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Predictive Torque and Stator Flux Control for $N*3$ -phase PMSM Drives with Parameter Robustness Improvement

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Abstract—In this study, a novel predictive torque and stator flux control (PTSF) method is proposed for N -segment three-phase PMSM ($N*3$ -phase PMSM), which can effectively eliminate the influence of parameter mismatch on stator flux and torque. The proposed PTSF method has two loops, stator flux loop and torque loop, both implemented with robust predictive control algorithm. First, the stator flux predictive controller with parameter robustness is designed, and the sensitivity of its parameters is analyzed. In the dq stator flux space, the robust stator flux predictive control with one-step delay compensation is then proposed, which can effectively enhance robustness against parameters mismatch. Moreover, the robust torque predictive control based on unknown torque disturbance observer is developed, which can effectively strengthen the robustness against the load torque disturbance and parameter mismatch. Finally, the validity and feasibility of proposed PTSF method are verified by simulation and experiment. Excellent simulation and experimental results were achieved with respect to the stator flux tracking error, torque /flux ripple reduction, and stator current distortion over conventional predictive torque control (PTC).

Index Terms — N -segment three-phase PMSM, parameters mismatch, predictive torque control, unknown torque disturbance observer.

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I. INTRODUCTION

MULTI-PHASE permanent magnet synchronous motor (Multi-phase PMSM) has been widely adopted in many special applications ranging from high power traction vehicles to marine propulsion due to its outstanding advantages of the better fault-tolerant capability, high power and torque densities [1-3]. In order to optimize the performance of multi-phase PMSM, a novel multi three-phase PMSM ($N*3$ -phase PMSM) with special mechanical degree winding placement is investigated in [4-6]. The main benefit of this $N*3$ -phase PMSM has extremely lower cogging and higher torque quality compared with conventional multi-phase PMSM. Moreover, the controller design of $N*3$ -phase PMSM is more simple and can be fed by several independent three-phase motor drives [4, 5]. In [4-6], the PI controllers were adopted to regulate the $N*3$ -phase PMSM currents, which can reduce the stator current harmonic distortion and obtain low torque ripple under the steady-state. However, the dynamic and static control performance of $N*3$ -phase PMSM is highly dependent on the PI controller gains.

Predictive torque control (PTC) owns the advantages of excellent dynamic and steady-state performance compared with the conventional PI controller, which makes it a powerful $N*3$ -phase PMSM control strategies under the whole operating range [7-10]. In the PTC method, all possible voltage vectors of the converter can be calculated by using the predicted values of stator flux and torque. These predictions are determined by the mathematical model of the $N*3$ -phase PMSM that depends on the rotor flux linkage parameter and inductance parameter. However, these parameters utilized in the PTC may not match with the actual $N*3$ -phase PMSM parameters due to the parameter identification errors or the parameter may change in the process of motor operation [11]. Model parameters mismatch results in unstable switching frequency and inaccurate prediction behavior, which affects the performance of the PTC, i.e., torque/flux ripple increases [12].

Designing disturbance observer is the main approach to improve the robustness of the model predictive controller under the parameters mismatch. In [13, 14], a robust cascaded nonlinear predictive controller based on disturbance observer was proposed, which can achieve precise speed trajectory tracking and disturbance rejection in the presence of parameters mismatch and external perturbations. In [15, 16], an improved deadbeat-based predictive current control based on the estimated disturbances was designed, which can obtain good dynamic performance and model parameters mismatch suppression performance. In [17], an improved predictive current control method with incremental model was developed, which can operate without knowing the rotor flux and improve robustness against the inductance parameter mismatch by using the inductance disturbance observer. In

[18], a novel linear quadratic model predictive control method based on nonlinear disturbance observer was proposed, which can minimize the torque ripple caused by parameters mismatch. In [19], a robust fault-tolerant predictive current control method based on a composite observer was proposed, which can eliminate the influence of permanent magnet demagnetization and parameter perturbation. In [13-19], the predictive control method based on disturbance observer compensation relies on the accurate measurement of voltage and current.

Some modified predictive control methods without observer compensation have been presented to eliminate the influence of parameters mismatch. Instead of using disturbance observer-based voltage feedforward methods, a deadbeat predictive current control method with closed form error compensation was proposed in [20], which can directly compensate the current tracking errors caused by parameter mismatch. In [21], the robust model predictive control scheme with simplified repetitive control was presented, which can guarantee stability and maintain high accuracy tracking under parameter mismatch and current distortion. In [22, 23], a modified predictive torque control algorithm with proportional controller was proposed, which can improve the prediction accuracy and obtain the perfect prediction performance under inductance mismatch. In [24], a model-free predictive current control methods based on a current difference detection technique was proposed, which can easy to implement due to its simplicity and free of any motor parameters. Although the model-free predictive control is quite independent from motor model, it has a strongly dependent on the accuracy of currents references and current responses.

Since the parameters mismatch lead to inaccurate predictions of torque/flux and deteriorate the control performance, a novel parameters immunity predictive torque and stator flux control method is proposed in this paper to guarantee the control performance of the $N*3$ -phase PMSM drives with parameter mismatch. The proposed PTSF method can enhance robustness against parameters mismatch by adopting the closed-loop control of d-q axis stator flux. Compare with the conventional PTC method, the proposed PTSF method can effectively eliminate the stator flux tracking errors and reduce the torque/flux ripple under parameters mismatch. This paper is organized as follows. The mathematical model of $N*3$ -phase PMSM is established in Section II. In the Section III, the robust predictive stator flux controller with parameter robustness improvement is designed. The robust predictive torque controller with parameter robustness improvement is proposed in Section IV. The effectiveness of the proposed PTSF algorithm through simulations and experimentations is validated in Section V and Section VI, respectively. Section VII concludes this paper.

II. MATHEMATICAL MODEL OF $N*3$ -PHASE PMSM

Taking 6*3-phase PMSM as an example, the structure diagram of the $N*3$ -phase PMSM studied in this paper is shown in Fig. 1. The $N*3$ -phase PMSM is composed of six three-phase PMSM units with the same characteristics. The stator winding of 6*3-phase PMSM consists of six segment three-phase PMSM winding, and each set of three-phase winding has a phase difference of 180 electrical angles, as

shown in Fig. 2. Note that there are electrical, magnetic and thermal isolation characteristics among the three-phase PMSM units winding.

Fig.1 The 6*3 phase PMSM structure diagram.

Fig.2 Star of slots for 6*3 phase PMSM.

The voltage equations of the $N*3$ -phase PMSM in a synchronous rotating frame are as follows [4-6]:

$$\begin{cases} \frac{di_{dj}}{dt} = -\frac{R}{L}i_{dj} + \omega_e i_{qj} + \frac{1}{L}u_{dj} \\ \frac{di_{qj}}{dt} = -\frac{R}{L}i_{qj} - \omega_e i_{dj} - \frac{\psi_r}{L}\omega_e + \frac{1}{L}u_{qj} \end{cases} \quad (1)$$

where j stands for any unit of the motor; u_{dj} and u_{qj} are the d - and q -axis voltages of the j th winding; i_{dj} and i_{qj} are the d - and q -axis currents of the j th winding; R and L are the stator resistance and stator inductance; ω_e is the electrical rotor speed; ψ_r is the flux linkage of permanent magnets.

The electromagnetic torque equation of $N*3$ -phase PMSM can be expressed as [4-6]

$$T_e = \frac{3n_p}{2} \sum_{j=1}^N (\psi_r i_{qj}) = \frac{3Nn_p}{2} \psi_r i_{qj} \quad (2)$$

The mechanical dynamic model of $N*3$ -phase PMSM can be described as

$$T_e - T_L = \frac{J}{n_p} \frac{d\omega_e}{dt} \quad (3)$$

where n_p is the number of pole pairs; J is the moment of inertia; T_e and T_L are the electromagnetic torque and load torque of the $N*3$ -phase PMSM, respectively.

III. ROBUST PREDICTIVE STATOR FLUX CONTROL WITH PARAMETER ROBUSTNESS IMPROVEMENT

According to the special winding structure and characteristics of 6*3-phase PMSM, the drive system structure is designed, as shown in Fig. 3. The 6*3-phase PMSM is driven by six three-phase voltage source inverters structures (VSIs), and each set of windings is independently excited by a three-phase VSI structure. Through the characteristics analysis of the 6*3-phase PMSM, it can be known that the vector control of the 6*3-phase PMSM is similar to that of the conventional three-phase PMSM [4-6]. In this paper, a novel PTSF

method is used to send out six groups of the same PWM driving signals, which can realize the control of six three-phase VSI structures under the parameters mismatch.

Fig. 3 Integral drive system diagram of 6*3-phase PMSM.

A. Design of Predictive Stator Flux Controller

This paper mainly studies the novel robust predictive control under inductance and flux parameters mismatch. Introducing $L = L_o + \Delta L$ and $\psi_r = \psi_{ro} + \Delta\psi_r$, where L_o and ψ_{ro} are actual values, ΔL and $\Delta\psi_r$ are uncertain components of the corresponding parameter. According to (1), we can express the voltage equations of the N^*3 -phase PMSM as follows under inductance and flux parameters mismatch.

$$\begin{cases} \frac{di_{dj}}{dt} = -\frac{R_o}{L_o + \Delta L} i_{dj} + \omega_e i_{qj} + \frac{1}{L_o + \Delta L} u_{dj} \\ \frac{di_{qj}}{dt} = -\frac{R_o}{L_o + \Delta L} i_{qj} - \omega_e i_{dj} - \frac{\psi_{ro} + \Delta\psi_r}{L_o + \Delta L} \omega_e + \frac{1}{L_o + \Delta L} u_{qj} \end{cases} \quad (4)$$

According to (4), the stator flux equations of the N^*3 -phase PMSM can be obtained as follows under inductance and flux parameters mismatch.

$$\begin{cases} \frac{d\psi_{dj}}{dt} = u_{dj} - \frac{R_o}{L_o} \psi_{dj} + \omega_e \psi_{qj} + \frac{R_o}{L_o} \psi_{ro} + f_{dj} \\ \frac{d\psi_{qj}}{dt} = u_{qj} - \omega_e \psi_{dj} - \frac{R_o}{L_o} \psi_{qj} + f_{qj} \\ \psi_{dj} = L_o i_{dj} + \psi_{ro} \\ \psi_{qj} = L_o i_{qj} \end{cases} \quad (5)$$

where f_{dj} and f_{qj} are the d- and q-axis unknown disturbance, respectively

$$\begin{cases} f_{dj} = \frac{\Delta L R_o}{(L_o + \Delta L)} i_{dj} - \frac{\Delta L}{(L_o + \Delta L)} u_{dj} \\ f_{qj} = \frac{\Delta L}{(L_o + \Delta L)} (-u_{qj} + R_o i_{qj} + \psi_{ro} \omega_e + \Delta\psi_r \omega_e) - \Delta\psi_r \omega_e \end{cases} \quad (6)$$

By discretizing the flux equations (5), the discrete flux equations of N^*3 -phase PMSM can be obtained as

$$\begin{cases} \psi_j(k+1) = \mathbf{u}_j(k)T_s - \mathbf{A}(k)\psi_j(k)T_s + \mathbf{B}T_s + \mathbf{f}(k)T_s \\ \psi_j(k) = \mathbf{D}i_j(k) + \mathbf{f}_\psi \end{cases} \quad (7)$$

$$\text{where } \mathbf{u}_j(k) = \begin{bmatrix} u_{dj}(k) \\ u_{qj}(k) \end{bmatrix}, \quad \psi_j(k) = \begin{bmatrix} \psi_{dj}(k) \\ \psi_{qj}(k) \end{bmatrix}, \quad i_j(k) = \begin{bmatrix} i_{dj}(k) \\ i_{qj}(k) \end{bmatrix},$$

$$\mathbf{A}(k) = \begin{bmatrix} \left(\frac{R_o}{L_o} - \frac{1}{T_s}\right) & -\omega_e(k) \\ \omega_e(k) & \left(\frac{R_o}{L_o} - \frac{1}{T_s}\right) \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} \frac{R_o}{L_o} \psi_{ro} \\ 0 \end{bmatrix}, \quad \mathbf{f}_\psi = \begin{bmatrix} \psi_{ro} \\ 0 \end{bmatrix},$$

$$\mathbf{f}(k) = \begin{bmatrix} f_{dj}(k) \\ f_{qj}(k) \end{bmatrix}, \quad \mathbf{D} = \begin{bmatrix} L_o & 0 \\ 0 & L_o \end{bmatrix}.$$

According to (7), the ideal output voltage vector of the predictive stator flux controller can be expressed as follows when inductance and flux parameters mismatch are considered.

$$\mathbf{u}_j^{ref}(k) = \frac{1}{T_s} \psi_j^{ref}(k+1) + \mathbf{A}(k)\psi_j(k) - \mathbf{B} - \mathbf{f}(k) \quad (8)$$

where $\mathbf{u}_j^{ref}(k)$ is the ideal output voltage vector of predictive stator flux controller, $\psi_j^{ref}(k+1)$ is the stator flux reference.

B. Parameter Sensitivity Analysis of Predictive Stator Flux Controller

The parameters sensitivity of predictive stator flux controller is evaluated in this section. According to (7), the actual flux equations of N^*3 -phase PMSM can be expressed as

$$\psi_j(k+1) = \mathbf{u}_j(k)T_s - \mathbf{A}(k)\psi_j(k)T_s + \mathbf{B}T_s \quad (9)$$

The ideal output voltage vector of the predictive stator flux controller is applied to the actual flux equations of N^*3 -phase PMSM under parameters mismatch. Therefore, substituting (8) into (9) yields

$$\psi_j(k+1) = \psi_j^{ref}(k+1) - \mathbf{f}(k)T_s \quad (10)$$

The stator flux predictive error is defined as

$$\Delta\psi_j(k+1) = \psi_j(k+1) - \psi_j^{ref}(k+1) = -\mathbf{f}(k)T_s \quad (11)$$

with

$$\mathbf{f}(k)T_s = \begin{bmatrix} \Delta\psi_d \\ \Delta\psi_q \end{bmatrix} = \begin{bmatrix} f_{dj}(k)T_s \\ f_{qj}(k)T_s \end{bmatrix} \quad (12)$$

According to (1) and (2), the voltage stationary equations can be expressed as follows when the N^*3 -phase PMSM reaches stable state [19]

$$\begin{cases} u_{dj}(k) = R_o i_{dj}(k) - L_o \omega_e i_{qj}(k) \\ u_{qj}(k) = R_o i_{qj}(k) + L_o \omega_e i_{dj}(k) + \omega_e \psi_{ro} \\ T_e(k) = \frac{3Nn_p}{2} \psi_{ro} i_{qj}(k) \end{cases} \quad (13)$$

Under $i_{dj}^{ref}(k) = 0$, combining with (6), (12), and (13), the d- and q-axis stator flux predictive errors can be simplified as follows

$$\begin{cases} \Delta\psi_d = -\frac{\chi_1 \cdot L_o}{(1+\chi_1)} \cdot \frac{2T_e(k)}{3Nn_p \psi_{ro}} \omega_e(k)T_s \\ \Delta\psi_q = \frac{\chi_2 \cdot \psi_{ro}}{(1+\chi_1)} \omega_e(k)T_s \\ \chi_1 = \Delta L / L_o \\ \chi_2 = \Delta\psi_r / \psi_{ro} \end{cases} \quad (14)$$

The N^*3 -phase PMSM parameters are shown in Table I. The basic value of the flux linkage is 0.892Wb. The simulation condition is: the load torque increases from 0 to 2000N.m; the speed increases from 0 to 800rad/s. According to (14), the d- and q-axis stator flux errors are plotted under parameters mismatch, as shown in Figs. 4 and 5. Fig. 4 shows the d- and q-axis stator flux predictive errors under inductance parameter mismatch. The q-axis stator flux predictive error is 0Wb under inductance parameter mismatch, which means that the inductance parameter mismatch does not affect the q-axis stator flux. At a speed of 800rad/s, the d-axis stator flux predictive error is as low as $-8e^{-3}$ Wb under inductance parameter mismatch ($\chi_1 = 100\%$). It can

be inferred that the influence of the inductance parameter mismatch on the q-axis stator flux is very small. Fig. 5 shows the d-and q-axis stator flux predictive errors under flux parameter mismatch. When flux parameter mismatch occurs, the d-axis stator flux predictive error is always 0Wb, and the q-axis stator flux predictive error is 0.08Wb. It can be inferred that flux parameter mismatch does not have considerable effect on d-and q-axis stator flux. Through the results described above, it is known that the predictive stator flux controller is immune to the $N*3$ -phase PMSM parameters, and the parameters mismatch has little effect on the control performance of $N*3$ -phase PMSM.

Fig. 4 d-and q-axis stator flux predictive errors under inductance parameter mismatch.

Fig. 5 d-and q-axis stator flux predictive errors under flux linkage parameter mismatch.

C. Predictive Stator Flux Control with One-step Delay Compensation

Due to the control delay in the actual digital system, there is a one beat delay between the actual output voltage vector and the ideal voltage vector of the predictive controller. In order to eliminate the influence of one beat delay on the predictive controller, it is necessary to calculate the output voltage vector one control cycle in advance. Fig. 6 demonstrates the one beat delay compensation in the predictive controller. The specific steps are as follows: Firstly, at the kT_s moment, the stator flux value $\psi_j(k+1)$ of $(k+1)T_s$ moment should be predicted. Then, the predicted stator flux value $\psi_j(k+1)$ is used to calculate the ideal voltage vector $u_j(k+1)$ at the kT_s moment.

Fig. 6 Logic diagram of a beat delay compensation.

At the kT_s moment, the stator flux value $\psi_j(k+1)$ of $(k+1)T_s$ moment can be predicted according to (9).

$$\psi_j(k+1) = u_j(k)T_s - A(k)\psi_j(k)T_s + BT_s \quad (15)$$

According to (15), the voltage vector of $(k+1)T_s$ moment can be calculated as

$$u_j(k+1) = \frac{1}{T_s} \psi_j(k+2) + A(k+1)\psi_j(k+1) - B \quad (16)$$

Note that the electrical time is far less than the mechanical time, the speed can be considered as unchanged in a control cycle (i.e., $\omega_e(k) \approx \omega_e(k+1)$), then $A(k) \approx A(k+1)$. At the kT_s moment, the output voltage vector of predictive stator flux controller can be obtained by substituting (15) into (16).

$$u_j(k+1) = \frac{1}{T_s} \psi_j(k+2) + T_s A(k) [u_j(k) - A(k)\psi_j(k) + B] - B \quad (17)$$

From (17), it is known that $u_j(k)$, $\psi_j(k)$, $A(k)$ and B are known quantities at the kT_s moment. Therefore, the voltage vector of $(k+1)T_s$ moment can be calculated by equation (17).

There exists $\frac{R}{L} \ll \frac{1}{T_s}$, then $A(k) \approx \begin{bmatrix} \frac{1}{T_s} & -\omega_e(k) \\ \omega_e(k) & -\frac{1}{T_s} \end{bmatrix}$, one yields

$$\begin{bmatrix} u_{dj}(k+1) \\ u_{qj}(k+1) \end{bmatrix} = \begin{bmatrix} -1 & -\omega_e(k)T_s \\ \omega_e(k)T_s & -1 \end{bmatrix} \begin{bmatrix} u_{dj}(k) \\ u_{qj}(k) \end{bmatrix} - \frac{1}{T_s} \begin{bmatrix} -1 & -T_s\omega_e(k) \\ T_s\omega_e(k) & -1 \end{bmatrix} \begin{bmatrix} \psi_{dj}(k) \\ \psi_{qj}(k) \end{bmatrix} + \begin{bmatrix} \frac{R_o}{L_o}\psi_{ro} \\ 0 \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \psi_{dj}(k+2) \\ \psi_{qj}(k+2) \end{bmatrix} - \begin{bmatrix} \frac{R_o}{L_o}\psi_{ro} \\ 0 \end{bmatrix} \quad (18)$$

In addition, the amplitude of the voltage source is much larger than the value of stator resistance voltage when the $N*3$ -phase PMSM operates stably (i.e., $\frac{R_o}{L_o}\psi_{ro} \ll u_{dj}(k)$). Therefore, Eq. (18) is modified to

$$\begin{cases} \begin{bmatrix} u_{dj}^{ref}(k+1) \\ u_{qj}^{ref}(k+1) \end{bmatrix} = T_s \begin{bmatrix} -\frac{1}{T_s} & -\omega_e(k) \\ \omega_e(k) & -\frac{1}{T_s} \end{bmatrix} \begin{bmatrix} u_{dj}(k) \\ u_{qj}(k) \end{bmatrix} - \begin{bmatrix} -\frac{1}{T_s} & -\omega_e(k) \\ \omega_e(k) & -\frac{1}{T_s} \end{bmatrix} \begin{bmatrix} \psi_{dj}(k) \\ \psi_{qj}(k) \end{bmatrix} \\ \quad + \frac{1}{T_s} \begin{bmatrix} \psi_{dj}^{ref}(k+2) \\ \psi_{qj}^{ref}(k+2) \end{bmatrix} \\ \begin{bmatrix} \psi_{dj}(k) \\ \psi_{qj}(k) \end{bmatrix} = \begin{bmatrix} L_o & 0 \\ 0 & L_o \end{bmatrix} \begin{bmatrix} i_{dj}(k) \\ i_{qj}(k) \end{bmatrix} + \begin{bmatrix} \psi_{ro} \\ 0 \end{bmatrix} \end{cases} \quad (19)$$

IV. ROBUST PREDICTIVE TORQUE CONTROL WITH PARAMETER ROBUSTNESS IMPROVEMENT

A. Design of Robust Predictive Torque Controller

According to (19), it is known that the key to realizing robust predictive stator flux control method is to obtain the d-and q-axis stator flux references value (i.e., $\psi_{dj}^{ref}(k+2)$, $\psi_{qj}^{ref}(k+2)$). In order to calculate

the accurate stator flux references value, the robust predictive torque controller is designed.

According to (2), the d-and q- axis stator flux linkage references can be calculated by using the Lagrange extrapolation under $i_{dj}^{ref}(k)=0$, expressed as follows

$$\begin{cases} \psi_{dj}^{ref}(k+2) = \psi_{dj}^{ref}(k+1) = \psi_{dj}^{ref}(k) = \psi_{ro} + L_o i_{dj}^{ref}(k) = \psi_{ro} \\ \psi_{qj}^{ref}(k+2) = 3\psi_{qj}^{ref}(k) - 3\psi_{qj}^{ref}(k-1) + \psi_{qj}^{ref}(k-2) = \frac{2L_o T_e^{ref}}{3Nn_p \psi_{ro}} \end{cases} \quad (20)$$

where T_e^{ref} is the torque reference.

According to (2), the electromagnetic torque equation can be rewritten as follow, when parameters mismatch are considered:

$$\tilde{T}_e = \frac{3n_p}{2} \sum_{j=1}^N ((\psi_{ro} + \Delta\psi_r) i_{qj}) = T_e + \Delta T_e \quad (21)$$

where ΔT_e is the unknown torque disturbance caused by flux parameter mismatch, and $\Delta T_e = \frac{3Nn_p \Delta\psi_r}{2} i_{qj}$.

Substituting (21) into (3), we can get

$$\frac{d\omega_e}{dt} = \frac{n_p}{J} (T_e + \Delta T_e - T_L) \quad (22)$$

According to (22), the speed equation of discrete-time form can be expressed as follow under parameter mismatch.

$$\omega_e(k+1) = \omega_e(k) + \frac{n_p}{J} T_d T_e(k) + \frac{n_p}{J} T_d \Delta T_e(k) - \frac{n_p}{J} T_d T_L \quad (23)$$

where T_d represents the speed sampling time, and $T_d = 10T_s$.

The torque reference can be given based on (23) as

$$T_e^{ref} = \frac{\omega_e^{ref} - \omega_e(k) - \frac{n_p}{J} T_d \Delta T_e(k) + \frac{n_p}{J} T_d T_L}{\frac{n_p}{J} T_d} \quad (24)$$

According to (2) and (3), the actual speed equation can be expressed as

$$\omega_e(k+1) = \omega_e(k) + \frac{n_p}{J} T_d T_e(k) - \frac{n_p}{J} T_d T_L \quad (25)$$

Substituting (24) into (25), the speed predictive error can be obtained as under steady-state operation.

$$\Delta\omega_e = \omega_e(k+1) - \omega_e^{ref} = -\frac{n_p}{J} T_d \Delta T_e(k) = -\frac{\chi_2 n_p}{J} T_d T_e(k) \quad (26)$$

The $N*3$ -phase PMSM parameters are shown in Table I. According to (26), the speed predictive error is plotted under flux linkage parameter mismatch, as shown in Fig. 7. From Fig. 7, it is known that the influence of flux parameter mismatch on predictive torque control is very small, and the maximum speed error is only 0.08rad/s.

Fig. 7 Speed predictive error under flux parameter mismatch.

In order to enhance robustness against the sudden change of load torque, the speed feedback value is replaced by the observed speed value. Thus, the robust predictive torque controller is designed as

$$T_e^{ref} = \frac{\omega_e^{ref} - \hat{\omega}_e(k) + \frac{n_p}{J} T_d \hat{T}_L}{\frac{n_p}{J} T_d} \quad (27)$$

where $\hat{T}_L$ is the observed values of T_L ; $\hat{\omega}_e$ is the observed value of ω_e .

The structural diagram of the 6*3-phase PMSM drive system with proposed PTSF method is illustrated in Fig. 8. Through the above analysis, it is known that the stator flux loop and torque loop of the $N*3$ -phase PMSM control system are immune to the motor parameters, which can obtain good control performance under parameters mismatch.

Fig. 8 Structural diagram of the 6*3-phase PMSM drive system with proposed PTSF method.

B. Design of Unknown Torque Disturbance Observer

In this section, the unknown torque disturbance observer based on integral terminal sliding surface is designed, which can simultaneously observer the unknown load torque and speed response.

The speed equation can be obtained by substituting (2) into (3).

$$\frac{d\omega_e}{dt} = \frac{3Nn_p^2 \psi_{ro}}{2J} i_{qj} - \frac{n_p}{J} T_L \quad (28)$$

The integral terminal sliding surface is defined as

$$s_\omega = e_\omega + \lambda \int_0^t \text{sgn}(e_\omega) d\tau \quad (29)$$

with

$$e_\omega = \hat{\omega}_e - \omega_e \quad (30)$$

where λ is the designed parameters, and $\lambda > 0$; $\text{sgn}()$ is sign function; $\hat{\omega}_e$ is the observed value of ω_e .

In order to achieve chattering suppression in the sliding mode observer, a hyperbolic tangent sign function is given as

$$\text{sgn}(v) = \tanh(v) = \frac{e^{2v} - 1}{e^{2v} + 1} \quad (31)$$

where v is any function variable.

According to (28), the sliding mode observer based on integral terminal sliding surface is designed as follows

$$\frac{d\hat{\omega}_e}{dt} = \frac{3Nn_p\psi_{ro}}{2J}i_{qj} - U_\omega \quad (32)$$

where U_ω is the sliding mode control function.

Combining with (28) and (32), and taking time derivative for (29), one yields

$$\dot{s}_\omega = \frac{n_p}{J}T_L - U_\omega + \lambda \operatorname{sgn}(e_\omega) \quad (33)$$

According to (33), the sliding mode control function is designed as follows to guarantee the convergence of sliding mode observer.

$$U_\omega = \lambda \operatorname{sgn}(e_\omega) + \eta s_\omega + \delta \operatorname{sgn}(s_\omega) \quad (34)$$

where η and δ are the designed observer sliding gain.

The following Lyapunov function is considered to prove the asymptotic stability of the proposed sliding mode observer.

$$\dot{V} = s_\omega \cdot \dot{s}_\omega \quad (35)$$

Substituting (33) and (34) into (35) yields

$$\begin{aligned} \dot{V} &= s_\omega \cdot \dot{s}_\omega = s_\omega \left(\frac{n_p}{J}T_L - U_\omega + \lambda \operatorname{sgn}(e_\omega) \right) \\ &= s_\omega \left(\frac{n_p}{J}T_L - \eta s_\omega - \delta \operatorname{sgn}(s_\omega) \right) \\ &= -\eta s_\omega^2 + \left(\frac{n_p}{J}T_L - \delta \|s_\omega\| \right) s_\omega \end{aligned} \quad (36)$$

The unknown load torque disturbance is bounded in engineer, that are $\|T_L\| \leq N_\omega$, where N_ω is the boundary values. When η and δ are set to $\eta \geq 0$ and $\delta \geq \frac{n_p}{J}N_\omega$, then $\dot{V} \leq 0$. Therefore, the proposed sliding-mode observer can reach the integral terminal sliding surface and maintain stable, that is, $e_\omega = \dot{e}_\omega = 0$, $s_\omega = \dot{s}_\omega = 0$. According to (33), the observed value of unknown disturbance can be obtained as

$$\hat{T}_L = \frac{J}{n_p}U_\omega(k) \quad (37)$$

If the sampling period τ_s is short enough, the discrete forms of (32) can be deduced as according to (34)

$$\begin{aligned} \hat{\omega}_e(k+1) &= \hat{\omega}_e(k) + \frac{3Nn_p\psi_{ro}}{2J}T_s i_{qj}(k) - \lambda T_s \operatorname{sgn}(e_\omega(k)) \\ &\quad - \eta T_s s_\omega(k) - \delta T_s \operatorname{sgn}(s_\omega(k)) \end{aligned} \quad (38)$$

Table I: MAIN PARAMETERS OF 6*3-PHASE PMSM

Parameters	Value
Stator phase resistance(R_s)	0.02Ω
Number of pole pairs (n_p)	4
Inductances(L_s)	3.572mH
Flux linkage of PM (ψ_{ro})	0.892Wb
Rotational inertia (J)	100kg.m ²

V. SIMULATIONS

The 6*3-phase PMSM drive system based on PTSF method is established in MATLAB/Simulink. The main parameters of 6*3-phase PMSM used in the MATLAB/Simulink and experiment are given in Table I, and the sampling frequency is set as 10 kHz. In addition, in order to guaranteed the stability and convergence of the unknown torque disturbance observer, the observer gains are set as $\lambda=5$,

$\eta=1000$ and $\delta=200$. The speed reference is 100rad/s, and the load torque reference is 1600N.m.

A. Simulation for N*3-phase PMSM under inductance parameter mismatch

Speed simulation results of the proposed PTSF and conventional PI are illustrated under inductance parameter mismatch, as shown in Figs. 9. At 1s, the inductance value is step to 50% of its initial values, i.e., 1.786 mH. From 0 to 1s, the conventional PI has the same superior speed control performance as the proposed PTSF, and the speed static errors of the two methods are about 0.2rad/s. After 1s, the speed response value of conventional PI have obvious fluctuation under inductance parameter mismatch, as shown in Fig. 9 (a). Compared with the conventional PI method, the speed response fluctuation is drastically reduced by using the proposed PTSF method in case of inductance parameter mismatch, as shown in Fig. 9 (b).

The comparative simulation results of the α - β stator flux linkage under inductance parameter mismatch are shown in Figs. 10. From 0 to 1s, the conventional PTC has the same superior stator flux control performance as the proposed PTSF when the parameter mismatch does not occur. After 1s, the α - β stator flux linkage response values of conventional PTC have obvious fluctuation, the fluctuation value is up to ± 0.3 Wb, as presented in Figs. 10(a). Fig. 10(b) shows that the α - β stator flux linkage fluctuation value of the proposed PTSF algorithm is only ± 0.03 Wb, the stator flux linkage fluctuation range is reduced by 90%. The comparative simulation results of three-dimensional rotor flux trajectories are shown in Figs.11. The conventional PTC method has a large flux linkage fluctuation error between the α - β stator flux linkage response value and the reference value under inductance parameter mismatch. While in the proposed PTSF method, the flux linkage fluctuation error is drastically reduced for the reason that it can immune to inductance parameters.

Fig. 9. Simulation results of the speed response under inductance parameter mismatch.

Fig. 10. Comparative simulation results of the α - β stator flux linkage under inductance parameter mismatch.

Fig. 11. Comparative simulation results of three-dimensional rotor flux trajectories under inductance parameter mismatch.

Fig. 12. Comparative simulation results of the torque and current under inductance parameter mismatch.

Fig. 13. Stator current frequency spectra of the i_a and i_{α} under inductance parameter mismatch.

Figs. 12 show the comparative simulation results of the torque ripple with the two methods. The peak-to-peak torque ripple of conventional PTC method is up to ± 700 N.m, while the proposed PTSF method is reduced by more than 57%, from ± 700 N.m to ± 400 N.m. Fig. 12(b) shows the simulation result of the estimated load torque under inductance parameter mismatch. From 0 to 0.4s, it can be seen from

Fig. 12(b) that there is a large estimation error between the observed value of load torque and the actual value of electromagnetic torque. The reason for the large estimation error is that the unknown torque disturbance observer does not reach the convergence state in the process of motor start-up. After 0.4s, it can be seen that the unknown torque disturbance observer based on integral terminal sliding surface can accurately estimate the load torque no matter in the case of inductance parameter mismatch or in the normal condition.

Figs. 13 show several stator current frequency spectra of the 6*3-phase PMSM and the PMSM unit. From Fig. 13(a), it is known that the stator current fundamental value of the 6*3-phase PMSM is about 312.2A, while that of the PMSM unit is about 52.04A, which is exactly six times. Fig. 13(b) also satisfies this mathematical relationship. The reason of the mathematical relationship is that the 6*3-phase PMSM meets the motor superposition characteristics. From Fig. 13(a) and 13(b), the THD of the conventional PTC method and proposed PTSF method are 22.37% and 2.53%, respectively. In summary, the conventional PTC is very sensitive to the inductance parameter, which will lead to the flux linkage fluctuate, the stator current distort and the torque ripple under inductance parameter mismatch. However, the above defects can be completely avoided for the reason that the PTSF method is immune to the inductance parameter.

B. Simulation for N *3-phase PMSM under flux linkage parameter mismatch

Speed simulation results of the proposed PTSF and conventional PI are given under flux linkage parameter mismatch, as shown in Figs. 14. At 1s, the flux linkage value is step to 50% of its initial values, i.e., 0.446Wb. The conventional PI has the same superior speed control performance as the proposed PTSF under flux linkage parameter mismatch, which are shown clearly in Fig. 14. Figs. 15 show the comparative simulation results of the α - β stator flux linkage under flux linkage parameter mismatch. The stator flux linkage response value of the conventional PTC is 1.32Wb, which is bigger than its reference value, that is, 0.892Wb, and the stator flux linkage static error is as high as 0.428Wb. From Fig. 15(b), the stator flux linkage static error of the proposed PTSF method is only 0.02Wb, which is reduced by 95.3%. From Fig. 16(a), it can be seen that there is a huge α - β stator flux linkage track error between the stator flux linkage response value and the reference value under flux linkage parameter mismatch. However, the α - β stator flux linkage track error can be effectively suppressed by using the proposed PTSF method, which are shown clearly in Fig. 16(b).

Figs. 17 show the comparative simulation results of the torque ripple under flux linkage parameter mismatch. Compared with the conventional PTC method, the peak-to-peak torque ripple of proposed PTSF method decreases from ± 500 N.m to ± 400 N.m, and the torque ripple is reduced by 20%. The simulation results of the estimated load torque under flux linkage parameter mismatch is illustrated in Fig. 17(b). From 0 to 0.4s, there is a large estimation error because the unknown torque disturbance observer does not reach the convergence state in the process of motor start-up. Note that there is a large overshoot when the flux linkage parameter are mismatched suddenly at 1s; the reason is that the unknown torque disturbance observer is affected by the parameter mismatch. After 1s, it can be known that the unknown torque disturbance observer can accurately estimate the load torque under flux linkage parameter mismatch. Figs. 18 show the stator current frequency spectra of the 6*3-phase PMSM and the PMSM unit with two methods. From Fig. 18(a) and Fig. 18(b), it can

be seen that the THD of the 6*3-phase PMSM is the same as that of PMSM unit. The THD of the proposed PTSF method and conventional PTC method are 0.91% and 1.96%, respectively. The reason for the large stator flux static error and torque ripple is that the conventional PTC must depend on the precise flux linkage parameter strictly.

Fig. 14. Simulation results of the speed response under flux linkage parameter mismatch.

Fig. 15. Comparative simulation results of the α - β stator flux linkage under flux linkage parameter mismatch.

Fig. 16. Comparative simulation results of three-dimensional rotor flux trajectories under flux linkage parameter mismatch.

Fig. 17. Comparative simulation results of the torque and current under flux linkage parameter mismatch.

Fig. 18. Stator current frequency spectra of the i_a and i_{aj} under flux linkage parameter mismatch.

C. Simulation results of the unknown torque disturbance observer under the low speed and high speed.

The speed estimation accuracy of the unknown torque disturbance observer under the low speed and high speed is shown in Figs. 19. The speed reference are respectively set as 50 rad/s and 500 rad/s under the low speed and high speed. The load torque suddenly increases from 800 to 1600 N. m, and it suddenly decreases to 800 N. m. From Figs. 19, it can be known that the unknown torque disturbance observer can accurately estimate the speed at low speed or high speed. In addition, the observed speed value does not overshoot under the sudden change of load torque, which are shown clearly in Figs. 19. However, the speed feedback value has an overshoot under the sudden change of load torque, but the overshoot value is very small. The simulation results show that the predictive control of speed loop will not lead to the degradation of the overall performance.

Fig. 19. Simulation results of the unknown torque disturbance observer under the low speed and high speed.

The simulation results of N *3-phase PMSM at low speed and high speed are shown in Figs. 20 and Figs. 21, respectively. From Fig. 20(a) and Fig. 21(a), it can be known that the peak-to-peak torque ripple of N *3-phase PMSM will increase by using the speed feedback value. In addition, it can be seen from Fig. 20(a) and Fig. 21(a) that the electromagnetic torque of N *3-phase PMSM has obvious overshoot under the sudden change of load torque, when the speed feedback value is adopted in predictive controller. The torque ripple and overshoot can be obviously suppressed by using the observed speed value, which are shown clearly in Fig. 20(b) and Fig. 21(b). In

summary, it can be known that the estimated speed value of unknown torque disturbance observer does not overshoot, and the predictive controller can obtain better control performance by using observed speed value. Therefore, in order to enhance robustness against the sudden change of load torque, the speed feedback value is replaced by the observed speed value.

Fig. 20. Simulation results of the $N*3$ -phase PMSM under the low speed.

Fig. 21. Simulation results of the $N*3$ -phase PMSM under the high speed.

D. Control performance of the load torque change under parameter mismatch

Simulation results of the load torque change under flux linkage parameter mismatch are shown in Figs. 22. The load torque suddenly increases from 800 to 1600 N. m at 0.8 s, and it suddenly decreases to 800 N. m at 1.2 s. The speed reference is set as 100 rad/s at 0 s. Fig. 22(a) shows the simulation result of the unknown torque disturbance observer under flux linkage parameter mismatch. From 0 to 0.35s, the load torque estimated by the unknown torque disturbance observer has a large estimation error. The reason for the large estimation error is that the speed does not run to a stable value in the process of motor start-up. After 0.35s, it can be seen from Fig. 22(a) that the unknown torque disturbance observer can accurately estimate the change of load torque under flux linkage parameter mismatch. Moreover, it can be known from Fig. 22(a) that the torque ripple of $N*3$ -phase PMSM is still very small by using the proposed PTSF method in case of the load torque change. Fig. 22(b) shows the simulation result of the α - β stator flux linkage under flux linkage parameter mismatch. From Fig. 22(b), it is known that the α - β stator flux response value of $N*3$ -phase PMSM can accurately track the reference value in case of load torque change, and the stator flux ripple is only ± 0.06 Wb.

Simulation results of the load torque change under inductance parameter mismatch is illustrated in Figs. 23. From Fig. 23(a), it can be known that the load torque estimated by the unknown torque disturbance observer can accurately estimate in case of inductance parameter mismatch. In addition, it can be seen from Fig. 23(a) that the torque ripple caused by inductance parameter mismatch is well suppressed by adopting the proposed PTSF method in case of the load torque change. The simulation result of the α - β stator flux linkage under inductance parameter mismatch is presented in Fig. 23(b). From

Fig. 23(b), it is known that the α - β stator flux ripple of the $N*3$ -phase PMSM is less than ± 0.04 N.m by using the proposed PTSF method in case of load torque change. Through the results analysis, it can be known that the $N*3$ -phase PMSM with parameter mismatch can also achieve excellent control performance by using the proposed PTSF method in case of load torque change.

Fig. 22. Simulation results of the load torque change under flux linkage parameter mismatch.

Fig. 23. Simulation results of the load torque change under inductance parameter mismatch.

E. Control performance of the $N*3$ -phase PMSM with different proportional parameters mismatch

Figs. 24 show the control performance of the $N*3$ -phase PMSM with 30% parameters mismatch. From Fig. 24, it can be known that the $N*3$ -phase PMSM with different proportional parameter mismatch can also obtain lower torque ripple and accurate load torque observation values by using the same parameters. Note that there is a large overshoot when the flux linkage parameter are mismatched suddenly at 1s; the reason is that the mismatching moment of flux linkage parameter will affect the observation accuracy of unknown torque disturbance observer.

Fig. 24. Simulation results of the proposed PTSF method under different proportional parameter mismatch.

F. Simulation for $N*3$ -phase PMSM under resistance parameter mismatch

Simulation results of the $N*3$ -phase PMSM under resistance parameter mismatch are shown in Figs. 25. At 1s, the resistance value is step to 50% of its initial values, i.e., 0.01 Ω . From Figs. 25(a), it can be known that the peak-to-peak torque ripple of the $N*3$ -phase PMSM

is both $\pm 400\text{N.m}$ in case of normal operation and resistance parameter mismatch. Moreover, it can be seen from Figs. 25(b) that the α - β stator flux linkage ripple of the N *3-phase PMSM is both $\pm 0.06\text{Wb}$ under normal operation and resistance parameter mismatch. In case of resistance parameter mismatch, the tracking performance of stator flux linkage is as accurate as that under normal condition, which are shown clearly in Fig. 25(b). Through the simulation analysis, it can be known that the influence of resistance parameter mismatch on the predictive controller is very small.

Fig. 25. Simulation results of the N *3-phase PMSM under resistance parameter mismatch.

VI. EXPERIMENT RESULTS

In this paper, the control hardware-in-the-loop experiments of the 6 *3-phase PMSM drive system is carried out to demonstrate the effectiveness of the proposed PTSF method. The RT-Lab hardware-in-the-loop simulation (HILS) platform is shown in Fig. 26 [19, 25]. The RT-Lab HILS platform is composed of the DSP controller, OP5600 simulation motor, and 6 *3-phase PMSM system model as software component. The TMS320F2812 is selected as the DSP controller. The HILS of 6 *3-phase PMSM drive system can be simulated by establishing the compiled code of 6 *3-phase PMSM and using the inverter model in OP5600. The time consumption of the proposed PTSF method in real time has been considered. The proposed PTSF method is performed on the platform of the TMS320F2812 processor. The system parameters are consistent with the simulation parameters.

Fig. 26 hardware-in-the loop simulation platform.

Figs. 27 and 28 show the comparative experimental results of the conventional PTC and proposed PTSF under inductance parameter mismatch. It can be seen from Fig. 27(a) and Fig. 28(a) that the obvious oscillation exists in the α - β stator flux linkage and torque wave form under inductance parameter mismatch, which indicates that the conventional PTC is mostly sensitive to inductance parameter. Note that the stator current using conventional PTC method has obvious distortion in the presence of inductance parameter mismatch, as shown in Fig. 28(a). While using the proposed PTSF method, since the PTSF method is immune to the inductance parameter of N *3-phase

PMSM, the stator flux linkage and torque wave form can maintain smooth without oscillation, as shown in Fig. 27(b) and Fig. 28(b). Note that the torque using the proposed PTSF method has a small overshoot when the inductance parameters are mismatched, as shown in Fig. 28(b). The reason for this phenomenon is that the proposed PTSF is compensated by unknown torque disturbance observer.

Fig. 27. Experimental results of the α - β stator flux linkage under inductance parameter mismatch.

Fig. 28. Experimental results of the torque and stator current under inductance parameter mismatch.

Fig. 29. Experimental results of the α - β stator flux linkage under flux linkage parameter mismatch.

Fig. 30. Experimental results of the torque and stator current under flux linkage parameter mismatch.

Fig. 29 and 30 show the comparative experimental results of the conventional PTC and proposed PTSF under flux linkage parameter mismatch. Comparative experimental results illustrate that the conventional PTC produces the obvious static tracking error in the α - β stator flux linkage, which are shown clearly in Fig. 29(a). The stator flux linkage response value of the proposed PTSF strictly follows the reference value, and the flux linkage parameter are not needed in the control operation process, thus, the proposed PTSF method can significantly suppress the influence of flux linkage parameter mismatch, as shown in Fig. 29(b). Fig. 30(b) demonstrates that the

proposed PTSF method provides satisfactory torque wave form with small torque ripples in the presence of flux linkage parameter mismatch. Compared with the proposed PTSF method, the conventional PTC method presents obvious torque ripple, as shown in Fig. 30(a). The reason of torque ripple is that the predictions of conventional PTC method must depend on the precise mathematical model of the $N*3$ -phase PMSM. In addition, it can be known from Fig. 30, that the stator current will suddenly increase when the flux linkage parameters are mismatched. The reason for this phenomenon is that the stator current of $N*3$ -phase PMSM should be large enough to keep the torque constant under flux parameter mismatch. In Figs. 31, the experimental results of the load torque change under the parameters mismatch are given by using the unknown torque disturbance observer. The experimental results show that the change of load torque can accurately be estimated by using the unknown torque disturbance observer in the case of flux linkage parameter mismatch or inductance parameter mismatch.

(a) Flux linkage parameter mismatch (b) Inductance parameter mismatch
Fig. 31. Experimental results of the load torque change under parameter mismatch.

(a) Conventional PTC (b) Proposed PTSF
Fig. 32. Experimental results of the $N*3$ -phase PMSM under inductance parameters mismatch.

(a) Conventional PTC (b) Proposed PTSF
Fig. 33. Experimental results of the $N*3$ -phase PMSM under flux linkage parameters mismatch.

(a) Flux parameter mismatch (30%) (b) Inductance parameter mismatch (30%)
Fig. 34. Experimental results of the proposed PTSF method under different proportional parameter mismatch.

In Figs. 32, the experimental results of the conventional PTC and proposed PTSF under inductance parameter mismatch are given. The experimental results of the conventional PTC and proposed PTSF under flux linkage parameter mismatch are illustrated in Figs. 33. It can be seen from Figs. 32 and Figs. 33 that the response speed value of conventional PTC is as good as that of the proposed PTSF under parameters mismatch. However, compared with the conventional PTC method, the torque ripple and stator current distortion of $N*3$ -phase PMSM is drastically reduced by adopting the proposed PTSF method in case of parameters mismatch, as shown in Figs. 32(b) and Figs. 33(b). In Figs. 34, the experimental results of the proposed PTSF method under different proportional parameters mismatch are given. In the case of different proportional parameters mismatch, it can be known from Figs. 34 that the proposed unknown torque disturbance observer can also achieve excellent estimate performance by using the same parameters.

VII. CONCLUSION

A novel robust predictive torque and stator flux control method for $N*3$ -phase PMSM, called PTSF, has been proposed in this paper, which can effectively enhance robustness against parameter mismatch. The robust predictive stator flux controller is formulated in the dq stator flux space, which can be independent of the motor parameters and compensate the one-step delay. Furthermore, the robust predictive torque controller is designed based on unknown torque disturbance observer, which can enhance robustness against parameters mismatch and load torque disturbance. Compared with the conventional PTC, the proposed PTSF method can effectively enhance robustness against parameters mismatch by adopting the closed-loop control of d - q axis stator flux. Comparative simulation and experimental results demonstrate that the proposed PTSF method outperforms the conventional PTC in yielding high stator flux precision tracking and satisfactory parameter immunity under parameters mismatch while achieving smaller stator current THD and lower torque/flux ripple.

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