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A novel direct torque and flux control of permanent magnet synchronous motor with analytically-tuned PI controllers

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ABSTRACT

This work presents a novel direct torque and flux control (DTFC) of permanent magnet synchronous motor (PMSM) with analytically-tuned proportional integral (PI) controllers. The proportional (K_p) and integral (K_i) gains of the PI controllers were accurately determined, from first principle, using the model of the control system. The PI flux and torque controllers were then developed in rotor reference frame. The designed PI controllers, together with the torque and flux controllers, were tested on a permanent magnet synchronous motor (PMSM). The results obtained were compared with results from conventional DTFC system using manually-tuned PI controllers. The total harmonic distortion (THD) of motor phase currents is 18.80% and 4.81% for the conventional and proposed models respectively. This confirms a significant reduction in torque ripples. The control system was tested for step torque loading and found to offer excellent performance both during load changes, speed reversal, and constant load conditions.

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1. INTRODUCTION

Permanent magnet synchronous motors (PMSM) have very wide field of application owing to their many advantages such as higher efficiency, power factor, power density, and torque-to-inertia ratio compared to the traditional induction motor often referred to as the workhorse of the industry [1]-[3]. This has resulted in intensified research activities in the design, analysis, and control of PMSM. In most industrial applications where PMSMs are used, high precision in speed and torque is usually required [4]-[9]. It has been established that direct torque control (DTC) offers better dynamic performance compared to field orientation control (FOC) especially for spplications requiring high precision, sensitivity, and minimized torque ripples [10]-[14].

Advancement of DTC has witnessed much attention in the development of control algorithms without considering analytical evaluation of PI controller gains and this has greatly affected precision and speed/torque distortion. Different DTC strategies based on hysteresis control were presented in [15]-[17]. In these works, inverter voltage vectors were selected from hysteresis-based switching tables but the problem of high torque ripples persists. Bo-Wen *et al.* in [18], a DTC method based on active disturbance rejection control was presented while [19] presented DTC in flux weakening region. In the hysteresis controllers were replaced with PI controllers without any analytical method for selection of the proportional and integral gain values of the PI controllers [20]-[22]. The results of these works show various levels of torque ripples which are not suitable for applications requiring minimal ripples in torque and speed. Recently, a model-based analytical evaluation of PI controller gains for field orientation control (FOC) was developed in [23] and tested on a PMSM. The results show that by accurately determining the gains of the PI controller, torque, and speed ripples as well as overshoots were greatly reduced.

The greatest drawback of the traditional DTC is its high torque and flux ripples. However, several researchers have tried different methods in an attempt to reduce these ripples and improve the overall performance of the traditional DTC. Some of these methods include: the hysteresis regulators-based stator voltage vector selection in [16], the active disturbance rejection controller DTC in [18], the MT-frame flux weakening DTC in [19], the SVM-based predictive torque control in [20], the PI based DTC with pole placement technique in [21], and sliding mode control with PI-based DTC in [22]. Apart from the predictive torque algorithm in [21] that significantly reduced the torque ripples, other algorithms did not achieve much in terms of torque and flux ripples reduction.

From the foregoing, it is seen that a clear gap exists in model-based analytical evaluation of PI controller gains for DTC systems. In this work therefore, a PI controller tuned by analytical means, using mathematical model of the control system, is developed and employed in a novel DTC algorithm. The effectiveness of the proposed scheme is demonstrated on a PMSM fed by a space vector pulsewidth modulated voltage source inverter (SV-PWM VSI). MATLAB/Simulink was used for modelling and simulation in this work.

This work is organised is being as: section 1 is Introduction while section 2 presents the PI controller design developed using the machine qd-axes model. Section 3 contains the design of the PI flux and torque controllers while section 4 presents the simulations and results. Section 5 is the conclusion.

2. PI CONTROLLER DESIGN

The dq-axis voltage of a permanent magnet synchronous motor can be written in terms of flux linkage as [23].

$$V_d = Ri_d + \frac{d\lambda_d}{dt} - \omega\lambda_q \tag{1}$$

$$V_q = Ri_q + \frac{d\lambda_q}{dt} + \omega\lambda_d \tag{2}$$

Where

$$\lambda_d = L_d i_d + \lambda_{pm} \tag{3}$$

And

$$\lambda_q = L_q i_q \tag{4}$$

From (3) and (4).

$$i_d = \frac{\lambda_d - \lambda_{pm}}{L_d} \tag{5}$$

$$i_q = \frac{\lambda_q}{L_q} \tag{6}$$

Substituting (5) and (6) into (1) and (2) respectively gives.

$$V_d = R\left(\frac{\lambda_d - \lambda_{pm}}{L_d}\right) + \frac{d\lambda_d}{dt} - \omega\lambda_q \tag{7}$$

$$V_q = R\frac{\lambda_q}{L_q} + \frac{d\lambda_q}{dt} + \omega\lambda_d \tag{8}$$

Laplace transform of (7) and (8) gives.

$$V_d(s) = (\frac{R}{L_d} + s)\lambda_d(s) - \omega\lambda_q(s) - \frac{R\lambda_{pm}}{L_d} \quad (9)$$

$$V_q(s) = (s + \frac{R}{L_q})\lambda_q(s) + \omega\lambda_d(s) \quad (10)$$

From (9)

$$\lambda_d(s) = [V_d(s) + \omega\lambda_q(s) + \frac{R\lambda_{pm}}{L_d}]\frac{1}{(\frac{R}{L_d} + s)} \quad (11)$$

In 11 is represented in block diagram as of Figure 1. It is seen from (1) and (2) that the dq flux control loops are not independent due to the back-emf terms in both equations. To make the dq flux control loops independent, a back-emf decoupling term is introduced. This decoupling term is introduced just after the PI controller to serve as a disturbance with the same value but but having opposite sign as the back-emf term in the motor model. The block diagram of the system is now as shown in Figure 2. Since the decoupling term will cancel the effect of the back emf term, Figure 2 reduces to Figure 3. There will be both sensor delay and computation time delay in the system; a first order time delay is introduced to modify Figure 3 as shown in Figure 4.

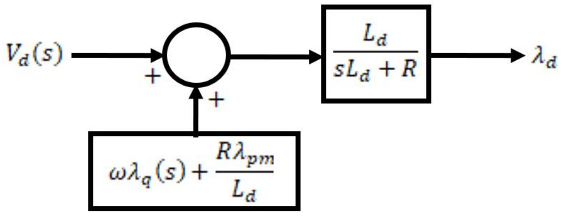

Figure 1. Open loop block diagram of the flux control model

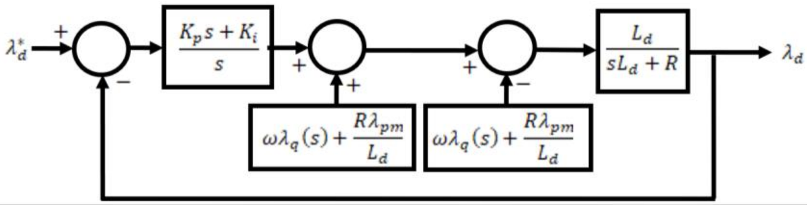

Figure 2. Closed-loop block diagram of decoupled flux control

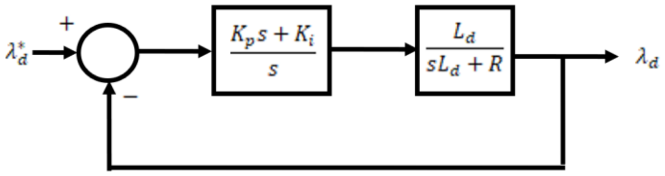

Figure 3. Closed-loop block diagram of the flux control model

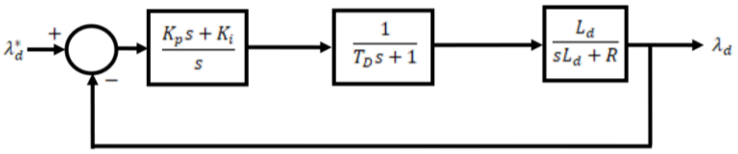

Figure 4. Closed-loop block diagram with time delay

The open loop transfer function, G_{OL}, is given by (12) as shown in.

$$G_{OL}(s) = \frac{K_p s + K_i}{s} . \frac{1}{T_D s + 1} . \frac{L_d}{L_d s + R} \tag{12}$$

If the zero of the PI controllers $\left(-K_i/K_p\right)$ is designed to cancel the pole $\left(-R_a/L_d\right)$ of the PMSM by pole-zero cancellation method, then we let:

$$K_{ip} = \frac{K_i}{K_p} = \frac{R}{L_d} \tag{13}$$

From (13),

$$K_i = K_p K_{ip} \tag{14}$$

Substituting (14) into (12) gives.

$$G_{OL}(s) = \frac{K_p s + K_p K_{ip}}{s} . \frac{1}{T_D s + 1} . \frac{1}{s + \frac{R}{L_d}} \tag{15}$$

$$G_{OL}(s) = \frac{K_p}{T_D} \frac{1}{s(s + \frac{1}{T_D})} \tag{16}$$

Let K = $\frac{K_p}{T_D}$, so that (16) becomes.

$$G_{OL}(s) = \frac{K}{s(s + \frac{1}{T_D})} \tag{17}$$

The closed-loop transfer function, G_{CL} is given as.

$$G_{CL} = \frac{G_{OL}}{1 + G_{OL}} = \frac{K}{s^2 + \frac{s}{T_D} + K} \tag{18}$$

The general equation of a second-order system is [24], [25].

$$H(s) = \frac{\omega_n^2}{s^2 + 2\delta\omega_n s + \omega_n^2} \tag{19}$$

By comparing (18) and (19), we obtain.

$$\omega_n^2 = K \ or \ \omega_n = \sqrt{K} \tag{20}$$

And

$$2\delta\omega_n = \frac{1}{T_D} \ or \ \delta = \frac{1}{2T_D\sqrt{K}} \tag{21}$$

Where ω_n is the natural frequency and δ is the damping ratio. The maximum overshoot of a second order system is given by [23].

$$M_p = e^{-\pi\delta/\sqrt{1-\delta^2}} \tag{22}$$

So, by choosing the maximum percentage overshoot allowed and the value of T_D; K_p and K_i can be calculated using (20)-(22).

3. DESIGN OF FLUX AND TORQUE CONTROLLERS

3.1. PI flux controller model

In rotor reference frame, the flux of a PMSM is controlled by d-axis current. So, the PI flux controller is designed to produce the accurate d-axis reference voltage from d-axis flux error which is its input. Thus, to obtain a flux controller model we make use of the d-axis voltage (7).

From (7).

$$V_d = \mathrm{R}\,(\frac{\lambda_d - \lambda_{pm}}{L_d}) + \frac{d\lambda_d}{dt} - \omega\lambda_q = \frac{R}{L_d}(\lambda_d - \lambda_{pm}) + \frac{d\lambda_d}{dt} - \omega\lambda_q \qquad (23)$$

Let the time rate of change of d-axis flux, $\frac{d\lambda_d}{dt}$, be equal to d-axis flux error per sample time, $\frac{e_d}{T_s}$, where e_d is the d-axis flux error and T_s is the sample time.
Therefore,

$$\frac{d\lambda_d}{dt} = \frac{e_d}{T_s} \qquad (24)$$

Taking integral of 24 gives 24.

$$\lambda_d = \frac{1}{T_s}\int e_d\, dt \qquad (25)$$

Substituting (24) and (25) into (23) gives.

$$V_d = \frac{R}{T_s L_d}\int e_d\, dt - \frac{R}{L_d}\lambda_{pm} + \frac{e_d}{T_s} - \omega\lambda_q \qquad (26)$$

Since $K_{ip} = \frac{K_i}{K_p} = \frac{R}{L_d}$

In (25) is re-written as.

$$V_d = \frac{K_i}{T_s K_p}\int e_d\, dt - \frac{K_i}{K_p}\lambda_{pm} + \frac{e_d}{T_s} - \omega\lambda_q \qquad (27)$$

In (26) is the required PI flux controller model.

3.2. PI torque controller model

The torque equation of a PMSM is given as [23].

$$T_e = 1.5p(\lambda_{pm} i_q + (L_d - L_q) i_d i_q) \qquad (28)$$

Where p is the pole pair.
Since $L_d = L_q$ for non-salient pole machine, then.

$$T_e = 1.5p\lambda_{pm} i_q \qquad (29)$$

In (28) shows that electromagnetic torque of a non-salient pole PMSM is directly proportional to the q-axis current, but from in (4),

$$\lambda_q = L_q i_q \Rightarrow i_q = \frac{\lambda_q}{L_q} \qquad (30)$$

Substituting (29) into (28).

$$T_e = 1.5p\lambda_{pm}\frac{\lambda_q}{L_q} \qquad (31)$$

Differentiating both sides of (30) with respect to time gives.

$$\frac{dT_e}{dt} = \frac{1.5p\lambda_{pm}}{L_q}\frac{d\lambda_q}{dt} \qquad (32)$$

Let the time rate of change of torque, $\frac{dT_e}{dt}$, be equal to the torque error per sample time, e_t. So

$$\frac{d\lambda_q}{dt} = \frac{L_q}{1.5p\lambda_{pm}} e_t \qquad (33)$$

Similarly, the PI torque controller is designed to produce the exact q-axis reference voltage from its input, which is torque error. Since the torque of a PMSM in rotor reference frame is controlled by q-axis current, the PI torque controller model is obtained by modifying the q-axis voltage (in (8)). From (8).

$$V_q = R\frac{\lambda_q}{L_q} + \frac{d\lambda_q}{dt} + \omega\lambda_d \tag{34}$$

Let the time rate of change of q-axis flux, $\frac{d\lambda_q}{dt}$, be equal to the q-axis flux error per sample time. That is;

$$\frac{d\lambda_q}{dt} = \frac{e_q}{T_s} \tag{35}$$

$$\lambda_q = \frac{1}{T_s}\int e_q\, dt \tag{36}$$

In (32) and (35) show that the q-axis flux error per sample time is directly proportional to torque error. So, by multiplying torque error by $\frac{L_q}{1.5p\lambda_{pm}}$, the q-axis flux error per sample time $\frac{e_q}{T_s}$ is obtained. Therefore, substituting (35) and (36) into (34), (35) is obtained.

$$V_q = \frac{K_i}{T_s K_p}\int e_q\, dt + \frac{e_q}{T_s} + \boldsymbol{\omega}\lambda_d \tag{37}$$

The block diagram of the conventional DTFC as reported in [21] is shown in Figure 5 while the block diagram of the proposed DTFC with analytically-tuned PI controller is shown in Figure 6.

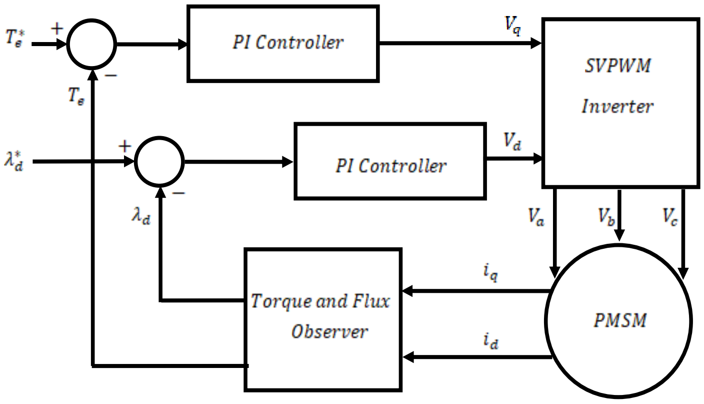

Figure 5. Block diagram of the normal or conventional DTFC system

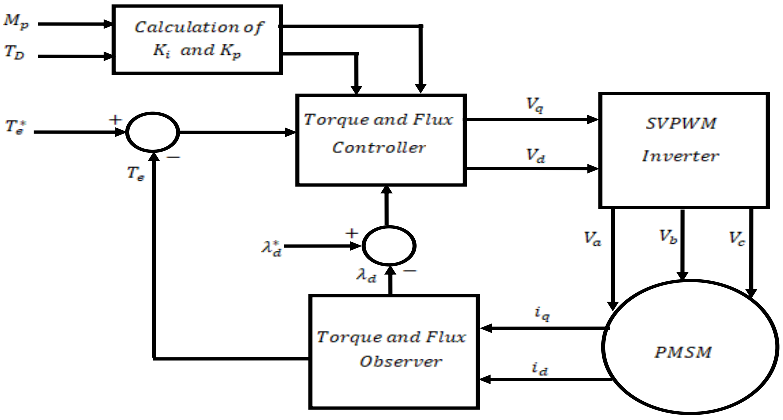

Figure 6. Complete block diagram of the proposed DTFC system

4. SIMULATIONS AND RESULTS

The conventional and the proposed models were modelled and simulated in MATLAB/Simulink environment and tested on a PMSM with the parameters shown in Tabel 1. The designed controller calculated gains of K_p = 326.72 and K_i = 31583 while the normal PI controller gains were tuned to K_p = 5.75 and K_i = 150 in the simulation. The reference torque varies from 5 Nm to -5 Nm and returns to 5 Nm during a total simulation time of 5 seconds. The results show that the proposed model effectively tracked the reference torque with almost no ripples while the conventional model produced much torque ripples.

Figure 7 compares the plot of torque response of the proposed and conventional models. It is seen that while the conventional PI controller produced much torque ripples, the proposed model gave almost no torque ripples. Another strength of the proposed model is its ability to calculate the gains of the controller rather than relying on the error-prone and time-wasting tuning methods used in normal PI controllers. Figure 8 compares the phase currents of the two models. A difference in phase currents, with that of the proposed model being higher and with lower total harmonic distortion (THD) is observed.

The disparity in phase currents is explained by Figure 9 which shows the i_q and i_d currents. It is seen in Figure 9 that the torque producing current (i_q) is essentially the same for both models. However, the d-axis current of the proposed model is higher than the normal one. This is because the two models are different and the proposed model uses higher gains. So, the control effort pushes any difference resulting from the higher gain to the d-axis current in order to ensure that the response effectively tracks the command torque. For these reasons, the phase currents which are combinations of d-axis and q-axis currents (by inverse Park transform) is different for the two models. The FFT for phase ‘a’ current for the conventional and the proposed DTFC are shown in Figures 10 and 11 respectively. The proposed model shows superior performance with THD of 4.81% compared to THD of 18.80% for the conventional DTFC.

Table 1. The models conventional and simulated in MATLAB/Simulink

Motor parameters	Value
Rated Power	9.4 kW
Frequency	50 Hz
Stator resistance (Rs)	0.203 Ω
Constamt rotor flux linkage(λ_f)	0.123Wb
Inductance d axis (L_d)	0.0021H
Inductance q axis (L_q)	0.0021H
Inertia constant (J)	0.0048Kgm^2
No. of poles (P)	4

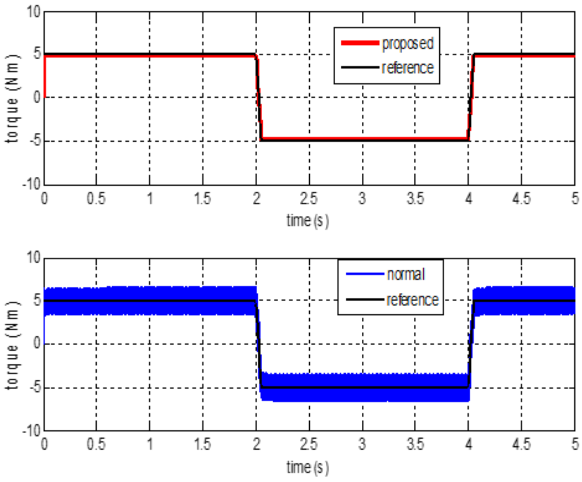

Figure 7. Torque response of the two model

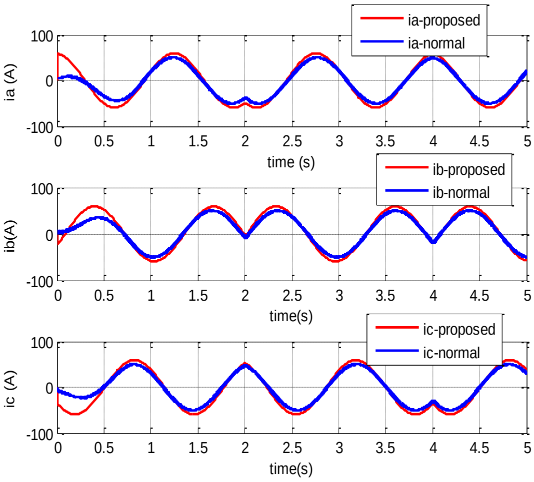

Figure 8. Stator phase currents

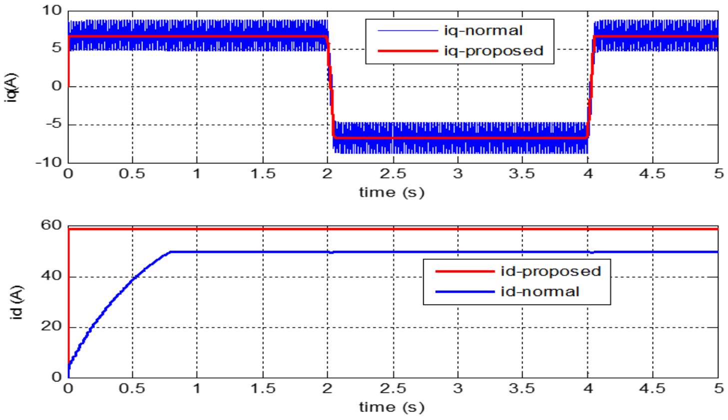

Figure 9. dq-axes currents

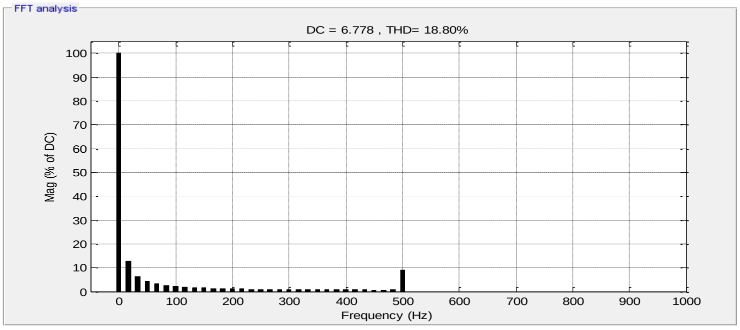

Figure 10. FFT analysis of phase ‘a’ current for conventional DTFC

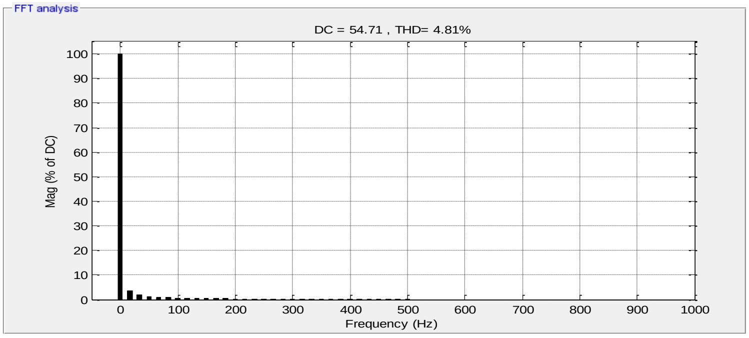

Figure 11. FFT analysis of phase ‘a’ current for proposed DTFC

5. CONCLUSION

This work has presented a novel direct torque and flux control in which the PI controllers were tuned analytically from first principle and compared with a conventional DTFC system that employs manually-tuned PI controllers. The accuracy of both the control model and the PI controller tuning method is evident in the results which show excellent tracking of the reference torque with no overshoot and less ripples than the one with normal PI controllers. By being able to calculate its controller gains; the developed algorithm saves the control engineer the trouble of tuning PI controllers by trial-and-error method and also performs better than conventional PI controllers. The developed model can be adopted for speed or torque control of PMSM. The overall objectives of the work have been achieved.

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