

ROBUST FEEDFORWARD TORQUE CONTROL OF SURFACE MOUNTED PMSM USING DISCRETE EVENT SYSTEM APPROACH

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ABSTRACT: *The proposed paper consider surface mounted permanent magnet synchronous motor drive (PMSM) as a discrete event system (DES), with discrete state and continuous dynamics at each state. This provides to express the quadrature voltages which are in terms of parks transformation to express explicitly in terms of discrete and continuous parameter of the system. More distantly by solving the d-q modeling equations of PMSM, a closed form expressions for flux and torque are derived which show precisely the effect of discrete state and the continuous system variables. This model is then used in an indirect torque control scheme (IDTC) of PMSM. Indirect Torque Control being in the nature of a feed forward control scheme through simulation results in fast response than conventional Direct Torque control (DTC) scheme in terms of transient speed response, load disturbance rejection, robustness and in dealing with uncertainty in machine parameters. In the proposed scheme, the flux and torque are estimated through switching parameters instead of using current measurement as in DTC thus eliminating the effect of noise. Also, a hardware implementation of the proposed IDTC is shown to perform favorable compared to conventional DTC. Another novel feature of the proposed IDTC scheme is that the current is not measured for load rejection as in conventional DTC thus circumventing the effect of noise in the current. The results can be extended to other model based control schemes, such as, internal model control, predictive control etc.*

Keywords: *Discrete event systems, Indirect Torque Control, Modeling, Permanent magnet motors, Space Vector Modulation.*

1. INTRODUCTION

Conventionally, in feedback control the effect of the control effort on the electrical drives is only felt after feedback is measured. On the other hand, a feedforward control strategy can use the model of the machine drive to provide, in advance, the estimated response of the machine to load and set point

changes thus providing a method for faster correction. Since feedback has always its merits of cancelling unknown disturbance, nonlinearity etc., feedforward- feedback control scheme is usually followed.

In this paper a model based indirect control scheme of PMSM is proposed. Being model based, the indirect control scheme is in the nature of feedforward control. There are many examples of application of indirect control. Sensorless control and Indirect field oriented control are examples of indirect control strategy and have found many successful applications in control of electrical drives [for example, 1, 2]. Indirect field oriented control has also been applied in grid connected [3] applications where the speed is estimated instead of directly measuring it thus reducing considerably the cost and complexity of the drive. In fault recovery control, speed response is an important criterion. Indirect control has been employed to control induction machine for fast recovery after faults [4]. Speed is estimated indirectly using Model Reference Adaptive Control Scheme (MRAS) in [5]. The drawback of employing MRAS alone is the inherent ripple when the model parameters are inaccurate and no rigorous proof of stability due to high system order and non-linear dynamics is available. Indirect torque control has also been used in switched reluctance motor [6] for the implementation of an iterative learning strategy. In indirect field oriented control of Induction motor, the instability caused by model errors has been corrected by using an adaptive fuzzy logic controller based on Levenberg-Marquardt algorithm [7] which is claimed to be more accurate and robust. In paper [8] the author proposes wavelet based fuzzy controller to control the speed of induction motor using indirect field oriented control where the PID controller is replaced by wavelet-fuzzy controller. The author models the source of

error in terms of low and high frequency components in order to minimize the effect of disturbance and noise and get accurate control of speed.

In this proposed paper, we consider the states of the DES to correspond to the sector position in a space vector pulse width modulation (SVPWM) scheme. Further, at each state of DES, the switching parameters of SVPWM are defined. The overall effect of the modeling thus, is to obtain expression for torque and flux in closed form as function of discrete state and continuous system variables of surface mounted PMSM ($L_s = L_d = L_q$). This solution is used to compute torque and flux of the machine using a look-up table for fast correction at the next switching state thus circumventing the problem state explosion and computation complexity issues in real time control using a DES approach as encountered in [9]-[13]. A very preliminary and incomplete version of the approach used in this paper appeared in [14].

Our closed form solution and the look-up table approach will help to improve the computation burden of the controller. However, one may ask that with the availability of software/hardware tools with enormous computational capabilities, is the closed form solution always needed? First, motor control systems operating in real time are relatively fast requiring a fast response controller. Our proposed scheme with the closed form solution and the state – based (DES) approach will certainly help in such a case. Further, the motivation for our modeling approach over the existing models is that with the proposed model we can actually predict the torque and flux generated for different switching conditions from a given state of the system simply by substituting the switching parameters into the closed form solution. This prediction can be used for fast correction in a feedback loop similar to a feedforward control scheme. Moreover, compared to the existing models, the proposed model explicitly exhibits the effect of the switching and continuous dynamics of the machine separately on the machine output in terms of flux and torque generated. The latter result may throw new insight into the problem of control of PMSM. Finally, the approach can also be considered as a generalization of sensorless control for PMSM which is a popularly applied technique for the controlling of PMSM.

The main contribution of the paper is as follows: i) Inverter and PMSM are modelled together as a DES/Hybrid system ii) Closed form solution of torque and flux is derived which explicitly indicates

the contribution of discrete and continuous variables iii) The closed form solution is used in the form of a lookup table for fast feedforward correction iv) Using the space vector of the drive for the state based approach avoids the state explosion problem typical in a DES approach and the closed form solution avoids the computation complexity issues in a real time context iv) The performance of the proposed indirect torque control scheme is compared favorably in terms of speed of response, fast disturbance rejection and robustness with respect to the conventional direct torque control scheme through simulation and hardware implementation.

The rest of the paper is organized as follows: The following section describes PMSM as a DES. Section 3 describes the theoretical analysis leading to the closed form expressions for torque and flux. Section 4 considers the development of the indirect torque control scheme for the speed control of PMSM. Section 5 describes the simulation results of the proposed scheme using Matlab/ Simulink and comparison of IDTC and DTC performances. Section 6 describes the experimental setup of both IDTC and DTC and their comparison. Section 7 concludes the paper.

2. PMSM AS A DES

The transition structure of PMSM as a DES [11, 12] is shown in Fig 1. Each state q_i , $i = 1, 2, \dots, 6$ corresponds to a 60° sector in which the space vector position lies. Transition from one state to another takes place as the space vector rotates through 360° . The event is denoted by $\sigma_i = 1, 2, \dots, 6$ in Fig.1. The cross line on a transition arrow denotes that the transition is controllable. There is also an uncontrollable event which corresponds to the effect of disturbance or fault conditions where the space vector may reverse its direction and the corresponding transition is denoted by γ_i , $i = 1, 2, \dots, 6$ as shown in Fig.1. Each state q_i , $i = 1, 2, \dots, 6$ in turn gives rise to the switching states (S_{abc}) as shown for state q_6 only in Fig.1.

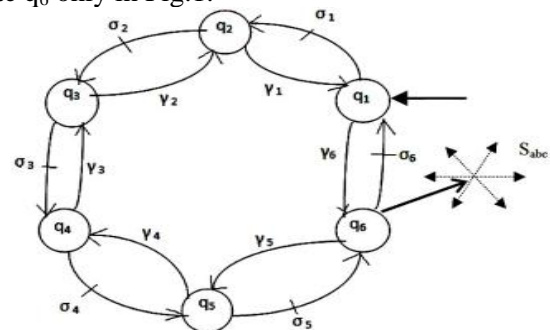


Fig.1. Transition Structure of PMSM as a DES

The formal structure of a DES representation of PMSM under space modulation scheme can be defined as a generator G [2] as follows

$$G = (Q, \Sigma, \delta, q_0)$$

where $Q = \{q_i\}_{i=1}^6$, $\Sigma = \{\sigma_i, \gamma_j\}_{i,j=1}^6$, $\delta: Q \times \Sigma \rightarrow Q$ is a partial transition function and q_0 is the initial state (shown by an arrow at q_1 in Fig.1).

3. ANALYSIS OF TORQUE AND FLUX

We formulate the first-order coupled differential equation of the PMSM. We then derive a closed form solution of the equation which exhibits explicitly the effect of switching and the continuous dynamics of the system.

3.1 Computation of Voltage input to PMSM:

We use the following notation for the phase and quadrature variables representing voltage, current, flux, switching states etc.

$$f_{abc} = [f_a \ f_b \ f_c]^t$$

$$f_{qd} = [f_q \ f_d]^t$$

where 't' stands for transpose/

The three phase voltages V_{abc} can be expressed in terms of switching states S_{abc} as follows,

$$V_{abc} = \frac{V_{dc}}{3} \begin{pmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{pmatrix} S_{abc} \quad (1)$$

Applying Parks transformation, $T_{qd}(\theta_e)$ given by

$$T_{qd}(\theta_e) = \begin{bmatrix} \cos \theta_e & \cos(\theta_e - \frac{2\pi}{3}) & \cos(\theta_e + \frac{2\pi}{3}) \\ \sin \theta_e & \sin(\theta_e - \frac{2\pi}{3}) & \sin(\theta_e + \frac{2\pi}{3}) \end{bmatrix} \quad (2)$$

to (1) yields, after simplification, including segregation into orthogonal terms,

$$V_q = V_{dc}(\alpha_1 \cos \theta_e + \beta_1 \sin \theta_e) \quad (3a)$$

$$V_d = V_{dc}(\alpha_2 \cos \theta_e + \beta_2 \sin \theta_e) \quad (3b)$$

where $\alpha_i, \beta_i, i=1, 2$ are such that

$$\alpha_1 = \beta_2, \alpha_2 = -\beta_1, \forall S_{abc}, a, b, c \in \{0,1\} \quad (4)$$

Using (4), (3a) and (3b) can be put as

$$V_q = V_{dc}(\alpha_1 \cos \theta_e - \alpha_2 \sin \theta_e) \quad (5a)$$

$$V_d = V_{dc}(\alpha_2 \cos \theta_e - \alpha_1 \sin \theta_e) \quad (5b)$$

where α_1 and α_2 are given in Table 1 for different switching states S_a, S_b, S_c .

Table 1 α_1 and α_2 values under various switching states

Remark I: V_q, V_d given in 5(a) and 5(b) are the machine supply voltage quadrature variables in terms of switching states (represented by α_1 and α_2) and the continuous variable θ_e , thus explicitly showing the influence of the discrete and the continuous parts of the dynamics. This is a new result not available in the literature.

3.2. Solution of Torque and Flux

For idealized surface mounted permanent magnet motor, the instantaneous voltage equation for the stator windings can be expressed as in (6) using simplified voltage source inverter and PMSM model.

$$V_{abc} = [R_s] i_{abc} + \frac{d}{dt} \lambda_{abc} \quad (6)$$

where

$$V_{abc} = [V_a \ V_b \ V_c]^t,$$

$$i_{abc} = [i_a \ i_b \ i_c]^t, \lambda_{abc} = [\lambda_a \ \lambda_b \ \lambda_c]^t$$

and

$$[R_s] = \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_s \end{bmatrix}$$

V_{abc} , i_{abc} and λ_{abc} represent phase voltages, phase currents and flux linkages respectively. R_s represent the equivalent resistance of a phase winding.

For the purpose of the mathematical analysis of PMSM we assume, for simplicity, that the flux linkages from the PM are pure sinusoids with constant amplitude. The flux linkage in equation (6) can be expressed as follows:

$$\lambda_{abc} = [L_s] i_{abc} + \lambda_m \theta_{abc} \quad (7)$$

where

$$[L_s] = \begin{bmatrix} L_s & 0 & 0 \\ 0 & L_s & 0 \\ 0 & 0 & L_s \end{bmatrix},$$

$$\theta_{abc} = \begin{bmatrix} \sin \theta_e \\ \sin(\theta_e - \frac{2\pi}{3}) \\ \sin(\theta_e + \frac{2\pi}{3}) \end{bmatrix}$$

L_s is the equivalent inductance of phase winding (for simplicity, mutual flux linkage is neglected), θ_e is the electrical angular position of the rotor, and λ_m is the amplitude of the flux linkage which is established by the permanent magnet.

Combining (6) and (7), and eliminating i_{abc} , we have

$$V_{abc} = [R_s][L_s]^{-1} \{\lambda_{abc} - \lambda_m \theta_{abc}\} + \frac{d}{dt} \lambda_{abc} \quad (8)$$

The phasor diagram of 3- ϕ PMSM in (α - β) coordinates and (d-q) coordinates is shown in Fig.2.

Switching States			α_1	α_2
S_a	S_b	S_c		
0	0	0	0	0
1	0	0	2/3	0
1	1	0	1/3	1/√3
0	1	0	-1/3	1/√3
0	1	1	-2/3	0
0	0	1	-1/3	-1/√3
1	0	1	1/3	-1/√3
1	1	1	0	0

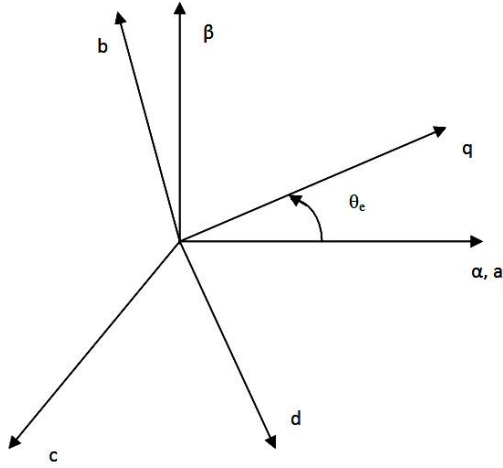


Fig.2. Phasor diagram of PMSM

Using Park's transformation, the voltage expression in (8) expressed in a rotating reference frame can be expressed in simplified form as

$$V_{qd} = [R_s][L_s]^{-1} \left\{ \lambda_{qd} - \lambda_m \begin{bmatrix} 0 \\ 1 \end{bmatrix} \right\} + T_{qd}(\theta_e) \frac{d}{dt} \lambda_{abc} \quad (9)$$

where

$$V_{qd} = [V_q \ V_d]^t; \lambda_{qd} = [\lambda_q \ \lambda_d]^t$$

Putting,

$$\frac{d}{dt} \{ T_{qd}(\theta_e) \} \lambda_{abc} = T_{qd}(\theta_e) \frac{d}{dt} \lambda_{abc} + \frac{d}{dt} T_{qd}(\theta_e) \lambda_{abc} \quad (10)$$

and substituting (10) into (9) and simplifying we can get, the quadrature flux variables in differential equation form as

$$V_{qd}' = \frac{d}{dt} \lambda_{qd} + K \lambda_{qd} \quad (11)$$

where

$$V_{qd}' = \left(V_q \ \left\{ V_d + \frac{\lambda_m}{\tau_s} \right\} \right)^t$$

$$K = \begin{bmatrix} 1/\tau_s & \omega_e \\ -\omega_e & 1/\tau_s \end{bmatrix}$$

$$\tau_s = \frac{L_s}{R_s}$$

Using 3(a) and 3(b), we can write

$$V_{qd}' = \underline{\alpha} \cos \theta_e + \underline{\beta} \sin \theta_e + \begin{bmatrix} 0 \\ \frac{\lambda_m}{\tau_s} \end{bmatrix} \quad (12)$$

where $\underline{\alpha} = (\alpha_1 \ \alpha_2)^t$; $\underline{\beta} = (\beta_1 \ \beta_2)^t$

Using the method of Undetermined Coefficients [19], the particular solution for λ_{qd} of (11) can be found, assuming ω_e to be a constant, as

$$V_{qd}' = \sin \theta_e (-\underline{m} \omega_e + K \underline{n}) + \cos \theta_e (\underline{n} \omega_e + K \underline{m}) + K \underline{c} \quad (13)$$

where

$$\underline{m} = M(K \underline{\alpha} - \omega_e \underline{\beta}) \quad (14)$$

$$\underline{n} = M(\omega_e \underline{\alpha} + K \underline{\beta}) \quad (15)$$

$$\underline{c} = \frac{\lambda_m}{(1 + \omega_e^2 \tau_s^2)} \begin{bmatrix} -\omega_e \tau_s \\ 1 \end{bmatrix} \quad (16)$$

and

$$M = \frac{\tau_s^2}{(1 + 4\omega_e^2 \tau_s^2)} \begin{bmatrix} 1 & -2\omega_e \tau_s \\ 2\omega_e \tau_s & 1 \end{bmatrix} \quad (17)$$

The solution for λ_{qd} of (11) can be expressed, after rearrangement, as

$$\lambda_{qd} = (MK \cos \theta_e + M \omega_e \sin \theta_e) \underline{\alpha} + (MK \sin \theta_e - M \omega_e \cos \theta_e) \underline{\beta} + \frac{\lambda_m}{(1 + \omega_e^2 \tau_s^2)} \begin{bmatrix} -\omega_e \tau_s \\ 1 \end{bmatrix} \quad (18)$$

By substituting $\beta_1 = -\alpha_2$, $\beta_2 = \alpha_1$ as in (4) into (18), we get after simplification

$$\begin{bmatrix} \lambda_q \\ \lambda_d \end{bmatrix} = \begin{bmatrix} \tau_s V_{dc} (\alpha_1 \cos \theta_e - \alpha_2 \sin \theta_e) - \frac{\lambda_m \omega_e \tau_s}{(1 + \omega_e^2 \tau_s^2)} \\ \tau_s V_{dc} (\alpha_1 \sin \theta_e + \alpha_2 \cos \theta_e) + \frac{\lambda_m}{(1 + \omega_e^2 \tau_s^2)} \end{bmatrix} \quad (19)$$

where α_1, α_2 are as given in Table 1 in section 3.1.

Equation (19) represents flux linkage in d-q frame. Converting the flux linkage to a stationary α - β frame as in Fig.2, we have

$$\begin{bmatrix} \lambda_\alpha \\ \lambda_\beta \end{bmatrix} = \begin{bmatrix} \tau_s V_{dc} \alpha_1 + \frac{\lambda_m}{(1 + \omega_e^2 \tau_s^2)} (\sin \theta_e - \omega_e \tau_s \cos \theta_e) \\ -(\tau_s V_{dc} \alpha_2 + \frac{\lambda_m}{(1 + \omega_e^2 \tau_s^2)} (\cos \theta_e + \omega_e \tau_s \sin \theta_e)) \end{bmatrix} \quad (20)$$

The stator flux linkage can be obtained as

$$\lambda = \sqrt{\lambda_\alpha^2 + \lambda_\beta^2} = \sqrt{a^2 + b^2 + 2ab \sin \Psi'} \quad (21)$$

where

$$a = \frac{2}{3} \tau_s V_{dc}, \quad b = \frac{\lambda_m}{\sqrt{(1 + \omega_e^2 \tau_s^2)}}$$

$$\Psi' = (\theta_e + \Psi)$$

$$\Psi = \tan^{-1} \left(\frac{\alpha_2 - \alpha_1 \omega_e \tau_s}{\alpha_1 + \alpha_2 \omega_e \tau_s} \right)$$

By applying Park's transformation to (7) and simplifying, the direct and quadrature axes current can be expressed as

$$\begin{bmatrix} I_q \\ I_d \end{bmatrix} = \frac{1}{L_s} \begin{bmatrix} \lambda_q \\ \lambda_d \end{bmatrix} - \frac{\lambda_m}{L_s} \begin{bmatrix} 0 \\ 1 \end{bmatrix} \quad (22)$$

Similarly stator current in stationary α - β frame can be expressed as

$$\begin{bmatrix} I_\alpha \\ I_\beta \end{bmatrix} = \frac{1}{L_s} \begin{bmatrix} \lambda_\alpha \\ \lambda_\beta \end{bmatrix} - \frac{\lambda_m}{L_s} \begin{bmatrix} \sin \theta_e \\ -\cos \theta_e \end{bmatrix} \quad (23)$$

Expressing torque [15] as

$$T_e = \left(\frac{3}{2} \right) \left(\frac{P}{2} \right) (\lambda_\alpha I_\beta - \lambda_\beta I_\alpha) \quad (24)$$

and using (20) and (23), we get

$$T_e = \left(\frac{3}{2}\right) \left(\frac{P}{2}\right)_{L_S} \lambda_m (\tau_s V_{dc} (\alpha_1 \cos \theta_e - \alpha_2 \sin \theta_e) - \frac{\lambda_m \omega_e \tau_s}{(1 + \omega_e^2 \tau_s^2)}) \quad (25)$$

Remark II: Equation (20) & (25) express the flux and torque in closed form explicitly showing the effect of switching variables α_1 and α_2 and continuous variable θ_e which is valid under transient and steady state condition. This is a new result not available in literature. The result is made possible by considering the inverter – machine drive as a DES/hybrid system and solution of the dynamic equations involved. Also, the closed form solution for flux and torque can be used for a given DES state q_i , $i=1, 2, \dots, 6$ and switching condition S_{abc} for fast computation in a model based control scheme.

4. INDIRECT TORQUE CONTROL OF PMSM

The block diagram Fig.3 describes the indirect torque control of PMSM using the proposed approach. The method described in section 3 is used to determine the voltage input at every sector as per switching parameter selected. The estimator block then computes the flux and torque as a function of the state and switching parameter.

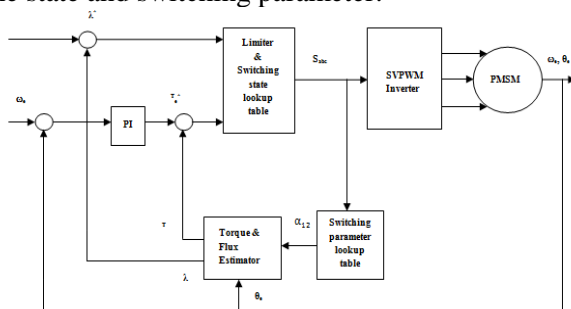


Fig.3. Block diagram of Indirect Torque Control Scheme

The specific algorithmic steps involved in the computation of control inputs are given below:

1. Based on flux error and torque error, the switching states S_a, S_b, S_c are determined using look up table as in conventional DTC.
2. The switching state is given to (the new additional) look up Table 1 where switching parameters α_1 and α_2 values are obtained.
3. Quadrature voltages V_q and V_d using (5a) and (5b) respectively are then estimated.
4. Flux and torque are also estimated using (20) and (25) respectively.
5. The estimated flux and torque are compared with the setpoint flux and torque and the flux and torque error are determined and the cycle repeats.

5. SIMULATION RESULTS & DISCUSSIONS

5.1 Motor Specification

Simulation has been carried to evaluate the proposed scheme using Matlab/Simulink package.

A PMSM machine with following parameters is simulated.

Stator Resistance	= 4.765 Ω
Direct axis inductance (L_d)	= 0.014H
Quadrature axis inductance (L_q)	= 0.014 H
No. of pole pairs	=2
Flux linkage	= 0.1848Wb

5.2 Closed loop responses

Fig.4&5 show the various responses like flux angle, generated machine torque, estimated torque, speed and flux linkage for DTC and IDTC respectively. Considering the closed loop speed control system response using DTC and IDTC approaches, fig.6 & 7 show the speed and the corresponding torque response. The optimum PI controller settings based on smooth and fast response with minimum overshoot are set as $K_p=0.008$, $K_i=0.1$ for DTC and $K_p = 0.05$, $K_i = 0.1$ for IDTC respectively. Fig.6 & 7 clearly illustrate that IDTC responds for sudden speed variation faster than DTC under no load condition.

Fig.8 & 9 compare the performances of DTC and IDTC in terms of speed and torque responses respectively under load disturbance of 1.7 Nm . It is clearly seen that IDTC responses are faster than DTC.

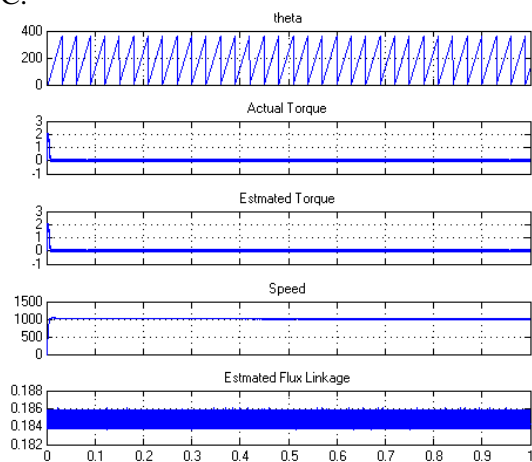


Fig.4 Various responses for no load condition using DTC approach

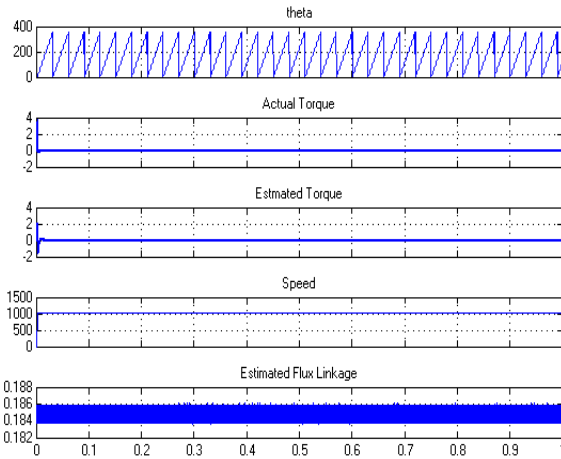


Fig.5 Various responses for no load condition using IDTC approach

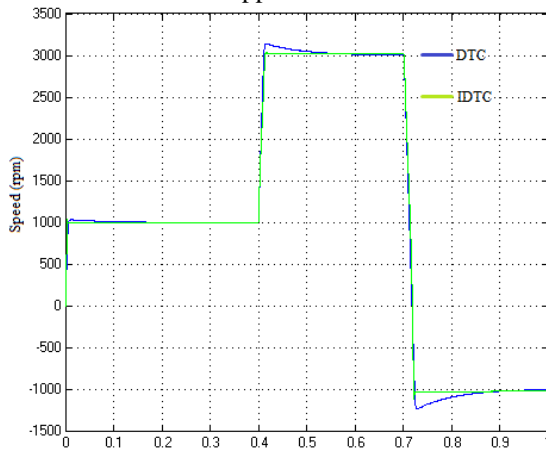


Fig.6 Speed response for different speed condition under no load

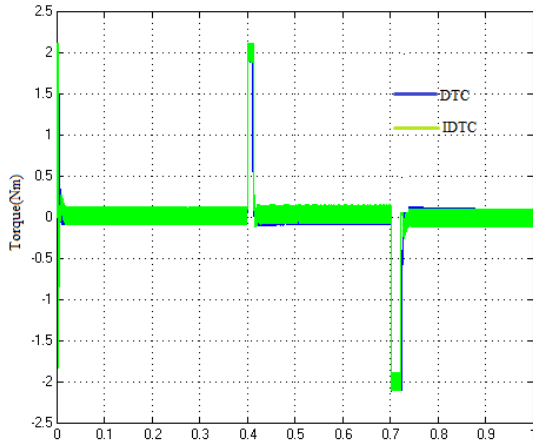


Fig.7 Torque response for different speed condition under no load

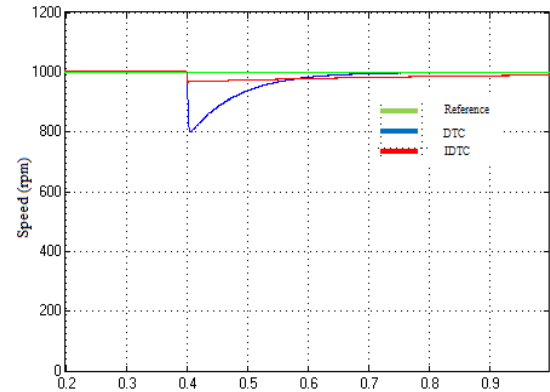


Fig.8 Speed Response for rated load condition

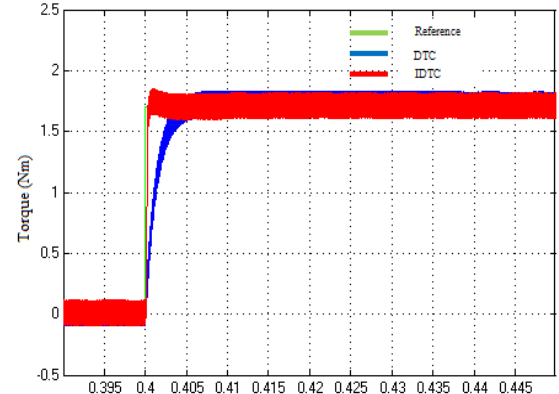


Fig.9 Torque response for rated load condition

5.3 Closed loop performance

Further to get a quantitative comparison the response of DTC and IDTC under start-up condition are compared in Table 2 in terms of rise time, settling time and peak over shoot at different speeds. Again IDTC is shown to provide better performance compared to DTC.

Table 2. Comparison of Transient responses

Response Parameters	N=1000rpm		N=2000rpm		N=3000rpm	
	DTC	IDTC	DTC	IDTC	DTC	IDTC
Rise Time(ms)	4.65	2.68	9.53	7.71	13.52	13.45
Settling time(ms)	6.075	3.28	10.7	8.75	16.5	16.1
Peak overshoot (rpm)	1034	1030	2123	2035	3235	3040

Table 3 and 4 provide a comparative error performance of DTC and IDTC against the various transient speed error criteria like Integral absolute error (IAE), Integral squared error (ISE), Integral time- weighted absolute error (ITAE) and integral time squared error at no load and rated load of 1.7Nm for both DTC and IDTC under different speed conditions. It is clearly seen that proposed IDTC approach outperforms the conventional DTC approach in terms of error performance.

Table 3. Comparative Transient speed error performance under no load condition

Error criteria	N=1000rpm		N=2000rpm		N=3000rpm	
	DTC	IDTC	DTC	IDTC	DTC	IDTC
IAE	5.883	3.37	22.52	12.9	45.7	34.72
ISE	2033	1493	1.6×10^4	1.05×10^4	5.334×10^4	3.905×10^4
ITAE	0.187	0.014	0.842	0.347	1.626	0.884
ISAE	5.412	2.1	84.22	29.4	355.2	209

Table 4. Comparative Transient speed error performance under rated load torque =1.7Nm

Error criteria	N=1000rpm		N=2000rpm		N=3000rpm	
	DTC	IDTC	DTC	IDTC	DTC	IDTC
IAE	22.78	12.46	39.43	21.35	61.95	41.02
ISE	3839	1709	1.784×10^4	1.075×10^4	5.515×10^4	3.916×10^4
ITAE	11.71	7.328	12.5	7.596	13.49	8.2
ITSE	1164	163.3	1241	158.5	1512	282.8

6. EXPERIMENTAL SETUP

The proposed IDTC scheme is implemented on an experimental platform which is based on hardware in loop (HIL) and Simulink/Xilinx as shown in below fig.10. The experimental hardware setup consists of a PMSM with load, a speed sensor, a power module, and a computer running with real time control software. The parameters of PMSM are as follows: DC link voltage 300V, rated power 3.7 kW, rated current 2.5A, rated torque 2.2 Nm, rated speed 1700 rpm, number of pole pairs 4, rotor permanent magnet flux 0.1875wb, direct and quadrature axes inductance 0.0024H, and stator resistance 9.8Ω. The sensor used is a Quadrature encoder pulse sensor which generates 2000 pulses for one rotation. The power module used is PEC16PSM01 which produces pulses based on SVPWM technique.

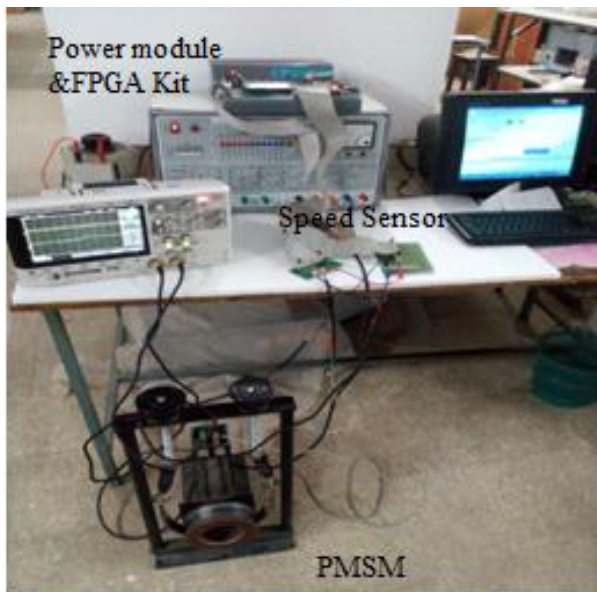


Fig.10 Experimental set up of the proposed scheme

The host computer runs Simulink/Xilinx real time control software. The control algorithm is implemented in Xilinx Spartan 3E FPGA XC3SD1800d-4FG676. The coding is written using VHDL coding. The PI controller used in both DTC and IDTC have the following settings. For IDTC, the settings are $K_p = 0.08$, $K_i = 0.1$ and for DTC, $K_p = 0.06$, $K_i = 0.1$.

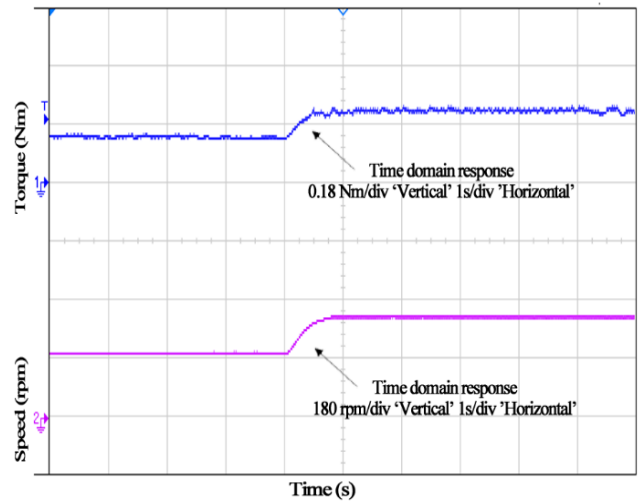


Fig.11. Response of torque and speed for increase in speed from 500r/m to 1000r/m in conventional DTC.

Fig.11 – 16 illustrate a comparison of set point response of speed for both conventional DTC and IDTC under no load conditions. As can be seen from these figures, when the speed is increased or decreased, the settling time is much faster for the proposed IDTC compared to conventional DTC.

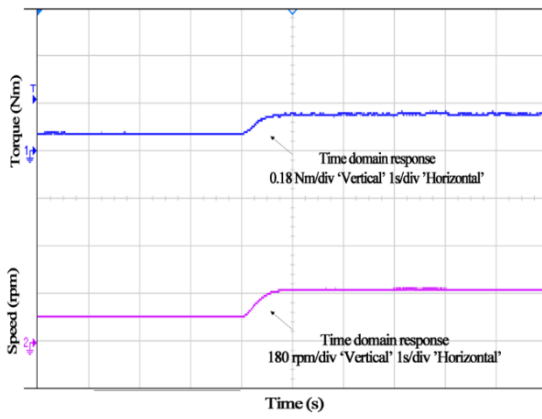


Fig.12. Response of torque and speed for increase in speed from 500r/m to 1000r/m in the proposed IDTC.

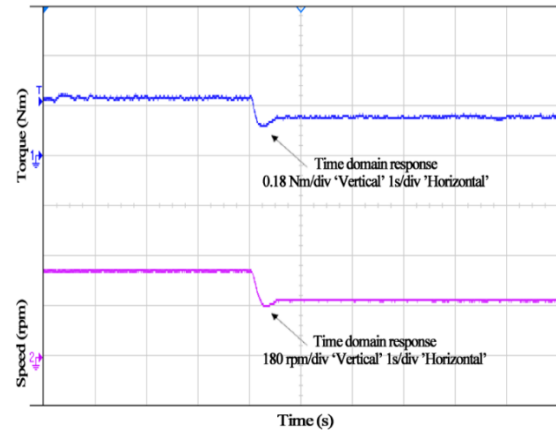


Fig.15. Response of torque and speed for decrease in speed from 1700r/m to 1000r/m in the proposed DTC.

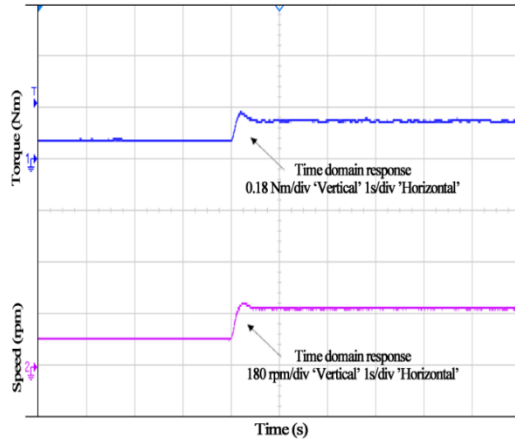


Fig.13. Response of torque and speed for increase in speed from 1000r/m to 1700r/m in the proposed DTC.

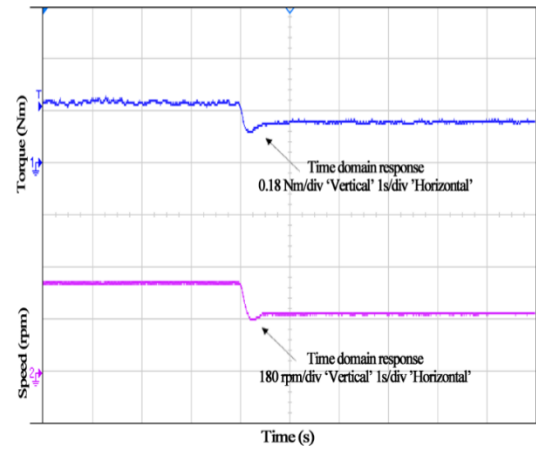


Fig.16. Response of torque and speed for decrease in speed from 1700r/m to 1000r/m in the proposed IDTC.

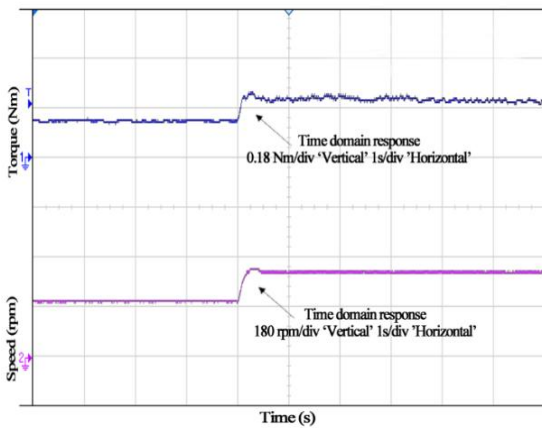


Fig.14. Response of torque and speed for increase in speed from 1000r/m to 1700r/m in the proposed IDTC

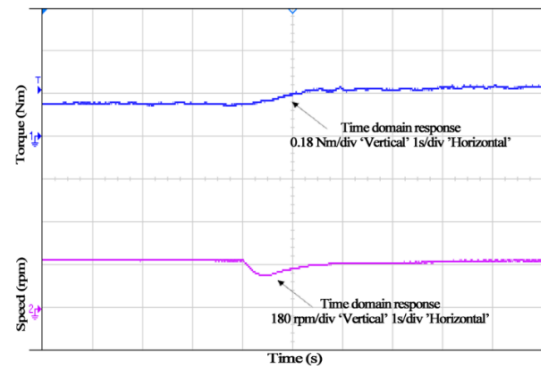


Fig 17. Response of torque and speed for change in torque from no load to 1Nm in the conventional DTC.

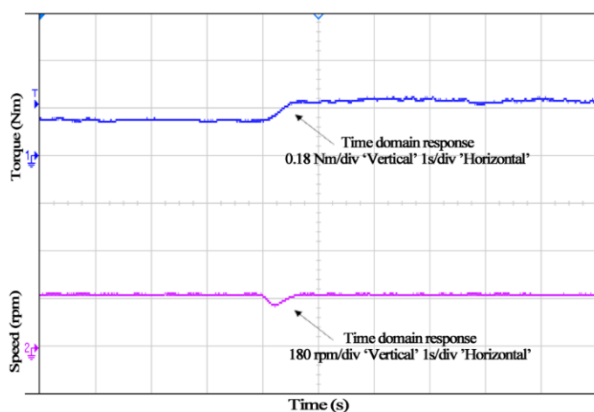


Fig.18. Response of torque and speed for change in torque from no load to 1Nm the proposed IDTC

7. CONCLUSION

This paper proposes a indirect torque control scheme for the control of PMSM. The proposal considers the PMSM under SVPWM as a DES and models voltage applied, torque and flux generated explicitly in terms of discrete states and continuous parameters of the system. The proposed scheme, being in the nature of feedforward control, has a natural speed of response advantage over DTC which is mainly feedback based. As expected, the Matlab / Simulink simulation confirms this conclusion showing a fast response for IDTC compared to DTC under set point and load changes system parameter uncertainty.

To further ensure the performance of the proposed control scheme, an experimental setup has been implemented using a PMSM drive which confirms the effectiveness of the proposed control scheme over conventional DTC. Besides, it is to be noted that the proposed IDTC scheme does not need current measurement for disturbance rejection as is required for DTC thus circumventing the attendant noise considerations. However, the measurement of position and speed are required in proposed IDTC scheme just as in the case of DTC.

As a future work, DES approach proposed in this paper can be extended to develop other model based control schemes, such as, Internal model control, Predictive control etc. applied to electrical drives.

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