# Model-Free Predictive Current and Speed Control for Modular Drive of a Non-Sinusoidal, Six-Phase Permanent Magnet Synchronous Motor with Double-Winding Stator based on Extended State Observers

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## Abstract

Multi-phase PMSMs are considered to increase the reliability of the high-power propulsion systems. This paper presents a model-free current and speed predictive control (MFPCSC) method based on ultra-local model for an asymmetric six-phase PMSM with non-sinusoidal back-EMFs. The proposed MFPCSC method is robust to model uncertainties and disturbances. To maximize the reliability, each phase has two windings and each winding has an open-end connection, in which the drive topology is fully modular. Due to the modular structure, there are limitations in the modeling and control of the motor, in which the conventional methods in *dq* reference frames cannot be employed. Also, due to torque ripples caused from non-sinusoidal back-EMFs, the current shaping method has been used in the six-axis stationary reference frame. To make the control method robust to changes in motor model parameters and reduce sensitivity to current controllers, six extended state observers (ESOs) based on an ultra-local model are used to estimate current and total disturbances. The superiority of the proposed control method over conventional control methods based on hysteresis, PI and proportional-resonant (PR) current regulators is verified by simulation in Simulink, and some experimental test results are presented to validate the proposed theories.



## Keywords

Asymmetric six-phase PMSM with double winding, model-free predictive control, extended state observer, torque ripple, current shaping method.

## 1. Introduction

The use of multi-phase electric motors has gained wide attention and application among industrialists and researchers in recent years [1, 2]. Multi-phase motors are typically used in medium- to high-power applications to increase reliability and reduce phase current stress, such as in ships, submarines, and mills [3, 4]. These refer to the motors with more than three phases, such as 5, 6, 7, 9, 12, and more, in which the six-phase motors being the most common type among them. The advantages of these motors over other multi-phase motors include their simple control, similarity to three-phase motors, as well as ease of design, and construction. Nowadays, six-phase motors are widely used in sensitive applications where a higher level of redundancy, reliability, and fault-tolerant control is required, especially in sub-marine propulsion systems [5]. In these applications, high reliability is crucial, and for this purpose, redundancy in designs, both in hardware and control systems, is definitely considered. For example, increasing the number of windings per phase, using open-end type connection for each winding, using independent H-bridge inverters for each winding, and redundancy in the number of microcontrollers are among these considerations [6].

Six-phase PMSMs can be divided into two main categories: symmetrical and asymmetrical. In the asymmetrical type, as shown in Fig. 1, the distance between two adjacent windings axes is usually 30 electrical degrees, while in the symmetrical type, it is 60 degrees [7]. In both types, two set windings with a star connection can be created using six phases, as shown in Fig. 1(a). In this case, the six-phase motor is known as a three-phase motor with double star windings [8]. For powering the motor, two independent three-phase inverters are used. To enhance reliability, the open-end winding connection in the stator can be utilized as shown in Fig. 1(b). In this configuration, each winding is powered by an independent single-phase H-bridge inverter.

The reliability of open-end topology is much higher than other six-phase even double-star topology and offers significantly greater flexibility in using fault-tolerant control methods [9]. To extend the reliability of the drive for open-end six-phase motors, in some designs, each stator phase winding is divided into two sections that are aligned with each other, with each occupying one half of the stator circumference. The schematic of this winding structure, known as the asymmetric double six-phase motor, is shown in Fig. 2. To maximize the reliability, each winding is powered up by a single-phase inverter as shown Fig. 3 and both inverters of each phase are controlled via independent local control system (microcontroller), whereas a central microcontroller supervises the local microcontrollers. This drive structure, where each phase has separate windings, independent inverters, and control circuit, is known as a fully modular structure and has the highest degree of reliability.

Note, the control signals sent by the central microcontroller to the local microcontroller that is applied to both inverters of each phase are exactly the same. One limitation of this structure is that, due to the modular control system, there is no data communication between local microcontrollers, and they are blind of other phases' voltage and current data. To control this specific type of PMSM, the dynamic model must be developed. For dynamic modeling of a three-phase sinusoidal PMSM, the motor model is usually represented in the rotating *dq* reference frame. For the six-phase PMSM, similar to the three-phase motor modeling method, several modeling methods exist, including:

(1) Modeling in independent two-axis reference frames [10]

(2) Modeling in independent, decoupled two-axis reference frames [11]

(3) Modeling using vector space decomposition (VSD) method [12]

(4) Modeling in six-axis stationary reference frame [13]

For six-phase PMSM with sinusoidal back EMFs, the first three modeling methods mentioned above are usually employed, and there is no significant difference in computational complexity between them. However, for modeling non-sinusoidal PMSMs, there are many considerations when using these three methods. Proposed methods for modeling of non-sinusoidal PMSMs are including:

(a) Non-sinusoidal vectorial or extended Park method [14]

(b) Modeling method in one *dq* reference frame considering harmonic components [15]

(c) Modeling in multiple harmonic *dq* reference frames (MRF) [16]

(d) Vector space decomposition (VSD) method [17]

The VSD method is one of the most powerful modeling methods. In this method, using appropriate 3-to-2 transformations, the main component of the motor signals is transferred to the p $\boldsymbol{z}_1\boldsymbol{z}_2$ rimary subspace (αβ), while harmonics of the 5th and 7th orders are transferred to the secondary subspace (), and the 3rd order harmonics are mapped to the zero-sequence subspace ( $\boldsymbol{o}_1\boldsymbol{o}_2$ ). Advantages of the VSD modeling method include working with simpler models in three two-dimensional subspaces, easier controller design, and more straightforward harmonic control. Using these methods for non-sinusoidal three-phase PMSMs involve significant computational complexity. Using these methods for six-phase PMSMs increases the computational volume even further.

However, the use of these and other modeling methods based on *dq* transformations is feasible only when a unified control structure is employed for the motor. In the drive system used in this research, shown in Fig. 3, the control calculations for each phase are performed in the local microcontroller. Each local microcontroller has access only to the voltage and current of its phase windings. Therefore, performing *dq* transformations are not possible. As a result, for modeling the modular six-phase PMSM in this research, modeling in six-axis stationary reference frames should be utilized.

As mentioned, there are many different control methods for three-phase PMSMs. Some of these methods can be applied to six-phase PMSMs as well. However, due to the increased number of phases, the computational volume, processing limitations in microcontrollers, and the poor performance of some methods in high-power motors, most of them are practically used less frequently [18]. Vector control is the common and main method for controlling six-phase PMSMs. For six-phase motors, vector control methods are applied in independent *dq*, decoupled *dq* reference frames and VSD [19]. The high computational requirements of Park transformations, the difficulty of tuning multiple PI controller gains, and the sensitivity of vector control methods to changes in motor model parameters are disadvantages of vector control for six-phase motors. The use of direct torque control (DTC) is an attempt to address some of the disadvantages of vector control, such as the number and tuning of PI controllers [20,21]. However, DTC also suffers from inherent drawbacks such as constant ripple in current and torque and significant high-frequency noise. In recent years, model predictive control (MPC) methods have been widely used for controlling three-phase PMSM drives are mainly classified into two categories: model predictive current control (MPCC) and model predictive torque control (MPTC). The performance of MPCC is sensitive to the accuracy of the motor parameters and uncertainties [22]. In [23], a model-free control method based on an ultra-local model is proposed to compensate for the effects of parameter changes, inverter nonlinearity. However, tuning the coefficients of this method is difficult, the estimated disturbances have high fluctuations, and the estimation accuracy is low, especially at low sampling frequencies [24].

Using observer-based control methods, such as the nonlinear disturbance observer (NDO), and the extended state observer (ESO), and sliding mode observer (SMO) can achieve robust control against motor parameter changes by observing and compensating for disturbances caused by parameter variations [25,26,27]. In [28], two NDOs are used for model-independent predictive control of both current and speed simultaneously, eliminating the need for any PI controllers for either current components or speed. Additionally, in [29], an ESO is used to estimate model uncertainties and unmodeled dynamics of the three-phase PMSM motor in the *dq0* frame. Results show that this method performs better in terms of tracking error, current harmonics, and transient overshoot [29, 30]. Presented methods have been for three-phase PMSMs. However, for six-phase PMSMs, few studies have been conducted using two-level and three-level inverters [31, 32]. Furthermore, all the aforementioned predictive control methods use the motor model in the ***dq*** reference frames and thus require the use of transformations. In the modular drive structure of this article, none of the above modeling and control methods for three-phase and six-phase PMSMs based on *dq* theories are not applicable. Therefore, in this article, a new model-free predictive current and speed control method based on the ESO observer in the six-axis stationary reference based on ultra-local) model is presented.

In the continuation of this article, section 2 focuses on modeling the six-phase non-sinusoidal PMSM in the stationary six-axis frame and then presents the control of this motor using the current shaping method. In section 3, the mathematical modeling of model-free predictive control based on ultra-local model is presented. Then, the ESO is designed in the stationary reference frame and subsequently used for motor control. Section 4 presents the simulation results of controlling this motor using the ESO observer and compares them with quasi-proportional-resonant current controllers. In section 5 some experimental results are presented and section 6 gives some conclusion.

## 2. Modeling and Control of Non-Sinusoidal Six-Phase PMSM with Double Windings

### 2.1. Modeling in the stationary six-axis frame

Given the modular topology of this article as shown in Fig. 3, there is no data exchange between local microcontrollers, none of the modeling methods based on ***dq*** transformations can be used. Additionally, the non-sinusoidal nature of the back EMFs further complicates the issue. Therefore, for a six-phase motor with the stator winding structure shown in Fig. 1(b), modeling in the six-axis stationary reference frame must be used, where the six axes align with the winding.

Corresponding to the windings shown in Fig. 2, the voltage-current relationship for each phase is as follows:

$$v_{x_i} = R_s i_{x_i} + \frac{d}{dt}\psi_{x_i} + e_{x_i} \tag{1}$$

where $R_s$ is the resistance of each stator phase, $\boldsymbol{v}_{x_i}$ , $\boldsymbol{i}_{x_i}$ , $\boldsymbol{\psi}_{x_i}$ , and $\boldsymbol{e}_{x_i}$ , are the voltage, current, flux linkage, and induced back-EMF voltage of winding $\boldsymbol{x_i}$ , respectively (where $x = A, B, C, X, Y, Z$ and i=1,2). The flux linkage of each winding depends on the current of the same winding and the current of the other 11 windings, that can be calculated as follow:

$$\psi_{x_i} = L_{x_i} i_{x_i} + \sum_{\substack{i,j=1,2 \\ (x_i \neq y_j)}}^{x,y=A,B,C,X,Y,Z} M_{x_i y_j} i_{y_j} \tag{2}$$

where $\boldsymbol{L}_{x_i}$ is the self-inductance of winding $\boldsymbol{x_i}$ and $\boldsymbol{M}_{x_i y_j}$ is the mutual inductance between windings $\boldsymbol{x_i}$ and $\boldsymbol{y_j}$. The stator inductance matrix, when the two windings of each phase (e.g. $A_1$ and $A_2$) are separate, is a 12×12 matrix. For enhanced performance of the motor, at low speed, they are connected in series and so the inductance matrix becomes a 6×6 matrix. The electromagnetic torque $\boldsymbol{T_e}$ of the asymmetric six-phase motor with double windings and the motor speed $\omega_m$ can be calculated as follows:

$$T_e = \sum_{\substack{x=A,B,C,X,Y,Z \\ i=1,2}} \frac{e_{x_i} i_{x_i}}{\omega_r} \tag{3}$$

$$\omega_m = \frac{1}{J}\int \left(T_e - B\omega_m - T_l\right) dt \tag{4}$$

where $\omega_r$, $J$, $\boldsymbol{B}$ , and $T_l$ are the electrical rotor speed, rotor inertia moment, friction coefficient and load torque, respectively. To implement of the model in Simulink, a 12-phase winding can be used, that to each phase, a dependent voltage source with desired harmonic waveform function of rotor position $\theta_r$ and amplitude as function of rotor speed, can be added in series.

### 2.2. Torque ripple reduction control strategy based on reference current shaping method

Due to the lack of data communication between the microcontrollers of each phase and the other phases, it is not possible to use control methods based on *dq* theory. In addition, the high number of motor windings and the non-sinusoidal nature of the back-EMFs are more than just a reason. In this section, to reduce the torque ripple caused by the non-sinusoidal back-EMFs, the harmonic current injection or the reference current shaping in a six-axis stationary reference frame is utilized. In the proposed method, corresponding to the harmonics content of the phase back-EMF waveform, the reference current is shaped using harmonic components of the current so that the torque ripple becomes zero and only the constant part of the torque remains. For the six-phase motor of this research, due to experimental measurements using embedded search coils beside the stator windings, only the harmonics with orders 1, 3, and 5 are present in the back-EMF voltage of each phase. On this way, the back-EMF voltage of the winding $\boldsymbol{x_i} = \mathbf{A}_1$ is considered as:

$$e_{x_i}(t) = E_1 \sin \omega_r t + E_3 \sin 3\omega_r + E_5 \sin 5\omega_r t \tag{5}$$

where $E_1$, $E_3$ and $E_5$ are the amplitudes of phase back-EMF voltage harmonics, which are 1, 0.07, and -0.03 per unit, respectively. These values have been extracted via experimental tests using embedded search coils embedded alongside the main stator windings and they are almost constant in different load torque and speeds. Consider the injected current into the winding $\boldsymbol{x_i} = \mathbf{A}_1$ as follows:

$$i_{x_i}(t) = I_1 \sin \omega_r t + I_3 \sin 3\omega_r t + I_5 \sin 5\omega_r t \tag{6}$$

For other windings, the argument of the sinusoidal functions changes according to Fig. 1(b). The air gap power of the winding $\boldsymbol{x_i} = \mathbf{A}_1$ will only include even order harmonics up to the $10^{\text{th}}$ order as follows:

$$P_{x_i}(t) = P_0 + P_2 \sin 2\omega_r t + \ldots + P_{10} \sin 10\omega_r t \tag{7}$$

The total air gap power of all windings in a PMSM with phase number multiple of 3, contains only harmonic with orders 6, 12, and 18. Assuming current and voltage harmonics are limited to the 5$^{th}$ order for the motor in this study, the total instantaneous air gap power $P_e$ is obtained as follows:

$$P_e(t) = P_0 + P_6 \sin 6\omega_r t \tag{8}$$

The instantaneous electromagnetic torque can also be calculated as:

$$T_e(t) = \frac{P_e(t)}{\omega_r} = T_0 + T_6 \sin 6\omega_r t \tag{9}$$

where:

$$T_0 = \frac{3}{2\omega_r}\left[E_1 I_1 + E_3 I_3 + E_5 I_5\right] \tag{10}$$

$$T_6 = \frac{3}{2\omega_r}\left[-I_1 E_5 + I_3 E_3 - I_5 E_1\right] \tag{11}$$

To determine the amplitudes of the current harmonics for each phase, by setting $\boldsymbol{T}_0$ equal to the reference torque $\boldsymbol{T_e^*}$ (the output of the speed controller) and setting $\boldsymbol{T}_6$ to zero, the following matrix equation must be solved:

$$\begin{bmatrix} E_1 & E_3 & E_5 \\ -E_5 & E_3 & -E_1 \end{bmatrix} \times \begin{bmatrix} I_1 \\ I_3 \\ I_5 \end{bmatrix} = \frac{2\omega_r}{3}\begin{bmatrix} T_e^* \\ 0 \end{bmatrix} \tag{12}$$

In above matrix equation, the number of unknown currents is 3 that is more than the number of equations. This refer that it has not a unique solution. Hence, an optimization must be performed to extract the best solution. An optimal method to solve this equation is such that, in addition to satisfying the above equations, the rms value of the current vector $\boldsymbol{I} * \boldsymbol{I}^{\boldsymbol{t}}$ should be also minimized, thereby obtaining the optimal harmonic values of the current. Thus, consider the following optimization problem:

$$Minimize \;\; J = xx^T \tag{13}$$

$$x = [I_1 \quad I_3 \quad I_5].$$

$$if\,;\,Ax = B$$

$$B = \frac{2\omega_r}{3}\begin{bmatrix} T_e^* \\ 0 \end{bmatrix}, A = \begin{bmatrix} E_1 & E_3 & E_5 \\ -E_5 & E_3 & -E_1 \end{bmatrix}$$

The optimal solution of above problem is obtained as follows [38]:

$$x_{opt} = A^T \times \left(A \times A^T\right)^{-1} \times B. \tag{14}$$

Using the abe solution, and substituting the given amplitudes of the $E_1$, $E_3$ and $E_5$ by 1, 0.07, and -0.03, the optimal values of the current harmonic amplitudes are obtained as follows:

$$\begin{bmatrix} I_1^* \\ I_3^* \\ I_5^* \end{bmatrix} = \begin{bmatrix} 0.9956 \\ 0.0736 \\ 0.0247 \end{bmatrix} \frac{2\omega_r T_e^*}{3} \tag{15}$$

**2.3. Phase current regulation**

Using Eq. (6) and Eq. (15), the time relation of the reference current for both windings of the phase A, based on the reference torque, can be determined as follows:

$$i_A^*(t) = I_1^* \sin \omega_r t + I_3^* \sin 3\omega_r t + I_5^* \sin 5\omega_r t \tag{16}$$

The current of phase A windings, in all transient and steady states, must accurately track the aforementioned reference value. For this purpose, appropriate current controllers should be used. The simplest type of controller is the hysteresis controller, which, due to its steady-state error in current and high, and variable switching frequency, is not suitable for high-power applications sensitive to acoustic noise. On the other hand, the traditional PI controller is also not suitable for this application because the PI controller, due to its small bandwidth, is not capable of tracking AC signals containing high-order harmonics [34]. An effective and suitable solution for tracking harmonic reference signals is the using of proportional-resonant (PR) controllers or quasi-proportional-resonant (QPR) controllers. The transfer function of the harmonic QPR controller for the harmonic reference current given by relation (16) is as follows:

$$\begin{aligned} G_{QPR,H}(s) = K_P + K_{R_1}\frac{2\omega_{c_1}s}{s^2+2\omega_{c_1}s+\omega_o{}^2} + \\ K_{R_3}\frac{2\omega_{c_3}s}{s^2+2\omega_{c_3}s+(3\omega_o)^2} + \\ K_{R_5}\frac{2\omega_{c_5}s}{s^2+2\omega_{c_5}s+(5\omega_o)^2} \end{aligned} \quad (17)$$

In which the coefficients $\boldsymbol{K_P}$, $\boldsymbol{K_{R_1}}$, $\boldsymbol{K_{R_3}}, \boldsymbol{K_{R_5}}$ and the frequency $\boldsymbol{\omega_{c_1}}, \boldsymbol{\omega_{c_3}}, \boldsymbol{\omega_{c_5}}$ must be accurately adjusted [35,36]. Various methods exist for designing and determining QPR controller parameters, the main ones being: (1) trial-and-error method, (2) forced oscillation method (time domain) and (3) frequency response method (frequency domain). The trial-and-error method is a straightforward and simple method that involves manually adjusting the controller gains and parameters while observing the motor's response to different input signals. The procedure for this method is detailed in the appendix.

## 3. Control of Six-phase PMSM motor Using Extended State Observer (ESO)

In this section, an ultra-local model of the six-phase PMSM is firstly developed, and then the design of the extended state observer (ESO) is presented. Subsequently, ESO is used for regulation of non-sinusoidal reference current determined by Eq. (16).

### 3.1. Ultra-local mathematical model theory

An ultra-local model for model-free predictive control of a single-input, single-output system using only the system's input and output while disregarding its mathematical model can be expressed as follows [37]:

$$\dot{y} = \alpha u + F \quad (18)$$

where $\boldsymbol{y}$ and $\boldsymbol{u}$ are the output and input of the system respectively, $\boldsymbol{F}$ is the sum of the known and unknown disturbances of the system, and  is the non-physical scaling coefficient of the designed model. Considering $\boldsymbol{y}^*$ as the reference output value of the system, $\hat{\boldsymbol{F}}$ as the estimated value of $F$, and $\boldsymbol{\xi}$ as control effort, the control law of the model-free controller is expressed as follows:

$$u = \frac{\dot{y}^* - \hat{F} + \xi}{\alpha} \tag{19}$$

where $\boldsymbol{\xi}$ can be the output of a controller such as a proportional-integral (PI) controller. Assuming $\hat{\boldsymbol{F}}$ is estimated with high accuracy, meaning $\hat{\boldsymbol{F}} \approx F$, from Eq. (18) and Eq. (19) it results:

$$\dot{e} + \xi = 0 \tag{20}$$

where $\boldsymbol{e} = \boldsymbol{y}^* - \boldsymbol{y}$ represents the output tracking error. Assuming that $\boldsymbol{\xi}$, the output of the proportional controller, is in the form $\boldsymbol{\xi} = \boldsymbol{K_p e}$, the control law of the closed-loop nonlinear system is described as follows:

$$u = \frac{\dot{y}^* - \hat{F} + k_p e}{\alpha} \quad \dot{e} + k_p e = 0 \tag{21}$$

**3.2. Ultra-local model of the current in six-phase PMSM**

To obtain the ultra-local model of the modular six-phase PMSM in the stationary stator frame, the following steps are taken. By rewriting Eq. (1) and using Eq. (2) for a typical winding such as winding $A_1$ of phase A, we have got:

$$v_{A_1} = R_s i_{A_1} + \frac{d}{dt}\psi_{A_1} + e_{A_1} \tag{22}$$

or

$$v_{A_1} = R_s i_{A_1} + L_s \frac{di_{A_1}}{dt} + \frac{d}{dt}\sum\nolimits_{x_j = B_1 \ldots Z_2} M_{A_1 x_j} i_{x_j} + \omega_r \psi_{PM} \sin(\theta_r) \tag{23}$$

It is observed that, in Eq. (23), the voltage of winding $\boldsymbol{A}_1$ depends on the currents of other windings. However, due to the modular topology of the power and control system, their values are not available, and the induced voltage due to the currents of other windings can be considered the unmodeled dynamics of the system. Considering parameter variations, unknown disturbances, and unmodeled dynamics, the mathematical model of winding $\boldsymbol{A}_1$ can be expressed as follows:

$$v_{A_1} = \left(R_s + \Delta R\right) i_{A_1} + \left(L_s + \Delta L\right) \frac{di_{A_1}}{dt} + \omega_r \left(\psi_{PM} + \Delta \psi_{PM}\right) \sin(\theta_r) + f_{A_1} \tag{24}$$

that, $\Delta R$, $\Delta L$, and $\boldsymbol{\Delta\psi_{pm}}$ represent the parameter variations of each phase, and $\boldsymbol{f_{A1}}$ indicates the unknown disturbances and unmodeled dynamics in winding $\boldsymbol{A}_1$. Eq. (24) can be rewritten as follows:

$$\frac{di_{A_1}}{dt} = \frac{v_{A_1}}{L_s} + \left( -\frac{\left(R_s + \Delta R\right) i_{A_1}}{L_s} - \frac{\Delta L}{L_s}\frac{di_{A_1}}{dt} - \frac{\omega_r \left(\psi_{PM} + \Delta \psi_{PM}\right) \sin(\theta_r)}{L_s} - \frac{f_{A_1}}{L_s} \right) \tag{25}$$

If the currents of each of the 12 windings of the motor are considered the system outputs, the applied voltages to the motor windings are considered the system inputs, and the other terms are considered the sum of the known and unknown disturbances of the system. Based on Eq. (18), the ultra-local model of the PMSM drive for winding $\boldsymbol{A}_1$ can be expressed as follows:

$$\frac{di_{A_1}}{dt} = \alpha v_{A_1} + F_{A_1} \tag{26}$$

where $\alpha$ is the controller coefficient equal to the inverse of the synchronous inductance ($\alpha = 1/L_S$), and the sum of the disturbances for winding $\boldsymbol{A}_1$ is described as follows:

$$F_{A_1} = -\frac{\left(R_s + \Delta R\right) i_{A_1}}{L_s} - \frac{\Delta L}{L_s}\frac{di_{A_1}}{dt} - \frac{\omega_r \left(\psi_{PM} + \Delta \psi_{PM}\right) \sin(\theta_r)}{L_s} - \frac{f_{A_1}}{L_s} \tag{27}$$

For other windings, the ultra-local model can also be presented in the same manner, only with changes to the winding index and the argument of the sinusoidal back EMF function, as shown in

Fig. 2. In the next section, an extended state observer (ESO) is designed to estimate the sum of the disturbances for each winding.

### 3.3. Extended state observer (ESO) design for current control

Based on the ultra-local model described by Eq. (24), a linear ESO observer can be designed, with the state variables estimating the values of $\boldsymbol{i}_{A1}$ and $\boldsymbol{F}_{A1}$. This observer can be designed using the feedback of estimation error of current $\boldsymbol{i}_{A1}$ as follows [29, 30]:

$$\begin{aligned} &\boldsymbol{e} = \boldsymbol{x}_1 - \boldsymbol{i}_{A_1} \\ &\dot{\boldsymbol{x}}_1 = \boldsymbol{x}_2 + \boldsymbol{\alpha} \boldsymbol{v}_{A_1} - \boldsymbol{\beta}_1 \boldsymbol{e} \\ &\dot{\boldsymbol{x}}_2 = -\boldsymbol{\beta}_2 \boldsymbol{e} \end{aligned} \tag{28}$$

where $\boldsymbol{x}_1 = \hat{\boldsymbol{i}}_{A_1}$ and $\boldsymbol{x}_2 = \hat{\boldsymbol{F}}_{A_1}$ are the state variables of the observer, which are the estimated values $\boldsymbol{i}_{A1}$ and $\boldsymbol{F}_{A1}$ respectively. The coefficients $\boldsymbol{\beta}_1$ and $\boldsymbol{\beta}_2$ are the feedback gains of the observer's error, which influence the estimation quality and must be appropriately determined. To design the observer and determine the appropriate values for $\boldsymbol{\beta}_1$ and $\boldsymbol{\beta}_2$, the state-space representation of the linear ESO can be expressed as follows:

$$\begin{aligned} &\dot{x} = Ax + Bv_{A_1} + D\left(y - \hat{y}\right) \\ &\hat{y} = Cx \end{aligned} \tag{29}$$

where $\hat{\boldsymbol{y}}$ is the estimated output $y$, and the vector $x$ and observer matrices are as follows:

$$\begin{aligned} &x = \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}, \\ &A = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix}, B = \begin{bmatrix} \alpha \\ 0 \end{bmatrix}, C^T = \begin{bmatrix} 1 \\ 0 \end{bmatrix}, D = \begin{bmatrix} \beta_1 \\ \beta_2 \end{bmatrix} \end{aligned} \tag{30}$$

The characteristic equation of the ESO is expressed as follows:

$$\left| sI - \left(A - DC\right) \right| = s^2 + \beta_1 s + \beta_2 \tag{31}$$

In order for the roots of the above characteristic equation to lie in the left half of the real axis of the complex plane, the following conditions must be met:

$$\beta_1 = 2\omega_0$$
$$\beta_2 = \omega_0^2 \tag{32}$$

The bandwidth $\boldsymbol{\omega}_0$ of the ESO determines the stability and dynamic performance of the observer. Given the digital implementation of the drive, the design of $\boldsymbol{\omega}_0$ is performed in the discrete-time domain $z$. Eq. (28) can be expressed in discrete-time form as follows:

$$\begin{aligned} e(k) &= \hat{i}_{A_1}(k) - i_{A_1}(k) \\ \hat{i}_{A_1}(k+1) &= \hat{i}_{A_1}(k) + T_s\left(\hat{F}_{A_1}(k) + \alpha v_{A_1}(k)\right) - \beta_{01} e(k) \\ \hat{F}_{A_1}(k+1) &= \hat{F}_{A_1}(k) - \beta_{02} e(k) \end{aligned} \tag{33}$$

where $\boldsymbol{\beta}_{02} = \boldsymbol{T_s\beta}_2$, $\boldsymbol{\beta}_{01} = \boldsymbol{T_s\beta}_1$ are the observer gains in the discrete-time domain, and $\boldsymbol{T_s}$ is the sampling time. The values of $\boldsymbol{\beta}_{01}$ and $\boldsymbol{\beta}_{02}$ affect the location of the closed-loop system poles and, consequently, the stability of the observer. Therefore, to ensure the system's stability and to achieve the desired control performance, the values of $\boldsymbol{\beta}_{01}$ and $\boldsymbol{\beta}_{02}$ should be selected appropriately. Based on Eq. (33), the transfer function of the ESO in the $z$ domain can be obtained as follows:

$$G(z) = \frac{\hat{i}_{A_1}(z)}{i_{A_1}(z)} = \frac{\beta_{01} z - \beta_{01} + \beta_{02} T_s}{(z-1)^2 + \beta_{01} z - \beta_{01} + \beta_{02} T_s} \tag{34}$$

whose characteristic equation is as follows:

$$(z-1)^2 + \beta_{01} z - \beta_{01} + \beta_{02} T_s = 0 \tag{35}$$

Considering $\boldsymbol{\beta}_{02} = \boldsymbol{\omega}_0^2 \boldsymbol{T_s}$ and $\boldsymbol{\beta}_{01} = 2\boldsymbol{\omega}_0 \boldsymbol{T_s}$, the poles of the transfer function will be obtained by solving Eq. (33) as follows:

$$z_{1,2} = 1 - \omega_0 T_s \tag{36}$$

where $\boldsymbol{\omega}_0$ can be calculated from the poles of the characteristic equation in the z-domain as follows:

$$\omega_0 = \frac{1 - z_{1,2}}{T_s} \tag{37}$$

$\omega_0$ is chosen such that the poles $z_{1,2}$ lie within the unit circle in the *z*-domain. Generally, if $\omega_0$ is too small ($\boldsymbol{z}_{1,2}$ close to 1), the dynamic performance of the observer will deteriorate. Conversely, if $\boldsymbol{\omega}_0$ is too large ($\boldsymbol{z}_{1,2}$ close to 0), the system's robustness is affected and may lead to divergence. For PMSM drives, the current control loop (inner loop) typically requires a high bandwidth to meet dynamic response requirements. Thus, in this paper, $\boldsymbol{z}_{1,2}$ is set to 8, and when the sampling frequency is 8 kHz, $\boldsymbol{\omega}_0$ is 1600. For other sampling frequencies, such as 16 kHz, $\boldsymbol{\omega}_0$ is chosen 3200.

**3.4. Current regulation based on ESO**

The relationship between voltage and current for each winding of the motor follows the ultra-local model by Eq. (26). Using the Euler discretization approximation, the voltage of motor winding $\boldsymbol{A}_1$ can be calculated as follows:

$$v_{A_1}(k) = \frac{i_{A_1}(k+1) - i_{A_1}(k)}{\alpha T_s} - \frac{\hat{F}_{A_1}(k)}{\alpha} \tag{38}$$

where $\boldsymbol{F}_{A_1}$ is replaced by $\hat{\boldsymbol{F}}_{A_1}$, which is obtained by the ESO. By substituting $\boldsymbol{i}_{A_1}(\boldsymbol{k}+1)$ with the reference current of the phase A windings, i.e., $\boldsymbol{i}_A^*$ determined by Eq. (16), the reference voltage value of the motor winding $\boldsymbol{A}_1$ is obtained as follows:

$$v_{A_1}^*(k) = \frac{i_A^*(k) - i_{A_1}(k)}{\alpha T_s} - \frac{\hat{F}_{A_1}(k)}{\alpha} \tag{39}$$

By modulating this voltage vector using fixed switching frequency modulation methods such as space vector (SVM) or sinusoidal modulation, the switching signals for each H-bridge inverter, as shown in Fig. 3 are determined.

### 3.5. Extended state observer (ESO) design for speed control

To determine the reference torque value $T_e^*$ in Eq. (15) and consequently the reference current of each phase in Eq. (16), speed control can be performed in the same way as current control without using a PI controller and based on another ESO observer. On this way, according to Eq. (4), the dynamic equation of the speed of PMSM is as follows:

$$\frac{d\omega_m}{dt} = \frac{1}{J}\left(T_e - B\omega_m - T_l\right) \tag{40}$$

Considering $\omega_m$ as the output and $T_e$ as the input of the model, an ultra-local model for speed control can be rewritten as same as Eq. (18) and Eq. (19) as follows:

$$\frac{d\omega_m}{dt} = \alpha_m T_e + F_m \tag{41}$$

$$F_m = \frac{1}{J}\left(-B\omega_m - T_l\right). \tag{42}$$

where $F_m$. shows the total mechanical disturbances of the system and $\alpha_m = 1/J$ is the input gain. Based on the ultra-local model with Eq. (41), it is possible to repeat the process similar to Eq. (28) to Eq. (37) to construct a linear ESO to estimate $\hat{F}_m$. by considering $z_1 = \omega_m$ and $z_2 = \hat{F}_m$ as state variables and with velocity error feedback $e_\omega$ as follows:

$$e_\omega = z_1 - \omega_m$$

$$\dot{z}_1 = z_2 + \alpha_m T_e - \beta_{1m} e_\omega \tag{43}$$

$$\dot{z}_2 = -\beta_{2m} e_\omega$$

Using Euler's approximation, Eq. (41) can be rewritten as follows:

$$\frac{\omega_m(k+1) - \omega_m(k)}{T_s} = \alpha_m T_e(k) + \hat{F}_m(k) \tag{44}$$

Considering $\omega_m(k+1) = \omega_m^*$, the reference value of torque $T_e$ can be calculated as follows:

$$T_e^* = T_e(k+1) = \frac{\omega_m^* - \omega_m(k)}{\alpha_m T_s} - \frac{\hat{F}_m(k)}{\alpha_m}. \tag{45}$$

Fig. 4 illustrates the block diagram of the six-phase PMSM control system in this paper using the MFPCSC strategy based on ESO current observers. It is obvious that the modeling and current control of each phase is implemented using independent local microcontrollers using corresponding phase current feedback. Actually, the QPR current controllers commonly used in similar studies have been replaced with the proposed MFPCSC controllers in this paper. Also, in central microcontroller, an ESO speed observer determines the reference current for each phase based on information of speed and rotor position of the motor.

## 4. Simulation Results

In this section, for a double-winding six-phase 8-pole PMSM with nominal specifications of 50 kW, 245 V, 220 rpm, 50 mΩ per phase, 2320 μH phase self- inductance, 1.34 Wb flux, and 1.37 rad/s back EMF constant, the drive behavior is examined using the current shaping (or harmonic current injection strategy) based on the speed and current ESO observers. The results are compared with other current regulation methods, including hysteresis controllers, PI, and QPR controllers. The load torque is of the propeller type and proportional to the square of the speed, reaching 2000 N·m at nominal speed. Fig. 5 shows the drive behavior when using PI current controllers. The $K_P$ and $K_I$ coefficients of the PI controllers for all 12 windings are optimally tuned to 10 and 50, respectively. According to the Fig. 4, current tracking due to the low bandwidth of this controller is inadequate during rapid reference current changes at its peaks and creates a torque ripple of about 400 N·m, which is not suitable for the intended application.

Fig. 6 shows the drive behavior when using hysteresis current controllers. Due to the variable switching frequency caused by this controller, the hysteresis band is chosen such that the maximum switching frequency does not exceed 16 kHz. For this purpose, the hysteresis band should not be less than 4 amps. It can be observed that the current tracking response of this controller is better than that of the PI controller but produces a torque ripple of 200 N·m.

Fig. 7 shows the drive behavior when using quasi-proportional resonant (QPR) controllers for current regulation. The parameters for all winding controllers are the same and, according to (17), are $K_P = 15$, $K_{R1} = 15$, $K_{R3} = 10$, $K_{R5} = 10$, and $\omega_{ci} = 20$. The torque ripple produced by the motor has significantly decreased compared to the current and hysteresis controllers, reaching about 65 N·m.

Fig. 8 shows the drive response when using the model-free predictive current and speed control (MFPCSC) method based on the ESO observers according to Fig. 4. The torque ripple has further decreased compared to the QPR controllers, reaching about 40 N·m. The observer bandwidth $\omega_o$ is chosen to 1000 rad/sec.

Fig. 9 displays the estimation of state variables $F_A$, $F_B$, and $F_C$, each representing the total known and unknown disturbances of phase windings A, B, and C, respectively.

Fig. 10 shows the total disturbance of rotor dynamic and its estimated value. The disturbance estimation is performed accurately.

In Table 1, the performance of four current controllers in reducing the torque ripple of the six-phase non-sinusoidal PMSM is compared, that indicating the superior performance of the current regulation based on ultra-local model and ESOs.

## 5. Experimental Results

To validate the performance of the proposed MFPCSC strategy, some practical tests were conducted on the laboratory setup. A schematic overview of the experimental setup is presented in Fig. 11, where seven microcontrollers type STM32F407VGT6 have been used as central and local controllers. The drive of each phase consists of a microcontroller board as well as double H-bridge single-phase inverters and a separate excited DC generator is employed as load.

Fig. 12 shows the back EMF voltage of each winding of phase A the injected current. Due to the non-sinusoidal back EMF voltage waveform of the motor, the injected current to the motor, is harmonic and is well tracked by the model-free predictive current controller based on the ESO.

Fig. 13 shows three-phase currents of the six-phase motor and Fig. 14 shows the speed response of the motor using the proposed control method, where the reference speed tracking is satisfactorily performed.

## 6. Conclusion

In this paper, a new method for reducing the torque ripple of a six-phase non-sinusoidal PMSM using the model-free predictive current control (MFPCSC) method based on the current and speed extended state observers in the stationary reference frame were presented. Due to modular structure of control system of the drive, motor modeling was performed in the six-axis stationary reference frame, and the current control of each phase was also performed independently in this reference frame. The harmonic current injection method was used to eliminate the torque ripple caused by the non-sinusoidal back EMF of the phases. In this method, it is assumed that the harmonic content of the back EMF voltages should be known and, in this research, their harmonic contents have been considered constant in different speed and load conditions. To adjust the phase currents around the corresponding harmonic reference values, the MFPCSC method was introduced based on an ultra-local model in the stationary reference frame. All modeled and unmodeled system uncertainties were estimated using the extended state observers, and these estimations were used in determining the reference voltages. The simulation results, as well as laboratory tests, indicate the effectiveness of the proposed control method and its superiority over other conventional control methods such as PI, hysteresis, and QPR controllers. Further research can be carried out to estimate the phases' back-EMF voltages to update the coefficient matrix in Eq. (12).

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## APPENDIX: QPR CONTROLLER ADJUSTMENT

The trial-and-error method is an approach involving the manual adjustment of gains and controller parameters while observing the motor's response to different input signals. To determine the optimal values, it is recommended to follow the tuning sequence as outlined below:

1.**Identifying key system performance criteria**: Before tuning the controller, it is essential to identify the key performance criteria you aim to optimize. For example, you may want to minimize steady-state error, maximize tracking accuracy, or reduce settling time.

2. **Selecting initial values:** Choose initial values for the parameters $K_P$, $K_R$ and $\omega_c$. based on prior experience or knowledge of the SM motor drive system. These initial values serve as the starting point for the tuning process.

3. **Tuning the proportional gain**: First, adjust the proportional gain $K_P$ to achieve the desired steady-state performance. This can be done using methods such as trial and error or systematic tuning approaches like the Ziegler-Nichols method.

4. **Tuning the resonance gain**: After setting the proportional gain, adjust the resonance gain $K_{R_h}$ to achieve optimal tracking performance. This can also be performed using trial and error or systematic tuning methods such as the Internal Model Control (IMC) approach.

5. **Tuning the cut-off frequency:** Finally, adjust the cut-off frequency $\omega_{c_h}$ to achieve the desired compromise between dynamic response and stability. This can be done by observing the system's response to a step change in the reference current and tuning $\omega_{c_h}$ to achieve the desired settling time and overshoot.

6. **Validating the tuned parameters**: After tuning the parameters, evaluate the controller's performance under various operational conditions and disturbances to ensure it meets the desired performance criteria.

**Figure and Table captions:**

Fig. 1. Two common types of asymmetrical six-phase PMSMs (a) Double-star type, (b) Open-end type

Fig. 2. Schematic of asymmetric six-phase PMSM motor with double stator windings

Fig. 3. Schematic of the six-phase PMSM drive system with double stator windings used in this research

Fig. 4. Block diagram MFPCSC method for the six-phase with double windings using the speed and current extended space observers

Fig. 5. Drive behavior using PI current controller at nominal speed and load torque

Fig. 6. Drive behavior using hysteresis current controllers at nominal speed and load torque

Fig. 7. drive behavior using QPR controllers at nominal speed and nominal load torque

Fig. 8. Drive behavior using state feedback based on ESO at nominal speed and nominal load torque

Fig. 9. Estimation of total disturbances in each motor phase by an ESO of current in a local microcontroller

Fig. 10. Estimation of total disturbances in each motor phase by the ESO of speed in the central microcontroller

Fig. 11. Experimental setup of modular six-phase PMSM drive

Fig. 12. Waveforms of back emf voltage and single-phase current

Fig. 13. Currents of three windings from the twelve windings of the motor

Fig. 14. Motor speed variations using the presented MFPCSC method

(Orange colour: speed; Green: rotor position; Red and Blue: current of windings A1 and A2 of phase A)

Table 1. Comparative analysis of torque ripple using different methods

## Biographies

Davood Maleki, was born in 1980 in Malayer, Iran. He received a Bachelor's degree in Electronics and a Master's degree in Electrical Machines from Malek Ashtar Industrial University in Shahin Shahr, Isfahan. His Bachelor's thesis was on the design of a buck-boost regulator for a 400 to 24 converter, and His Master's thesis focused on the construction of an LSPMSM drive. He is currently a PhD student at University of Kashan. His interests are control and implementation of multi-phase PMSM drives.

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