IM drive systems has gained significant attention due to its simple structure, rapid dynamic response,

flexible multi-objective capability, and effective handling of nonlinear constraints [4]–[6]. In variable-speed drive systems, accurate knowledge of state variables such as currents, flux linkages, and angular speed is essential for effective control. Traditionally, these quantities are measured using sensors. However, such sensors require additional installation space, increase system cost, and may reduce reliability in compact environments. Therefore, sensorless techniques have been developed to simplify the system structure and reduce overall manufacturing costs [7], [8]. For IM, rotor speed and position can be estimated from stator voltage and current measurements. Among sensorless approaches, the adaptive Luenberger observer is commonly utilized to obtain these quantities based on the motor model. This observer offers strong robustness to parameter variations, straightforward hardware implementation, and low computational complexity [9]–[11].

This paper proposes a sensorless predictive torque control scheme for an induction motor drive system. The method is based on the dynamic model of the induction motor to predict the stator currents corresponding to all possible inverter switching states. The model is formulated in the stationary reference frame, which eliminates the need for accurate rotor position information. A cost function is then designed to minimize the error between the predicted torque and its reference value, as well as the stator flux, thereby enabling the selection of the optimal switching state. To remove the requirement for a mechanical speed sensor and reduce system cost, an adaptive Luenberger observer is employed to estimate the rotor speed. Simulation results obtained in MATLAB demonstrate that the proposed approach achieves both steady-state and transient performance comparable to that of conventional predictive torque control methods.

The remainder of the paper is organized as follows. Section II presents the IM dynamic model and the proposed sensorless predictive control strategy. Section III provides and discusses the simulation results. Finally, the conclusions are given in

Section IV.

II. SENSORLESS PREDICTIVE TORQUE CONTROL OF INDUCTION MOTOR

A. Dynamic model of induction motor fed by a voltage source inverter

The dynamic model of the induction motor can be expressed in the $\alpha\beta$ reference frame as follows:

$$\begin{aligned} \frac{di_{s\alpha}}{dt} &= -\left(\frac{1}{\sigma T_s} + \frac{1-\sigma}{\sigma T_r}\right)i_{s\alpha} + \frac{1-\sigma}{\sigma L_m T_r}\psi_{r\alpha} + \frac{1-\sigma}{\sigma L_m}\omega_r\psi_{r\beta} \\ &\quad + \frac{1}{\sigma L_s}u_{s\alpha}, \\ \frac{di_{s\beta}}{dt} &= -\left(\frac{1}{\sigma T_s} + \frac{1-\sigma}{\sigma T_r}\right)i_{s\beta} + \frac{1-\sigma}{\sigma L_m}\omega_r\psi_{r\alpha} + \frac{1-\sigma}{\sigma L_m T_r}\psi_{r\beta} \\ &\quad + \frac{1}{\sigma L_s}u_{s\beta}, \\ \frac{d\psi_{r\alpha}}{dt} &= \frac{L_m}{T_r}i_{s\alpha} - \frac{1}{T_r}\psi_{r\alpha} - \omega_r\psi_{r\beta}, \\ \frac{d\psi_{r\beta}}{dt} &= \frac{L_m}{T_r}i_{s\beta} + \omega_r\psi_{r\alpha} - \frac{1}{T_r}\psi_{r\beta}, \end{aligned} \quad (1)$$

where $T_s = \frac{L_s}{R_s}$, $T_r = \frac{L_r}{R_r}$ represent the stator and rotor time constant. R_s and R_r denote the stator and rotor resistances, while L_s and L_r represent the stator and rotor inductances, respectively. ω_r is the electrical speed of rotor. The inductances are derived from the magnetizing inductance (L_m), stator leakage inductance ($L_{s\sigma}$), and rotor leakage inductance ($L_{r\sigma}$) as follows:

$$\begin{aligned} L_s &= L_{s\sigma} + L_m, \\ L_r &= L_{r\sigma} + L_m, \end{aligned} \quad (2)$$

where $\sigma = 1 - \frac{L_m^2}{L_s L_r}$ is the leakage coefficient.

Moreover, the voltage supplied to the induction motor's stator is generated by the inverter output. This voltage can be determined based on the switching states of the semiconductor devices and the DC-link voltage (V_{dc}), expressed as follows:

$$\begin{aligned} u_{s\alpha} &= \frac{V_{dc}}{3}(2S_a - S_b - S_c), \\ u_{s\beta} &= \frac{\sqrt{3}V_{dc}}{3}(S_b - S_c), \end{aligned} \quad (3)$$

where S_a , S_b , and S_c denote the switching states of the switches in phases a , b , and c , respectively, with values of 0 or 1.

The determination of stator flux is predicated on the equation of stator voltage by given:

$$\begin{aligned} \frac{d\psi_{s\alpha}}{dt} &= u_{s\alpha} - R_s i_{s\alpha}, \\ \frac{d\psi_{s\beta}}{dt} &= u_{s\beta} - R_s i_{s\beta}. \end{aligned} \quad (4)$$

By applying the first-order Euler approximation to (1) over the sampling interval T_{sp} , the discrete-time stator current dynamic model is given by:

$$\begin{aligned} i_{s\alpha}(k+1) &= i_{s\alpha}(k) \left(1 + \frac{T_{sp}}{\tau_\sigma}\right) \\ &\quad + \frac{T_{sp}}{\tau_\sigma + T_{sp}} \frac{1}{R_\sigma} \left(\left(\frac{k_r}{T_r} + k_r \omega_r\right) \psi_{r\beta}(k) + u_{s\alpha}(k)\right), \\ i_{s\beta}(k+1) &= i_{s\beta}(k) \left(1 + \frac{T_{sp}}{\tau_\sigma}\right) \\ &\quad + \frac{T_{sp}}{\tau_\sigma + T_{sp}} \frac{1}{R_\sigma} \left(\left(\frac{k_r}{T_r} - k_r \omega_r\right) \psi_{r\alpha}(k) + u_{s\beta}(k)\right), \end{aligned} \quad (5)$$

where $k_r = \frac{L_m}{L_r}$, $R_\sigma = R_s + k_r^2 R_r$ and $\tau_\sigma = \frac{\sigma L_s}{R_\sigma}$. The rotor flux is estimated based on the stator flux and stator current according to the following equation:

$$\begin{aligned} \psi_{r\alpha}(k) &= \frac{L_r}{L_m} \psi_{s\alpha}(k) + i_{s\alpha}(k) \left(L_m - \frac{L_r L_s}{L_m}\right), \\ \psi_{r\beta}(k) &= \frac{L_r}{L_m} \psi_{s\beta}(k) + i_{s\beta}(k) \left(L_m - \frac{L_r L_s}{L_m}\right), \end{aligned} \quad (6)$$

where the stator flux $\psi_{s\alpha}(k)$, $\psi_{s\beta}(k)$ is obtained from (4) by applying the Euler approximation.

Similarly, the predicted stator flux at instant $k+1$ is expressed as follows:

$$\begin{aligned} \psi_{s\alpha}(k+1) &= \psi_{s\alpha}(k) + T_{sp}(u_{s\alpha}(k) - R_s i_{s\alpha}(k)), \\ \psi_{s\beta}(k+1) &= \psi_{s\beta}(k) + T_{sp}(u_{s\beta}(k) - R_s i_{s\beta}(k)), \end{aligned} \quad (7)$$

where the stator voltages $u_{s\alpha}(k)$ and $u_{s\beta}(k)$ are derived from (3).

The predicted torque is calculated based on the stator flux and stator current as:

$$T_e(k+1) = \frac{3p}{2} (\psi_{s\alpha}(k+1)i_{s\beta}(k+1) - \psi_{s\beta}(k+1)i_{s\alpha}(k+1)) \quad (8)$$

B. Rotor speed estimation via an extended Luenberger observer

The adaptive Luenberger observer (ALO) employs the stator current and rotor flux as state variables to estimate the rotor speed, as illustrated in Fig. 1. The linearized model of the induction motor employed in the ALO is derived from (1) and can be expressed as follows:

$$\begin{aligned} \frac{d}{dt} \underbrace{\begin{bmatrix} \mathbf{i}_s \\ \psi_r \end{bmatrix}}_{\mathbf{x}} &= \underbrace{\begin{bmatrix} a_{11}\mathbf{I} + a_{12}\omega_r\mathbf{J} & a_{21}\mathbf{I} + a_{22}\omega_r\mathbf{J} \\ a_{31}\mathbf{I} & a_{41}\mathbf{I} + a_{42}\omega_r\mathbf{J} \end{bmatrix}}_{\mathbf{A}} \begin{bmatrix} \mathbf{i}_s \\ \psi_r \end{bmatrix} \\ &\quad + \underbrace{\begin{bmatrix} b_1\mathbf{I} \\ \mathbf{I} \end{bmatrix}}_{\mathbf{B}} \mathbf{u}_s, \\ y &= \underbrace{\begin{bmatrix} \mathbf{I} & \mathbf{0} \end{bmatrix}}_{\mathbf{C}} \begin{bmatrix} \mathbf{i}_s \\ \psi_r \end{bmatrix} \end{aligned} \quad (9)$$

where $\mathbf{I}$ and $\mathbf{0}$ denote the 2×2 identity and zero matrices, respectively. $\mathbf{J}$ represents a linear transformation.

$$\mathbf{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}, a_{11} = -\left(\frac{1}{\sigma T_s} + \frac{1-\sigma}{\sigma T_r}\right),$$

$$a_{12} = 1, a_{21} = \frac{1-\sigma}{\sigma L_m T_r}, a_{22} = \frac{1-\sigma}{\sigma L_m},$$

$$a_{31} = \frac{L_m}{T_r}, a_{41} = -\frac{1}{T_r}, a_{42} = 1, b_1 = \frac{1}{\sigma L_s}.$$

The state observer can be defined as:

$$\frac{d\hat{\mathbf{x}}}{dt} = \hat{\mathbf{A}}\hat{\mathbf{x}} + \mathbf{B}\mathbf{u} + \mathbf{L}(\hat{\mathbf{i}}_s - \mathbf{i}_s), \quad (10)$$

where the superscript $\wedge$ indicates estimated quantities and $\mathbf{L}$ denotes the observer gain matrix which can be determined as follows:

$$\mathbf{L} = \begin{bmatrix} l_1 & l_2 & l_3 & l_4 \\ -l_2 & l_1 & -l_4 & l_3 \end{bmatrix}^T, \quad (11)$$

where $l_1 = (k-1)(a_{11} + a_{41})$, $l_2 = (k-1)a_{42}\hat{\omega}_r$,
 $l_3 = (k^2-1)(ca_{11}\hat{\omega}_r + a_{41}) - c(k-1)(a_{11} + a_{41})$,
 $l_4 = -c(k-1)a_{42}\hat{\omega}_r$, $c = \frac{\sigma L_s L_r}{L_m}$.

$\hat{\omega}_r$ represents the estimated rotor speed. The gain coefficient k is selected as a pole-placement factor ($k > 1$) to ensure that the observer error dynamics matrix is Hurwitz. In this study, $k = 2$ is adopted as it provides a suitable trade-off between estimation convergence speed and robustness to measurement noise. Setting $k = 2$, the observer dynamics are sufficiently faster than the motor dynamics in the predictive torque control loop to meet the high-speed performance requirement without sacrificing the stability of the system under signal noise.

The error dynamics of the ALO can be estimated as follows:

$$\frac{de}{dt} = (\mathbf{A} + \mathbf{LC})e - \Delta\mathbf{A}\hat{\mathbf{x}}, \quad (12)$$

where $e = x - \hat{x}$, $\Delta\omega_r = \hat{\omega}_r - \omega_r$,
 $\Delta\mathbf{A} = \hat{\mathbf{A}} - \mathbf{A} = \begin{bmatrix} a_{11}\Delta\omega_r\mathbf{J} & a_{22}\Delta\omega_r\mathbf{J} \\ \mathbf{0} & \mathbf{0} \end{bmatrix}$.

The Lyapunov function for the speed-adaptive observer is given by:

$$V = e^T e + \frac{(\hat{\omega}_r - \omega_r)^2}{\lambda}, \quad (13)$$

where λ is a positive constant.

In deriving the adaptive mechanism, the unknown parameter $\hat{\omega}_r$ is assumed constant. Accordingly, the time derivative of V can be expressed as:

$$\frac{d}{dt}V = e^T \left\{ (\mathbf{A} + \mathbf{LC})^T + (\mathbf{A} + \mathbf{LC}) \right\} e$$

$$- 2a_{22}\Delta\omega_r \left(e_{i\alpha}\hat{\psi}_{r\beta} - e_{i\beta}\hat{\psi}_{r\alpha} \right) + 2\Delta\omega_r \frac{d\hat{\omega}_r}{dt} \frac{1}{\lambda}$$

where $e_{i\alpha} = i_{s\alpha} - \hat{i}_{s\alpha}$, $e_{i\beta} = i_{s\beta} - \hat{i}_{s\beta}$.

The adaptive speed law is formulated so that the the second and third components of (14) eliminate each other, yielding:

$$\frac{d\hat{\omega}_r}{dt} = \lambda a_{22} \left(e_{i\alpha}\hat{\psi}_{r\beta} - e_{i\beta}\hat{\psi}_{r\alpha} \right) \quad (15)$$

the time derivative of the Lyapunov function is rendered negative semi-definite, which guarantees that the estimation errors e and $(\hat{\omega}_r - \omega_r)$ converge to zero.

To enhance the convergence of the speed adaptation mechanism, a PI regulator is utilized. Accordingly, the estimated speed can be formulated as follows:

$$\hat{\omega}_r = K_p \left(\psi_{r\beta} (i_{s\alpha} - \hat{i}_{s\alpha}) - \psi_{r\alpha} (i_{s\beta} - \hat{i}_{s\beta}) \right) \quad (16)$$

$$+ K_i \int \left(\psi_{r\beta} (i_{s\alpha} - \hat{i}_{s\alpha}) - \psi_{r\alpha} (i_{s\beta} - \hat{i}_{s\beta}) \right) dt,$$

where K_p and K_i represent the PI gains.

For the practical realization related to what was seen in (16), it is important to clarify that the stator currents and rotor fluxes are AC in nature in the stationary $\alpha\beta$ frame but $e_{i\alpha}\hat{\psi}_{r\beta} - e_{i\beta}\hat{\psi}_{r\alpha}$ represents a vector cross-product. In steady-state, the current error vector and the flux vector rotate at the same synchronous speed. The product of these provides a scalar (DC component), proportional to the angular separation of each. This DC signal enables the PI controller to set the steady-state speed error to zero, while the integral action smooths out high-frequency switching noise.

C. Sensorless speed control of IM using model predictive torque control

The principle of the torque predictive control method for induction motors is to predict the torque and stator flux errors in the $\alpha\beta$ coordinate system during the next control interval, based on the measured stator current and the discrete voltage vectors of the inverter. The optimal voltage vector is then selected by minimizing a predefined cost function among the eight possible switching states. To meet the control objectives, conventional predictive torque control schemes for induction motors typically employ a cascaded structure, consisting of an outer speed control loop based on a proportional–integral (PI) controller and an inner torque control loop using predictive control. Accordingly, the optimal switching state of the inverter supplying the induction motor is determined by solving an optimization problem as follows:

$$\mathbf{S}_{opt} = [S_a \ S_b \ S_c] = \arg \min \{J(\mathbf{S}_k)\} \quad (17)$$

such as $\mathbf{S}_k \in \mathbf{S} = \{S_1, \dots, S_8\}$, $S_i \in \{0, 1\}^3$,

$$J(\mathbf{S}_k) = (T_e^*(k+1) - T_e(k+1))^2$$

$$+ \gamma(\|\psi_s^*(k+1)\| - \|\psi_s(k+1)\|)^2,$$

where $T_e^*(k+1)$, $\psi_s^*(k+1)$ denote the reference torque and stator flux, respectively, and $\|\cdot\|$ represents the Euclidean norm of a vector. A weighting factor γ is used to cope with the effect of torque and flux contributions in the cost function. Since torque and flux are measured in different units with different magnitudes, γ is set empirically using simulation studies to achieve good trade-offs between torque ripple reduction and flux regulation performance.

The future values of torque and stator flux are anticipated using the Lagrange extrapolation method as:

$$T_e^*(k+1) = 3T_e^*(k) - 3T_e^*(k-1) + T_e^*(k-2), \quad (18)$$

$$\psi_s^*(k+1) = 3\psi_s^*(k) - 3\psi_s^*(k-1) + \psi_s^*(k-2).$$

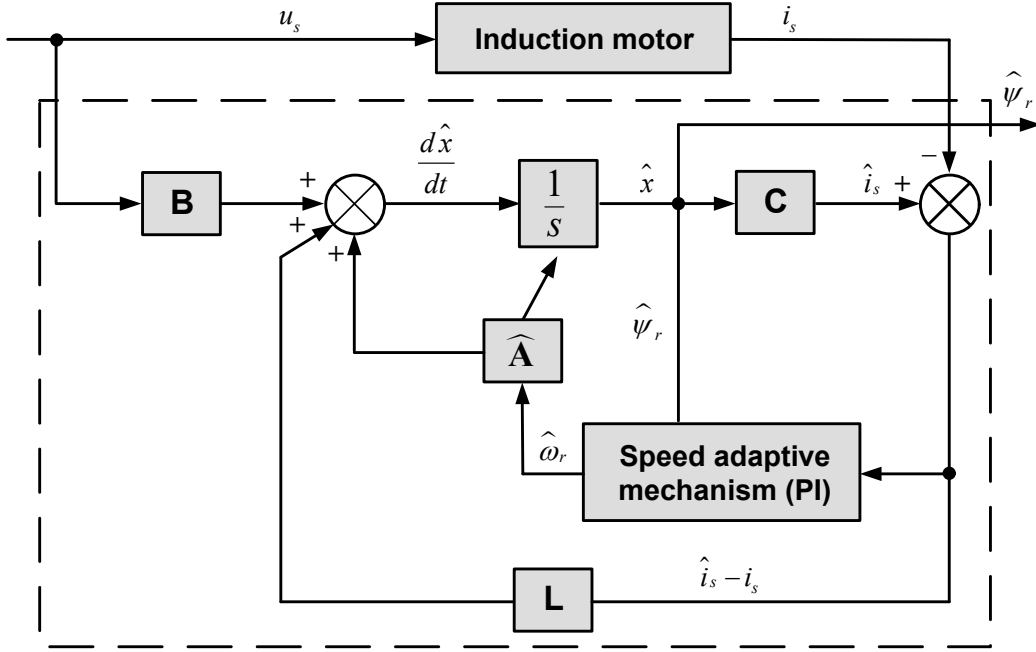


Fig. 1: Diagrammatic representation of the adaptive Luenberger observer.

In this paper, an adaptive Luenberger observer is employed to estimate the rotor speed, thereby eliminating the need for a mechanical speed sensor. The block diagram of the proposed control method is illustrated in Fig. 2.

III. SIMULATION RESULTS

The Matlab environment was used to validate the proposed system architecture. The inverter supplying the IM was modeled using the SimPowerSystems toolbox, with the rated parameters listed in Table I. A PI controller was employed for speed regulation, and the predictive torque control algorithm was implemented with a sampling time of $40 \mu s$.

To validate the effectiveness of the proposed strategy, the method was evaluated under both steady-state conditions at various speeds and transient scenarios, including abrupt changes in reference speed and load torque. Moreover, a comparative analysis was conducted to assess the performance of the proposed method against conventional model predictive torque control (PTC) [2] under identical operating conditions. The ripples in torque was quantified using the root mean square error (RMSE), defined as follows:

$$RMSE = \sqrt{\frac{1}{N} \sum_{t=1}^N (x(k) - \hat{x}(k))^2}, \quad (19)$$

where x denotes the reference value, while $\hat{x}$ represents the corresponding simulated value. N is the number of samples.

A. Steady-state analysis

To validate the feasibility and effectiveness of the proposed method, two test scenarios were conducted. In the first scenario, the system was evaluated under steady-state conditions

TABLE I: Parameters of the IM.

Parameters	Value
Power rating	7.5 kW
Line-to-line voltage	380 V
Frequency	50 Hz
Stator resistance	0.729 Ω
Rotor resistance	0.4 Ω
Stator leakage inductance	0.0013 H
Rotor leakage inductance	0.0027 H
Mutual inductance	0.1125 H
Number of pole pairs	2

with a load torque of 45 Nm and a reference rotor speed of 1445 rpm. The corresponding results for both the conventional PTC method and the proposed approach are presented in Fig. 3. As illustrated in Fig. 3(a), the proposed control method achieves a steady-state speed error comparable to that of the conventional PTC strategy, demonstrating its accuracy in speed tracking. In terms of torque performance, the RMSE of the electromagnetic torque obtained with the proposed method is only slightly higher than that of the traditional predictive control technique, indicating a minor trade-off while maintaining acceptable performance. Furthermore, the proposed control strategy produces a high-quality sinusoidal stator current waveform with low distortion, as shown in Fig. 3(c). Fig. 3(e) depicts the error between the estimated and actual rotor speeds. It can be observed that the speed estimation error remains within a narrow range, approximately from 12 rpm to 19 rpm, highlighting the accuracy and reliability of the employed observer. Overall, these results demonstrate that the proposed method ensures stable steady-state operation,

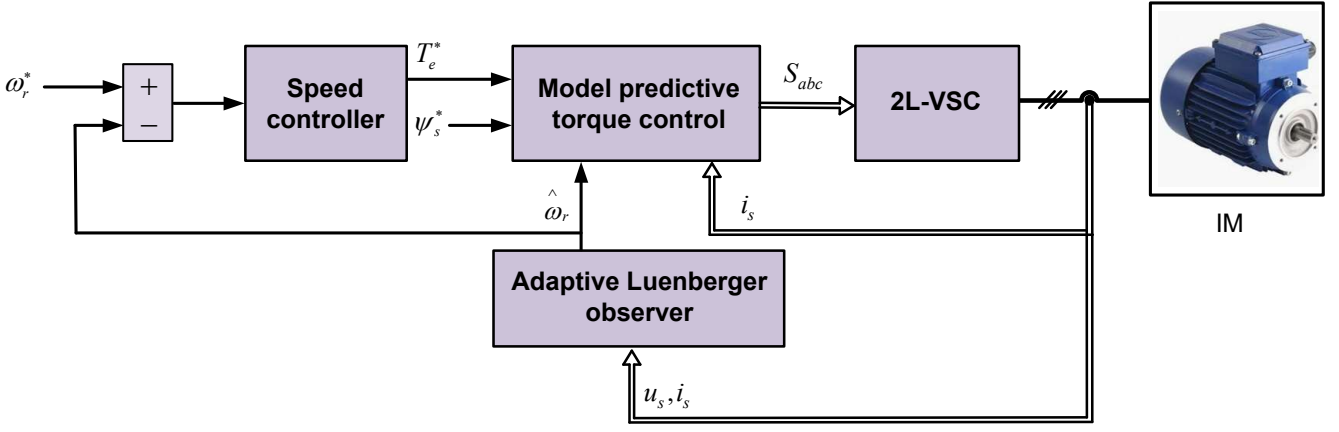


Fig. 2: Block diagram of the sensorless speed predictive control scheme for IM.

accurate speed tracking, and satisfactory current and torque performance.

B. Dynamic performance evaluation

The second scenario demonstrates the capability of the proposed method under dynamic conditions. The rotor speed is increased from 700 rpm to 1440 rpm at $t = 0.5s$, then reversed to -800 rpm at $t = 1s$, and subsequently increased to 200 rpm at $t = 1.5s$, as shown in Fig. 4. Meanwhile, the load torque is increased from 10 Nm to 45 Nm at $t = 0.8s$, further intensifying the dynamic operating conditions. The results highlight the fast dynamic response of the system. The IM accelerates from 700 rpm to 1440 rpm in approximately 8.8 ms, maintains this speed for about 0.41 s, and then undergoes braking and reversal, reaching -800 rpm in approximately 0.17 s. These results demonstrate the system's ability to rapidly adapt to significant speed variations. In summary, the proposed method effectively tracks the reference signal with minimal torque and rotor speed overshoot, achieving performance comparable to the conventional PTC method, as shown in Figs. 4(a) and 4(c). Notably, this performance is achieved without the use of a mechanical speed sensor, thereby simplifying the system structure.

IV. CONCLUSIONS

In this paper, a sensorless predictive torque control scheme for an induction motor drive system has been proposed. The approach utilizes the dynamic model of the induction motor in the fixed coordinate system to predict stator currents for all possible inverter switching states, thereby eliminating the need for precise rotor position information. A cost function is formulated to minimize the errors between the predicted torque, its reference value, and the stator flux, enabling the selection of the optimal switching state. Furthermore, an adaptive Luenberger observer is employed to estimate the rotor speed, removing the requirement for a mechanical speed sensor and reducing overall system cost. Simulation results obtained in Matlab confirm that the proposed method achieves both steady-state accuracy and transient response comparable

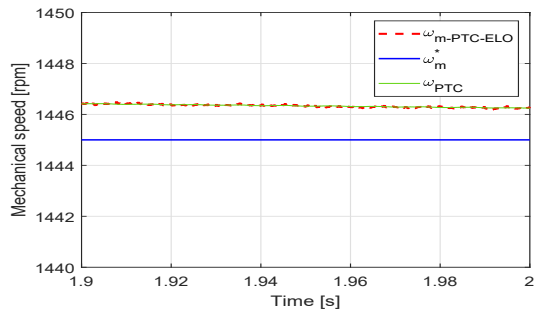
to conventional predictive torque control strategies, demonstrating its effectiveness and practical applicability. Further studies will focus on real-time implementation and experimental verification using DSP or FPGA-based platforms, as well as hardware-in-loop testing (HIL) to further assess the practical feasibility of the proposed approach.

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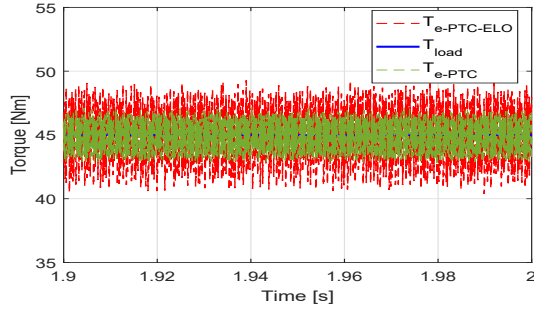
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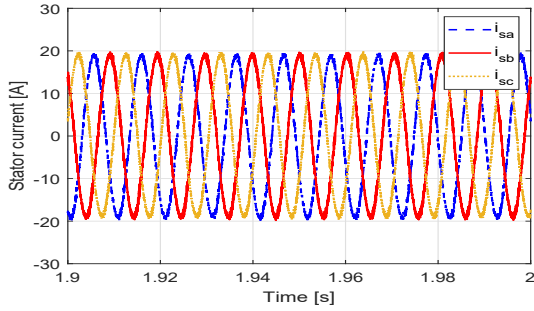
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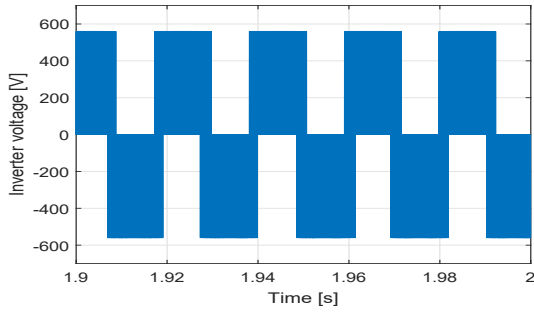
(a) Rotor speed of two methods



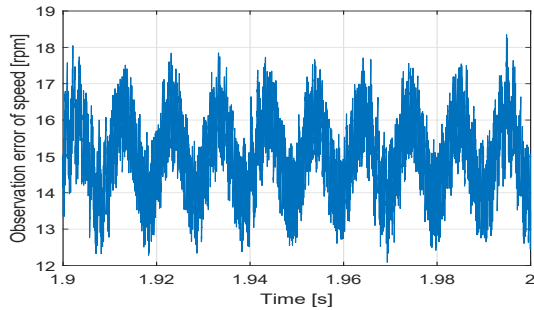
(b) Torque of two methods



(c) Stator current of the proposed method

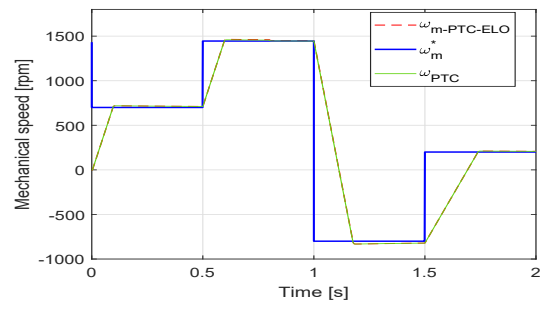


(d) Inverter voltage

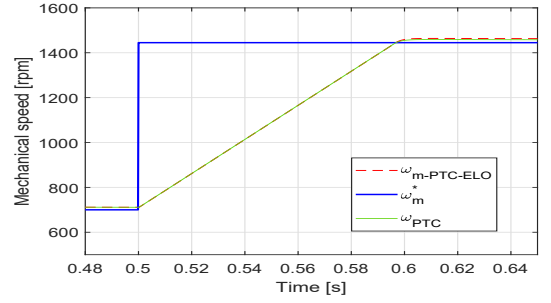


(e) Rotor speed estimation error

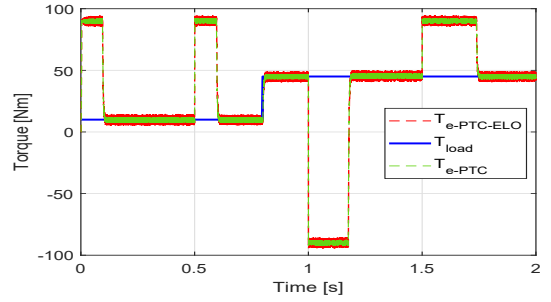
Fig. 3: Simulation results at steady-state.



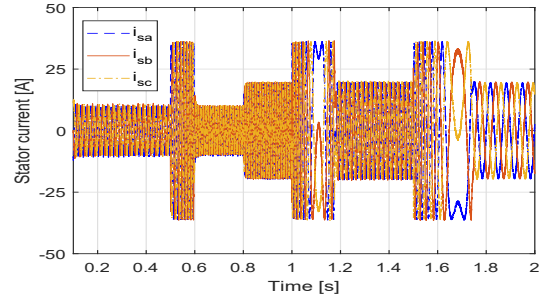
(a) Rotor speed of two methods



(b) Zoom of rotor speed



(c) Torque for two controllers



(d) Three-phase stator current

Fig. 4: Simulation results under dynamic conditions.

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