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Discrete-time Current Regulator for AC Machine Drives

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Abstract—In a practical alternating current (AC) machine drive system, a direct discrete-time current regulator model based on a zero order hold (ZOH) equivalent discrete-time machine model should be the first choice to be established compared with the discrete-time current regulator model based on continuous-time machine model when designing a digital current regulator. However, it is complicated to establish a direct discrete-time current regulator for asymmetric machine¹ model due to the involvement of matrix exponential related to unequal machine inductance parameters. In order to overcome this issue, a simple direct discrete-time current regulator based on flux model is proposed, where a novel AC machine model is developed by employing a new feed-forward voltage compensation. In addition, a direct discrete-time flux observer with low parameter sensitivity is proposed to compensate the inherent one-step delay in practical systems. Simulation results and theoretical analysis have shown that the proposed discrete-time current regulator and flux observer have good parameter robustness and stability. Also, the experimental verification is performed on a AC machine drive test rig.

Index Terms—Alternating current (AC) machine, Parameter robustness, Current regulator, Discrete-time model.

NOMENCLATURE

U_{dq}	dq-axis stator voltage
U'_{dq}	dq-axis distorted voltage caused by inverter nonlinearity
U_{dq}^{ref}	Inverter output dq-axis reference voltage
i_{dq}	dq-axis stator current
i_{dq}^{ref}	Reference dq-axis stator current
λ_{dq}	dq-axis flux linkage
ω_e	Electrical angular speed
θ_e	Electrical angle
T_s	System sampling and control period
L_d	Actual d-axis stator inductance
L_q	Actual q-axis stator inductance
R_s	Actual stator resistance
ψ_m	Actual rotor flux linkage
$\frac{L_d}{L_q}$	Nominal d-axis inductance in controller
$\frac{L_q}{L_d}$	Nominal q-axis inductance in controller
$\frac{R_s}{L_d}$	Nominal stator resistance in controller
$\frac{\psi_m}{L_d}$	Nominal rotor flux linkage in controller

I. INTRODUCTION

A. Literature Review

IN the field of alternating current (AC) machine control, many current regulators have been developed. Generally, based on Clarke and Park transform [1], the three phase AC stator current of machines can be transformed to dc current at steady state, namely, excitation and torque current, which can be easily controlled in the linear system [2]. The design of current regulators can be carried out based on AC machine models. Based on linear system control theory, the relationship between output current i_{dq} and input reference voltage U_{dq}^{ref} can be considered as a transfer function of the plant, and the transfer function can own low-pass filter characteristic. Based on the transfer function of the plant, a state-feedback decoupling proportional-integral (PI) current regulator was developed [3]. In this PI structure, low frequency disturbances can be suppressed. Adding an imaginary part in the pole of the plant, a complex vector PI current regulator is proposed [4], which can exhibits little dependency on the synchronous frequency compared with the state-feedback decoupling PI current regulator. To enhance the bandwidth in the transfer function from disturbance to steady-state error, active resistance term is added as a feed-forward disturbance by adjusting the plant model [5].

Apart from the above mentioned model-based PI current regulators, model-based P current regulators are also applied in AC machine control. The typical representation is dead beat predictive current control (DPCC) [6], [7]. Based on the deadbeat principle, the dynamic response is quite fast by using accurate plant parameters. However, the low frequency disturbances can not be suppressed based on the model-based P current regulator, and the disturbances caused by resistance, rotor flux linkage and inductance parameter mismatch can induce the steady-state error [8], [9]. In order to solve this issue, an observer based on nonlinear sliding mode control is designed [10]. Besides that, based on linear control theory, a disturbance observer and extended state observer have been developed [11]–[13]. However, high frequency noises will negatively affect the observer performance with the increase of the disturbance observer bandwidth. There is trade-off between disturbance observer bandwidth and measurement noises suppression.

Therefore, the model-based PI current regulators with an ability to suppress the low frequency disturbances are also getting widely attention in industrial applications. As the rapid development of digital systems, digital implementation of current regulators occupies in most of applications compared with current regulators based on analog continuous systems. Thus, current regulators designed in continuous time domain

¹The asymmetric machine denotes a AC salient machine in this paper, whereas the symmetric machine denotes a AC non-salient machine.

need to be transformed to discrete-time current regulators by employing approximation methods such as Euler, Trapezoidal, and Tustin methods [14], [15]. In order to achieve the high current performance, especially in the condition that the large current bandwidth that is relative to sampling frequency and fundamental excitation frequencies that are relative to sampling frequency, a direct discrete-time current regulator is proposed [16], [17]. The brief procedure is that an exact zero order hold (ZOH) equivalent discrete-time plant model can be established directly and internal model control (IMC) principle can be employed to constitute the first-order closed-loop system to acquire the current regulator model. To enhance the disturbance rejection of discrete-time current regulators, literature in [18] has developed a simple two-degrees-of-freedom (2-DOF) digital current controller, which can optimize the current settling time and overshoot. Apart from employing the active resistance feedback term, a digital current controller by adding active inductance feedback is developed to enhance the disturbance rejection but stability is not relatively good [19].

Considering computational and modulation delays model also can significantly enhance the current performance [16], [20]. A modified discrete-time current regulator with computational and modulation delay model is proposed [18], [21]. In this case, the inner active resistance feedback delay issue can be solved. Another method for solving this issue is to employ a current prediction scheme, but the AC machine parameters should be accurate. [20]. However, the above discrete-time current regulator only can be applied in symmetric AC machines [16], [18], [21].

In order to develop a discrete-time current regulator applied in different types of AC machines considering computational and modulation delays, Hinkkanen et al. derived an exact ZOH equivalent discrete salient AC machines based on a matrix exponential e^{AT_s} (A is a matrix related to AC machine parameters) and proposed a 2-DOF PI current regulator based on an analytical discrete-time pole-location design method [22]. However, the matrix exponential e^{AT_s} is much more complicated and the related coefficient can only be computed offline first. In addition, a discrete-time current regulator based on flux with the assumption that $\frac{\sin \omega_e T_s}{\omega_e} \approx T_s$ is established, and the current regulator complexity is relatively reduced compared with the current regulator in [22], but the matrix exponential e^{AT_s} still needs to be utilized [23]. To avoid the involvement of e^{AT_s} in the discrete-time current regulator model, a discrete-time salient machines model based on flux without resistance term is proposed [24], [25].

B. Motivation and Innovation

In order to reduce the complexity of the above mentioned discrete-time current regulators (see e.g., [22], [23]), this paper proposes a direct discrete-time current regulator without using the matrix exponential e^{AT_s} . The contribution of this work can be listed as follows.

1) Compared with the discrete-time current regulators in [22], [23] that need to employ Putzer algorithm to simplify the matrix exponential e^{AT_s} , a direct discrete-time current

regulator based on flux model is proposed. In the proposed current regulator, a novel AC machine model can be established due to the involvement of a new feed-forward voltage compensation, which means that there is no need to obtain the matrix exponential e^{AT_s} . Apart from that, in order to illustrate the superiority of the proposed current regulator, the discrete-time current regulators without resistance term in [24], [25], state-feedback decoupling discrete-time current regulator and DPCC with one-step delay compensation are also compared with the proposed current regulator.

2) A direct discrete-time flux observer is proposed to compensate the one-step delay issue in practical application, and the current bandwidth of the proposed current regulator can be increased further (equal to the DPCC dynamic response). Apart from that, the convergence and criterion for the related coefficients of the proposed flux observer has been presented in this paper.

3) Parameter sensitivity of current regulators using Bode plot has been analyzed in detail. In addition, the stability analyzed by discrete-time pole-zero location is presented. In this part, literature in [24], [25] and DPCC will be compared.

C. Paper Organization

Since an AC PMSM is considered as a control plant in this paper the content of this paper is organized as follows: the discrete-time PMSM model and direct discrete-time current regulators are introduced in Section II. The proposed current regulator, parameter sensitivity, stability and proposed direct discrete-time flux observer are presented in Section III. Experiment results of the proposed current regulators are carried out in Section IV. Finally, the conclusions are derived in Section V.

II. DISCRETE-TIME CURRENT REGULATOR

In this section, the analysis on the complexity of discrete-time PMSM model with e^{AT_s} in [22], [23] will be carried out. In addition, the discrete-time PMSM model based on Flux without R_s in [24], [25] will be presented herein.

A. Complexity of Discrete-time PMSM Model with e^{AT_s}

The PMSM flux matrix equation can be described as:

$$\dot{\lambda}_{dq} = A\lambda_{dq} + U_{dq} + C \quad (1)$$

where $A = \begin{bmatrix} -\frac{R_s}{L_d} & \omega_e \\ -\omega_e & -\frac{R_s}{L_q} \end{bmatrix}$, $C = \begin{bmatrix} -U'_d \\ -U'_q - \psi_m \omega_e \end{bmatrix}$, $U_{dq} = [U_d, U_q]^T$, $\lambda_{dq} = [\lambda_d, \lambda_q]^T$, $U'_{dq} = [U'_d, U'_q]^T$. The superscript “.” means the derivative with respect to time. Considering the matrix C as disturbances, based on (1), the continuous-time current matrix equation can be presented as:

$$\lambda_{dq}(t) = e^{At} \lambda_{dq}(t_0) - (I - e^{At})A^{-1}U_{dq}(t) \quad (2)$$

where t and t_0 mean the current time and initial time; I is an identity matrix. Since there exists a one-step delay issue in practical systems and U_{dq} is delayed for one sampling period.

Based on the above equation, the discrete-time PMSM model considering the computational time can be expressed as [23]:

$$\lambda_{dq}(k+1) = e^{AT_s} \lambda_{dq}(k) - (I - e^{AT_s}) A^{-1} U_{dq}(k-1) \quad (3)$$

Based on (3), the discrete-time PMSM model G_{p0} in z domain can be expressed as follows:

$$G_{p0}(z) = \frac{\lambda_{dq}(z)}{U_{dq}(z)} = \frac{z^{-1} A^{-1} (e^{AT_s} - I)}{z - e^{AT_s}} \quad (4)$$

However, it is difficult to establish the current digital based on IMC because the calculation is relatively larger. The reason behind this is that the matrix exponential e^{AT_s} needs to be obtained based on Putzer algorithm [22]:

$$e^{AT_s} = e^{-cT_s} \begin{bmatrix} \cosh(aT_s) - b \frac{\sinh(aT_s)}{a} & \omega_e \frac{\sinh(aT_s)}{a} \\ -\omega_e \frac{\sinh(aT_s)}{a} & \cosh(aT_s) + b \frac{\sinh(aT_s)}{a} \end{bmatrix} \quad (5)$$

where $a = \sqrt{b^2 - \omega_e^2}$, $b = \frac{R_s}{2} (\frac{1}{L_d} - \frac{1}{L_q})$ and $c = \frac{R_s}{2} (\frac{1}{L_d} + \frac{1}{L_q})$. The functions $\sinh$ and $\cosh$ are sine and cosine hyperbolic function, respectively. Thus, it can be seen that this type of discrete-time PMSM model is quite complex, which is difficult to be widely applied in industrial applications without employing the approximation of the hyperbolic function or offline self-commissioning.

B. Discrete-time Current Regulator Based on Flux without R_s

In order to simplify the PMSM model, another discrete-time PMSM model based on flux model without R_s is developed in literature [24]. Using this equation, the matrix exponential e^{AT_s} can be avoided but the R_s term is considered as disturbances. A straightforward solution is to compensate $R_s i_{dq}$ at the current regulator output. Similarly, the terms U'_{dq} and $\psi_m \omega_e$ are considered as disturbances. In this case, the discrete-time PMSM flux model can be expressed as:

$$\lambda_{dq}(k+1) = (e^{-j\omega_e T_s}) \lambda_{dq}(k) + T_s e^{-2j\omega_e T_s} U_{dq}(k-1) \quad (6)$$

After that, the voltage to flux transfer function $G_{p2}(z)$ in z domain can be expressed as:

$$G_{p2}(z) = \frac{\lambda_{dq}(z)}{U_{dq}(z)} = \frac{T_s}{ze^{j\omega_e T_s} (ze^{j\omega_e T_s} - 1)} \quad (7)$$

Based on (7), IMC principle can be employed to obtain the current regulator model based on the first order closed-loop system. The flux λ_{dq} can be obtain based on the following equation:

$$\begin{aligned} \lambda_{dq}(z) G_l^{-1}(z) &= i_{dq}(z) \\ G_l^{-1}(z) &= \begin{bmatrix} \frac{1}{L_d} & 0 \\ 0 & \frac{1}{L_q} \end{bmatrix} \end{aligned} \quad (8)$$

where G_l means the relationship between current i_{dq} and flux λ_{dq} . The transfer function $G_{c2}(z)$ of the current regulator in z domain can be presented as:

$$\begin{aligned} G_{c2}(z) &= G_{l2}(z) G_{p2}^{-1}(z) L(z) \gamma(z) \\ L(z) &= \frac{z}{z-1} 2\pi f_{bw} T_s \\ G_{l2}(z) &= \begin{bmatrix} \frac{L_d}{0} & 0 \\ 0 & \frac{L_q}{0} \end{bmatrix} \end{aligned} \quad (9)$$

Fig. 1. The block diagram of direct discrete-time current regulator

where $\gamma(z)$ and $L(z)$ are an adjustable model and discrete-time integrator; $G_{l2}(z)$ means inductance parameters in the current regulator. $f_{bw} \in \mathbb{R}_+$ the desired bandwidth of the controller. Generally, the term $\gamma(z)$ needs to be set to $\frac{1}{z^2}$ to make a causal system. The current regulator can be expressed as:

$$G_{c2}(z) = G_{l2}(z) e^{j\omega_e T_s} (e^{j\omega_e T_s} - z^{-1}) \frac{z}{z-1} 2\pi f_{bw} T_s \quad (10)$$

Based on (10), the output of the current regulator can be obtained. The block of the discrete-time current regulator without R_s in z domain is shown in Fig.1.

III. PROPOSED DISCRETE-TIME CURRENT REGULATOR

In this section, a novel discrete-time current regulator is proposed to reduce the complexity of e^{AT_s} . In order to analyze the stability and parameter sensitivity of the proposed current regulator, pole-zero location and Bode plots are carried out in the subsection-B. In order to compensate the control delay in practical systems, a discrete-time flux observer is proposed herein. Finally, the digital implementation of the proposed current regulator is presented.

A. Proposed Discrete-time Current Regulator Based on Flux Model

To avoid the involvement of the matrix exponential e^{AT_s} in current regulator, a simple discrete-time PMSM model combined with inverter model on synchronous reference frame is established herein. In addition, the proposed direct discrete-time current regulator is presented in detail.

1) *Delay Compensation from Stationary to Synchronous Reference Frame:* In order to compensate the delay caused by Ipark transform from synchronous reference frame to stationary reference frame [26], an effective compensation method is employed herein, which is shown as:

$$\begin{aligned} K &= k_1 e^{j\theta_e} \\ k_1 &= \frac{2}{\omega_e T_s} \sin\left(\frac{\omega_e T_s}{2}\right) e^{j1.5T_s \omega_e} \end{aligned} \quad (11)$$

where K is compensated Park transform and k_1 is compensated coefficient.

2) *Proposed Discrete-time PMSM Model Based on Flux Model:* Adding the term $R_s i_q (\frac{L_d}{L_q} - 1)j$ in the feed-forward voltage compensation, a simple discrete-time PMSM model can be expressed as:

$$\dot{\lambda}_{dq} = A_1 \lambda_{dq} + \frac{U_{dq}}{k_1} + D \quad (12)$$

Fig. 2. The block diagram of the proposed overall control loops

where $A_1 = \begin{bmatrix} -\frac{R_s}{L_d} & \omega_e \\ -\omega_e & -\frac{R_s}{L_d} \end{bmatrix}$, $D = \begin{bmatrix} -U'_d \\ -U'_q - \psi_m \omega_e j + R_s i_q (\frac{L_q}{L_d} - 1)j \end{bmatrix}$. From (12), it can be seen that the previous A is changed to A_1 , where the value of the diagonal in A_1 is the same. In this case, the proposed discrete-time PMSM model can be expressed as:

$$\dot{\lambda}_{dq} = (-\frac{R_s}{L_d} - j\omega_e)\lambda_{dq} + \frac{U_{dq}}{k_1} + D \quad (13)$$

Considering the matrix D as disturbances, based on (13), the discrete-time flux at the $(k+1)$ th instant can be solved as:

$$\lambda_{dq}(k+1) = (1 - e^{(-\frac{R_s}{L_d} - j\omega_e)T_s}) \frac{U_{dq}(k-1)}{(\frac{R_s}{L_d} + j\omega_e)k_1} + e^{(-\frac{R_s}{L_d} - j\omega_e)T_s} \lambda_{dq}(k) \quad (14)$$

Transforming (14) from discrete-time domain to z domain, the voltage to flux transfer function $G_{n3}(z)$ can be expressed as:

$$G_{p3}(z) = \frac{\lambda_{dq}(z)}{U_{dq}(z)} = \frac{(1 - e^{(-\frac{R_s}{L_d} - j\omega_e)T_s})}{z(z - e^{(-\frac{R_s}{L_d} - j\omega_e)T_s})(\frac{R_s}{L_d} + j\omega_e)k_1} \quad (15)$$

3) *Proposed Direct Discrete-time Current Regulator Based on Flux Model*: Since parameters in the current regulator may not be equal to the parameters in the plant, the proposed current regulator based on the IMC principle can be shown as:

$$G_{c3}(z) = G_{l2}(z) \frac{(e^{j\omega_e T_s} - z^{-1} e^{-\frac{\overline{R_s}}{L_d} T_s})(\frac{\overline{R_s}}{L_d} + j\omega_e)k_1}{e^{j\omega_e T_s} - e^{-\frac{\overline{R_s}}{L_d} T_s}} L(z) \quad (16)$$

Because the disturbance D needs to be transformed to stationary reference frame, k_1 also should be added herein. Overall, the block diagram of the proposed direct discrete-time current regulator is presented in Fig.2. After obtaining the reference voltage, based on Ipark transform, the reference voltage on stationary reference frame through a space vector pulse width modulation (SVPWM) can be applied in the AC machine [27]. The block diagram of the proposed overall control loops is shown in Fig.3.

B. Stability and Parameter Sensitivity of Discrete-time Current Regulator

Based on the closed-loop structure, the current steady state error e_{dq} are affected by reference current, disturbances caused by parameter mismatch and measurement noises caused by

Fig. 3. The block diagram of the proposed overall control loops

Fig. 4. The block diagram of three current regulators. (a)Method4; (b) Method2; (c) Method3

sensors. For example, AC machine inductance mismatch can affect the overall system pole-zero cancellation, which may get systems unstable. In a disturbances path, rotor flux linkage parameter mismatch not only can affect the pole location of plant but also affect the current steady state error. It should be noted that since the transfer function of the measurement noises to e_{dq} (also called complementary sensitivity transfer function) is complementary with the transfer function of the reference current to e_{dq} (also called sensitivity transfer function) [2], there is no need to analyze both. It is a trade-off without using a 2-DOF controller. In this case, the measurement noises will not be analyzed in this paper.

In this paper, in order to illustrate the superiority of the proposed method, the three current regulators will be analyzed, namely DPCC [28] (Method4) which is actually a model-based P current controller, discrete-time current regulator without R_s

[24] (Method2) and proposed discrete-time current regulator (Method3), respectively. To simplify models, the distorted voltage caused by nonlinearity is assumed to be compensated well in the three methods. The block diagram of the three methods can be shown in Fig.4.

Based on (7), (9), (15) and (16), the transfer function of the open loop system in Method2 and Method3 can be expressed as:

$$\begin{aligned} G_{ol2}(z) &= G_{c2}(z)G_{p2}(z)G_l^{-1}(z) \\ G_{ol3}(z) &= G_{c3}(z)G_{p3}(z)G_l^{-1}(z) \end{aligned} \quad (17)$$

where subscript “ol” means the open-loop of transfer functions. In Method4, the transfer function of the plant and open loop system can be respectively expressed as:

$$\begin{aligned} G_{p1}(z) &= \frac{\lambda_{dq}(z)}{U_{dq}(z)} = \frac{T_s}{z(z-1)} \\ G_{ol1}(z) &= \frac{1}{z(z-1)} 2\pi f_{bw} T_s G_{l2}(z) G_l^{-1}(z) \end{aligned} \quad (18)$$

Assuming that there exists parameters mismatch, the feed-forward disturbances in the controller cannot cancel the disturbance in the plant, and the disturbances are defined as d_1, d_2, d_3 for Method 4, Method2, and Method3, respectively. In Fig.4, the current error $e_{dq1}, e_{dq2}, e_{dq3}$ of Method 4, Method2, and Method3 can be respectively expressed as:

$$\begin{aligned} e_{dq1}(z) &= i_{dq}^{ref}(z) \frac{\overbrace{1}^{G_{err1}(z)}}{1 + G_{ol1}(z)} + d_1(z) \frac{\overbrace{-G_{p1}(z)G_l^{-1}(z)}^{G_{dis1}(z)}}{1 + G_{ol1}(z)} \\ e_{dq2}(z) &= i_{dq}^{ref}(z) \frac{\overbrace{1}^{G_{err2}(z)}}{1 + G_{ol2}(z)} + d_2(z) \frac{\overbrace{-G_{p2}(z)G_l^{-1}(z)}^{G_{dis2}(z)}}{1 + G_{ol2}(z)} \\ e_{dq3}(z) &= i_{dq}^{ref}(z) \frac{\overbrace{1}^{G_{err3}(z)}}{1 + G_{ol3}(z)} + d_3(z) \frac{\overbrace{-G_{p3}(z)G_l^{-1}(z)}^{G_{dis3}(z)}}{1 + G_{ol3}(z)} \end{aligned} \quad (19)$$

where $G_{err1}(z), G_{err2}(z), G_{err3}(z)$ denote the reference current to current error transfer functions of Method 4, Method2, and Method3, respectively; $G_{dis1}(z), G_{dis2}(z), G_{dis3}(z)$ denote the disturbances to current error transfer functions of Method 4, Method2, and Method3, respectively.

1) *Stability Comparison:* Without parameters mismatch for the three methods, the transfer function of $G_{err1}(z), G_{err2}(z), G_{err3}(z)$ are the same due to the zero-pole cancellation between the regulator and plant, but $G_{dis1}(z), G_{dis2}(z), G_{dis3}(z)$ are totally different from (19). The pole-zero location is shown in Fig.5. From Fig.5(b), it can be seen that one of poles is out of the unit circle in Method2, which means that the current regulator goes unstable as long as the existence of disturbances. In Fig.5(a) and Fig.5(c), it can be seen that all of poles are located in the unit circle, which means that the current regulators of Method4 and Method3 are stable.

2) *Parameter Sensitivity Comparison:* When the AC machine parameter mismatch occurs ($\bar{L}_d, \bar{L}_q, \bar{R}_s, \bar{\psi}_m$ of the current regulators G_{c1}, G_{c2}, G_{c3} are unequal to L_d, L_q, R_s, ψ_m of

Fig. 5. The zero-pole location of $G_{dis1}(z), G_{dis2}(z), G_{dis3}(z)$. (a) Method4; (b) Method2; (c) Method3. (The condition is $\bar{L}_d = L_d, \bar{L}_q = L_q, \bar{\psi}_m = \psi_m, \bar{R} = R_s, f_{bw} = 100\text{Hz}, \omega_e = 314\text{rad/s}$ and $T_s = 200\mu\text{s}$)

the actual plants G_{p1}, G_{p2}, G_{p3} , the performance of the current regulators is deteriorated. In this subsection, the Bode plot analysis of the parameter sensitivity based on (19) for the three methods under the parameter variation ($\bar{L}_d = 0.3L_d \sim 3L_d, \bar{L}_q = 0.3L_q \sim 3L_q, \bar{R} = 0.3R_s \sim 3R_s$) will be carried out.

First, i_{dq}^{ref} effect on the current error is analyzed. Bode plot of $G_{err1}(z), G_{err3}(z)$ for the two methods² is shown in Fig.6. It can be seen that the lowest magnitude of the current error is about -38dB in Method3 while the lowest magnitude of Method4 is about -30dB at low frequency i_{dq}^{ref} when $\bar{L}_q = 0.3L_q$ and $\bar{L}_d = 0.3L_d$, which can testify that the proposed method has the lowest current error at low frequency i_{dq}^{ref} . The magnitude of $G_{err3}(z)$ can keep lowest values (dB) under different resistance mismatch.

Second, the analysis of the disturbances d_1, d_2, d_3 effect on the current error is presented. The Bode plot of $G_{dis1}(z), G_{dis3}(z)$ for the two methods² is shown in Fig.7. In Method4, it can be seen that, it can not suppress the

²From the zero-pole location plot in Fig.5, it can be seen that Method2 is unstable and cannot be applied herein. In this case, this paper will only analyze the parameter sensitivity using Bode plot for Method4 and Method3.

Fig. 6. The Bode plot of $G_{err1}(z), G_{err3}(z)$. (a) Method4; (b) Method3. (The condition is $\bar{L}_d = 0.3L_d \sim 3L_d, \bar{L}_q = 0.3L_q \sim 3L_q, \bar{\psi}_m = \psi_m, \bar{R} = 0.3R_s \sim 3R_s, f_{bw} = 100\text{Hz}, \omega_e = 314\text{rad/s}$ and $T_s = 200\mu\text{s}$)

disturbances in low frequency (especially for dc disturbances). This is a reason why the resistance and rotor flux linkage mismatch can induce the dc offset in DPCC. On the contrary, Method3 has the best suppression ability for the disturbance caused by low frequency under different parameter mismatch. When increasing the inductance parameters in controller, the suppression ability is increased in (b) of Fig.7. It should be noted that the active resistance term and disturbance observer also can be employed to enhance the disturbance rejection of the proposed current controller further, but they are not considered in this paper since the previous work for the analysis of the disturbance rejection has been carried out in [5], [10]–[13], [18], [21], [22].

Overall, based on the above analysis, the proposed discrete-time current regulator has the lower parameter sensitivity under different parameter mismatch conditions compared with DPCC.

C. Proposed Direct Discrete-time Flux Observer

In (9), it can be seen that the adjustable model $\gamma(z)$ is set to $\frac{1}{z^2}$ to achieve a casual system [29]. However, it can induce one-step delay in systems, and the performance of current regulators without one-step delay compensation can be negatively affected. Based on the last instant known AC machine stator voltage, this paper proposes a simple discrete-time flux observer. The detail procedure can be presented as follows.

1) Defining the lumped disturbances caused by resistance voltage and voltage deviation caused by AC machine parameter mismatch as the controller G_{c4} output r_{dq} , and the external disturbances $U_{dq} - U'_{dq} - \bar{\psi}_m \omega_e - j\omega_e G_{12} i_{dq}$ as d_{2dq} , the predicted flux equation of the plant G_{p4} can be simply expressed as:

$$\dot{\lambda}_{dq} = d_{2dq} + r_{dq} \quad (20)$$

Fig. 7. The Bode plot of $G_{dis1}(z), G_{dis3}(z)$. (a) Method4; (b) Method3. (The condition is $\bar{L}_d = 0.3L_d \sim 3L_d, \bar{L}_q = 0.3L_q \sim 3L_q, \bar{\psi}_m = \psi_m, \bar{R} = 0.3R_s \sim 3R_s, f_{bw} = 100\text{Hz}, \omega_e = 314\text{rad/s}$ and $T_s = 200\mu\text{s}$)

Fig. 8. The block diagram of discrete-time flux observer

Based on the above equation, the flux equation in z domain can be shown as:

$$G_{p4}(z) = \frac{\hat{\lambda}_{dq}(z)}{d_{2dq}(z) + r_{dq}(z)} = \frac{1}{z-1} T_s \quad (21)$$

2) Since the transfer function of the plant is integral, the second order closed-loop should be established to eliminate the steady state error of the observer caused by disturbances d_{2dq} generally. Thus, the block diagram of the proposed discrete-time flux observer can be shown in Fig.8 and the proposed discrete-time controller G_{c4} can be designed as:

$$\begin{aligned} G_{c4}(z) &= G_{p4}^{-1}(z) L_2(z) \\ L_2(z) &= \frac{k_2 z + k_3}{(z-1)^2} \end{aligned} \quad (22)$$

where $k_2, k_3 \in \mathbb{R}_+$ the proposed flux observer coefficients, which can be theoretically designed as follows.

$$\begin{aligned} w_{ob} &= 2\pi f_{ob} T_s \\ k_2 &= 2w_{ob} \\ k_3 &= w_{ob}^2 - 2w_{ob} \end{aligned} \quad (23)$$

where $w_{ob} \in \mathbb{R}_+$ the observer undamped natural frequency; $f_{ob} \in \mathbb{R}_+$ the related to the observer bandwidth.

3) The convergence of the proposed flux observer can be shown as follows: The transfer function of the open loop

system G_{ol4} and flux error e_{dq4} can be respectively expressed as:

$$G_{ol4}(z) = G_{c4}(z)G_{p4}(z) = \frac{k_2 z + k_3}{(z-1)^2}$$

$$e_{dq4}(z) = i_{dq}(z)G_{l2}(z) \frac{1}{1+G_{ol4}(z)} + d_{2dq}(z) \frac{G_{dis4}(z)}{1+G_{ol4}(z)} \quad (24)$$

From this equation, it can be seen that $G_{err4}(z)$ possesses the high pass filter characteristic while $G_{dis4}(z)$ owns the bandwidth pass filter characteristic, which means that e_{dq4} can get reduced to zero under the system steady state. In this case, $\hat{i}_{dq}$ can be convergent to $i_{dq}G_{l2}$.

In terms of the analysis of the parameter sensitivity, it can be seen that the initial resistance parameter of controller is not involved in (24) and cannot affect the proposed observer estimation performance. Although the initial inductance parameter is used, it will be canceled to obtain the next instant current $\hat{i}_{dq}$. It should be noted that only the initial rotor flux linkage parameter is utilized but it can be taken as disturbances d_{2dq} and the disturbances caused by the rotor flux linkage parameter can be suppressed by G_{dis4} . Overall, it can be seen that the parameter sensitivity is low for the proposed discrete-time flux observer.

4) The criterion for the selection of the controller gains is carried out. Based on Fig.8, the closed-loop observer transfer function G_{cl4} can be expressed as:

$$G_{cl4}(z) = \frac{G_{ol4}(z)}{1+G_{ol4}(z)} = \frac{k_2 z + k_3}{(z-1)^2 + k_2 z + k_3} \quad (25)$$

In order to make a stable second order system, $G_{cl4}(z)$ can be regarded as $\frac{k_2 z + k_3}{(z-1+w_{ob})^2}$, where (23) can be acquired. Apart from that, in order to meet the stability of the observer, the coefficient f_{ob} needs to be satisfied as $|1 - 2\pi f_{ob} T_s| < 1$ (see, e.g., [30]).

Based on (22)-(23), the flux at the next instant can be acquired. The digital implementation of the flux observer will be presented in the next subsection.

D. Digital Implementation of Discrete-time Current Regulator

In practical digital systems, the digital implementation of the proposed discrete-time current regulator needs to be designed. Based on the proposed current regulator transfer function in (16), the transfer function can be obtained as based on the first, second order Taylor expansion or without approximation method from s to discrete-time domain:

$$G_{c3}(z) = G_{l2}(z)(e^{j\omega_e T_s} - z^{-1}e^{-\frac{R_s}{L_d}T_s})\zeta k_1 L(z)$$

$$\zeta_1 = \frac{1}{T_s}$$

$$\zeta_2 = \frac{\frac{j\omega_e T_s}{4} + \frac{\overline{R_s}T_s}{4L_d} - 1}{T_s \left[-\frac{\omega_e^2 T_s^2}{16} - \left(\frac{\overline{R_s}T_s}{4L_d} - 1 \right)^2 \right]} \quad (26)$$

$$\zeta_3 = \frac{\frac{\overline{R_s}}{L_d} + j\omega_e}{e^{j\omega_e T_s} - e^{-\frac{\overline{R_s}}{L_d}T_s}}$$

Fig. 9. The diagram of the proposed discrete-time current regulator with flux observer.

where the coefficient ζ can be chosen as ζ_1 , ζ_2 or ζ_3 ($\zeta \in (\zeta_1, \zeta_2, \zeta_3)$) based on different accuracy requirements. The current regulator with ζ_2 and ζ_3 is more accurate, but the calculation of the current regulator with ζ_2 and ζ_3 is increased. In this paper, ζ_1 is chosen and the output U_{dq}^{ref} of the discrete-time current regulator can be expressed as:

$$e_{dq}(k) = i_{dq}^{ref}(k) - i_{dq}(k)$$

$$e_{1dq}(k) = \left[e_{dq}(k)e^{j\omega_e(k)T_s} - e_{dq}(k-1)e^{-\frac{\overline{R_s}}{L_d}T_s} \right] \quad (27)$$

$$U_{dq1}(k) = e_{1dq}(k)2\pi f_{bw}G_{l2} + U_{dq1}(k-1)$$

$$U_{dq}^{ref}(k) = [U_{dq1}(k) - D(k)]k_1(k)$$

In order to improve the current performance caused by the one-step control delay, the proposed discrete-time flux observer also can be employed to work with the proposed current regulator. In this case, the measured dq-axis currents $i_{dq}(k)$ in (27) can be replaced by the predicted dq-axis currents $\hat{i}_{dq}(k+1)$. The $\hat{\lambda}_{dq}(k+1)$ of the discrete-time flux observer can be obtained as:

$$e_{2dq}(k) = [i_{dq}(k) - \hat{i}_{dq}(k)]G_{l2}$$

$$r_{dq}(k) = \frac{k_2 e_{2dq}(k)}{T_s} + \frac{k_3 e_{2dq}(k-1)}{T_s} + r_{dq}(k-1) \quad (28)$$

$$\hat{\lambda}_{dq}(k+1) = \hat{\lambda}_{dq}(k) + T_s [r_{dq}(k) + d_{2dq}(k)]$$

After obtaining the predicted flux $\hat{\lambda}_{dq}(k+1)$, the predicted current $\hat{i}_{dq}(k+1)$ can be acquired, which can be shown as:

$$\hat{i}_{dq}(k+1) = \hat{\lambda}_{dq}(k+1)G_{l2}^{-1} \quad (29)$$

Substitute $\hat{i}_{dq}(k+1)$ into the proposed current regulator in (27), the diagram of the proposed discrete-time current regulator with flux observer can be presented in Fig.9.

IV. EXPERIMENTAL RESULTS

The AC machine drive test rig is constructed as shown in Fig. 10. It includes a XCUBE control board (the control chip is DSP TMS320F28377d), drive board, sensor board, two PMSMs, load system and main power supply. The AC machine and inverter parameters are listed in Table.I. To make a fair comparison, all of methods are added distorted voltage compensation caused by inverter nonlinearity and compensated coefficient caused by Park transform. It should be noted that there is no improvement for the part of the active resistance term, the active resistance term is not considered in

TABLE I
AC MACHINE AND INVERTER PARAMETERS

Parameters	Value	Unit
Stator resistance (R_s)	1.1	Ω
Flux linkage (ψ_m)	0.092	Wb
d axis inductance (L_d)	5.7e-3	H
q axis inductance (L_q)	5.7e-3	H
Rated current (I_N)	4.2	A
VSI deadtime (t_d)	2.5	μs

Fig. 10. AC machine drive test rig.

this paper. The active resistance term based on [18], [21] can be employed in this paper to enhance the disturbance rejection. Because the discrete-time current regulator without R_s [24] (Method2) is unstable, this method is not compared in the test results. In this paper, four methods are compared in the AC machine test rig, namely state-feedback decoupling discrete-time current regulator [3] (Method1), proposed discrete-time current regulator (Method3), DPCC with one-step delay compensation [28] (Method4) and proposed discrete-time current regulator with flux observer (Method5).

A. Proposed Flux Observer Performance

In order to validate the effectiveness of the proposed flux observer, the dq-axis flux observer performance is shown in Fig.11. The sampling and control periods are set to $200\mu s$, the load reference speed is set to 600rpm and the dq-axis currents are set to -1A and 4A, respectively. From this figure, it can be seen that the predicted flux is one step control period ahead than the present measured flux, which can validate that the proposed stator observer is able to predict the next instant flux. In addition, the proposed flux observer owns strong parameter robustness under resistance and rotor flux linkage parameters mismatch. It should be noted that the measured flux can be computed based on the measured current.

B. Disturbance Rejection Performance

The disturbance rejection performance of the four methods is shown in Fig.12. The AC machine speed is set to 700rpm and the dq-axis reference currents are set to 0A and 2A,

Fig. 11. dq-axis flux comparison of four methods under different conditions. (a) Without parameter mismatch under 600rpm; (b) The condition under $R_s = 2R_s$ and $\psi_m = 1.5\psi_m$

Fig. 12. The q-axis current performance of four methods under an abrupt disturbance. (a) Method1; (b) Method4; (c) Method3; (d) Method5.

respectively. In order to create the load disturbance, the sudden back-emf change is applied in the q-axis voltage equation of the AC machine. The sampling and control periods are set to $400\mu s$ and the bandwidth of Method1, Method3, Method4 and Method5 are set to 300Hz, 300Hz, 400Hz and 400Hz. It can be seen that Method3 has shorter settling time under a abrupt disturbance compared with Method1 in Fig.12(a) and (c). In addition, Method4 does not own the disturbance rejection in Fig.12(b). Through adding one-step flux prediction scheme, Method5 has the best disturbance rejection (the smallest value of the q-axis current overshoot and shortest settling time) in Fig.12(d).

C. Current Performance under Low Speed

In order to testify the proposed method performance, the sampling and control periods are set to $400\mu s$ in the four methods. Because the one-step delay prediction is adopted, the bandwidth f_{bw} of DPCC is $\frac{1}{2\pi T_s} \approx 400\text{Hz}$, and the bandwidth in Method5 is set to 400Hz. Since there is no one-step delay prediction, the bandwidth in Method1 and Method3 is all set to 200Hz. Utilizing the high frequency sine waveform reference speed, the output of speed regulator can generate trapezoid waveform q-axis reference current due to the limitation of the speed bandwidth. In this case, the PMSM speed can be controlled at low speed condition (about $-40 \sim 40\text{rpm}$).

Fig. 13. The q-axis current performance of four methods without parameter mismatch.

Fig. 14. The q-axis current performance of four methods. (a)-(b) The condition under $L_d = 1.5L_d$, $L_q = 1.5L_q$; (c)-(d) The condition under $L_d = 0.5L_d$, $L_q = 0.5L_q$.

Without parameter mismatch, the q-axis current performance of the four methods is shown in Fig.13. It can be seen that there is no distinct overshoot in Method4 and Method5 due to one-step delay prediction. Method3 has the lower overshoot than Method1, which can show the proposed direct discrete-time current regulator merits. In Fig.14(c)-(d), it can be seen that the dynamic current response speed is reduced and there is an overshoot in Method5 compared with Method4 but the settling time is almost the same. In Fig.14(a)-(b), Method3 has the lower current overshoot than Method1. In Fig.15, it can be seen that there is a dc offset in q-axis current of Method4, which can validate that it cannot suppress the low frequency disturbances. Compared with Method1, Method3 has the lower current overshoot under resistance parameter mismatch.

D. Current Performance under Rated Speed

In order to testify the current performance of the proposed method under AC machine rated speed condition (3000rpm), the d-axis current is set to -1A and q-axis current is increased from 1A to 3A, which is shown in Fig.16. It can be found that compared with Method3, there exists large current overshoot in Method1. Meanwhile, there exists dq-axis dc offset in Method4 and the current performance is almost the same in the two proposed methods. It should be noted that the AC machine dq-axis inductance parameters are set to different values, which can validate that the proposed current regulator can be applied in asymmetric AC machines.

Fig. 15. The q-axis current performance of four methods. (a)-(b) The condition under $R_s = 2R_s$; (c)-(d) The condition under $R_s = 0.5R_s$.

Fig. 16. The dq-axis currents performance of four methods under 3000rpm. (a) Method1; (b) Method4; (c) Method3; (d) Method5. (The condition: The current bandwidth of Method1 and Method3 is set to 150Hz and the current bandwidth of Method5 is set to 200Hz; $L_d = 1.2L_d$, $L_q = 1.5L_q$)

E. Current Performance under Variable Speed

The dq-axis currents are set to -1A and 4A, respectively. In Fig.17, it can be seen that there are larger dq-axis current offsets in Method4 with the speed increased. In Fig.17(a), it can be seen that large current vibration occurs. The reason is that the inductance in regulator is larger than the actual inductance, and the current bandwidth is forced to be increased, which can lead to the instability. Due to the merit of direct discrete-time structure, the system stability is getting better in Method3. In addition, owing to the involvement of the one-step flux prediction, the best current performance is in Method5. In Fig.18, it can be seen that all of systems are stable since the inductance in regulator is lower than the actual inductance. Since Method4 cannot suppress the low frequency disturbances caused by resistance and rotor flux linkage mismatch, there are large current variation between dq-axis measured and reference currents. In Fig.18(c), it can be seen that the actual d-axis current is larger than the d-axis reference current due to the resistance mismatch. When the speed is abruptly changed, the current fluctuation of Method4

Fig. 17. The dq-axis current performance of four methods from 400rpm to 1500rpm. (a) Method1; (b) Method3; (c) Method4; (d) Method5. (The condition: The current bandwidth of Method1 and Method3 is set to 300Hz and the current bandwidth of Method5 is set to 400Hz; $\bar{L}_d = 1.5L_d$, $\bar{L}_q = 1.5L_q$, $\bar{R}_s = 2R_s$, $\bar{\psi}_m = 1.5\psi_m$)

Fig. 18. The dq-axis current performance of four methods from 150rpm to 1000rpm. (a) Method1; (b) Method3; (c) Method4; (d) Method5. (The condition: The current bandwidth of Method1 and Method3 is set to 300 Hz and the current bandwidth of Method5 is set to 400Hz; $\bar{L}_d = 0.75L_d$, $\bar{L}_q = 0.75L_q$, $\bar{R}_s = 0.5R_s$, $\bar{\psi}_m = 1.8\psi_m$)

is larger than in Method3 and Method5. A detailed comparison of the five methods is listed in TableII.

V. CONCLUSION

A direct discrete-time current regulator which includes symmetric and asymmetric machines is proposed. Based on the proposed method, it can possess the merit of the disturbance rejection compared with DPCC in Bode plot. Adding the resistance parameter in the plant, it can be seen that the proposed system is stable under different conditions compared with discrete-time current regulator without R_s in zero-pole location plot. In addition, to compensate the one step delay

TABLE II
RESULT COMPARISON OF FIVE METHODS

Method types	Dynamic response	Parameter robustness	Steady state offset	Calculation	Stability
1 (Paper in [3])	slow	medium	zero	low	stable
2 (Fig.1)	slow	low	zero	medium	unstable
3 (Fig.2)	slow	high	zero	medium	stable
4 (Fig.4(a))	fast	low	exist	low	stable
5 (Fig.9)	fast	high	zero	medium	stable

issue, a direct discrete-time flux observer is proposed which can suppress the disturbances caused by rotor flux linkage and resistance mismatch.

The experimental results show that the proposed current regulator without one step delay compensation can reduce the current overshoot, settling time and have strong parameter robustness compared with state-feedback decoupling discrete-time current regulator and DPCC. Also, setting the same current bandwidth, it can validate that the proposed method with flux observer has better current performance than DPCC under different parameter mismatch. Therefore, the proposed discrete-time current regulator can be experimentally applied in practical application for different types of AC machines.

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