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A new FOC technique based on predictive current control for PMSM drive

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Abstract. This paper presents a new predictive current control (PCC) strategy using field oriented controller (FOC) for permanent magnet synchronous motor (PMSM). The suggested technique estimates the desirable electrical torque to track mechanical torque at a fixed speed operation of PMSM. The estimated torque is used to calculate the required current of motor based on conventional FOC technique. Finally, the duty cycle of the inverter is determined to feed the PMSM at a fixed switching frequency. The performance of the control schemes is evaluated in terms of torque and current ripple, and transient response to step variations of the torque command. Numerical simulations tests have been carried out to validate the proposed method in MATLAB.

Keywords: permanent magnet synchronous motors, predictive current control, field oriented control

1 Introduction

In recent years, permanent magnet synchronous motors (PMSM) are increasing applied in several areas as reaction, automobiles, robotics and aerospace technology[1]. In addition, many researchers have tried to reduce the torque pulses and harmonics in PMSM. There are many ways to use power converters to fix the voltage or current in a PMSM driver to desired set point. Among them are voltage control such as six-step, sinusoidal, and space-vector modulation, and current control methods such as hysteresis and delta modulation[13]. Since important improvements have been made regarding the control techniques of the special machines, adequate operation for any industrial application is obtained with the appropriate supplying-motor assembly. In the same time, two of the control techniques, which are widely used in this industrial environment, are the field oriented and the direct torque control (FOC and DTC, respectively)[3]. These control strategies are different on the operation principle but their objectives are the same. They aim both to control effectively the motor torque and flux in order to force the motor to accurately track the command trajectory regardless of the machine and load parameter variation or any extraneous disturbances. Both control strategies have been successfully implemented in industrial products[2].

DTC is able to produce very fast torque and flux control, if the torque and the flux are correctly estimated, and will be robust with respect to motor parameters and perturbations. By controlling stator current by FOC it reduces torque ripple for quieter motor operation. The supporters of field-oriented control and direct torque control claim the superiority of their strategy versus the other. Up to now, the question of which strategy is superior the other, has not been clearly answered[5]. Nonetheless these studies have been developed which propose alternative solutions to the FOC control of a PWM inverter-fed motor drive with two objectives: first, achievement of an accurate and fast response of the flux and the torque, and second, reduction in the complexity of the control system.

The idea of combining the advantages of DTC and PMSMs into a highly dynamic drive appeared in the literature in the late 1990's[4, 18]. In the past decade several authors have proposed ways to adapt DTC

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to work with PMSMs[15]. In 1989 Pillay and Krishnan[9] presented the PMSM which was one of several types of permanent magnet ac motor drives available in the drives industry. They did not consider the damper windings and designed the motor drive system in field-oriented control. Qianad and Rahman[10] developed FOC in PMSM they presented a microprocessor-based FOC for PM hysteresis synchronous motor.

In the current control for an inverter-fed PMSM drive, there are four main types of control schemes: the hysteresis control, the ramp comparison control, the synchronous frame proportional integral (PI) control and the predictive control[6] Among the current control methods, a current regulated delta modulation method has been presented in[7] applying the symmetrical hysteresis band seems to be a best way to reduce current ripple, but due to delta modulation technique is not reduced sufficiently. In spite of the fast transient response of hysteresis current control technique, the maximum switching frequency is limited by the sampling frequency[17].

According to the mentioned techniques above, both FOC and PCC techniques have good performance in PMSM drive. On basis of the researches done before, the DTC based controllers are very little sensible to the parameters detuning in comparison with FOC[14], so we will present a FOC based control technique in PMSM drives. Also, due to increase the conventional FOC and PCC controllers, a new method based on FOC and PCC (FOC-PCC) is presented. The proposed method uses the d-q form of motor equations and discritized the equations to have controlled torque with lower error. The electromechanical relations of PMSM are used to estimate the desired torque and a fix speed. The applied electrical torque is applied to proposed controller to drive duty cycle of the inverter. In contrast with this method, other methods such as the ramp-comparison and synchronous PI controls overcome pro the hystresis current control. Although, it's transient response is relatively slow.

Due to this disadvantage of conventional current control, the current control of a PMSM by a predictive current control technique based on the space vector modulation is presented in [8]. Other references applied predictive current control technique to PMSM drive system and voltage source inverter in [6, 8, 16, 17] and [11, 12] respectively. Wipasuramonton and et al.[17] proposed a technique to reduce the current errors that exist due to the inaccuracies in the system parameters and the non ideal behavior of the inverter. In addition to these[6] is discretized the PMSM module to improve transient and steady state, better than the conventional predictive current controllers.

2 Analytic mode of PMSM

Permanent magnet synchronous motor has been studied for more than two decade. In majority of devoted method on PMSM there is a set of equations depended on rotor position. Representing the equations of the motor in rotor reference frame, cause to have a set of equations that does not depended on rotor position. Used transformation that refers to the equations in abc-frame into dq-frame is called "Park Transform". So K is the Park Transform as:

$$K = \frac{2}{3}\begin{bmatrix} \cos\theta & \cos\left(\theta - \frac{2\pi}{3}\right) & \cos\left(\theta + \frac{2\pi}{3}\right) \\ \sin\theta & \sin\left(\theta - \frac{2\pi}{3}\right) & \sin\left(\theta + \frac{2\pi}{3}\right) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix}; \tag{1}$$

$$\begin{bmatrix} S_q \\ S_d \\ S_o \end{bmatrix} = K \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix}. \tag{2}$$

According to the mentioned above, electromechanical behavior of the PMSM in the dq-frame is as below:

$$V_q = ri_q + \omega_r\lambda_d + \frac{\mathrm{d}\lambda_q}{\mathrm{d}t},\ V_d = ri_d - \omega_r\lambda_q + \frac{\mathrm{d}\lambda_d}{\mathrm{d}t},\ V_o = ri_o + \frac{\mathrm{d}\lambda_o}{\mathrm{d}t}, \tag{3}$$

$$\lambda_q = L_q i_q,\ \lambda_d = L_d i_d + \lambda_m,\ \lambda_o = L_o i_o, \tag{4}$$

$$Te = \frac{3}{4}P\left(\lambda_m i_q + (L_d - L_q)\, i_d i_q\right), \tag{5}$$

$$\frac{\mathrm{d}\omega_r}{\mathrm{d}t} = \frac{P}{2J}\left(T_e - F\omega_r - T_m\right). \tag{6}$$

If the angle between stator and rotor field flux be kept at 90° we have:

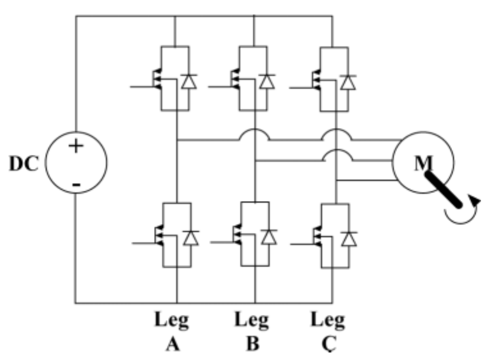
$$i_d = 0. \tag{7}$$

So, by this assumption and set $i_o = 0$, determining i_q leads to control the electrical torque directly. It should be mentioned that a predictive process will be present to have constant motor speed accompanied with low error torque tracking procedure.

3 FOC-PCC based control strategy

A simplified schematic of a voltage source inverter that is used to feed the three phase PMSM drive is shown in Fig. 1. Note that each leg of the described inverter has a Duty cycle Da, Db and Dc for Leg A, Leg B and Leg C, respectively.

Due to the basic concept in lack of DC offset for three phase output bipolar voltages of inverter, and specified model shown in Fig. 2, we assume that sum of mean values of the three phase voltages is equal to zero. It should be mentioned that this summation has zero value in very short time duration (T). In other



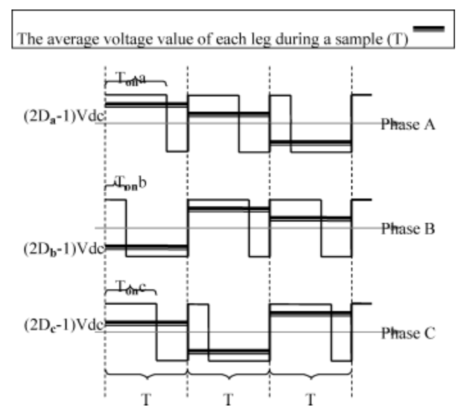


Fig. 1. Voltage source inverter model for PMSM drive **Fig. 2.** Bipolar three phase voltage of PMSM in each sample T

words, due to small values of T, the instantaneous values of the phase voltages are replaced by their mean values. Sum of the phase voltages $(\overline{V}_a, \overline{V}_b, \overline{V}_c)$ equals to zero. Therefore we have:

$$\overline{V}_a + \overline{V}_b + \overline{V}_c = 0, \tag{8}$$

where, $\overline{V}_{a,b,c}$ are mean values of the phase voltages in each switching period. According to Fig. 2 the mean value of each phase voltage (based on bipolar switching of each leg with duty cycle D) is as follow:

$$\overline{V}_a = (2D_a - 1)V_{dc}; \quad \overline{V}_b = (2D_b - 1)V_{dc}; \quad \overline{V}_c = (2D_c - 1)V_{dc}. \tag{9}$$

Substituting Eq. (9) into Eq. (8) yields:

$$(2D_a - 1)V_{dc} + (2D_b - 1)V_{dc} + (2D_c - 1)V_{dc} = 0 \rightarrow 2D_a + 2D_b + 2D_c = 3 \rightarrow D_a + D_b + D_c = \frac{3}{2}. \tag{10}$$

According to the result in Eq. (10), duty cycle of one leg calculates by duty cycle of two other legs, therefore:

$$D_b = 1.5 - (D_a + D_c). \tag{11}$$

With the values resulted in Eq. (9), Eq. (1) and Eq. (2), stator voltage represents in dq-frame as bellow:

$$\begin{vmatrix} V_q \\ V_d \\ V_o \end{vmatrix} = \frac{2}{3} \begin{vmatrix} \cos(\theta_r) & \cos\left(\theta_r - \frac{2\pi}{3}\right) & \cos\left(\theta_r + \frac{2\pi}{3}\right) \\ \sin(\theta_r) & \sin\left(\theta_r - \frac{2\pi}{3}\right) & \sin\left(\theta_r + \frac{2\pi}{3}\right) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{vmatrix} \begin{vmatrix} V_a \\ V_b \\ V_c \end{vmatrix} \rightarrow \begin{vmatrix} V_q \\ V_d \\ V_o \end{vmatrix} = \frac{2}{3} K \begin{vmatrix} (2D_a - 1) \\ (2 - 2(D_a + D_c)) \\ (2D_c - 1) \end{vmatrix} V_{dc}. \quad (12)$$

Now, we would be able to represent d, q and o values on the basis of rotor position, and duty cycle of Leg A and Leg C. In the other hand, if the duty cycle of two legs defines, then the values of these voltages would be reached.

PMSM control techniques can be dividing into scales and vector control. The problem with scalar is that motor flux and torque in general, are coupled. This inherent coupling affects the response and makes the system drone to instability if it is not considerable. In vector control, not only the magnitude of stator and rotor flux is considered but also their mutual angle.

The vector control of currents and voltages results in control of the spatial orientation of the electromagnetic fields in the machine and it has led to the field, called orientation-field. Oriented control usually refers to controllers which maintain a 90° orientation referred to as field angle control or angle control.

To have an independent controlled orthogonal special angle (electrical) between stator and rotor flux, the angle between stator and rotor field flux should be kept at 90° which leads to:

$$i_d = 0; \quad i_q = i_s. \tag{13}$$

In other words, the desired value of id is zero. Substituting 8-6 in 5-3, represents motor voltage equations in dq-frame as:

$$V_q = ri_q + L_d\omega_r i_q + \lambda_m\omega_r + L_q\frac{\mathrm{d}i_q}{\mathrm{d}t}; \tag{14}$$

$$V_d = ri_d + L_q\omega_r i_q + L_d\frac{\mathrm{d}i_d}{\mathrm{d}t} + \frac{\mathrm{d}\lambda_m}{\mathrm{d}t}. \tag{15}$$

We can write Eq. (14) and Eq. (15) in discreet time, by considering the concept of derivation in discreet time, as:

$$\overline{V_d} = ri_d - L_q\omega_r i_q(n) + L_d\frac{i_d(n+1) - i_d(n)}{T}; \tag{16}$$

$$\overline{V_q} = ri_q(n) - L_d\omega_r i_d(n) + \omega_r\lambda_m + L_q\frac{i_q(n+1) - i_q(n)}{T}. \tag{17}$$

That subscript (n) represents the present time values and $(n + 1)$ values in the next sample. Note that the duration between two samples is equal with T. By considering Eq. (13), desired values of id in the next sample $i_d(n+1)$, considered to be zero. So the value of $\overline{V}_d$ on the basis of d and q axis current in this sample achieves:

$$\overline{V}_d = ri_d(n) - L_q\omega_r i_q(n) - L_d\frac{i_d(n)}{T}. \tag{18}$$

By achieving V_d from Eq. (12), we have:

$$\overline{V}_d = \frac{2}{3}V_{dc}\left[\sin\theta_r\,(2D_a - 1) + \sin\left(\theta_r - \frac{2\pi}{3}\right)2\left(1 - (D_a + D_c)\right) + \sin\left(\theta_r + \frac{2\pi}{3}\right)(2D_c - 1)\right]. \tag{19}$$

By equalize Eq. (18) and Eq. (19) we would have an equation on the basis of D_a and D_c:

$$K_1D_a + K_2D_c = K_3. \tag{20}$$

In the following, by considering the form of equations in discreet time, after a derivation convert the mechanical equation of motor (Eq. (6)) to discreet time, as:

$$\frac{\mathrm{d}^2\omega_r}{\mathrm{d}t^2} = \frac{P}{2J}\left(\frac{\mathrm{d}T_e}{\mathrm{d}t} - F\frac{\mathrm{d}\omega_r}{\mathrm{d}t} - \frac{\mathrm{d}T_m}{\mathrm{d}t}\right); \tag{21}$$

$$\frac{\mathrm{d}T_e}{\mathrm{d}t} = \frac{T_e(n+1) - T_e(n)}{T}; \tag{22}$$

$$\frac{\mathrm{d}^2\omega_r}{\mathrm{d}t^2} = \frac{\mathrm{d}}{\mathrm{d}t}\left(\frac{\mathrm{d}\omega_r}{\mathrm{d}t}\right) = \frac{\mathrm{d}}{\mathrm{d}t}\left(\frac{\omega_r(n+1) - \omega_r(n)}{T}\right) = \frac{\omega_r(n+1) - 2\omega_r(n) + \omega_r(n-1)}{T^2}; \tag{23}$$

$$\frac{\mathrm{d}T_m}{\mathrm{d}t} = \frac{T_m(n) - T_m(n-1)}{T}. \tag{24}$$

By considering this fact that the synchronous motor has a fixed speed, then we would have $\omega_r(n+1)$ as a known parameter of the next sample condition. So, the desired value of Te at next sample $(T_e(n+1))$ will be derived from Eq. (21) to Eq. (24). Also, on the basis of controlling strategy type chosen for FOC, electrical torque of motor could be representing from Eq. (5) as:

$$T_e = \frac{3}{4}P\lambda_m i_q. \tag{25}$$

That derivation of Eq. (25) yields:

$$\begin{aligned} \frac{\mathrm{d}T_e}{\mathrm{d}t} &= \frac{3}{4}P\lambda_m\frac{\mathrm{d}i_q}{\mathrm{d}t} \rightarrow \frac{T_e(n+1) - T_e(n)}{T} = \frac{3}{4}P\lambda_m\frac{i_q(n+1) - i_q(n)}{T}; \\ \Rightarrow i_q(n+1) &= \frac{4}{3P\lambda_m}\left[T_e(n+1) - T_e(n)\right] + i_q(n). \end{aligned} \tag{26}$$

With this fact, if we have the value of desired electrical torque in the next cycle, then the required current of stator q axis to reach that torque represents by Eq. (26). Replacing Eq. (22) to Eq. (24) in Eq. (21) yields:

$$\frac{\omega_r(n+1) - 2\omega_r(n) - \omega_r(n-1)}{T}\frac{2J}{P} = T_e n + 1 - T_e(n) - F\omega_r(n+1) + F\omega_r(n) - T_m(n) + T_m(n-1); \tag{27}$$

$$T_e(n+1) = T_e(n) + T_m(n) - T_m(n-1) + F\omega_r(n+1) - F\omega_r(n) + \frac{2J}{P}\frac{\omega_r(n+1) - 2\omega_r(n) + \omega_r(n-1)}{T}. \tag{28}$$

As it seen, all right side values of Eq. (28) are known. So, by considering the speed of motor and mechanical torque in the next cycle, with a simple calculation we can calculate the required electrical torque. When the electrical torque values in the next cycle obtained, then the q axis current of stator in the next cycle achieves by Eq. (26). On the basis of obtained values in the calculations above, and extracting voltage $\overline{V_q}$ from Eq. (12) we have:

$$\overline{V_q} = \frac{2}{3}V_{dc}\left[\cos\theta_r(2D_a - 1) + \cos\left(\theta_r - \frac{2\pi}{3}\right)2(1 - (D_a + D_c)) + \cos\left(\theta_r + \frac{2\pi}{3}\right)(2D_c - 1)\right]. \tag{29}$$

By equalize two sides of Eq. (17) and Eq. (29) we would have an equation like Eq. (20) on the basis of D_a and D_c that are the unknowns of the system:

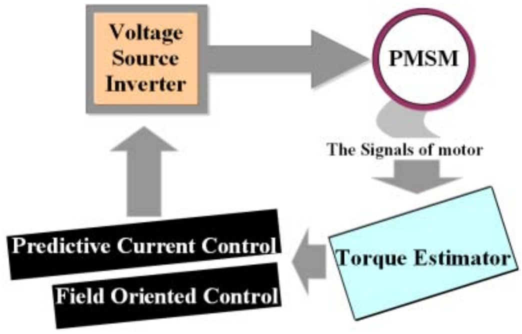
$$K_1^{'}D_a + K_2^{'}D_c = K_3^{'}. \tag{30}$$

By solving the set of Eq. (20) and Eq. (30) as bellow we can calculate Duty cycle values of Legs A and C on the basis of D_a and D_c:

$$K_1D_a + K_2D_c = K_3; \quad K_1^{'}D_a + K_2^{'}D_c = K_3^{'}. \tag{31}$$

Finally, by having D_a and D_c we can calculate Duty cycle values of Leg B with Eq. (10), too:

$$D_b = 1.5 - (D_a + D_c). \tag{32}$$



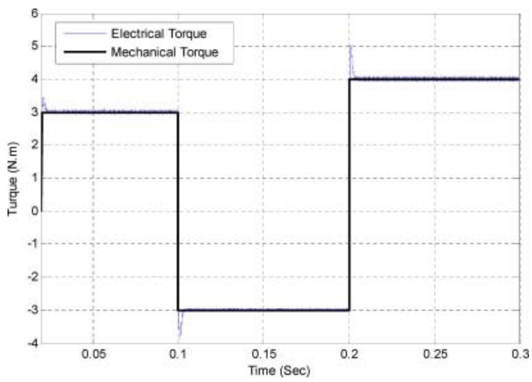


Fig. 3. PMSM electrical torque response to the mechanical torque changes

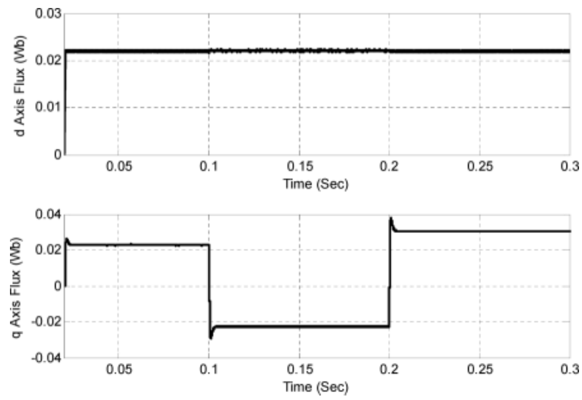


Fig. 4. d-q axis stator flux

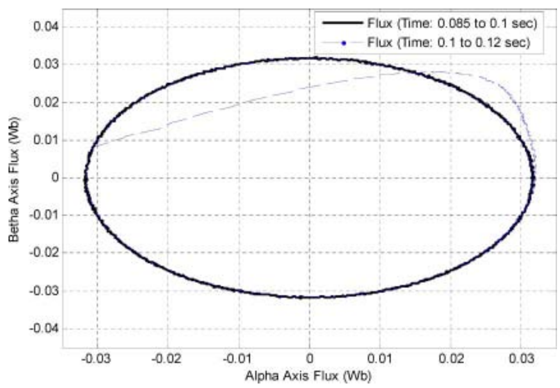


Fig. 5. Stator flux in α-β reference frame during a-6 N.m change of the mechanical torque (step time: 0.1 sec)

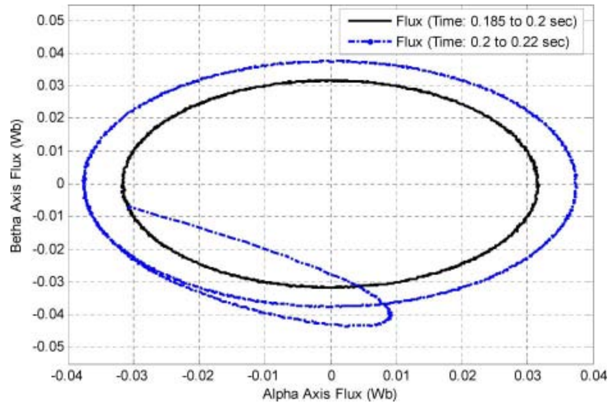


Fig. 6. Stator flux in α-β reference frame during a+7 N.m change of the mechanical torque (step time: 0.2 sec)

So, by having a Duty cycle of each legs and using them on the inverter we can estimate torque of motor predictive on the basis of the motor speed, the torque of mechanical load and stator current situation. The function blocks of FOC-PCC system and the main circuit of PMSM drive are shown in Fig. 3, respectively. FOC-PCC is based on the traditional field oriented control system and the predictive current control technique. The main circuit of Fig. 3 consists of three main parts: inverter permanent magnet synchronous motor and FOC-PCC based controller. As shown in this figure, the proposed controller calculates the duty cycles of the inverter to regulate the torque and flux of the PMSM in two steps:

- Estimation of electrical torque of motor at next sample based on the Eq. (21) to Eq. (24);
- Calculating the duty cycles of the inverter based on the mentioned equations (summarized in Eq. (31), Eq. (32)).

Finally, it should be mentioned that, the applied controller operates in fixed switching frequency.

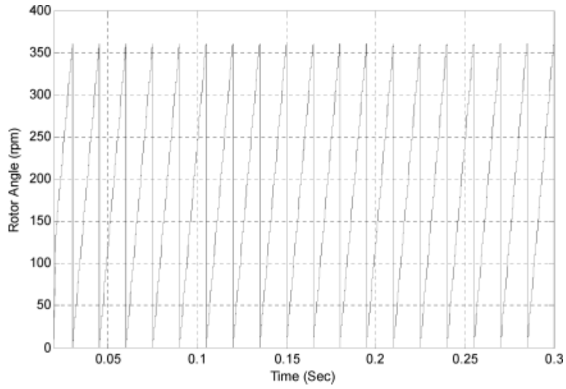


Fig. 7. Rotor position (electrical angle)

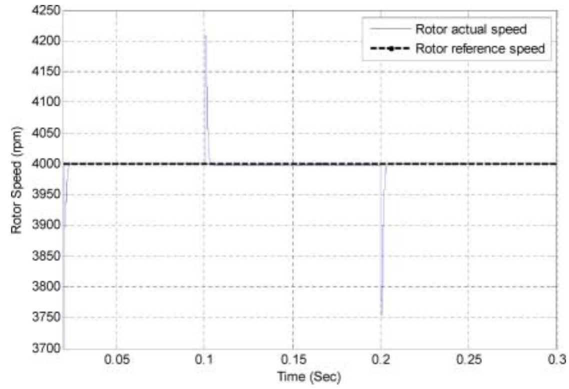


Fig. 8. Mechanical speed response

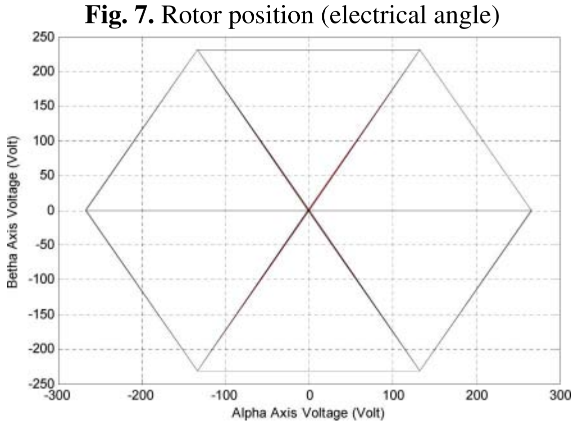


Fig. 9. Stator voltage space vectors

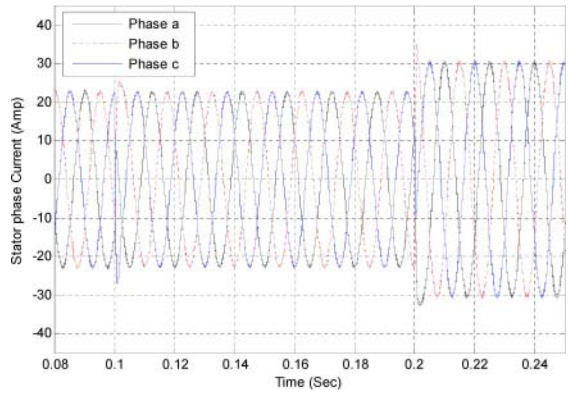


Fig. 10. Stator three phase current

4 Simulation result

In order to validate the proposed model, the dynamic simulation was also carried out using the PMSM model embedded in the Simulink library of MATLAB. The results of simulation of the proposed PMSM drive are shown in Fig. 3 $\sim$ Fig. 10 respectively. All Figures are the responses to step torque command with amplitude of 3, -6 and 7 N.m., which is applied at 0.02, 0.1 and 0.2 sec.

The simulation result at Fig. 3 and Fig. 4 show that flux and torque responses are desirable. Due to basic principle of two-phase stationary α-β reference frame, it is expected that the stator flux has a circular direction. By the way, exiting of a transient state in motor torque is unavoidable. So we want to reduce this transient state. As shown in Fig. 5 and Fig. 6 the proposed FOC-PCC based technique is able to reduce transient states at torque variation times 0.1 and 0.2 sec, respectively.

The rotor speed and its electrical angle are shown in Fig. 7 and Fig. 8, respectively. In addition, the stator voltage space vectors are illustrated in Fig. 9. According to the Fig. 10, the stator current has a considerable variation at mechanical torque changes.

5 Conclusion

The aim of this paper was to give a simple and accurate method in PMSM drive. A new control technique based on conventional FOC is suggested. Also, the proposed control strategy uses predictive current control to estimate the required electrical torque and track the mechanical torque accurately. Several numerical simulation using MATLAB have been carried out in steady-state and transient-state operation conditions. According to the results, the proposed technique is able to reduce torque ripple and its transient-states at mechanical torque variations. The advantages of the proposed method are fixed switching frequency, fast response and low harmonic current.

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