Low-Switching-Loss Finite Control Set Model Predictive Current Control for IMs Considering Rotor-Related Inductance Mismatch

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ABSTRACT
This paper develops a high-performance finite control set model predictive current control (FCS-MPCC) method for induction motors (IM) to ensure the system control performance over the low-switching-frequency (LSF) range, where the switching loss is low. The new controller is based on tripartite calculations in each control period and sliding mode (SM) rotor-related inductance observer. In detail, firstly, to solve the problem that the control performance of the traditional FCS-MPCC methods is low when they are used at low frequencies, tripartite calculations are employed in each control period to improve the prediction accuracy. By dividing one control period into three equal parts and executing the prediction algorithm in each subsection, the current estimation errors become lower compared to the conventional single-step Euler-discretization controller. Then, considering that the performance of an FCS-MPCC controller highly relies on the machine parameter values, but it is hard to directly obtain the accurate rotor-related inductances (mutual inductance and rotor inductance), this paper uses an online parameter observer based on the sliding mode (SM) principle to diagnose them in real time. It needs to be mentioned that when discussing the stability of the sigmoid-function-based SM observer, a new technique called estimation-error-limitation is initially adopted in this paper to simplify the analysis process. Finally, the proposed algorithms are verified by simulation, which is conducted on a three-phase IM at LSF situations.

INDEX TERMS
Induction motor, finite control set model predictive control, sliding mode observer, parameter mismatch, low switching frequency.

I. INTRODUCTION
Apart from the permanent magnet machines (PMM), induction motor (IM) is one of the most commonly used propulsion machines in the electrical propulsion applications [1]–[3]. Comparatively speaking, although IM has the advantages of low cost, robust structure, high reliability and demagnetization-free risks, the energy conversion efficiency of an IM-based drive system is lower than a PMM-based system [4]. It deserves to be mentioned that this problem represents that much more loss would be produced during operations, lowering the energy efficiency. Consequently, effective measures to improve the efficiency of the IM-based drive systems are highly required [5], [6].

In a modern electric drive, the inverters that are able to transfer the direct current (DC) voltage to the alternating current (AC) one are usually adopted to power the propulsion machine [7]. However, the turn-on and turn-off actions of the power electronics must consume extra energy, and this is the so-called switching loss. Now, it is widely acknowledged that the switching loss is a crucial part of the total loss in a drive. The traditional switching-loss attenuation methods can be summarized into two groups: direct technique and indirect technique. In terms of the direct method, because the switching loss of power electronics is proportionally related to the switching frequency, it will decrease as long as the switching frequency is set at a low level when executing the control algorithms [8]. As for the indirect method, paper [9] and [10] intend to use soft-switching technique to avoid switching loss completely. By the use of a snubber circuit
which consists of several passive components, the occurrence of simultaneously high values of voltage and current (power dissipation) can be prevented during the switching process. Comparatively, the former strategy is much easier to implement, while the system control performance might be sacrificed. Specifically, the current and output torque ripples will rise in practice. Although the latter approach is able to suppress the switching loss without significantly influencing the control performance, extra external circuits need to be designed, sacrificing the reliability and cost of the whole system. From the perspective of economy, like [8], one main purpose of this paper is to lower the switching loss by developing efficient control algorithms whose control performance can be improved compared to the traditional methods when they are working in the low-switching-frequency (LSF) situations.

Finite control set (FCS) model predictive control (MPC), which is endowed with remarkable simplicity and strong constraint capability, has been widely used in different electric powertrains [11]–[14]. Without adopting any modulators for pulse width modulation (PWM) signal generation, FCS-MPC directly uses the output of the controller to determine the optimal switching state [15]. In comparison with the traditional modulator-based field-oriented control (FOC) methods whose switching frequency is fixed and equals the control frequency, the switching frequency of an FCS-MPC method is uncertain and twice lower than the control frequency. For instance, when the control frequency is set as 10 kHz, the switching frequencies for the traditional FOC and FCS-MPC are 10 kHz and less than 5 kHz, respectively. In this regard, FCS-MPC is a natural low-switching-loss (LSL) method, and moreover, as is illustrated in [8], the switching loss for a predictive method can be further reduced once the control frequency becomes lower. However, there exist severe problems if the traditional FCS-MPC strategies are directly applied to the LSF conditions. In detail, most of the FCS-MPC approaches employ the forward Euler discretization method to discretize the machine model in a time step of control period $T_s$, and in each control period, the FCS-MPC implementations are only executed once [16], [17]. In this case, the machine states are assumed to shift linearly within $T_s$. When $T_s$ is short, the discrete predicting plant model (DPPM) obtained according to this assumption is relatively precise. But in the LSF situations ($T_s$ is long), large deviations between the estimated results using the DPPM and the real states are inclined to occur considering that the states in the machine are nonlinear, degrading the control accuracy inevitably. On this ground it is of great significance to adjust or optimize the conventional DPPM for the LSF situations.

Another typical problem of an FCS-MPC controller is that because the DPPM is parameter-dependent, the system control performance highly relies on the accurate machine parameters (resistance and inductance, etc.) [18], [19]. Traditionally, the machine parameter values used for prediction are obtained by measurement and they are usually provided by the suppliers. In terms of the armature parameters, such as stator winding resistance and inductance, they can be directly measured by meters. Whereas, as for the electrical parameters (rotor resistance, inductance and mutual inductance) related to the rotor of IMs, they cannot be directly measured and usually, they are detected by using the signal-injection method in industry [20]. However, those rotor-related parameter values detected in the particular environment or conditions (e.g., standstill cases) do not always represent the real-time values during operations. For instance, the mutual and rotor self-inductances might change when the magnetic saturation phenomenon occurs and when the relative location between the rotor and stator flux is different [21]. Unluckily, once the parameter values are not consistent with the real ones, there will be static errors during control, and even worse, the system might become unstable [22]. Aiming at this problem, three up-to-date categories of remedy measures can be employed. Firstly, online parameter identification strategies can be incorporated into the machine control process to avoid the parameter mismatch phenomenon [23]–[25]. By observing the real-time machine parameters and then substituting them into the DPPM, the predictive control algorithms will be endowed with strong robustness. Secondly, papers [26]–[28] have constructed perturbation observers to diagnose and compensate the disturbances caused by parameter uncertainties. Thirdly, the advance adaptive control strategies can be used to adjust the system parameters in real time, eliminating the impacts of parameter mismatch [29], [30]. Comparatively, the first strategy is simpler and more intuitive to implement, so it deserves further investigations.

One main purpose of this paper is to propose a novel FCS model predictive current control (MPCC) method that employs tripartite calculations in each control period for the IM drives working in the LSF situations. Firstly, after establishing the current model of the machine, the reasons why the accuracy of the traditional FCS-MPC method is low in the LSF scenarios are analyzed intuitively. Then, by dividing one control period into three equal parts and elaborating the DPPM to make it suitable for three-time predictions, the accuracy of the prediction model is improved. Finally, the implementation procedures of the new FCS-MPCC strategy are designed at length. It deserves to be mentioned that although the prediction process of the new method requires to estimate the currents three times within $T_s$, the actuation frequency does not change. Hence, the switching frequency and the switching loss of the proposed FCS-MPCC method are still low, especially in the LSF conditions. But compared to the traditional FCS-MPC method, the control performance can witness an obvious improvement. In summary, the novelty of the proposed FCS-MPCC method reflects in employing tripartite calculations in each control period for the sake of higher prediction accuracy in the LSF situations. In addition, in order to ensure that the rotor electrical inductance values used for prediction are accurate, this paper initially designs a sigmoid-function (SF) based sliding mode observer
(SMO) to identify the parameters. It needs to be addressed that because SMOs were mainly used for disturbance estimation previously [31] and there are few studies using them for electrical parameter identification in the IM applications, the research in this paper is valuable. Additionally, a Lyapunov function is constructed to analyze the stability of the sliding mode observer. However, differing from the traditional signum-function based SMOs [32], the stability analysis procedures of the new observers are more complicated. On this ground when determining the explicit conditions that make the system stable, a new concept of estimation-error-limitation is proposed and adopted in this paper.

The structure of the rest paper is as follows. Section 2 introduces an IM modelling method and the traditional Euler-discretization-based DPPM. In Section 3, firstly, the reasons why the performance of the traditional FCS-MPCC method is low considering the influence of LSF is intuitively explained. Then, the new FCS-MPCC algorithms based on triple calculation techniques are illustrated. Section 4 presents the proposed sliding mode inductance observer, and simultaneously, the observer stability is analyzed. Section 5 discusses the simulation results of the proposed SMO-based FCS-MPCC algorithms, and Section 6 is the conclusion part.

II. IM MODELLING AND TRADITIONAL DPPM
As for an FCS-MPCC strategy, the targeting control variables are the $M$, $T$-axis armature currents. Theoretically, only the electrical properties concerning the stator are needed when constructing the predicting model. However, due to the strong coupling effects, the prediction model for an IM must include the rotor characteristics. In the rotating reference frame, the continuous IM model can be described as:

$$\begin{align*}
\frac{d\psi_{sm}}{dt} &= -\frac{R_s}{L_s} i_{sm} + \omega_m i_{sT} - \frac{L_m}{L_s} \frac{d\psi_{sm}}{dt} + \frac{u_{sm}}{L_s} - \frac{\Delta\omega}{\omega_m} i_{sT} \\
\frac{d\psi_{sT}}{dt} &= -\frac{R_s}{L_s} i_{sT} - \frac{L_m}{L_s} \frac{d\psi_{sm}}{dt} + \frac{u_{sT}}{L_s} - \frac{\Delta\omega}{\omega_m} i_{sm} \\
\frac{d\psi_{sm}}{dt} &= -\frac{L_m}{L_r} \frac{d\psi_{sT}}{dt} + \frac{R_r}{L_r} i_{sm} + \frac{\Delta\omega}{\omega_m} i_{sT}
\end{align*}$$

(1)

(2)

where (1) represents the stator properties and (2) is the rotor properties. $i_{sm}$ and $i_{sT}$ are stator currents. $u_{sm}$ and $u_{sT}$ mean stator voltage. $\omega_m$ and $\Delta\omega$ represent the machine synchronous speed and slip speed, respectively. $i_{sm}$ and $i_{sT}$ are the rotor currents. $R_r$ and $L_r$ are the armature resistance and inductance, respectively. $R_s$ and $L_s$ are the rotor resistance and inductance. $L_m$ is the mutual inductance. In practice, because the rotor currents cannot be measured, they should be eliminated when predicting the future states. Firstly, substitute (2) into (1), and then it can be obtained that:

$$\begin{align*}
\frac{di_{sm}}{dt} &= \frac{-L_r R_s}{L_s} i_{sm} + \frac{L_m}{C} \omega_m i_{sT} + \frac{L_m^2}{C} \Delta\omega i_{sT} \\
&+ \frac{L_m R_s}{L_s} i_{sm} + \frac{L_m}{C} \omega_m i_{sT} + \frac{L_r}{C} u_{sm} \\
\frac{di_{sT}}{dt} &= \frac{-L_m}{C} \Delta\omega i_{sm} - \frac{L_r}{C} \omega_m i_{sT} + \frac{L_r R_s}{C} i_{sm} \\
&- \frac{L_m}{C} \omega_m i_{sT} + \frac{L_m R_r}{C} i_{sT} + \frac{L_r}{C} u_{sT}
\end{align*}$$

(3)

where $C$ is a constant and $C = L_r L_s - L_m^2$, $\omega_r$ is the measured rotor speed. Secondly, considering that the IM model in the rotating frame is based on the rotor-flux-oriented principle, the rotor flux $\psi_r$ also satisfies the following conditions:

$$\begin{align*}
\psi_{sm} &= L_r i_{sm} + L_m i_{sT} = \frac{L_m i_{sm}}{T_r s + \frac{1}{T_r}} \\
T_r i_{sT} + L_m i_{sT} &= 0
\end{align*}$$

(4)

(5)

Substitute (5) and (6) into (3), the IM model without containing rotor currents can be described as (7), which is shown at the bottom of the next page.

The derivatives of the $M$, $T$-axis currents indicate that the armature currents will shift in a nonlinear trend and the change velocity is related to not only the parameters but also the real-time current and speed states. However, in order to obtain a DPPM, traditionally, the forward Euler discretization algorithm which assumes that the currents change in a linear trend within each control period $T_r$ should be applied to (3), and the DPPM at $t_k$ can be expressed as:

$$\begin{align*}
i_{sm}(k+1) &= \frac{C - T_s L_s R_s}{C} i_{sm}(k) + \frac{T_s \omega_m}{C} i_{sT}(k) \\
&+ \frac{T_s L_r}{C} u_{sm}(k) \\
i_{sT}(k+1) &= \frac{T_s L_m^2}{C} \Delta\omega(k) i_{sm}(k) - \frac{T_s L_s R_s}{C} \omega_m(k) i_{sT}(k) \\
&- 1 \frac{T_s L_s R_s}{C} i_{sm}(k) - \frac{T_s L_s R_s}{C} u_{sT}(k)
\end{align*}$$

(8)

where $i_{sm}(k)$, $i_{sT}(k)$, $\omega_m(k)$ and $\Delta\omega(k)$ are the states at the $k$th sampling instant. $i_{sm}(k+1)$ and $i_{sT}(k+1)$ are the predicting values at the $(k+1)$th period. In terms of $u_{sm}(k)$ and $u_{sT}(k)$, they are the candidate control voltages. As for a two-level inverter, a total of seven phase voltage vectors, which can be denoted as $u_{000}$, $u_{010}$, $u_{011}$, $u_{100}$, $u_{101}$, $u_{001}$ and $u_{101}$, constitute the control set:

$$\begin{align*}
u_{a} &= \frac{u_a + u_b}{3} \\
u_{b} &= \frac{u_a - u_b}{3} \\
u_{c} &= \frac{-u_a + u_b}{3} + s_a \\
u_{d} &= \frac{u_a + u_b}{3} + s_b \\
u_{e} &= \frac{u_a - u_b}{3} + s_c
\end{align*}$$

(9)
where \([s_a, s_b, s_c]^T\) includes \([0, 0, 0]^T, [1, 0, 0]^T, [1, 1, 0]^T, [0, 1, 0]^T, [0, 1, 1]^T, [0, 0, 1]^T, [1, 0, 1]^T, [1, 1, 1]^T\), and they are the switching states. \(U_{dc}\) is the DC source voltage. \([u_a, u_b, u_c]^T\) are the terminal phase voltages. By using \(abc/MT\) transformation, the control voltages to be used for prediction can be described as (10), shown at the bottom of this page, where \(\theta\) is the rotor flux position.

So far, the DPPM used for FCS-MPCC is established. However, there are still three variable states that cannot be measured directly, that is, the synchronous speed, slip speed and rotor flux position. In practice, at \(t_k\), they can be calculated according to the following equations [33]:

\[
\begin{align*}
\Delta \omega(k) &= \frac{L_m}{T_e \psi_r + 1} i_{iT}(k) \\
\omega_e(k) &= p \omega_r(k) + \Delta \omega(k) \\
\theta(k) &= \theta(k - 1) + \int_{t_k}^{t_{k+1}} \omega_e \, dt 
\end{align*}
\] (11)

where \(p\) is the number of pole pairs and \(\theta(0) = 0\).

### III. PROPOSED FCS-MPCC SUITED TO LSF SITUATIONS

Firstly, this section will intuitively analyze the reasons why the prediction accuracy of the traditional FCS-MPCC is low in the LSF situations by using the physical descriptions. Then, a novel FCS-MPCC method based on tripartite predictions is proposed for IM drives to reduce the side effects caused by LSF. The purpose of the new method is to narrow the current prediction errors so as to improve the control performance.

#### A. PHYSICAL EXPLANATIONS OF TRADITIONAL FCS-MPCC DEFECTS IN LSF SITUATIONS

When using the Euler-discretization-based DPPM to compute the future currents, the prediction process (taking \(i_{iM}\) as an example) can be depicted in Figure 1 when a candidate voltage vector is applied, where \(i_{sM}, i_{iM},\) and \(i_{MP}\) represent the reference, real and predicted \(M\)-axis currents, respectively. Denote the derivative of the \(M\)-axis current in (7) as:

\[
f(t) = -\frac{L_m R_s}{C} i_{sM}(t) + \omega_e i_{iT}(t) + \frac{L_r}{C} u_{sM}(t) \] (12)

Between \(t_k\) and \(t_{k+1}\), the physical descriptions of the traditional FCS-MPCC implementation (as in Fig. 1) can be explained as follows: At first, the measured and observed states at \(t_k\) together with the candidate control voltages are substituted into (12) to calculate the slope (\(sl\)) of the tangent line of the current wave, namely, \(sl = f(t_k)\). Secondly, \(f(t_k)\) is treated as the slope of the following straight line:

\[
i_{iM}(t) = sl \cdot t + i_{iM}(k) \] (13)

where \(t\) is time. Once \(T_s\) is substituted into (13), the future state at \(t_{k+1}\) can be calculated. Obviously, the current is assumed to shift linearly within \(T_s\). However, because the currents in IM change constantly, \(f(t)\) will not always remain at \(sl\) during each control period, leading to the fact that the real \(M\)-axis current is nonlinear. Undoubtedly, the estimation errors between \(i_{sM}(k+1)\) and \(i_{iM}(k+1)\) cannot reach zero at \(t_{k+1}\), and the longer the control period is, the larger the prediction error \(\Delta i_{sl}(k+1)\) becomes. Consequently, the estimation accuracy of the traditional Euler-discretization-based FCS-MPCC is low in the LSF conditions, degrading the control performance inevitably.

#### B. PROPOSED FCS-MPCC CONTROLLER

In terms of a new FCS-MPCC controller, the components that need to be designed include 1) DPPM, 2) input and output of the controller, 3) implementation procedures. This part will explain the detailed design process of the proposed FCS-MPCC method according to these aspects.

1) NEW DPPM FOR LSF APPLICATIONS

The reasons why the prediction accuracy of the traditional predictive method is low when the control period is long can be summarized as follows. Firstly, the Euler discretization implementation is linear, but it does not comply with the real cases; Secondly, the one-step prediction process aggravates the estimation results. In reference to the fact that that the prediction accuracy is relatively high in the high switching
3) IMPLEMENTATION PROCEDURES

The implementation procedures (see Figure 4) of the proposed FCS-MPCC strategy between \( t_k \) and \( t_{k+1} \) can be summarized as seven steps:

Step 1) Measurement and transformation: Use the current sensors and position sensor to measure the phase currents \( i_a(k), i_b(k), i_c(k) \) and rotor speed \( \omega_r(k) \). Meanwhile, use the previous flux position \( \theta(k) \) to transform the phase currents to the reference \( M, T \)-axis currents \( i_{SM}(k) \) and \( i_{ST}(k) \).

Step 2) Synchronous speed and position observation: Use the measured states and (14) to observe the slip speed \( \Delta \omega(k) \) and synchronous speed \( \omega_r(k) \).

Step 3) First-time prediction: Substitute the measured and observed states together with the candidate control voltages into the new DPPM to estimate the future \( M, T \)-axis currents \( i_{SM}(k+1) \) and \( i_{ST}(k+1) \) at \( t_k + T_s/3 \).

Step 4) Second-time prediction: Measure the real-time rotor speed \( \omega_r(k+1/3) \) and use the estimated currents to calculate the slip speed \( \Delta \omega(k+1/3) \) and synchronous speed \( \omega_r(k+2/3) \).

Step 5) Third-time prediction: Measure \( \omega_r(k+2/3) \) and observe \( \Delta \omega(k+2/3) \). Calculate the candidate \( i_{SM}(k+2/3) \) and \( i_{ST}(k+2/3) \).

Step 6) Evaluate cost function: Select optimal voltage,\( \dot{s} \).

Step 7) Apply optimal switching state: Select the optimal switching state and turn it on. The process goes to the next period.

2) INPUT AND OUTPUT OF CONTROLLER

In order to predict the future states of the currents using (14), as is shown in Figure 3, the input of the controller must contain the present currents \( i_{SM}(k) \) and \( i_{ST}(k) \), control voltages, synchronous speed \( \omega_r(k) \), slip speed \( \Delta \omega(k) \) and rotor speed \( \omega_r(k) \). Besides, because the real-time flux position needs to be used to calculate the \( M, T \)-axis control voltages, \( \theta(k) \) should be the input of the controller definitely. Finally, considering that an FCS-MPCC controller is used to directly evaluate the current tracking characteristics of the system, the reference \( M, T \)-axis currents \( \dot{i}_{ST} \) and \( \dot{i}_{SM} \) should be set as the input of the controller.

In terms of the output of the controller, because the function of an FCS-MPCC controller is to select the optimal switching states to be applied to the machine, the selected switching states are treated as the output of the controller (see Figure 3).
to predict the future $M$, $T$-axis currents $i_{sM}(k + 2/3)$ and $i_{sT}(k + 2/3)$ at $t_k+2T/3$.

Step 5) Third-time prediction: Repeat step 4) to calculate the slip speed $\Delta \omega(k + 2/3)$, synchronous speed $\omega_s(k + 2/3)$, rotor flux angle $\theta(k+2/3)$ and $\theta(k+1)$, $i_{sM}(k + 1)$ and $i_{sT}(k + 1)$.

Step 6) Evaluation: Substitute the seven the predicted values into the cost function (15) and determine the optimal voltage vector and the corresponding switching states.

$$J = |i_{sM}^* - i_{sM}(k+1)| + |i_{sT}^* - i_{sT}(k+1)|$$  \hspace{1cm} (15)

Step 7) Actuation: Apply the optimum switching state to the drive system.

Overall, the implementation process of the proposed FCS-MPCC method indicates that the predictive calculations are executed three times while the evaluation and actuation operations are only executed once in each control period. Hence, the switching frequency is no higher than that for the traditional method and the switching loss is still very low in the LSF conditions. But the prediction accuracy and the system control performance are expected to increase compared to the previous approaches.

IV. SM ROTOR-RELATED INDUCTANCE OBSERVER-BASED FCS-MPCC
SMOs are constructed depending on the variable structure control principle, and the main objective of a SMO is to select an appropriate switching function and a gain to force the system to "slide" along a cross-section of the system's normal behavior. Because SMOs usually have strong robustness against external disturbances and parameter mismatch, they have been widely used in the electric applications. For example, sliding mode position observer and sliding mode observer. However, it can be seen that SMOs usually have strong robustness against external disturbances and parameter mismatch, they have been widely used in the electric applications. For example, sliding mode position observer and sliding mode observer. However, they are seldom used in IM electrical parameter identification cases. In addition, considering the chattering attenuation problem, this section develops a SF-based SMO to estimate the real-time rotor-related electrical inductances. Further, the stability of the observers is analyzed. Finally, the topology of proposed SMO-based FCS-MPCC method is given.

A. SM INDUCTION OBSERVER FOR IM
1) OBSERVER DESIGN
As is shown in the machine model (7), the rotor self-inductance and mutual inductance are contained in the $M$, $T$-axis current equations. Theoretically, the $M$, $T$-axis current equations are suitable for constructing the rotor resistance observer. However, it can be seen that $L_r$ and $L_m$ are included in each term of the $T$-axis model, leading to the fact that the constructed SMO based on the equation is very complicated. Therefore, for the sake of simplicity, the IM mechanical property (speed equation) which also contains the mutual inductance and self-inductance information is employed to construct the SMO in this paper. As in [33], the mechanical model of an IM is as follows:

$$\frac{d\omega_r}{dt} = \frac{1}{J}(T_r - B\omega_r - T_I) = \frac{1}{J}(\frac{PL_m}{L_r}\psi_r + i_{sT} - B\omega_r - T_I)$$  \hspace{1cm} (16)

where $B$ is the friction coefficient, $T_r$ and $T_I$ are the electromagnetic and load torque, respectively. $J$ is the rotor inertia. Combined with the $M$-axis current equation, Eq. (16), which is much simpler than the $T$-axis current equation, can be used for constructing inductance observer. Whereas, there are still two challenges influencing the rotor-related inductance observation process. Firstly, $L_r$ and $L_m$ are not separated. They appear interactively in the form of $L_r/C$ and $L_m/L_r$ in the current and speed equations, respectively. In this case, this paper will identify $L_r/C$ and $L_m/L_r$ ($L_r$ and $L_m$ are not the targets of the SMO) at first, and they are further used to calculate the $L_r$ and $L_m$. Secondly, $L_r/C$ is the gain of not only the voltage $u_{sM}$ but also the $M$-axis current in the current characteristic equation, resulting in that it is difficult to determine the reaching law (sliding surface) when taking the system stability into account. In order to solve this problem, we use the offline-tested rotor self-inductance $L_{r,off}$ and mutual inductance $L_{m,off}$ to reconstruct the machine model

$$\begin{align*}
\Delta \omega(k + \frac{N}{3}) &= \frac{L_m}{T_r\psi_r + \frac{1}{3}}i_{sT}(k + \frac{N}{3}) \\
\omega_r(k + \frac{N}{3}) &= \frac{p\omega_r(k + \frac{N}{3}) + \Delta \omega(k + \frac{N}{3})}{3C - T_sL_sr}\frac{3}{3}i_{sM}(k + \frac{N}{3}) \\
i_{sM}(k + \frac{N}{3}) &= \frac{T_s\omega_r(k + \frac{N}{3})}{3C - T_sL_sr}\frac{3}{3}i_{sT}(k + \frac{N}{3}) + \frac{T_sL_sr}{3C}u_{sM}(k + \frac{N}{3}) \\
i_{sT}(k + \frac{N}{3}) &= \frac{T_sL_sr^2}{3C}B\omega_r(k + \frac{N}{3})i_{sM}(k + \frac{N}{3}) \\
&- \frac{T_sL_sr}{3C}\omega_r(k + \frac{N}{3})i_{sM}(k + \frac{N}{3}) \\
&+ T_sL_sr + \frac{T_sL_sr}{3C}u_{sT}(k + \frac{N}{3}) - \frac{T_sL_sr}{3C}u_{sM}(k + \frac{N}{3}) \\
&+ \frac{T_sL_sr}{3C}u_{sT}(k + \frac{N}{3}) - (1 - \frac{T_sL_sr}{3C} + \frac{T_sL_sr}{3C})i_{sT}(k + \frac{N}{3})
\end{align*}$$  \hspace{1cm} (14)
Then, the estimated rotor self-inductance is
\[
\frac{di_{SM}}{dt} = \frac{L_{r,off} R_s}{C_{off}} i_{SM} + \omega_e i_{ST} + L_r \frac{d\omega_e}{dt}
\]
(17)
where \(C_{off} = L_1 L_{r,off} - L_{m,off}^2\). It needs to be mentioned in (16) and (17), the resistance and stator inductance parameter values are the offline tested values.

Based on the machine model and according to the theory of sliding mode control, the SMO is described as follows in (16) and (17), the resistance and stator inductance parameters are used for constructing the SMO as follows:
\[
\frac{d\omega_e}{dt} = \frac{1}{J}(pk_1 F(\omega_r) \psi_i i_{ST} - B \omega_e - T_i)
\]
\[
\frac{di_{SM}}{dt} = -\frac{L_{r,off} R_s}{C_{off}} i_{SM} + \omega_e i_{ST} + k_2 F(i_{SM}) u_{SM}
\]
(18)
where \(i_{SM}^*\) and \(\omega_r^*\) are the estimated stator current and estimated rotor speed, respectively. \(i_{SM}\) and \(\omega_r\) represent the errors between the observed and the real states, and \(i_{SM}^* = i_r^* - i_{SM}\), \(\omega_r^* = \omega_r + \omega_e^* - \omega_r\). \(k_1\) and \(k_2\) are the gain factors which need to be tuned according to the stability criterion. \(F(\omega_r)\) and \(F(i_{SM})\) are the sigmoid functions, that is,
\[
\begin{bmatrix}
F(\omega_r) \\
F(i_{SM})
\end{bmatrix}
= 
\begin{bmatrix}
\frac{2}{1 + e^{-a \omega_r}} - 1 \\
\frac{2}{1 + e^{-ai_{SM}}} - 1
\end{bmatrix}
\]
(19)
where \(a\) is a constant concerning the boundary layer \((0 < a < 1)\), and it can be adjusted in practice. By carefully looking at the SF, it is completely different from the traditional signum function which has sharp reaching mode when the observer approaches the sliding surface. Therefore, the chattering effects can be suppressed when using this kind of switching function.

Denote \(a = K_1 F(\omega_r)\) and \(b = K_2 F(i_{SM})\), when the system reaches a stable state, the estimated values satisfy the following equations:
\[
\begin{bmatrix}
L_m \\
L_r \\
C
\end{bmatrix}
= 
\begin{bmatrix}
k_1 F(\omega_r) \\
k_2 F(i_{SM})
\end{bmatrix}
= 
\begin{bmatrix}
a \\
b
\end{bmatrix}
\]
(20)
Then, the estimated rotor self-inductance \(\hat{L}_r\) and mutual inductance \(\hat{L}_m\) can be computed as:
\[
\begin{bmatrix}
\hat{L}_r \\
\hat{L}_m
\end{bmatrix}
= 
\begin{bmatrix}
\frac{L_r b - 1}{a^2 b} \\
\frac{L_r b - 1}{ab}
\end{bmatrix}
\]
(21)

The block diagram of the proposed observer is shown in Figure 5.

2) STABILITY ANALYSIS

Before constructing the Lyapunov function for stability analysis, the speed and current sliding surfaces \((S_\omega\) and \(S_M)\) need to be defined, and they are:
\[
S = \begin{bmatrix} S_\omega \\ S_M \end{bmatrix} = \begin{bmatrix} \omega_r \\ i_{SM} \end{bmatrix} = 0
\]
(22)

Construct the Lyapunov function as:
\[
V = \frac{1}{2} S \cdot S^T = \frac{1}{2} \omega_r^2 + \frac{1}{2} i_{SM}^2
\]
(23)

Obviously, \(V > 0\), and only when \(dV/dt\) meets the requirement that it is less than zero can we reach the conclusion that the observer is asymptotically stable. Take the derivative of \(V\) and we can obtain:
\[
\frac{dV}{dt} = -B \omega_e \frac{d\omega_r}{dt} + i_{SM} \frac{d\omega_r}{dt} + \frac{L_{r,off} R_s}{C_{off}} i_{SM}^2
\]
(24)

Substitute (16)–(17) into (24), and it can be derived that:
\[
\frac{dV}{dt} = -B \omega_e \frac{d\omega_r}{dt} - \frac{L_{r,off} R_s}{C_{off}} i_{SM}^2
\]
\[
+ (k_1 F(\omega_r) \cdot \frac{\psi_i i_{ST}}{J} - L_m \cdot \frac{\psi_i i_{ST}}{J}) \omega_r
\]
\[
+ (k_2 F(i_{SM}) u_{SM} - \frac{L_r C_{off}}{C u_{SM}} )^2
\]
(25)

Undoubtedly, \(term 1\) is smaller than zero. Consequently, to ensure the system to be stable, the following conditions need to be satisfied:
\[
\begin{cases}
(k_1 F(\omega_r) \cdot \frac{\psi_i i_{ST}}{J} - L_m \cdot \frac{\psi_i i_{ST}}{J}) \omega_r < 0 \\
k_2 F(i_{SM}) u_{SM} - \frac{L_r C_{off}}{C u_{SM}} i_{SM}^2 < 0
\end{cases}
\]
(26)

Empirically, the \(M\)-axis voltage is usually less than zero when IM works as a motor, and then, according to the signs of the speed and current estimation errors, (26) can be written as (27) and (28):
\[
\begin{cases}
\frac{i_{ST} \cdot (k_1 F(\omega_r) - \frac{L_m}{L_r})}{L_r} < 0, & \omega_r > 0 \\
\frac{i_{ST} \cdot (k_1 F(\omega_r) - \frac{L_m}{L_r})}{L_r} > 0, & \omega_r < 0
\end{cases}
\]
(27)
\[
\begin{cases}
k_2 F(i_{SM}) - \frac{L_r}{C} i_{SM}^2 > 0, & i_{SM} > 0 \\
k_2 F(i_{SM}) - \frac{L_r}{C} i_{SM}^2 < 0, & \frac{L_r}{C} i_{SM} < 0
\end{cases}
\]
(28)

In (27), because \(i_{ST}\) is used to generate torque in an IM, it must be greater than zero. Therefore, according to (27)
and (28), the system stability conditions can be deduced as:

\[ k_1 < - \frac{L_m}{L_r |F(\omega_r)|}, \quad k_2 > \frac{L_r}{C |F(i_{M})|} \]  

(29)

Because the magnitudes of \( F(\omega_r) \) and \( F(i_{M}) \) range from \(-1\) to \(1\), to make the system stable, \( k_1 \) and \( k_2 \) should be negatively infinite in theory. Definitely, this is impossible in engineering. Therefore, a new concept of estimation-error-limitation is proposed and used for stability analysis in this paper.

In detail, the physical descriptions of \( \omega_r \) and \( i_M \) are estimation errors. Practically, the estimation errors cannot remain zero, and they will fluctuate mostly around the sliding surfaces, which can be regarded as the intrinsic property of a SMO. However, we can decide the limitations of the estimation errors when designing the observer. Assume the lower limit of the magnitude of estimation errors is denoted as \( \varepsilon \):

\[ \varepsilon = \min(|\omega_r|, |i_M|) \]  

(30)

Then, the minimum absolute value of the switching function is:

\[ F_{\min} = \left| \frac{2}{1 + e^{-\varepsilon \cdot \varepsilon}} - 1 \right| \]  

(31)

So far, it can be concluded that if the observer gains meet the following condition, the observer will keep stable with a minimum estimation error of \( \varepsilon \).

\[ k_1 < - \frac{L_m}{L_r F_{\min}}, \quad k_2 > \frac{L_r}{CF_{\min}} \]  

(32)

Interestingly, it can be noticed that after using the estimation-error-limitation method to tune the observer gains, as long as the estimation errors are greater than the pre-set value, the system will be stable during operations. However, it is possible that there are moments when the estimation errors are lower than \( \varepsilon \). In this case, the system might become unstable momentarily and the estimation errors will increase, leading to the fact the system returns to the stable state again.

B. DESIGN OF SMO-BASED FCS-MPCC

After obtaining the real-time rotor self-inductance and mutual inductance, they can be substituted into the DPPM to predict the future states when using the proposed FCS-MPCC method for control. By using the accurate parameters, the prediction accuracy of the predictive model is expected to increase in the LSF situations, eliminating the impacts of parameter mismatch on the system control performance.

The block diagram of the SMO-based FCS-MPCC strategy is illustrated in Figure 6 (a). It can be noted that there are four specific features concerning the design scheme. First of all, compared to the traditional field-oriented control (FOC) that employs two proportional-integral-derivative (PID) controllers for \( M \), \( T \)-axis current regulation, only one single FCS-MPCC regulator is used to simultaneously control the currents. Secondly, the rotating speed and the rotor flux are regulated by a speed regulator and a flux regulator, respectively. It needs to be mentioned that the outputs of the speed and flux regulators are the reference \( M \), \( T \)-axis currents, respectively. Thirdly, when controlling an IM by the use of the proposed FCS-MPCC controller, the information about rotor flux, synchronous speed, slip speed and flux position should be diagnosed by an observer established according to (4) and (7) due to the lack of corresponding sensors. Finally, the rotor-related inductances estimated by the SMO are directly used in the proposed FCS-MPCC controller, so they can be treated as the inputs of the proposed FCS-MPCC controller. When integrating the SMO into the predictive controller, the detailed implementation procedures in Chapter III-B-3) can be modified as:

Step 1) Measurement and transformation: Obtain the information of rotor speed \( \omega_r(k) \), flux position \( \theta(k) \) and \( M \), \( T \)-axis currents \( i_{M}(k) \) and \( i_{T}(k) \).

Step 2) Synchronous speed and position observation: Observe the slip speed \( \Delta \omega(k) \) and synchronous speed \( \omega_s(k) \).

Step 3) Rotor-related inductance estimation: Substitute the speed, flux, current and voltage information into the SMO to calculate the real-time inductance \( \hat{L}_m \) and \( \hat{L}_r \).

Step 4) First-time prediction: Use the DPPM whose parameters have been updated using the estimated inductances to estimate the future \( M \), \( T \)-axis currents \( i_{M}(k+1/3) \) and \( i_{T}(k+1/3) \) at \( t_k + T_s/3 \).

Step 5) Second-time prediction: Calculate the slip speed \( \Delta \omega(k + 1/3) \), synchronous speed \( \omega_s(k + 1/3) \) and flux

![FIGURE 6. Diagram of proposed SMO-based FCS-MPCC strategy. (a) Block diagram. (b) Simulation setup diagram.](image-url)
position θ(k + 1/3). Then, use the DPPM whose parameters have been updated using the estimated inductances to predict the future M, T-axis currents \( i_{SM}(k + 2/3) \) and \( i_{ST}(k + 2/3) \) at \( t_k + 2T_s/3 \).

Step 6) Third-time prediction: Repeat step 4) to obtain \( i_{SM}(k + 1) \) and \( i_{ST}(k + 1) \).

Step 7) Evaluation: Use the cost function (15) to select the optimal switching states.

Step 8) Actuation: Apply the optimum switching states to the drive system.

V. SIMULATION VERIFICATIONS

In this part, simulation is conducted on a three-phase IM to verify the effectiveness of the proposed algorithms. The actual parameters of the system are shown in Table 1. The simulation setup diagram (established in MATLAB/Simulink 2018b) is shown in Figure 6 (b). It can be seen that for the sake of simplicity, compared to Figure 6 (a), the variable states that need to be measured are directly from the IM model in Figure 6 (b), which is acceptable in simulation. More professional setup details concerning the simulation include: Solver–discrete, Solver Type–fixed-step, Powergui is used for control frequency regulation and Scope is used to display the simulation curves.

| TABLE 1. Actual IM parameters. |
|--------------------------------|
| Parameters | Value | Unit |
| Stat inductance \( L_s \) | 0.134 | Ω |
| Stator resistance \( R_s \) | 0.076 | Ω |
| Mutual inductance \( L_m \) | 0.068 | Ω |
| Rotor inductance \( L_r \) | 0.095 | Ω |
| Rotor resistance \( R_r \) | 0.05 | Ω |
| Number of pole pairs \( p \) | 2 | |
| Friction coefficient \( B \) | 0.003 | |
| Rated load \( T_{load} \) | 5 | Nm |
| Rated speed \( \omega_{rated} \) | 100 | rad/s |

To validate that the proposed three-time-calculation based predictive strategy is superior in the LSF situations, firstly, a comparative study concerning the control performance of the new and traditional strategies is carried out at the control frequencies of 5 kHz and 2.5 kHz. In these two cases, the switching frequencies are lower than 2.5 kHz and 1.25 kHz, respectively. Secondly, in order to verify that the sliding mode inductance observer is effective, the parameter estimation results in the non-mismatch and mismatch situations are given. Simultaneously, when inductance mismatch phenomenon occurs, the control performance characteristics (before and after using the observed parameter values for prediction) are compared. According to the above analysis, the performance measures of the proposed strategies are summarized as follows: 1) The tripartite calculation-based predictive control strategy should show better performance than the traditional one in the LSF conditions; 2) The SMO is able to detect the real-time rotor inductance and mutual inductance even parameter mismatch phenomenon occurs; 3) After combining the FCS-MPCC controller with the SMO, the predictive control strategy should be robust against the inductance uncertainties so as to show better performance.

A. COMPARATIVE RESULTS BETWEEN PROPOSED FCS-MPCC AND TRADITIONAL FCS-MPCC IN LSF CONDITIONS

To compare the performance of the controllers at low control frequencies, the simulation setup is as follows: Between 0 and 5 s, the reference speed is set as 30 rad/s (low speed) under no load. At 5 s, a sudden load of 5 Nm is imposed on the machine rotor. From 10 s, the machine is controlled to run at the speed of 100 rad/s (high speed). Figure 7 illustrates the dynamics of the traditional and the proposed FCS-MPCC methods when the control frequency is 5 kHz. Firstly, both the traditional and the new methods are able to control the motor to stabilize at the desired speed and flux. Secondly, in Figure 7 (a) and (b), the settling time is about 0.7 s when the machine speed rises from zero to 30 rad/s. In terms of the speed overshoot, it is 2% for the traditional method, and it is slightly lower for the new method (1.33%). Thirdly, after the load is imposed on the rotor, the machine speed witnesses a small decrease (1.85 rad/s and 1.75 rad/s, respectively) at first, after which it quickly recovers to the normal level within 0.5 s. Simultaneously, the flux experiences a similar trend when the load shifts. These represent that the FCS-MPCC strategies have strong robustness against the external load changes. Finally, before leaving Figure 7, it is interesting to see that the electromagnetic torque ripples of the new method are lower than those of the traditional FCS-MPCC, meaning that the proposed scheme has better steady-state control performance.
In order to further explain the capability of the proposed FCS-MPCC in improving the system steady-state performance characteristics, Figure 8 clearly depicts the flux and current properties when the machine rotates at 30 rad/s under load when the control frequency is 5 kHz. As for the traditional method, the flux ripples (FR), M-axis current ripples (MCR) and T-axis current ripples (TCR) are 0.001 Wb (1.2%), 3.3 A and 5.3 A, respectively. However, the FR, MCR and TCR for the proposed strategy are smaller, with 0.0008 Wb, 3.0 A and 4.83 A. In addition, the total harmonic distortion (THD) also indicates that the tripartite-calculation-based predictive method is superior in the LSF conditions. In detail, the THD of \(a\)-phase current \(i_a\) in Figure 8 (b) is 16.2%, which is 17.3% lower than that in Figure 8 (a). In Figure 9, the steady-state control performance at the speed of 100 rad/s is illustrated. Compared to the low-speed case, the FR, MCR and TCR have experienced an obvious increase regardless of the traditional and the novel strategies. In detail, the FR, MCR and TCR are 0.0013 Wb, 3.8 A and 4.8 A respectively for the conventional FCS-MPCC, while they are only 0.001 Wb, 3.3 A and 5.5 A for the proposed technique. When it comes to the THD of the phase current, nearly 33% harmonics are witnessed for the proposed FCS-MPCC, which is much smaller than that (38.5%) for the traditional strategy. Together with the torque properties in Figure 7, the flux and current characteristics have proven that the new FCS-MPCC is suitable for the LSF situations because it has better control performance.

For the sake of comprehensiveness, this paper will further compare the system steady-state performance when the control performance is lower (2.5 kHz). Figure 10 illustrates the dynamics of the traditional and the proposed FCS-MPCC methods when the control frequency is 2.5 kHz. The simulations setups are the same to those in Figure 7. First of all, the machine can be controlled to work at the desired statuses at such a low frequency. Secondly, although the settling time in Figure 10 (a) and (b) is larger than that in Figure 7 (a) and (b), respectively, the speed overshoots get slightly larger. Moreover, the system still shows strong robustness against the external load variations.

Figure 11 shows the steady-state flux and current performance at the speed of 30 rad/s under load. Obviously, the flux and current ripples have increased greatly when the switching frequency is lower. In detail, as for the traditional control strategy, the FR, MCR and TCR are 0.003 Wb (3.5%), 7.8 A and 15.5 A, respectively. Meanwhile, the THD of the phase current rises to 56.3%. Comparatively speaking, the flux and current steady-state performance characteristics are much better, with the FR, MCR and TCR of 0.002 Wb, 7.0 A and 13.5 A, respectively.

Figure 12 demonstrates the system control performance at the speed of 100 rad/s under load. In comparison with Figure 11, the flux and current ripples and the phase current THD experience a sharp increase. Combined with the results in the 5 kHz situation, it can be concluded that the steady-state performance of an FCS-MPCC strategy is low...
over the high-speed range. This is reasonable because when the sampling period gets longer, there will be very limited sampling points within one rotation cycle if the machine speed is high. This will degrade the control performance inevitably. Compared to the traditional control strategy of which FR, MCR and TCR are 0.005 Wb (5.8%), 9.0 A and 20 A, the flux and current performance of the new method is better. Specifically, the FR, MCR and TCR of the new approach are 0.004 Wb, 8.1 A and 18 A, respectively.

Overall, the simulation results indicate that the proposed FCS-MPCC is superior to the traditional method in the low control frequency situations. Hence, when we use the control-frequency-reduction method to lower the switching loss of an IM drive system, the proposed control method is more suitable than the traditional one.

B. EFFECTIVENESS AND ROBUSTNESS OF SMO

According to the aforementioned analysis, the control performance of the proposed controller at 5 kHz is much greater than that at 2.5 kHz, especially in the high-speed cases.

In practice, 5 kHz can be definitely treated as a low-frequency scenario considering the actual switching frequency is lower than 2.5 kHz. On this ground this part will verify the proposed inductance observer when the machine is controlled by the proposed method to rotate at rated speed (100 rad/s) under rated load with the control frequency of 5 kHz (SMO is used in stable states for the sake of accuracy). In order to comprehensively verify the effectiveness and robustness of the SMO, the parameter identification results in four different working conditions are discussed. Firstly, the mutual and rotor inductance values used for predictive control and inductance observation (denoted as $L_{r_{off}}$ and $L_{m_{off}}$, respectively) are totally consistent with the real ones $L_r$ and $L_m$. Secondly, both the mutual inductance and rotor inductance values are 40% higher than the real ones ($L_{r_{off}} = 1.4L_r$, $L_{m_{off}} = 1.4L_m$). Thirdly, both the mutual inductance and rotor inductance values are 60% lower than the real ones ($L_{r_{off}} = 0.4L_r$, $L_{m_{off}} = 0.4L_m$). Fourthly, the mutual inductance value is 40% higher and the rotor inductance value is 60% lower than the real one ($L_{r_{off}} = 1.4L_r$, $L_{m_{off}} = 0.4L_m$).

Figure 13 shows the inductance identification results in the four situations. It can be seen that regardless of whether or not the parameter mismatch phenomenon occurs, the estimated inductance values are consistent with the real ones. This represents firstly, the proposed observer is able to accurately detect the system inductance, and secondly, the SMO is endowed with strong robustness against the parameter uncertainties. This proves that the approximation in (17) is reasonable. Undoubtedly, the observed parameters can also be substituted into (18) to update the SMO in real time practically.

C. PERFORMANCE OF SMO-BASED FCS-MPCC

In this part, the effects of inductance uncertainties on the control performance (speed, flux and torque) of the proposed FCS-MPCC method (without using SMO) are given at first,
FIGURE 14. Effects of inductance uncertainties on speed, flux and torque characteristics. (a) $L_{r\text{off}} = L_r$, $L_{m\text{off}} = L_m$; (b) $L_{r\text{off}} = 0.4L_r$, $L_{m\text{off}} = 0.4L_m$; (c) $L_{r\text{off}} = 1.4L_r$, $L_{m\text{off}} = 0.4L_m$.

In simulation, the control performances are very similar after incorporating the SMO into the predictive control process even if the degree parameter uncertainty is different. On this ground this paper demonstrates the speed, flux and torque performance characteristics in a typical case (see Figure 15), where $L_{r\text{off}} = 0.4L_r$, $L_{m\text{off}} = 0.4L_m$. Compared to Figure 14 (b), the control performances have been improved greatly. In detail, the speed overshoot gets lower, and the stead-state errors disappear nearly. These represent that the SMO-based FCS-MPCC is robust to the inductance uncertainties. Moreover, the system still shows strong robustness against the external load changes. These represent that the proposed SMO is able to improve the system control performance in the inductance mismatch conditions.

Moreover, one principal purpose of the proposed SMO-based FCS-MPCC is to improve the current control performance when parameter mismatch arises, especially in the stable states. Figure 16 compares the steady-state current control performance before and after using the observed parameter values for prediction in two inductance mismatch situations (working at 100 rad/s under load). It needs to be mentioned that between 0 and 5 s, the proposed FCS-MPCC witnesses a sudden drop and then returns to the normal level in nearly 0.4 s, and the flux hardly changes, indicating that the proposed FCS-MPCC has strong robustness against load variations. However, compared to Figure 14 (a), the dynamics of the system in Figure 14 (b) and (c) show remarkable differences. In detail, firstly, although the settling time is still about 1.4 s when parameter mismatch arises, the speed overshoots become larger (3.5 rad/s and 2.8 rad/s, respectively). Secondl, steady-state errors in speed and flux curves are witnessed in Figure 14 (b) and (c) regardless of the load conditions. In other words, when the parameter mismatch phenomenon emerges, it hard to control the speed and flux to maintain at the desired levels. Finally, when the rotor-related inductances are mismatched, the system still shows strong robustness against load changes. Overall, the parameter mismatch issue can degrade not only the dynamics of the system but also the steady-state performance characteristics of the proposed FCS-MPCC controller, so it is essential to employ effective solutions to the problem.
method using the offline parameter values for prediction is adopted, while after 5 s, the SMO-based FCS-MPCC algorithms are executed. It can be seen that when $L_{r,off} = 0.4L_r$ and $L_{m,off} = 0.4L_m$, the $M$-axis and $T$-axis current ripples are 4.2 and 6.3 A when the mismatched parameters are used for predictive control, while they get down to 3.3 and 5.5 A respectively after compensation. In terms of the situation in which $L_{r,off} = 1.4L_r$ and $L_{m,off} = 0.4L_m$, the current ripples also witness an obvious decrease when the observed inductances are employed for control. These comply with the conclusion in [15], meaning that the proposed SMO-based FCS-MPCC method has strong robustness against parameter uncertainties.

VI. CONCLUSION

This paper proposes a novel SMO–based FCS-MPCC method for IMs to improve the system control performance in the LSF situations and enhance the robustness against the mutual and rotor inductance mismatches. The main novelties and contributions are as follows:

1. According to the discretization procedures of the traditional FCS-MPCC method, the reasons why the prediction accuracy of Euler strategy is low are intuitively explained by the use of physical descriptions. On this ground a DPPM based on three-time calculations in each control period is proposed to achieve the new predictive control strategy. Because the FCS-MPCC algorithms are only executed once in each control period, the switching frequency as well as the switching loss is still low in the LSF conditions, but the control performance is higher than the traditional FCS-MPCC method.

2. A SMO based on SF is designed to identify the real-time values of mutual and rotor inductances, making this research novel and valuable. By constructing a Lyapunov function, the observer stability is analyzed. In the process, the stability conditions are determined by using a new estimation-error limitation technique. It needs to be mentioned that the SMO can be applied to not only the LSF conditions but also the high switching frequency cases.

3. The SMO is incorporated into the DPPM of the FCS-MPCC controller to solve the parameter mismatch problem. The proposed SMO-based FCS-MPCC method shows better performance than the traditional scheme. Specifically, the static errors will disappear, and the steady-state ripples will become lower in the parameter mismatch situations.

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FIGURE 16. Current performance before and after using observed parameter values for prediction. (a) $L_{r,off} = 0.4L_r$, $L_{m,off} = 0.4L_m$; (b) $L_{r,off} = 1.4L_r$, $L_{m,off} = 0.4L_m$.\[108940\] VOLUME 8, 2020
1. Introduction

The rotor-related inductance mismatch problem in interior permanent-magnet synchronous motors (IPMSMs) has been a significant challenge in the development of modern electrical machines. This problem arises due to the variations in the magnetic properties of the rotor, which can be caused by various factors such as material defects, manufacturing tolerances, and aging. The presence of this mismatch can lead to adverse effects on the motor's performance, including decreased efficiency, increased losses, and reduced torque capabilities. To address these issues, various strategies have been proposed in the literature, ranging from hardware solutions, such as using adjustable rotor resistances, to software-based solutions that rely on precise modeling and control strategies.

2. Literature Review

2.1. Control Strategies for Inductance Mismatch

Several control strategies have been developed to mitigate the impact of rotor-related inductance mismatch on IPMSMs. These strategies typically involve either model-based techniques or observer-based methods. Model-based approaches, such as those described by [10] and [11], utilize mathematical models of the motor to predict and compensate for the inductance mismatch. On the other hand, observer-based methods, as discussed in [12] and [13], estimate the rotor position and angle using state observers, which can then be used to correct for the inductance mismatch.

2.2. Disturbance Observer Applications

Recent advancements in disturbance observers have been applied to address the inductance mismatch issue in various motor drives. The works by [14] and [15] demonstrate the effectiveness of disturbance observers in improving the transient response and reducing the ripple in the output torque. Moreover, the integration of a disturbance observer with predictive current control has been shown to be beneficial in [16], where it helps in mitigating the effects of inductance mismatch on the current ripple.

3. Methodology

To overcome the challenges posed by inductance mismatch, this paper proposes a novel approach that combines a predictive current control strategy with an adaptive disturbance observer. The objective is to ensure robust control performance under varying conditions, while minimizing the effects of inductance mismatch on the motor's operation. The methodology involves the development of a hybrid control system that dynamically adjusts the disturbance observer gain based on the detected inductance mismatch, thereby optimizing the performance of the motor drive.

4. Results

The proposed control strategy was tested on a laboratory prototype of an IPMSM drive system. The experimental results show a significant improvement in the motor's dynamic response and torque ripple compared to conventional control methods. The adaptive disturbance observer effectively compensates for the inductance mismatch, leading to smoother operation and increased efficiency.

5. Conclusion

This paper has presented a novel control strategy for IPMSMs that utilizes a combination of predictive current control and an adaptive disturbance observer. The approach demonstrates improved performance in terms of reduced torque ripple and better dynamic response, making it a promising solution for practical applications. Further research is needed to validate these findings across a wider range of motor drives, including those subjected to more severe operating conditions.