# A Robust Deadbeat Predictive Controller with Delay Compensation Based on Composite Sliding Mode Observer for PMSMs

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***Abstract*—This paper proposes an improved deadbeat predictive controller for permanent magnet synchronous motor (PMSM) drive systems. It can eliminate the influence of parameter mismatch of inductance, resistance and flux linkage. First, the performance of the conventional predictive current method is investigated to analyze sensitivities of the electric parameters. Then, a composite sliding mode disturbance observer (SMDO) based on stator current and lumped disturbance is proposed, which can simultaneously estimate the future current value and lumped disturbance caused by the parameter mismatch of inductance, resistance and flux linkage. Based on the discrete-time SMDO, currents are estimated and used to replace the sampled values to compensate one-step delay caused by calculation and sampling delay. Both simulation and experimental performances of the proposed method have been validated and compared with the conventional control methods under different conditions. The comparison results show the superiority of the proposed predictive current control method based on the composite SMDO.**

***Index Terms*—Permanent magnet synchronous motor (PMSM), deadbeat predictive control, sensitivity, sliding mode disturbance observer.**

## I. Introduction

### A. Motivation

DUE to the inherent features of high efficiency, high power density and fast control response, permanent magnet synchronous motors (PMSMs) have been widely applied in electric vehicles [1]-[5]. The classical field-oriented control (FOC) has been widely adopted in the PMSM drive system to achieve desired servo control performance for its effective and reliable control methodology [6]-[8].

In FOC-based PMSM drive control, a double cascade loop controller is typically employed. High-performance electrical drives require efficient inner current control loops due to the intrinsic relationship between current quality response and torque control. In order to achieve high transient performance and steady-state precision, many current control schemes have been investigated including proportional-integral (PI) control [9], hysteresis control [10], and predictive control [11]-[13].

Among them, the predictive control has coherent and advantageous characteristics such as improved rotational speed response. The main principle of predictive control is to predict the future behavior of the state variables through drive system model [14]-[16]. And hysteresis-based, trajectory-based, deadbeat predictive control (DBPC) and finite-set model predictive control are known as the main four types of predictive control, which are applied to motor control. Among them, the DBPC and model predictive control are the most widely researched predictive control schemes [17]-[18].

### B. Related Research

Model predictive control can predict the future behavior of states through the system discrete model and inherent discrete characteristics of inverter. Then it determines future voltage vector based on the optimization of the cost function. The selected voltage vector is one of seven basic voltage vectors and can minimize the cost function, which is used for the output of control system. Model predictive control has good ability to handle constraints of system variables, such as the maximum output voltage limits from the inverter. Furthermore, the advantages of robustness ensure that the model predictive control provides excellent control of overall system performance. However, an inevitable drawback of the model predictive control is that the switching frequency may vary, as the state of switches depends on the sequence of inputs, leading to a suboptimal current ripple [19]-[22].

Compared to model predictive control and other predictive control approaches, the DBPC combines good dynamic performance with constant switching frequency, which can force the control error to zero in a short time. Thus, DBPC has been widely used in many applications, including pulsewidth modulation (PWM) rectifiers, active filter control [23], and

Manuscript received Jun 22, 2020; revised Aug 13, 2020, Dec 9, 2020 and Jan 18, 2021; accepted Feb 23, 2021. This work was supported by the National Natural Science Foundation of China under Project 51875261, the Natural Science Foundation of Jiangsu Province of China under Project BK20180046, the "Qinglan project" of Jiangsu Province, and the Postgraduate Research & the Practice Innovation Program of Jiangsu Province under Project SJCX19_0574. *(Corresponding author: Gang Lei)*

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motor drive control [24]. Since the DBPC does not need a cost function minimization algorithm, it is less computationally complex than model predictive control [25]-[27]. The method forecasts voltage reference values based on the motor discrete model and then converts them into corresponding switch configurations by space vector PWM. Based on the above idea, the DBPC includes predictive current control and deadbeat control (DB) direct torque control. Accurate current control with working flux estimation may still lead to inaccurate torque control of the motor, which results in adoption of DB direct torque control. In [28], an advanced deadbeat direct torque and flux control strategy was proposed to improve the operation performance, in which a decouple control strategy of torque and stator flux was proposed by constructing the algorithm model in the stator flux oriented coordinate system, which shows superiority in torque control.

Both transient and steady-state performances of the DBPC method crucially depends on the accuracy of the stator resistance, inductance, as well as permanent magnet flux linkage parameter. However, the electric parameters, including flux linkage, resistance and inductance may change due to temperature rise and magnetic saturation, especially under high-temperature operation conditions [29]. In [19], a new discrete-time robust predictive current controller was presented for PMSM drives, in which a discrete-time integral term is added to the deadbeat current prediction to provide the robustness against normbounded parametric uncertainties and unmodeled dynamics. In [30], a composite control method combining deadbeat predictive current control part and current prediction and feedforward compensation part was proposed to promote control performance. In the designed stator current and disturbance observer, a novel sliding-mode exponential reaching law was introduced to further suppress the chattering of the observer. The composite control method is a creative deadbeat current control method, which solves parameters mismatch problem through feedforward compensation part and further suppresses the chattering of voltage compensation significantly. However, there are still some problems with the reference. The performance of the sliding mode disturbance observer is dependent on the design of the new approach law, which requires a smaller reaching time and sliding mode ripples. Additionally, it is necessary to make a reasonable balance between the computational load caused by the design of the complex reaching law and the performance of the law. Due to the digital implementation of predictive control, computing and sampling delay are inevitable [31]. However, the reference does not consider the delay in controller design, which will lead to system performance degradation.

In order to overcome the above problems, some methods have been proposed. In [21], a robust high bandwidth discrete-time predictive current control scheme for voltage-source pulsewidth-modulated (VS-PWM) converters was presented, in which a digital predictive current controller with delay compensation is adopted based on a current observer with an adaptive internal model. In [32], an oversampling deadbeat current control approach was presented to achieve a constant switching frequency and an optimal current ripple along with a high current loop bandwidth and robust behavior to parameter variation. In [33], a robust predictive current control was proposed to eliminate the influence of inductance parameter mismatch. However, the influence caused by the mismatch of permanent magnet flux linkage and resistance is ignored. In [34], an improved deadbeat-based predictive current control scheme based unified high-order sliding-mode disturbance observer was proposed to promote speed robustness and current tracking accuracy. It is a creative deadbeat control method, which essentially solves the mismatch problem of mechanical and electrical parameters through disturbance compensation with feedback to the designed control system.

### C. *Contributions*

An improved deadbeat predictive controller is developed to guarantee the performance of PMSM drives regardless of parameter mismatch and one-step delay in digital control in this work. Compared with the conventional predictive current controller, the proposed method can achieve lower control currents ripples based on current and lumped disturbance discrete-time sliding mode observer (SMO). Furthermore, compensation of calculation time delay is considered and the second-order expansion of the rotor speed is adopted to improve the current reference accuracy.

The main contributions of this paper are listed as follows.

1) This paper proposes a different disturbance observer to applied in the improved deadbeat predictive current controller, which is designed through analyzing the structure of current state space equations to effectively eliminate the coupling terms. Therefore, the design idea is completely different from the high-order sliding mode structure or new reaching law-based method in the mentioned references.
2) The design method based on the state space structure of the control plant reduces the calculation complexity of the observation value for the nonlinear model, which is conducive to achieving higher control accuracy.
3) In order to suppress the influence of the control delay, one-step delay compensation is considered to calculate the voltage vector $u_{dq}(k+1)$. Furthermore, the compensation inevitably causes a two-sample delay in the reference tracking of the reference currents $i^*_{dq}(k)$, which can be ignored by appropriate design of control parameters.
4) In order to improve the current reference accuracy, the second-order expansion of the rotor speed is adopted in the method, which further improves the control performance.

### D. *Paper Organization*

The remainder of this paper is organized as follows. Section II describes the mathematical model of a surface-mounted PMSM (SPMSM) and develops the model under parameter disturbance. In Section III, based on the discrete-time model of the SPMSM, parameter sensitivity of the conventional predictive current control to inductance, resistance and flux linkage is analyzed. In Section IV, a discrete-time SMDO based on stator current and disturbance is designed to eliminate the influence of parameter mismatch and compensate one-step delay. Simulation and experimental results are provided in Section V. Finally, Section VI gives the summary and draws the main conclusions of the study.

## II. MATHEMATICAL MODEL

The dynamical model of the SPMSM can be described in the *dq* synchronously rotating reference frame as follows:

$$\begin{cases} u_d = R_s i_d + \dfrac{d\psi_d}{dt} - \omega_e \psi_q \\ u_q = R_s i_q + \dfrac{d\psi_q}{dt} + \omega_e \psi_d \end{cases} \tag{1}$$

The flux linkage equation is

$$\begin{cases} \psi_d = L_d i_d + \psi_f \\ \psi_q = L_q i_q \end{cases} \tag{2}$$

where $u_d$, $u_q$, $i_d$, and $i_q$ are the *d*- and *q*-axis components of the stator voltage and current, respectively. $R_s$ and $L_s$ represent the stator resistance and inductance, respectively. $\psi_f$ is permanent magnet flux linkage. $\omega_e$ is electrical angular speed.

The system model can be written in the standard state-space form as

$$\frac{d}{dt}\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \boldsymbol{P}\cdot\begin{bmatrix} i_d \\ i_q \end{bmatrix} + \boldsymbol{Q}\cdot\begin{bmatrix} u_d \\ u_q \end{bmatrix} + \begin{bmatrix} 0 \\ -\omega_e \psi_f / L_s \end{bmatrix} \tag{3}$$

where $\boldsymbol{P} = \begin{bmatrix} -R_s/L_s & \omega_e \\ -\omega_e & -R_s/L_s \end{bmatrix}, \boldsymbol{Q} = \begin{bmatrix} 1/L_s & 0 \\ 0 & 1/L_s \end{bmatrix}$.

The electromagnetic torque and mechanical equations of SPMSM are shown as follows:

$$\begin{cases} T_e = \dfrac{3}{2} p_n \psi_f i_q \\ \dfrac{J}{p_n}\dfrac{d\omega_e}{dt} = T_e - T_L \end{cases} \tag{4}$$

where $J$ is the moment of inertia. $p_n$ is the number of pole pairs. $T_e$ and $T_L$ are the electromagnetic torque and the load torque, respectively.

According to (3) and (4), when the mismatch of three machine parameters exists, the dynamical model can be written as

$$\begin{cases} \dfrac{di_d}{dt} = -\dfrac{\tilde{R}_s}{\tilde{L}_s} i_d + \omega_e i_q + \dfrac{1}{\tilde{L}_s} u_d + r_d \\ \dfrac{di_q}{dt} = -\dfrac{\tilde{R}_s}{\tilde{L}_s} i_q - \omega_e i_d + \dfrac{1}{\tilde{L}_s} u_q - \dfrac{\tilde{\psi}_f \omega_e}{\tilde{L}_s} + r_q \\ \dfrac{d\omega_e}{dt} = \dfrac{3 p_n^2 \tilde{\psi}_f}{2J} i_q + r_{\omega_e} \end{cases} \tag{5}$$

where $r_d$, $r_q$ and $r_{\omega_e}$ represent the parameter perturbances, which consist of flux linkage error, resistance error and induction error. And the perturbances can be described as

$$\begin{cases} r_d = \dfrac{-\Delta L_s \tilde{R}_s + L_s \Delta \tilde{R}_s}{\tilde{L}_s(\tilde{L}_s - \Delta L_s)} i_d + \dfrac{\Delta L_s}{\tilde{L}_s(\tilde{L}_s - \Delta L_s)} u_d \\ r_q = \dfrac{-\Delta L_s \tilde{R}_s + \tilde{L}_s \Delta R_s}{\tilde{L}_s(\tilde{L}_s - \Delta L_s)} i_q + \dfrac{\Delta L_s}{\tilde{L}_s(\tilde{L}_s - \Delta L_s)} u_q + \dfrac{-\Delta L_s \tilde{\psi}_f + \tilde{L}_s \Delta \psi_f}{\tilde{L}_s(\tilde{L}_s - \Delta L_s)} \omega_e \\ r_{\omega_e} = -\dfrac{3 p_n^2 \Delta \psi_f}{2J} i_q - \dfrac{p_n}{J} T_L \end{cases} \tag{6}$$

And $\tilde{R}_s = R_s + \Delta R_s$, $\tilde{L}_s = L_s + \Delta L_s$ and $\tilde{\psi}_f = \psi_f + \Delta\psi_f$ are the nominal values of system parameters. $\Delta R_s$, $\Delta L_s$, and $\Delta\psi_f$ are the parameters errors between the nominal values and the actual values.

## III. PARAMETER SENSITIVITY ANALYSIS

### A. Conventional DBPC Method

In order to digitally implement the model-based controller, the discretization of the plant equations is necessary. The discretized state-space form of the plant (3) around a generic time instant $k$ is presented as

$$\begin{bmatrix} i_d(k+1) \\ i_q(k+1) \end{bmatrix} = \boldsymbol{M}(k)\cdot\begin{bmatrix} i_d(k) \\ i_q(k) \end{bmatrix} + \boldsymbol{N}\cdot\begin{bmatrix} u_d(k) \\ u_q(k) \end{bmatrix} + \boldsymbol{T}(k) \tag{7}$$

where

$$\boldsymbol{M}(k) = e^{\boldsymbol{P}T_s},\ \boldsymbol{N} = \int_0^{T_s} e^{\boldsymbol{P}t}\boldsymbol{Q}dt,\ \boldsymbol{T}(k) = \begin{bmatrix} 0 \\ -\omega_e(k)\psi_f T_s / L_s \end{bmatrix},$$

and $T_s$ denotes the sampling period of the control system.

Taylor expansion is carried out on $\boldsymbol{M}$ and $\boldsymbol{N}$, and the quadratic and higher order terms are ignored. Then the matrices of the state-space model are as follows:

$$\boldsymbol{M}(k) = \begin{bmatrix} 1 - T_s R_s / L_s & T_s \omega_e(k) \\ -T_s \omega_e(k) & 1 - T_s R_s / L_s \end{bmatrix},\ \boldsymbol{N} = \begin{bmatrix} T_s / L_s & 0 \\ 0 & T_s / L_s \end{bmatrix}.$$

According to (7), the output voltage vector of the DBPC is expressed as follows:

$$\begin{bmatrix} u_d(k) \\ u_q(k) \end{bmatrix} = \boldsymbol{N}^{-1}\left\{\begin{bmatrix} i_d^*(k+1) \\ i_q^*(k+1) \end{bmatrix} - \boldsymbol{M}(k)\cdot\begin{bmatrix} i_d(k) \\ i_q(k) \end{bmatrix} - \boldsymbol{T}(k)\right\} \tag{8}$$

where $i_d^*(k+1)$ and $i_q^*(k+1)$ are the reference currents in *dq*-axes reference rotor frame. It can be seen that the future reference current value $i^*(k+1)$ is needed, which can be assumed to be equal to the actual value $i^*(k)$, for $T_s$ is short enough compared with the dynamic behavior of the system. Therefore, the reference current value can be considered constant over $T_s$.

The output stator voltage vector enables the actual current vector to approach the expected values. The block diagram of the conventional DBPC is illustrated in Fig. 1. The voltage vector is obtained through the predictive mode (8), which is converted to switching signals through modulation process.

*B. Parameter Sensitivity Analysis*

The DBPC is a predictive control method based on mathematical model of PMSM due to the existence of three machine parameters in a current prediction model. This means that deadbeat predictive controller is subject to parameter sensitivity, and the accuracy of the prediction model will directly influence the control performance. In order to evaluate the relationship between the prediction error and three parameters mismatch, parameter sensitivity analysis is discussed.

According to (7), the discrete current model under parameters disturbance can be expressed as follows:

$$\begin{bmatrix}\tilde{i}_d(k+1)\\ \tilde{i}_q(k+1)\end{bmatrix}=\tilde{\boldsymbol{M}}(k)\cdot\begin{bmatrix}i_d(k)\\ i_q(k)\end{bmatrix}+\tilde{\boldsymbol{N}}\cdot\begin{bmatrix}u_d(k)\\ u_q(k)\end{bmatrix}+\tilde{\boldsymbol{T}}(k) \quad (9)$$

where

$$\tilde{\boldsymbol{M}}(k)=\begin{bmatrix}1-T_s\tilde{R}_s/\tilde{L}_s & T_s\omega_e(k)\\ -T_s\omega_e(k) & 1-T_s\tilde{R}_s/\tilde{L}_s\end{bmatrix},\ \tilde{\boldsymbol{N}}=\begin{bmatrix}T_s/\tilde{L}_s & 0\\ 0 & T_s/\tilde{L}_s\end{bmatrix},$$

$$\tilde{\boldsymbol{T}}(k)=\begin{bmatrix}0\\ -\omega_e(k)\tilde{\psi}_f T_s/\tilde{L}_s\end{bmatrix}.$$

Therefore, the errors between the current response values and the current reference values subjected to parameter perturbation can be obtained as

$$\begin{cases}e_d=\tilde{i}_d(k+1)-i_d(k+1)\\ e_q=\tilde{i}_q(k+1)-i_q(k+1)\end{cases} \quad (10)$$

with

$$\begin{cases}e_d=\dfrac{\Delta L_sR_s-L_s\Delta R_s}{L_s(\Delta L_s+L_s)}T_si_d(k)-\dfrac{\Delta L_s}{L_s(\Delta L_s+L_s)}T_su_d(k)\\ e_q=\dfrac{\Delta L_sR_s-L_s\Delta R_s}{L_s(\Delta L_s+L_s)}T_si_q(k)-\dfrac{\Delta L_s}{L_s(\Delta L_s+L_s)}T_su_q(k)\\ \qquad +\dfrac{\Delta L_s\psi_f-L_s\Delta\psi_f}{L_s(\Delta L_s+L_s)}T_s\omega_e(k)\end{cases}. \quad (11)$$

Differentiating (11) yields

$$\begin{cases}\dfrac{\partial e_d}{\partial\Delta L_s}=\dfrac{T_s\left[(R_s+\Delta R_s)i_d(k)-u_d(k)\right]}{(\Delta L_s+L_s)^2},\\ \dfrac{\partial e_q}{\partial\Delta L_s}=\dfrac{T_s\left[(R_s+\Delta R_s)i_q(k)-u_q(k)-\omega_e(k)(\psi_f+\Delta\psi_f)\right]}{(\Delta L_s+L_s)^2}\\ \dfrac{\partial e_d}{\partial\Delta R_s}=-\dfrac{T_si_d(k)}{(\Delta L_s+L_s)},\dfrac{\partial e_q}{\partial\Delta R_s}=-\dfrac{T_si_q(k)}{(\Delta L_s+L_s)}\\ \dfrac{\partial e_d}{\partial\Delta\psi_f}=0,\dfrac{\partial e_q}{\partial\Delta\psi_f}=-\dfrac{T_s\omega_e(k)}{(\Delta L_s+L_s)}\end{cases} \quad (12)$$

Equations (11) and (12) indicate that $d$-axis current is not affected by flux linkage error $\Delta\psi_f$, while $dq$-axes currents are under influence of resistance error $\Delta R_s$ and inductance error $\Delta L_s$. Therefore, the uncertainty of any system parameter will cause errors in the current response. Meanwhile, the current response errors and error variation rates are related to mechanical parameters, speed and currents.

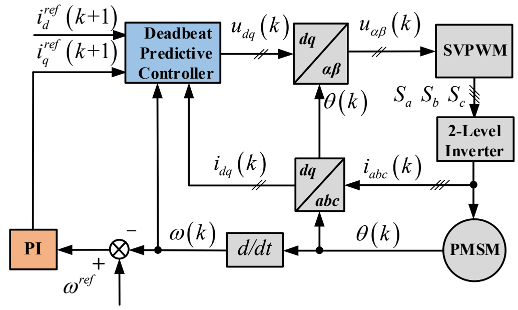
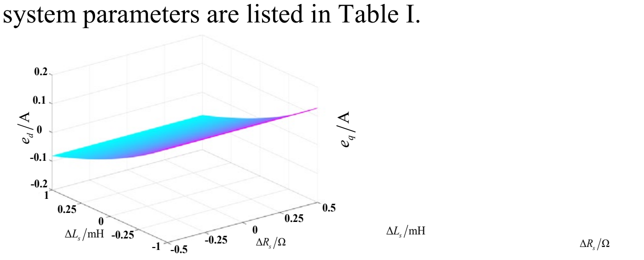


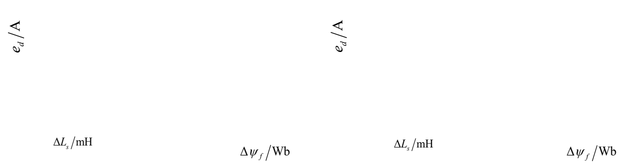

Fig. 1. Block diagram of the conventional DBPC.

The relationship of the $dq$-axes current response errors and parameter mismatch of inductance $L_s$, resistance $R_s$ and flux linkage $\psi_f$ are shown in Figs. 2 and 3. The speed reference is set as 360 rpm and the load torque is set as 0 The SPMSM and system parameters are listed in Table I.




(a) (b)

Fig. 2. Current prediction errors under inductance and resistance parameter mismatch. (a) d-axis current error, and (b) q-axis current error.




(a) (b)

Fig. 3. Current prediction errors under inductance and flux linkage parameter mismatch. (a) d-axis current error, and (b) q-axis current error.

Fig. 2 shows the $dq$-axes current response errors under inductance $L_s$ and resistance $R_s$ parameter mismatch. As shown, the resistance error has little effect on $dq$-axes current response errors when parameter mismatch of inductance remains unchanged. While inductance error has a great influence on the current prediction error. Fig. 3 illustrates the $dq$-axes current prediction errors under parameter mismatch of inductance $L_s$ and flux linkage $\psi_f$. It can be seen that the parameter mismatch of flux linkage has no effect on $d$-axis current response error. As shown in Fig. 3(b), the flux linkage error produces remarkable influence on $q$-axis current prediction errors.

According to the analysis above, the conclusion can be made that the conventional predictive current control method is sensitive to inductance, resistance and flux linkage parameters. Therefore, robustness design based on the predictive current controller should be considered.

## IV. DPCC Method with Perturbance Observer

### *A. Composite Sliding Mode Disturbance Observer (SMDO) Design*

According to (5), the state equation can be expressed as follows when parameter variations are taken into consideration:

$$\frac{d\boldsymbol{x}}{dt} = \boldsymbol{A}\boldsymbol{x} + \boldsymbol{B}\boldsymbol{u} + \boldsymbol{f}_a + \boldsymbol{f}_b \tag{13}$$

where $\boldsymbol{x} = \begin{bmatrix} i_d \\ i_q \\ \omega_e \end{bmatrix}$, $\boldsymbol{u} = \begin{bmatrix} u_d \\ u_q \\ i_{sq} \end{bmatrix}$, $\boldsymbol{f}_a = \begin{bmatrix} 0 \\ -\tilde{\psi}_f \omega_e / \tilde{L}_s \\ 0 \end{bmatrix}$, $\boldsymbol{f}_b = \begin{bmatrix} r_d \\ r_q \\ r_{\omega_e} \end{bmatrix}$ are the state variables, system output, flux linkage term and lumped disturbance, respectively.

The coefficient matrixes of the state equations are as follows:

$$\boldsymbol{A} = \begin{bmatrix} -\tilde{R}_s/\tilde{L}_s & \omega_e & 0 \\ -\omega_e & -\tilde{R}_s/\tilde{L}_s & 0 \\ 0 & 0 & 0 \end{bmatrix} \boldsymbol{B} = \begin{bmatrix} 1/\tilde{L}_s & 0 & 0 \\ 0 & 1/\tilde{L}_s & 0 \\ 0 & 0 & 3p_n{}^2\tilde{\psi}_f/2J \end{bmatrix}.$$

In order to implement the proposed deadbeat predictive current controller, it is necessary to estimate the disturbance caused by parameter variation. Since the sliding mode observer can estimate the uncertainty, the SMDO can be designed as follows based on (13)

$$\frac{d\hat{\boldsymbol{x}}}{dt} = \boldsymbol{A}\hat{\boldsymbol{x}} + \boldsymbol{B}\boldsymbol{u} + \boldsymbol{f}_a + \boldsymbol{F}\boldsymbol{e} + \boldsymbol{G}\operatorname{sgn}(\boldsymbol{e}) \tag{14}$$

where $\hat{\boldsymbol{x}}$ is the estimated value of state variables $\boldsymbol{x}$, sgn() is the symbolic function, and $\boldsymbol{F}$ and $\boldsymbol{G}$ are the gain matrixes.

According to sliding-mode control theory, sliding-mode design procedure can be divided into two steps: the first step is sliding-mode surface design, and the second step is to design the sliding-mode control function, which can force the state trajectory to converge to the sliding-mode surface. In this paper, the sliding-mode surface is defined as

$$\boldsymbol{e} = \boldsymbol{x} - \hat{\boldsymbol{x}} = \begin{bmatrix} e_1 \\ e_2 \\ e_3 \end{bmatrix} = \begin{bmatrix} i_d - \hat{i}_d \\ i_q - \hat{i}_q \\ \omega_e - \hat{\omega}_e \end{bmatrix}. \tag{15}$$

According to (13) and (14), the error equation can be obtained as follows:

$$\frac{d\boldsymbol{e}}{dt} = \boldsymbol{A}\boldsymbol{e} + \boldsymbol{f}_b - \boldsymbol{F}\boldsymbol{e} - \boldsymbol{G}\operatorname{sgn}(\boldsymbol{e}) \tag{16}$$

where $\boldsymbol{F}$ and $\boldsymbol{G}$ are designed as follows:

$$\boldsymbol{F} = \begin{bmatrix} 0 & \omega_e & 0 \\ -\omega_e & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}, \boldsymbol{G} = \begin{bmatrix} g_1 & 0 & 0 \\ 0 & g_2 & 0 \\ 0 & 0 & g_3 \end{bmatrix}.$$

In order to guarantee the convergence of current errors and disturbance estimation errors of the proposed composite SMDO, the Lyapunov candidate function is selected as follows:

$$\boldsymbol{V} = \frac{1}{2}\boldsymbol{e}^T\boldsymbol{e}. \tag{17}$$

Deriving (17) and combining with (16) yields

$$\begin{aligned} \frac{d\boldsymbol{V}}{dt} &= \boldsymbol{e}^T\left(\boldsymbol{A}\boldsymbol{e} + \boldsymbol{f}_b - \boldsymbol{F}\boldsymbol{e} - \boldsymbol{G}\operatorname{sgn}(\boldsymbol{e})\right) \\ &= \boldsymbol{e}^T\left(\boldsymbol{A} - \boldsymbol{F}\right)\boldsymbol{e} + \boldsymbol{e}^T\boldsymbol{f}_b - \boldsymbol{e}^T\boldsymbol{G}\operatorname{sgn}(\boldsymbol{e}) \end{aligned}. \tag{18}$$

Substituting the gain matrixes into (18) yields

$$\frac{d\boldsymbol{V}}{dt} = \boldsymbol{e}^T \begin{bmatrix} -\tilde{R}_s/\tilde{L}_s & 0 & 0 \\ 0 & -\tilde{R}_s/\tilde{L}_s & 0 \\ 0 & 0 & 0 \end{bmatrix} \boldsymbol{e} + \boldsymbol{e}^T \begin{bmatrix} r_d \\ r_q \\ r_{\omega_e} \end{bmatrix} - \boldsymbol{e}^T \begin{bmatrix} g_1 & 0 & 0 \\ 0 & g_2 & 0 \\ 0 & 0 & g_3 \end{bmatrix} \operatorname{sgn}(\boldsymbol{e}) \tag{19}$$

Simplifying (19), one can obtain

$$\begin{aligned} \frac{d\boldsymbol{V}}{dt} &= -\frac{\tilde{R}_s}{\tilde{L}_s}e_1^2 - \frac{\tilde{R}_s}{\tilde{L}_s}e_2^2 + e_1 r_d + e_2 r_q + e_3 r_{\omega_e} - g_1|e_1| - g_2|e_2| - g_3|e_3| \\ &\le |e_1|\left(|r_d| - g_1\right) + |e_2|\left(|r_q| - g_2\right) + |e_3|\left(|r_{\omega_e}| - g_3\right) \\ &\le |e_1|\left(G_1 - g_1\right) + |e_2|\left(G_2 - g_2\right) + |e_3|\left(G_3 - g_3\right) \end{aligned} \tag{20}$$

Since the disturbance function is bounded, the boundary values are introduced, that is, disturbance values satisfy $|r_d|<G_1$, $|r_q|<G_2$, $|r_{\omega_e}|<G_3$. And the elements of gain matrix $\boldsymbol{G}$ satisfy that $g_1>G_1$, $g_2>G_2$, $g_3>G_3$, then the Lyapunov candidate function satisfies $dV/dt\le 0$. Therefore, the proposed SMDO with matrix $\boldsymbol{G}$ can reach the sliding mode surface in finite time and remains stable, which ensures the asymptotic stability of the proposed composite SMDO and the parameter observations will approach the actual values.

### *B. Discrete Expression of Composite SMDO*

Because the proposed composite SMDO will be only computed with discrete form and applied to control system within the sampling period, it is necessary to deduce the discrete expression of the composite observer. Assuming that the sampling period is sufficiently small for the discrete time system, the composite discrete-time observer can be discretized as (21) according to (14)

$$\begin{cases}\hat{i}_d(k+1)=\left(1-\dfrac{T_s\tilde{R}_s}{\tilde{L}_s}\right)\hat{i}_d(k)+\dfrac{T_s}{\tilde{L}_s}u_d(k)+T_s\omega_e(k)\hat{i}_q(k)\\ \qquad +T_s\omega_e(k)e_2(k)+T_sg_1\,\mathrm{sgn}(e_1(k))\\ \hat{i}_q(k+1)=\left(1-\dfrac{T_s\tilde{R}_s}{\tilde{L}_s}\right)\hat{i}_q(k)+\dfrac{T_s}{\tilde{L}_s}u_q(k)-T_s\omega_e(k)\hat{i}_d(k)-\dfrac{T_s\tilde{\psi}_f}{L_s}\omega_e(k)\\ \qquad -T_s\omega_e(k)e_1(k)+T_sg_2\,\mathrm{sgn}(e_2(k))\\ \hat{\omega}_e(k+1)=\hat{\omega}_e(k)+\dfrac{3T_sp_n^2\tilde{\psi}_f}{2J}i_q(k)+T_sg_3\,\mathrm{sgn}(e_3(k))\end{cases} \tag{21}$$

where $\hat{i}_d(k+1)$ and $\hat{i}_q(k+1)$ are the predictive values of *dq*-axes currents, and $\hat{\omega}_e(k+1)$ represents the predictive values of electrical angular speed.

In order to suppress predictive current ripple caused by sign function, it is necessary to replace the sign function with smoothing function such as hyperbolic tangent function with smooth continuity. Then the discrete block diagram of the proposed composite SMDO is shown in Fig. 4.

When the system trajectory reaches the sliding-mode surface and enters the sliding-mode state, it can be obtained that

$$\frac{d\boldsymbol{e}}{dt}=\boldsymbol{e}=0\,. \tag{22}$$

Combining state equation of observer and (22) yields

$$\boldsymbol{f}_b=\boldsymbol{G}\tanh(\boldsymbol{e})\,. \tag{23}$$

Therefore, the variation rate of disturbance can be obtained as follows

$$\begin{cases}r_d(k)=g_1\tanh(e_1(k))\\ r_q(k)=g_2\tanh(e_2(k))\\ r_{\omega_e}(k)=g_3\tanh(e_3(k))\end{cases}. \tag{24}$$

Combining (6) and (24), the estimated flux linkage can be obtained as

$$\Delta\psi_f(k)=\frac{2Jg_3\tanh(e_3(k))+2p_nT_L}{3p_n^2 i_q(k)}\,. \tag{25}$$

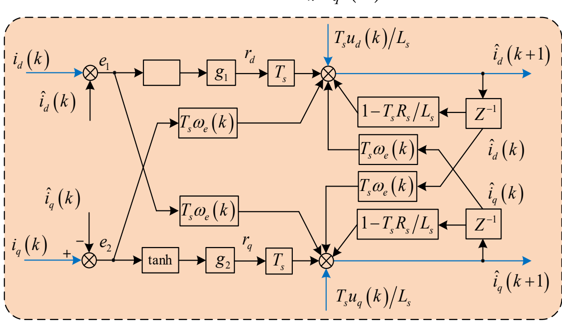



Fig. 4. Discrete block diagram of proposed composite SMDO.

### C. Predictive Current Control with Composite SMDO

According to (5) and (8), the discrete expression of voltage equation can be described as

$$\begin{bmatrix}u_d(k)\\u_q(k)\end{bmatrix}=\begin{bmatrix}i_d^*(k+1)\tilde{L}_s/T_s\\ i_q^*(k+1)\tilde{L}_s/T_s\end{bmatrix}-\begin{bmatrix}0\\-\tilde{\psi}_f\omega_e(k)\end{bmatrix}-\tilde{L}_s\begin{bmatrix}r_d\\r_q\end{bmatrix} -\begin{bmatrix}\tilde{L}_s/T_s-\tilde{R}_s & \tilde{L}_s\omega_e(k)\\ -\tilde{L}_s\omega_e(k) & \tilde{L}_s/T_s-\tilde{R}_s\end{bmatrix}\begin{bmatrix}i_d(k)\\i_q(k)\end{bmatrix}. \tag{26}$$

When the control scheme based on predictive current control is implemented experimentally, it is worth to highlight that the calculation and sampling delay is unavoidable due to the digital implementation mode of the prediction control. This delay can deteriorate the performance of the system if it is not considered in the design of the controller. In order to suppress the influence of the delay, it is necessary to calculate the voltage vector $u_{dq}(k+1)$ by using delay compensated currents $i_{dq}(k+1)$. Since the electromagnetic time constant is smaller than the mechanical time constant in the motor system, the rotor speed can be considered as constant during one sampling period. These compensated currents are employed to replace the measured currents $i_{dq}(k)$ of the model (26), which indicates that the voltage equation can be updated with one-step delay compensation as

$$\begin{bmatrix}u_d(k+1)\\u_q(k+1)\end{bmatrix}=\begin{bmatrix}i_d^*(k+2)\tilde{L}_s/T_s\\ i_q^*(k+2)\tilde{L}_s/T_s\end{bmatrix}-\begin{bmatrix}0\\-\tilde{\psi}_f\omega_e(k)\end{bmatrix}-\tilde{L}_s\begin{bmatrix}\hat{r}_d(k+1)\\\hat{r}_q(k+1)\end{bmatrix} -\begin{bmatrix}\tilde{L}_s/T_s-\tilde{R}_s & \tilde{L}_s\omega_e(k)\\ -\tilde{L}_s\omega_e(k) & \tilde{L}_s/T_s-\tilde{R}_s\end{bmatrix}\begin{bmatrix}\hat{i}_d(k+1)\\\hat{i}_q(k+1)\end{bmatrix}. \tag{27}$$

When compensation of calculation time delay is considered, the future reference can be assumed to be $i_{dq}^*(k+2) = i_{dq}^*(k)$. The current reference approximation will lead to a two-sampling delay in the reference tracking of the reference currents. Fortunately, the sampling time used in this paper is small enough that the delay introduced by the approximation of future references can be ignored.

In order to improve the current reference accuracy, the second-order expansion of the rotor speed is adopted in the method, which is expressed as

$$\frac{\omega_e(k+1)-\omega_e(k)}{T}=\frac{d\omega_e}{dt}+\frac{d^2\omega_e}{dt^2}\cdot\frac{T}{2} \tag{28}$$

where $T$ is the external loop sampling time, much larger than the sampling time $T_s$.

Combining (4) and (28) and considering flux linkage error, it can be obtained that

$$\frac{d^2\omega_e}{dt^2}=\frac{3p_n^2(\tilde{\psi}_f+\Delta\psi_f)}{2J}\frac{di_q}{dt}\,. \tag{29}$$

Substituting (4) and (29) into (28) yields

$$\frac{\omega_e(k+1)-\omega_e(k)}{T}=\frac{3p_n^2(\tilde{\psi}_f+\Delta\psi_f)}{2J}\frac{i_q(k)-i_q(k-1)}{2} +\frac{3p_n^2(\tilde{\psi}_f+\Delta\psi_f)}{2J}i_q(k)-\frac{p_n}{J}T_L\,. \tag{30}$$

The values for $\omega_e^*$ and $i_q^*(k)$ are used as the rotor speed and current reference in (30) by considering $\omega_e^* = \omega_e^*(k+1)$ and $i_q^*(k) = i_q(k)$.

From (30), the reference current can be obtained as

$$i_q^*(k) = \frac{4J\omega_e^* - 4J\omega_e(k) + 4Tp_n T_L + 3Tp_n^{\ 2}(\tilde{\psi}_f + \Delta\psi_f(k))i_q(k-1)}{9Tp_n^{\ 2}(\tilde{\psi}_f + \Delta\psi_f(k))} \tag{31}$$

The control diagram of the proposed deadbeat predictive current controller is shown in Fig. 5. The proposed SMDO is used to compensate parameter disturbances with the estimated lumped disturbance value, which overcomes the influence of parameter mismatch on DBPC method. In order to reduce the influence of sampling and calculation due to the digital implementation of the control system, one-step delay is compensated.

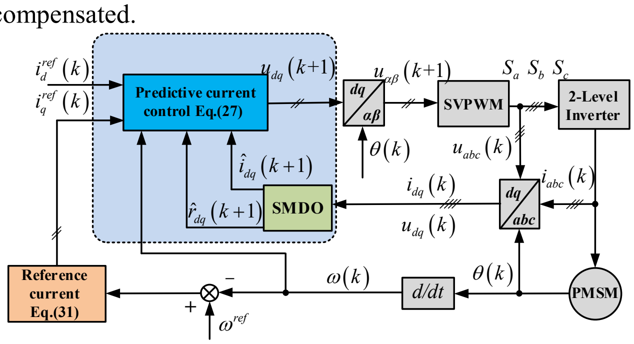



Fig. 5. Control diagram of the proposed deadbeat predictive current controller.

## V. Simulation and Experimental Results

For the purpose of verifying the effectiveness of the proposed method, simulations and experiments for comparison between the proposed method and the conventional DBPC have been conducted at a laboratory platform.

Simulations are established in MATLAB/Simulink. The experiments are performed on the platform of dSPACE 1401 test bench, through which the experimental measurements can be exported to MATLAB and plotted.

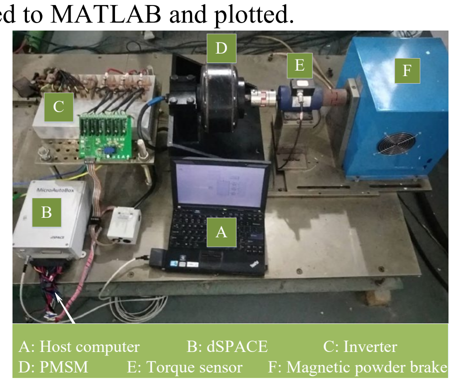



Fig. 6. Experimental setup used to verify the purposed DBPC method.

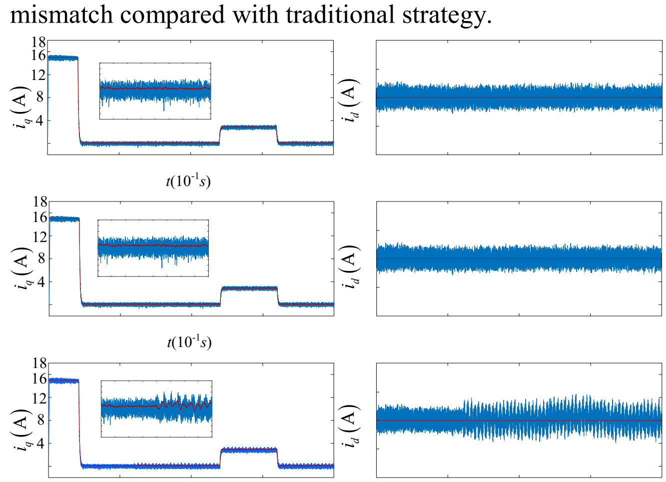
Since the parameters of the motor body cannot be set arbitrarily, the parameters in the control program are changed to achieve the corresponding parameter mismatch. The experimental setup consists of an SPMSM, a torque sensor, and a magnetic powder brake, as shown in Fig. 6. The parameters of the SPMSM system are listed in Table I.

TABLE I
SPMSM and System Parameters

| Parameters | Symbol | Value |
|---|---|---|
| Number of pole pairs | $P_n$ | 22 |
| Stator resistance | $R_s$ | 0.8 Ω |
| Stator inductance | $L_s$ | 4.5 mH |
| Permanent-magnet flux linkage | $\psi_f$ | 0.215 Wb |
| Inertia | $J$ | 0.03 kgm$^2$ |
| Viscous friction coefficient | $v$ | 0.0006 N·m·s |
| Rated speed | $N$ | 360 rpm |
| Rated power | $P_N$ | 30 kW |

The parameters of the composite SMDO are $g_1$=72000, $g_2$=72000, and $g_3$=48000, which ensure the asymptotic stability of composite SMDO. The speed reference is set as 360 rpm, and the sampling frequency used in simulation and experiment is 50 kHz while the external loop sampling frequency is 5 kHz. At the time instants $t_1$ =0.06 s and $t_2$ =0.08 s, step changes in the load torque from 0 to 20 Nm and from 20 Nm to 0 have been applied to the system, respectively. Parameter variations are set at 0.03 s to show the simulation results.

Under different parameter mismatch conditions, the simulation results of *dq*-axes current responses and current references are shown in Figs. 7 and 8. The blue line is the current response while the brown line is the current reference. In Fig. 7, the simulation results of the conventional DBPC are illustrated. It is seen that the parameter mismatch of inductance $L_s$ and flux linkage $\psi_f$ produces great influence on *dq*-axes current responses. In contrast, the variation of resistance $R_s$ has little effect on current response when other parameters remain unchanged, which is similar to the parameter sensitivity analysis in Section III. Furthermore, different parameter variations cause current response error, which can be calculated by (11). Additionally, Fig. 8 presents the simulation results of the proposed DBPC method with compensation voltage based on composite SMDO, which can exactly track the current references under parameters mismatch compared with traditional strategy.




(c)

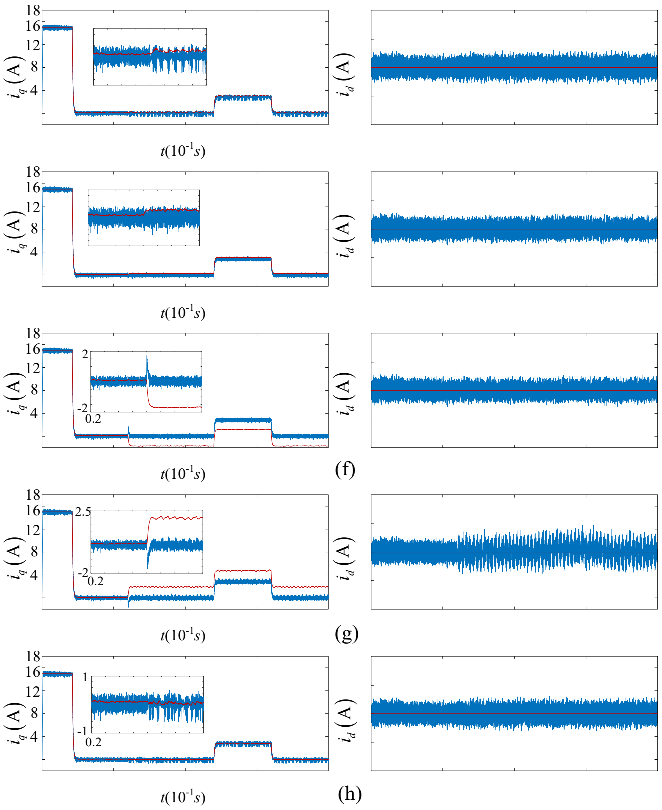
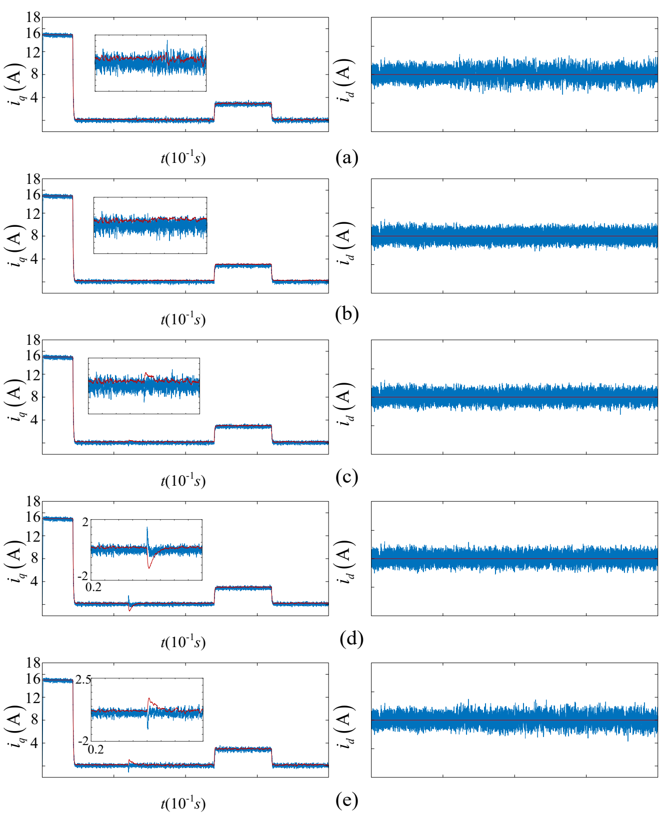



Fig. 7. Simulation performance of conventional DBPC method under different parameter mismatch conditions. (a) $\Delta R_s$=-0.9$R_s$ (b) $\Delta R_s$=9$R_s$ (c) $\Delta L_s$=-0.9$L_s$ (d)$\Delta L_s$=9$L_s$ (e) $\Delta \psi_f$=-0.9$\psi_f$ (f) $\Delta \psi_f$=9$\psi_f$ (g) $\Delta R_s$=-0.9$R_s$, $\Delta L_s$=-0.9$L_s$, $\Delta \psi_f$=-0.9$\psi_f$ (h) $\Delta R_s$=9$R_s$, $\Delta L_s$ =9$L_s$, $\Delta \psi_f$=9$\psi_f$.



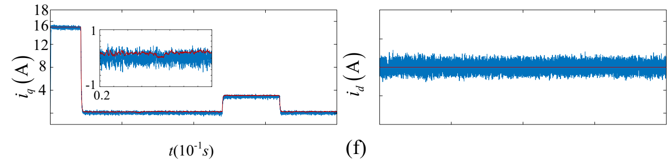




Fig. 8. Simulation performance of proposed DBPC method under different parameter mismatch condition. (a) $\Delta L_s$=-0.9$L_s$ (b) $\Delta L_s$=9$L_s$ (c) $\Delta \psi_f$=-0.9$\psi_f$ (d) $\Delta \psi_f$=9$\psi_f$ (e) $\Delta R_s$=-0.9$R_s$,$\Delta L_s$=-0.9$L_s$,$\Delta \psi_f$=-0.9$\psi_f$ (f)$\Delta R_s$=9$R_s$, $\Delta L_s$ =9$L_s$, $\Delta \psi_f$=9$\psi_f$.

Fig. 9 shows the compensation voltages of the proposed DBPC method under different parameter mismatch conditions, respectively. The blue line is the $d$-axis current compensation voltage while the red line is the $q$-axis compensation voltage. Due to little effect of resistance variation $\Delta R_s$ on current responses, parameter mismatch of resistance $R_s$ is not studied in following analysis. As shown in Fig. 9, compensation voltages are produced to eliminate the influence of parameters variation when parameter mismatch exists after 0.03 s. Furthermore, compensation voltage is related to load torque, because variation of load torque causes current response to change, which further affects the compensation voltage in parameter mismatch condition.

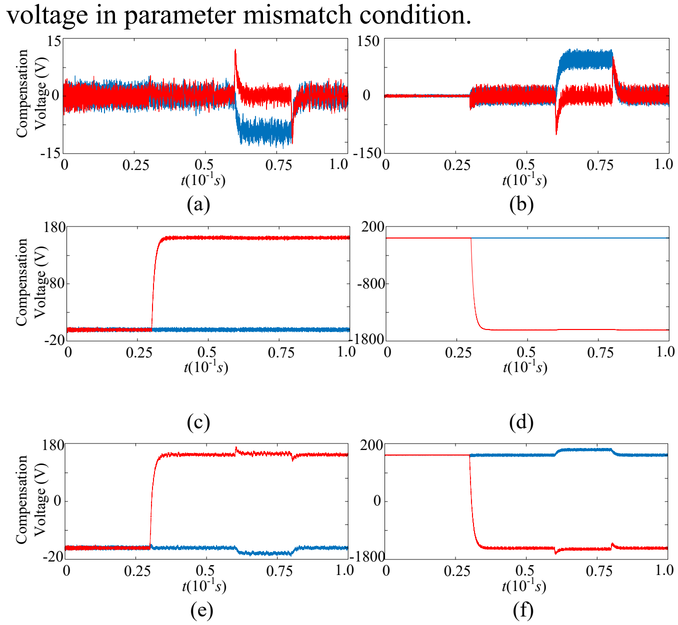
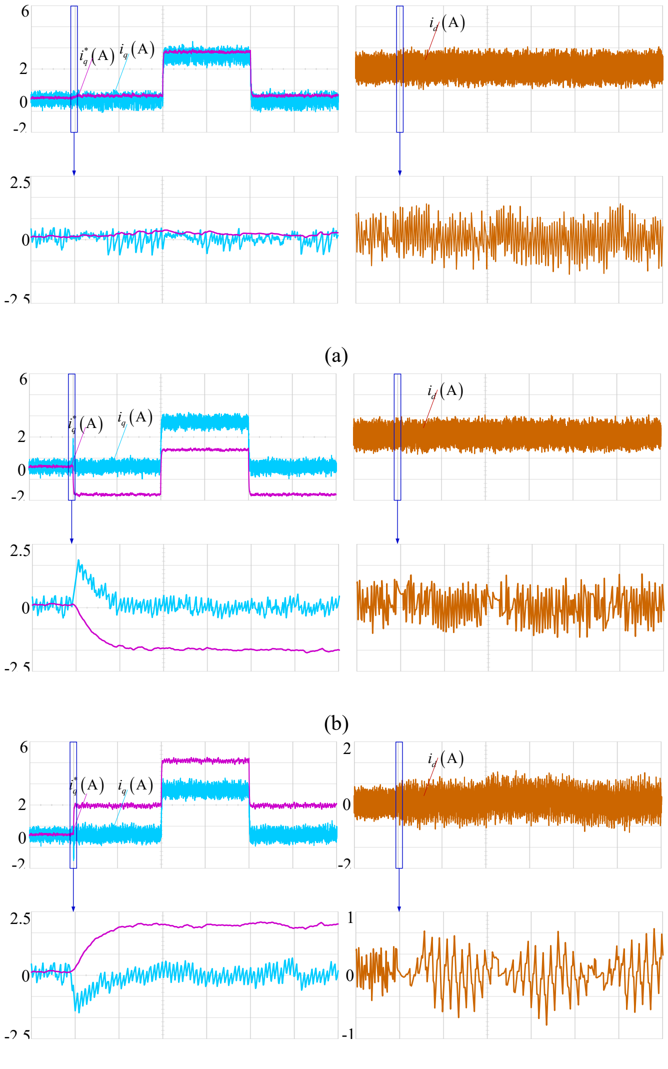



Fig. 9. Compensation voltage of proposed DBPC method under different parameter mismatch conditions. (a) $\Delta L_s$=-0.9$L_s$ (b) $\Delta L_s$=9$L_s$ (c) $\Delta \psi_f$=-0.9$\psi_f$ (d) $\Delta \psi_f$=9$\psi_f$ (e) $\Delta R_s$=-0.9$R_s$, $\Delta L_s$ =-0.9$L_s$, $\Delta \psi_f$=-0.9$\psi_f$ (f)$\Delta R_s$=9$R_s$, $\Delta L_s$ =9$L_s$, $\Delta \psi_f$=9$\psi_f$.

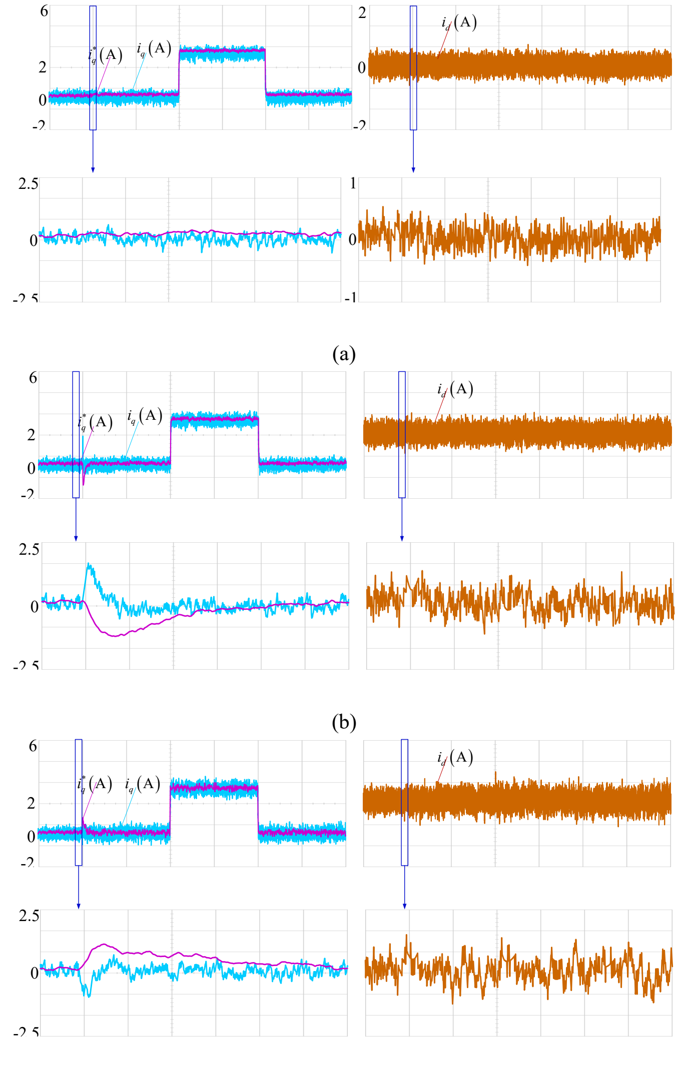
The control system is implemented on a dSPACE 1401 real-time platform with Control Desk. The Rapid Control Prototype (RCP) is realized based on the dSPACE. The experimentally measured currents are available via analog to digital converter board. The experimental results are shown in Figs. 10-13. Parameter mismatch is set at 0.05 s. At the time instants $t_1$ =0.1 s and $t_2$ =0.15 s, step changes in the load torque from 0 to 20 Nm and from 20 Nm to 0 are applied to the system, respectively.




(c)

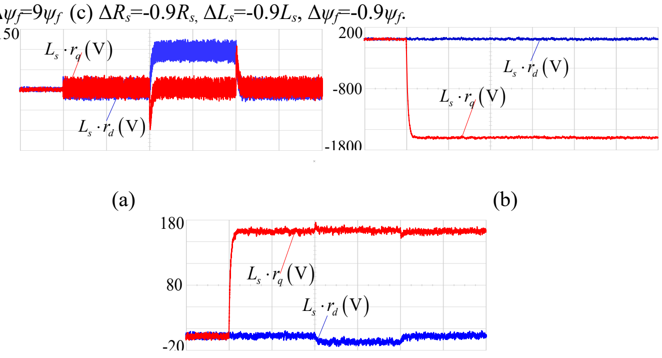
Fig. 10. Experimental results of the conventional DBPC method. (a) $\Delta L_s$=9$L_s$ (b)$\Delta\psi_f$=9$\psi_f$ (c) $\Delta R_s$=-0.9$R_s$, $\Delta L_s$=-0.9$L_s$, $\Delta\psi_f$=-0.9$\psi_f$.

As shown in Fig. 10, reference currents have a step change and the ripple of currents response becomes larger when parameter mismatch exists at 0.05 s. In particular, the variation of permanent flux produces significant effect on currents response, which results in $q$-axis current error of 1.8 A. Furthermore, when three parameters variation exists simultaneously, $q$-axis current of conventional DBPC fails to track the reference value accurately with an error about 2 A, while $d$-axis current response produces large fluctuations of 1.6 A thanks to the coupling in motor model.




(c)

Fig.11 Experimental results of the proposed DBPC method. (a) $\Delta L_s$=9$L_s$ (b) $\Delta\psi_f$=9$\psi_f$ (c) $\Delta R_s$=-0.9$R_s$, $\Delta L_s$=-0.9$L_s$, $\Delta\psi_f$=-0.9$\psi_f$.




(c)

Fig. 12. Compensation voltage of the proposed DBPC method. (a) $\Delta L_s$=9$L_s$ (b) $\Delta\psi_f$=9$\psi_f$ (c) $\Delta R_s$=-0.9$R_s$, $\Delta L_s$=-0.9$L_s$, $\Delta\psi_f$=-0.9$\psi_f$.

Compared with the conventional method, the currents tracking error is smaller except for the acceptable ripple at the original instant of parameter mismatch. From Figs. 11 (b) and (c), the transient ripples of current response are about -1.2 A and 0.9A, respectively, which converge to stable reference values quickly. Meanwhile the compensation voltages are given in Fig. 12, which is used to eliminate the influence of parameter mismatch. Different parameter mismatch conditions cause different perturbance, which results in different voltage compensation according to (6) and (13).

Figs. 13 illustrates the control performance of both methods when three parameters mismatch occurs during operation, in which step changes in the reference speed from 360 to 365 rpm and 365 to 360 rpm have been applied at 0.1 s and 0.15 s, respectively. Under the conventional predictive control method, the existing parameters mismatch affects the current response and produces permanent $q$-axis current tracking error of 4.8 A and $d$-axis current fluctuation of 1.1 A, which deteriorates the control performance. Compared with the conventional method, the proposed DBPC method produces obvious current errors only at 0.05 s, 0.1 s and 0.15 s, which converges to stable reference quickly. Therefore, the improved method ensures the control performance is not subject to parameter variation. The ripples and errors of the current response can be suppressed effectively.

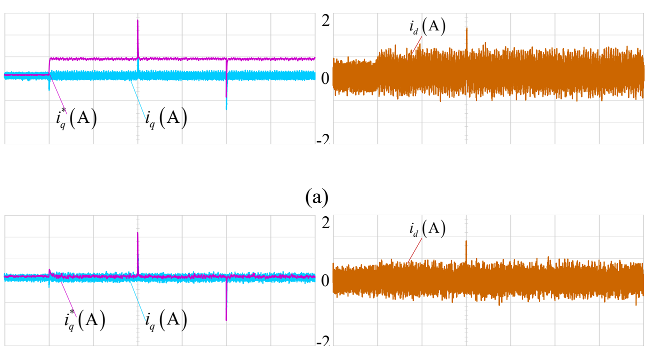



Fig. 13. Experimental results in reference speed step-change conditions. (a) conventional DBPC method, and (b) the proposed DBPC method.

## VI. Conclusion

In this paper, an improved deadbeat predictive current controller was proposed and experimentally applied to a PMSM drive system to enhance the robust performance of the drive system. In order to evaluate the relationship between the current response and mismatch of three parameters, parameter sensitivity analysis was discussed first. Then an SMDO based on stator current and lumped disturbance was designed to simultaneously estimate the future current value and lumped disturbance caused by the parameter mismatch, which can effectively eliminate the influence of parameters mismatch. Furthermore, considering the influence of the calculation and sampling delay, currents are estimated by discrete expression of composite SMDO and used to replace the sampled values to compensate one-step delay.

Compared with related references, which adopt high-order sliding mode structure or new reaching law-based method, this design method based on the state space structure of the control plant reduced the calculation complexity of the observation value for the nonlinear model. Additionally, the second-order expansion of the rotor speed was applied to improve the current reference accuracy. Simulation and experimental results confirmed the performance of the proposed control strategy.

In future work, the composite observation system will be further studied, which comprehensively considers the variation of mechanical parameters. Meanwhile, the balance of calculation burden and control performance will be considered.

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