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Nonlinear Controller for the Velocity Tracking and Rejection of Sinusoidal Disturbances in Permanent Magnet Stepper Motors

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Abstract - In this paper, a nonlinear controller is proposed to track the desired velocity and to cancel sinusoidal disturbances. The proposed method consists of a velocity tracking controller and internal model principles (IMPs). For the design of the velocity tracking controller, mechanical and electrical dynamic controllers are independently designed. For the mechanical dynamics, the velocity tracking controller generates the desired quadrature current to track the desired velocity. The current tracking controller is designed to guarantee the desired quadrature current and to regulate the direct current. Therefore, the proposed velocity tracking controller has a field-oriented control. Since the controllers of the mechanical and electrical dynamics are independently designed, the stability of the closed-loop system is demonstrated using passivity. Since both the cogging torque and DC current errors act as sinusoidal disturbances in PMSM, we use four add-on type IMPs that preserve the merits and performance of the pre-designed controller without sacrificing the closed-loop stability. The performance of the proposed method is validated via simulations.

Key Words : Stepper motor, Velocity tracking, Cogging torque

1. Introduction

Permanent magnet stepper motors (PMSMs) have been widely used for position and velocity control applications in low speed operations due to their durability, high efficiency, and power density, as well as high torque to inertia ratio and absence of rotor winding [1,2]. PMSMs are less expensive to produce compared with DC servo motors of the same size or their equivalent brushless motors. However, the control of PMSM is essentially governed by nonlinear dynamics, and they are, thus, difficult to design and implement for precision motion control.

Various control methods have been studied for velocity tracking in PMSMs [3, 4, 5]. In [3], a sensorless velocity tracking controller for a full-order PMSM model with actuated mechanical dynamics was designed. The electrical dynamics were considered to reconstruct the position and velocity from the phase current measurements and the phase voltage. Another velocity tracking controller based on the Lyapunov method has been proposed [4]. The load torque observer was designed to estimate the unknown load torque disturbance. A real-time non-linear adaptive speed control scheme based on a backstepping control technique was previously proposed [5]. In the controller design, input-output feedback linearization with adaptive backstepping control was used to compensate for the nonlinearities, parameter uncertainties, and load torque disturbance. Although these methods demonstrated good performances, the sinusoidal disturbances due to the cogging torque and DC current offset were neglected. To improve the velocity tracking performance, rejection of the sinusoidal disturbances is required.

In this paper, a nonlinear controller is proposed to track the desired velocity and to remove the sinusoidal disturbances. The proposed method consists of the velocity tracking controller and internal model principles (IMPs). In the design of the velocity tracking controller, the controllers of the mechanical and electrical dynamics are independently designed. The mechanical dynamics controller generates the desired quadrature current to track the desired velocity. Then, the current tracking controller is designed to the guarantee the desired quadrature current and to regulate the direct current. Therefore, the proposed velocity tracking controller utilizes field-oriented control (FOC) that maintains zero

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direct current to maximize the torque. Since the controllers of the mechanical and electrical dynamics are independently designed, the stability of the closed-loop system is demonstrated using passivity. The cogging torque is produced by the fourth harmonic component of the permeance distribution [6]. The cogging torque is manifested in the mechanical dynamics as a disturbance. In current measurements, a DC current offset may be present due to the DC offsets of the current sensors in the motor driver or to the D/A converters in the motion controller [7, 8]. The DC current offsets are the sinusoidal signal in the PMSM that is applied to direct quadrature (DQ) transformation. These offsets disturb the convergence to the desired quadrature current and regulate the direct current. Therefore, both cogging torque and DC current errors act as sinusoidal disturbances in a PMSM. To compensate for these sinusoidal disturbances, four add-on type IMPs are used [9, 10] to preserve the merits and performance of the pre-designed controller without destroying the closed-loop stability [11]. Simulation results show the performance of the proposed method.

This paper is organized as follows. In Section 2, a mathematical model is introduced. The controller development is explained in Section 3, and simulation results are presented in Section 4. Finally, the research conclusions are presented in Section 5.

2. Model

A PMSM consists of a slotted stator with two phases and a permanent magnet rotor which has north and south poles. A schematic of this engine is shown in Fig. 1. Detailed descriptions of the PMSMs are given in [12]. The dynamics of a PMSM can be represented as

$$\begin{aligned}\dot{\theta} &= \omega \\ \dot{\omega} &= \frac{1}{J}[-K_m i_a \sin(N_r\theta) + K_m i_b \cos(N_r\theta) - B\omega] \\ \dot{i}_a &= \frac{1}{L}[v_a - Ri_a + K_m \omega \sin(N_r\theta)] \\ \dot{i}_b &= \frac{1}{L}[v_b - Ri_b - K_m \omega \cos(N_r\theta)]\end{aligned} \tag{1}$$

where $x=[\theta,\omega,i_{a,}i_b]^T$ is the state, and $u=[v_{a,}v_b]^T$ is the input. $v_{a,}v_b$ and $i_{a,}i_b$ are the voltages [V] and currents [A] in phases A and B, respectively. ω is the rotor (angular) velocity [rad/s], θ is the rotor (angular) position [rad], B is the viscous friction coefficient [N・m・s/rad], J is the inertia of the motor [Kg・m^2], K_m is the motor torque constant [N・m/A], R is the resistance of phase winding [W], L is the inductance of phase winding [H], and N_r is the number of rotor teeth. The magnetic coupling between the phases is ignored in the model. This model

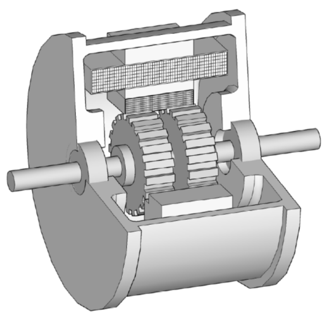

Fig. 1 A cutaway view of PMSM

also ignores the variation in inductance due to magnetic saturation. In addition, an ideal sinusoidal flux distribution is assumed. The DQ transformation for the phase voltages and currents is defined [13] as

$$\begin{aligned}\begin{bmatrix} v_d \\ v_q \end{bmatrix} &= \begin{bmatrix} \cos(N_r\theta) & \sin(N_r\theta) \\ -\sin(N_r\theta) & \cos(N_r\theta) \end{bmatrix} \begin{bmatrix} v_a \\ v_b \end{bmatrix}, \\ \begin{bmatrix} i_d \\ i_q \end{bmatrix} &= \begin{bmatrix} \cos(N_r\theta) & \sin(N_r\theta) \\ -\sin(N_r\theta) & \cos(N_r\theta) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \end{bmatrix}\end{aligned} \tag{2}$$

To linearize the mechanical dynamics, the application of the DQ transformation to the PMSM (1) yields the following equations,

$$\begin{aligned}\dot{\theta} &= \omega \\ \dot{\omega} &= \frac{1}{J}[-K_m i_q - B\omega)] \\ \dot{i}_d &= \frac{1}{L}[v_d - Ri_d + N_r L\omega i_q] \\ \dot{i}_q &= \frac{1}{L}[v_q - Ri_q - N_r L\omega i_d - K_m\omega]\end{aligned} \tag{3}$$

where $v_{d,}v_q$ are the direct and quadrature voltages [V] and $i_{d,}i_q$ are the direct and quadrature currents [A].

3. Controller Design

The proposed controller consists of a tracking module to track the desired velocity and an IMP to reject the sinusoidal disturbance. First, the FOC is developed with an assumption of no disturbance in PMSM, then an add-on type IMP is designed to include the cogging torque.

3.1. Velocity Tracking Controller Design

To track the desired velocity, the controllers of the mechanical and electrical dynamics are separately designed. In mechanical dynamics, the desired quadrature current is generated to track the desired velocity using the proposed controller. Then, the quadrature current tracks the desired quadrature current, and the direct current converges to zero.

In mechanical dynamics, i_q is the input as follows

$$\dot{\theta}=\omega \qquad \dot{\omega}=\frac{1}{J}[-K_m i_q - B\omega]. \tag{4}$$

The mechanical tracking errors are defined as

$$e_1=\int_0^t \omega^d-\omega dt, \qquad e_2=\omega^d-\omega. \tag{5}$$

where ω^d is the desired velocity. The mechanical tracking error dynamics is

$$\dot{e}_1=e_2 \qquad \dot{e}_2=\dot{\omega}^d-\frac{1}{J}[K_m i_q - B\omega]. \tag{6}$$

The mechanical dynamics controller for the desired quadrature current is designed as

$$i_q^d=\frac{1}{K_m}[k_I e_1+k_P e_2+B\omega^d+J\dot{\omega}^d] \tag{7}$$

where i_q^d is the desired quadrature current, and the controller gains, k_I and k_P are positive constants. Then if we assume that $i_q=i_q^d$, the mechanical tracking error dynamics becomes

$$\dot{e}_1=e_2 \qquad \dot{e}_2=-\frac{k_I}{J}e_1-\frac{k_P+B}{J}e_2. \tag{8}$$

Since the mechanical tracking error dynamics is linear and the system matrix is always Hurwitz with the positive controller gains k_I and k_P, the zero equilibrium point in (8) is asymptotically stable.

The electrical tracking errors are defined as

$$e_3=-i_d, \qquad e_4=i_a^d-i_a. \tag{9}$$

The electrical tracking error dynamics is

$$\dot{e}_3=-\frac{1}{L}[v_d-Ri_d+N_rL\omega i_q] \qquad \dot{e}_4=\dot{i}_q^d-\frac{1}{L}[v_q-Ri_q-N_rL\omega i_d-K_m\omega]. \tag{10}$$

To tracks the desired quadrature current and regulate the direct current, the electrical controller are defined as

$$v_d=Ri_d-N_rL\omega i_q+\gamma_d Le_3 \tag{11}$$

$$v_q=Ri_q+N_rL\omega i_d+K_m\omega+\gamma_q Le_4+L\dot{i}_q^d \tag{12}$$

where the controller gains γ_d, γ_q are positive constants. Then the electrical tracking error dynamics is

$$\dot{e}_3=-\gamma e_3 \qquad \dot{e}_4=-\gamma e_4. \tag{13}$$

Therefore, the zero equilibrium point in (13) is exponentially stable, and the proposed velocity tracking controller is in the form of field-oriented control (FOC) in order to maintain zero direct current to maximize the torque.

Since the desired quadrature current (7) and the control law (10) are designed separately, closed-loop stability should be demonstrated.

Theorem 1. Consider the PMSM (3). From the (5), (7), (9), (11), and (12), the tracking error dynamics of the closed-loop is given by

$$\underbrace{\begin{bmatrix}\dot{e}_1\\ \dot{e}_2\end{bmatrix}}_{\dot{e}_m}=\underbrace{\begin{bmatrix}0 & 1\\ -\frac{k_I}{J} & -\frac{k_P+B}{J}\end{bmatrix}}_{A_m}\underbrace{\begin{bmatrix}e_1\\ e_2\end{bmatrix}}_{e_m}+\underbrace{\begin{bmatrix}0 & 0\\ 0 & \frac{K_m}{J}\end{bmatrix}}_{B_m}\underbrace{\begin{bmatrix}e_3\\ e_4\end{bmatrix}}_{e_e}$$
$$\underbrace{\begin{bmatrix}\dot{e}_3\\ \dot{e}_4\end{bmatrix}}_{\dot{e}_e}=\underbrace{\begin{bmatrix}-\gamma_d & 0\\ 0 & -\gamma_q\end{bmatrix}}_{A_e}\underbrace{\begin{bmatrix}e_3\\ e_4\end{bmatrix}}_{e_e}. \tag{14}$$

If the controller gains k_I, k_P, γ_d, and γ_q are positive constants, the equilibrium point of (14) is asymptotically stable. ◇

Proof: In (14), If the controller gains k_I, k_P, γ_d, and γ_q are positive constants, then both A_m and A_e are Hurwitz, and B_m is bounded. By defining the Lyapunov candidate function, V_m, as

$$V_m=\frac{1}{2}e_m^TP_me_m \tag{15}$$

where P_m is positive definite such that $A_m^TP_m+P_mA_m=-I$, we obtain

$$\dot{V}_m=-e_m^Te_m+e_m^TP_mB_me_e. \tag{16}$$

If we define $P_mB_me_e$ as the input and e_m as the output in (15), (15) can be rewritten as

$$\underbrace{e_m^T}_{output}\underbrace{P_mB_me_e}_{input}=\dot{V}_m+\underbrace{e_m^Te_m}_{>0}. \tag{17}$$

Since $e_m^TP_mB_me_e \geq \dot{V}_m+e_m^Te_m$, the relationship between em and e_e is strictly output passive [14]. And $\dot{e}_m=A_me_m$ is zero-state observable. Therefore, the dynamics of e_m is bounded input bounded output (BIBO) stable. The equilibrium point of e_e dynamics is globally exponentially stable. Therefore, e_e globally exponentially converges to zero, and e_m globally asymptotically converges to zero. ■

Since the quadrature current tracks the desired quadrature current and the proposed controller causes the direct current to converge to zero, the controller described by (7), (11), and (12) utilizes FOC.

3.2. IMP Design for the Rejection of Sinusoidal Disturbance

In PMSM, the cogging torque and DC current offset error represent main disturbances. The cogging torque d_1 that is produced by the forth harmonic component of the permeance distribution is noted by

$$d_1 = -T_{dm}\sin(4N_r\theta) \tag{18}$$

where T_{dm} is the cogging torque constant [N/m] [9]. The DC current offset error due to the DC offsets of the current sensors in the motor driver or the D/A converters in the motion controller [6, 7]. Now we analyze the DC current offset. The DC current offset errors, i_{ad} and i_{bd} of phases A and B are defined as

$$\begin{aligned} i_{am} &= i_a + i_{ad}, \\ i_{bm} &= i_b + i_{bd} \end{aligned} \tag{19}$$

where i_{am} and i_{bm} are the measured currents. The application of DQ transformation to the DC current error is

$$\begin{aligned} i_{dd} &= i_{ad}\cos(N_r\theta) + i_{bd}\sin(N_r\theta) \\ i_{qd} &= -i_{ad}\sin(N_r\theta) + i_{bd}\cos(N_r\theta). \end{aligned} \tag{20}$$

where i_{dd} and i_{qd} are the DC current offset errors of the direct and the quadrature currents. The phase between i_{ad} and i_{bd} is defined as

$$\phi = \mathrm{atan}(i_{bd}/i_{ad}). \tag{21}$$

From (20) and (21), we get

$$\begin{aligned} i_{dd} &= \underbrace{\sqrt{i_{ad}^2 + i_{bd}^2}}_{I_d}\left(\frac{i_{ad}}{\sqrt{i_{ad}^2+i_{bd}^2}}\cos(N_r\theta) + \frac{i_{bd}}{\sqrt{i_{ad}^2+i_{bd}^2}}\sin(N_r\theta)\right) \\ &= I_d(\cos(\phi)\cos(N_r\theta) + \sin(\phi)\sin(N_r\theta)) \\ &= I_d\cos(N_r\theta - \phi), \\ i_{qd} &= \underbrace{\sqrt{i_{ad}^2 + i_{bd}^2}}_{I_d}\left(-\frac{i_{ad}}{\sqrt{i_{ad}^2+i_{bd}^2}}\sin(N_r\theta) + \frac{i_{bd}}{\sqrt{i_{ad}^2+i_{bd}^2}}\cos(N_r\theta)\right) \\ &= I_d(-\cos(\phi)\sin(N_r\theta) + \sin(\phi)\cos(N_r\theta)) \\ &= -I_d\sin(N_r\theta - \phi). \end{aligned} \tag{22}$$

The cogging torque (17) and the DC current offset errors (22) affect the mechanical dynamics (4) and the actual control voltage inputs (11), (12) Therefore, the mechanical tracking error dynamics (8) and the electrical tracking error dynamics (12) can be rewritten as

$$\begin{aligned} \dot{e}_1 &= e_2 \\ \dot{e}_2 &= -\frac{k_I}{J}e_1 - \frac{k_P + B}{J}e_2 + \frac{T_{dm}\sin(4N_r\theta)}{J} \\ &\quad + \frac{K_m I_d \sin(N_r\theta - \phi)}{J} \end{aligned} \tag{23}$$

$$\begin{aligned} \dot{e}_3 &= -\gamma e_3 + \left(\frac{R}{L} - \gamma_d\right)I_d\cos(N_r\theta - \phi) - N_r\omega I_d\sin(N_r\theta - \phi) \\ \dot{e}_4 &= -\gamma e_4 - \left(\frac{R}{L} - \gamma_q\right)I_d\sin(N_r\theta - \phi) - N_r\omega I_d\cos(N_r\theta - \phi). \end{aligned} \tag{24}$$

In order to eliminate the sinusoidal disturbance, an add-on type IMP [12, 13] is designed to preserve the merits and performance of the pre-designed controller without destroying the closed-loop stability [14]. In order to cancel the two sinusoidal disturbances in (23), we use two add-on type IMPs (IMP1, IMP4) as

$$\begin{aligned} \begin{bmatrix} \dot{x}_{11} \\ \dot{x}_{12} \end{bmatrix} &= \begin{bmatrix} 0 & N_r\omega \\ -N_r\omega & 0 \end{bmatrix}\begin{bmatrix} x_{11} \\ x_{12} \end{bmatrix} \\ u_{IMP1} &= [0\; k_{IMP1}]\begin{bmatrix} x_{11} \\ x_{12} \end{bmatrix}, \end{aligned} \tag{25}$$

$$\begin{aligned} \begin{bmatrix} \dot{x}_{41} \\ \dot{x}_{42} \end{bmatrix} &= \begin{bmatrix} 0 & 4N_r\omega \\ -4N_r\omega & 0 \end{bmatrix}\begin{bmatrix} x_{41} \\ x_{42} \end{bmatrix} \\ u_{IMP4} &= [0\; k_{IMP4}]\begin{bmatrix} x_{41} \\ x_{42} \end{bmatrix}. \end{aligned} \tag{26}$$

Futhermore, to compensate for the two sinusoidal disturbances in (24), two add-on type IMPs (IMPd, IMPq) are used as

$$\begin{aligned} \begin{bmatrix} \dot{x}_{d1} \\ \dot{x}_{d2} \end{bmatrix} &= \begin{bmatrix} 0 & N_r\omega \\ -N_r\omega & 0 \end{bmatrix}\begin{bmatrix} x_{d1} \\ x_{d2} \end{bmatrix} \\ u_{IMPd} &= [0\; k_{IMPd}]\begin{bmatrix} x_{d1} \\ x_{d2} \end{bmatrix}, \end{aligned} \tag{27}$$

$$\begin{aligned} \begin{bmatrix} \dot{x}_{q1} \\ \dot{x}_{q2} \end{bmatrix} &= \begin{bmatrix} 0 & N_r\omega \\ -N_r\omega & 0 \end{bmatrix}\begin{bmatrix} x_{q1} \\ x_{q2} \end{bmatrix} \\ u_{IMPq} &= [0\; k_{IMPq}]\begin{bmatrix} x_{q1} \\ x_{q2} \end{bmatrix}. \end{aligned} \tag{28}$$

From (25)-(28), the desired quadrature current (7) and the control law (11), (12) are modified as

$$i_q^d = \frac{1}{K_m}[k_I e_1 + k_P e_2 + B\omega^d + J\dot{\omega}^d] + u_{IMP1} + u_{IMP4} \tag{28}$$

$$v_d = Ri_d - N_r L\omega i_q + \gamma_d L i_d + u_{IMPd}, \tag{29}$$

$$v_q = Ri_q + N_r L\omega i_d + K_m\omega + \gamma_q L e_4 + L\dot{i}_q^d + u_{IMPq}. \tag{30}$$

We summarize our proposed method in the next lemma

Lemma 1. Consider the PMSM (3). If we design IMPS (25)-(28), the desired quadrature current (29) and the control law (30), (31), the velocity asymptotically converges to the desired velocity. Furthermore, the quadrature and the direct currents converge to the desired quadrature current and zero. ◇

The structure of the proposed method is depicted in Fig. 2.

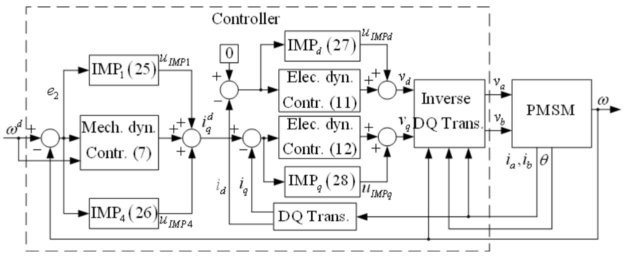

Fig. 2 Block diagram of the controller structure

4. Simulation Results

TABLE I

PARAMETERS OF PMSM AND CONTROLLER GAINS

Para.	Value	Para.	Value
L	40 mH	R	14.8 Ω
J	8×10^{-5} kgm^2	K_m	0.5 Nm/A
Nr	50	B	5×10^{-3} Nms/rad
k_I	1000	k_P	0.1
k_{IMP1}	100	k_{IMP4}	100
γ_d	0.1	γ_q	0.1
k_{IMPd}	1000	k_{IMPq}	1000

Simulations were performed to evaluate the performance of the proposed controller. The PMSM and controller parameters are listed in Table I. The cogging torque constant T_{dm} was 0.025 Nm, and the constant disturbances i_{ad} and i_{bd} were 0.002A and −0.002A, respectively. The desired velocity shown in Fig. 3 was used. Simulations were conducted for the following four cases: 1) Only the tracking controller (7), (11), (12) was used when there was no cogging torque or DC offset current error, 2) Only the tracking controller (7), (11), (12) was used when both disturbances were present, 3) The tracking controller (7), (11), (12) and IMPs (25), (26) of only the mechanical tracking error dynamics were used when both disturbances were present, and 4) The tracking controller (7), (11), (12) and IMPs (25)-(28) of both types of error were used when both disturbances were present.

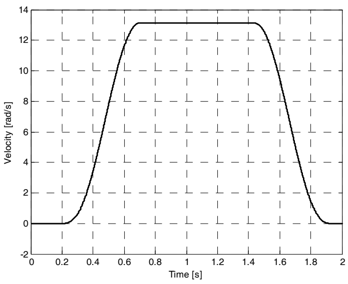

Fig. 3 Desired velocity

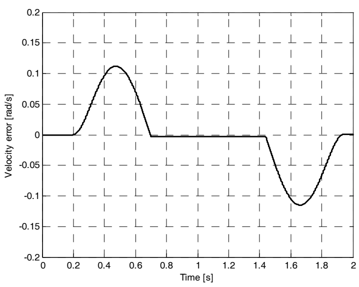

Fig. 4 Velocity tracking error of case 1

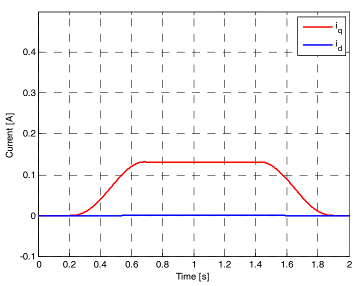

Fig. 5 i_d and i_q of case 1

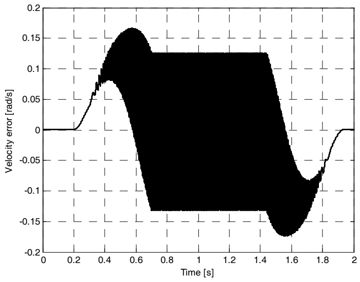

Fig. 6 Velocity tracking error of case 2

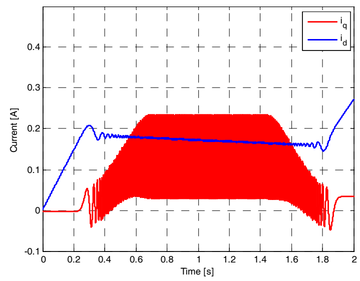

Fig. 7 i_d and i_q of case 2

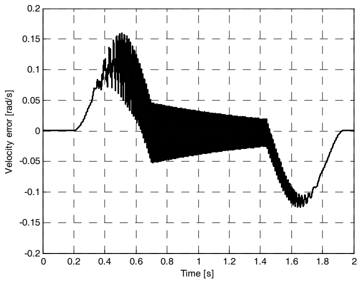

Fig. 8 Velocity tracking error of case 3

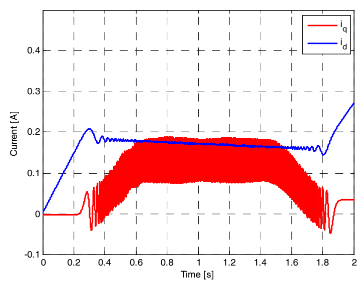

Fig. 9 i_d and i_q of case 3

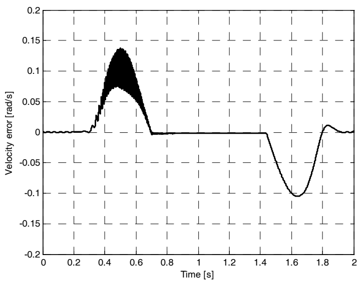

Fig. 10 Velocity tracking error of case 4

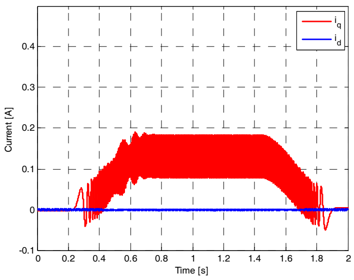

Fig. 11 i_d and i_q of case 4

The results of case 1 are shown in Figs. 4 and 5. Since there was no cogging torque or DC current offset error, the desired velocity was well tracked and the zero direct current was maintained. However, in case 2, ripples appeared in the velocity tracking error, id and iq, as shown in Figs. 6 and 7. Furthermore, the quadrature current tracking error experienced DC offset, and the direct current did not converge to zero. In the dynamics of e_a , a cosine term disturbance was present when the velocity was zero, indicating that the id diverged. On the other hand, in the dynamics of e_4 , a sine term disturbance was present when the velocity was zero. This implied that the remaining disturbance was suppressed when the velocity was zero. However, the dynamics of e_a also had a sine term disturbance in the non-zero velocity period. Therefore, iq experienced DC offset in the final stop period, and both id and iq had ripples. Therefore, FOC was not realized and the torque was not maximized. The results of case 3 are depicted in Figs. 8 and 9. Since the sinusoidal disturbances in the mechanical dynamics were compensated by the IMPs (25), (26), the ripple in the velocity tracking was cancelled. However, FOC was not still realized since the DC offset current errors in the current dynamics were not rejected. Therefore, the torque was not maximized. In case 4, since IMPs in the electrical dynamics were used to compensate for DC current offset errors, the ripple in the quadrature current was removed, and the zero direct current was maintained as shown in Fig. 10 and 11. Therefore, the ripple in the velocity tracking error was rapidly rejected in comparison with that in case 3.

4. Conclusion

In this paper, a nonlinear controller was proposed to track the desired velocity and to remove the sinusoidal disturbances. The proposed method consists of velocity tracking and internal model principles (IMPs). For the design of the velocity tracking controller, the controllers of the mechanical and electrical dynamics were independently developed. For the mechanical dynamics, the velocity tracking controller was designed to generate the desired quadrature current to track the desired velocity. Then, the current tracking controller was designed to guarantee the desired quadrature current and to regulate the direct current. Therefore, the proposed velocity tracking controller was in FOC form. Since the controllers of the mechanical dynamics and the electrical dynamics were independently designed, the stability of the closed-loop system was demonstrated using passivity. Since both cogging torque and DC current errors acted as sinusoidal disturbances in PMSM, an add-on type four IMP was used to compensate for the sinusoidal disturbances. In simulations, the desired velocity was quite well tracked, while adjustments were made to compensate for the sinusoidal disturbances. As such, the zero direct current was maintained.

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