A Modified DTC of Speed Sensorless IPMSM Drives Using Variable Structure Approach

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Abstract— In this paper a flexible direct torque control (DTC) of a sensorless interior permanent magnet synchronous motor drive is proposed based on variable structure approach and space vector modulation technique. According to this method, a sliding mode plus-PI controller is designed for torque and flux control respectively in stator flux reference frame. Simulation results obtained indicate improved performances for the proposed DTC controller and its robustness are preserved for the IPMSM drive.

Keywords- interior permanent magnet synchronous motor drives(IPMSM), Direct torque control(DTC), Variable structure control(VSC), Sliding Mode.

I. INTRODUCTION

Interior permanent magnet synchronous motors (IPMSM) offer many advantages over induction motors, such as overall efficiency, effective use of reluctance and magnet torques, smaller losses, and compact motor size. A multitude of solutions for control of IPMSM drives have been proposed [1-2]. Among them, the direct torque and stator flux control for IPMSM drives has been developed as direct torque control (DTC) [3]. The DTC features fast responses, structural simplicity and robustness to modeling uncertainty and disturbances. However, it still has some disadvantages that can be summarized in the following points: high ripple torque and current ripples; variable switching frequency behavior; difficulty to control torque and flux at very low speed.

In order to overcome these drawbacks of classical DTC, different methods such as discrete space vector control modulation [4,5], fuzzy logic based DTC [6], multi-level inverter [7], and variable structure (VSC) approach have been proposed. The VSC is an effective high-frequency switching control strategy for nonlinear systems with uncertainties. It features good robustness in the face of parameter uncertainties and other disturbances. The controller has fast response, but the controlled quantities exhibit chattering [8].

A VSC-DTC drive for IPMSM machine was proposed in stationary reference frame in reference [9]. Although that proposed controller assure the tracking of the torque and stator flux of IPMSM but the fast switching may generate undesirable chattering. To solve this chattering issue, authors of [9] have suggested a correcting function to take into account the saturation. This function reduces chattering phenomenon, however produces steady-state error and reduces the transient response.

A novel VSC-DTC for sensorless IPMSM drives is suggested in this paper. Two sliding mode-plus-PI controllers in stator flux reference frame are designed. The controller employs a switching component and a linear one so that has dual behavior. This flexible DTC scheme takes advantage of the best features of linear control, smooth operation and VSC. For the transient response, the linear component is dominant, and the PI gains are selected so as the linear control realizes the desired dynamic response of the drive. In the steady-state case, the VSC component is prevailing, and the gains are selected to obtain the desired performance in terms of steady-state robustness and chattering free responses. Thus, the DTC transient performance and robustness merits are preserved while the steady-state behavior is much improved.

II. ANALYSIS OF THE TORQUE IN THE STATOR FLUX REFERENCE FRAME

The stator flux linkage vector φ_S and rotor (magnet) flux Linkage vector φ_f can be drawn in the rotor flux (dq), stator Flux (xy), and stationary reference frames (DQ), as shown in Fig. 1. The angle between the stator and rotor flux linkages δ is the load angle when the stator resistance is neglected. In the Steady state, δ is constant corresponding to a load torque, and both stator and rotor flux rotate at the synchronous speed. In transient operation, δ varies and the stator and rotor flux rotate at different speeds.

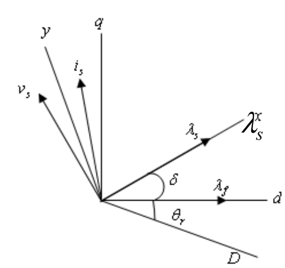

Figure 1. The stator and rotor flux linkages in different reference frames.

Since the electrical time constant is normally much smaller than the mechanical time constant, the rotating speed of stator flux, with respect to the rotor flux, can be easily changed. The well-known stator flux linkage, voltage, and electromagnetic torque equations in the rotor flux reference frame are given as follows:

$$\lambda_d = L_d i_d + \lambda_f \tag{1}$$

$$\lambda_q = L_q i_q \tag{2}$$

$$v_d = R_s i_d + p\lambda_d - \omega_r \lambda_q \tag{3}$$

$$v_q = R_s i_q + p\lambda_q + \omega_r \lambda_d \tag{4}$$

$$T = \frac{3}{2} P\left(\lambda_d i_q - \lambda_q i_d\right) \tag{5}$$

Where

λ_f rotor flux linkage due to permanent magnet;

L_d, L_q direct and quadrature inductance;

λ_d, λ_q direct and quadrature stator flux linkage;

R_s Stator resistance;

p Differential operator;

ω_r Rotor angular speed in electrical degree;

v_d, v_q, i_d, i_q Direct and quadrature voltage and current in rotor flux reference frame;

P Pole pairs;

T Electromagnetic torque;

When the back EMF and the variation of the stator inductances are sinusoidal. In addition, stator flux and torque control can be achieved by taking into account the IPMSM stator equation in stator flux reference frame. In this frame the stator per phase voltage can be expressed as,

$$v_s = R_s i_s + \frac{d\lambda_s}{dt} + j\omega_{\lambda s}\lambda_s \tag{6}$$

Where

v_s, i_s Stator voltage and current space vector in stator flux reference frame;

λ_s Stator flux space vector;

R_s Stator resistance;

$\omega_{\lambda s}$ Stator flux angular speed;

The relation between voltage and current in two frames is:

$$v_s = \sqrt{v_d^2 + v_q^2} \tag{7}$$

$$i_s = \sqrt{i_d^2 + i_q^2} \tag{8}$$

$$\lambda_s = \sqrt{\lambda_d^2 + \lambda_q^2} \tag{9}$$

The direct and quadrature components of (7) are

$$v_{sx} = R_s i_{sx} + \frac{d}{dt}\lambda_s \tag{10}$$

$$v_{sy} = R_s i_{sy} + \lambda_s \omega_{\lambda s} \tag{11}$$

Equation (10) indicates that the direct component of the stator voltage v_{sx} ,can be employed for flux control.

Under the assumed orientation, the developed torque is

$$T_e = 1.5P\lambda_s i_{sy} \tag{12}$$

Taking into account (12), the torque becomes

$$T_e = \frac{3P}{2R_s}\lambda_s(v_{sy} - \omega_{\lambda s}\lambda_s) \tag{13}$$

This last equation (13) indicates that the quadrate component of the voltage v_{sy} , can be employed for torque control.

III. PROBLEM ASSOCIATED WITH THE DTC OF IPMSM DRIVE AND THEIR REMEDIES

In a DTC scheme, the stator flux linkage is estimated by integrating difference between the input voltage and the voltage drop across the stator resistance given by (14).

$$\lambda_s = \int_0^t (v_s - i_s R)dt + \lambda_s\big|_{t=0} \tag{14}$$

There exist some limitations associated with this method of control, namely, 1) error in the estimation of stator flux linkage due to variation of stator resistance; and 2) requirement of a mechanical position sensor to detect the initial rotor position. Ways of mitigating these problems are given below.

A. Effect of Stator Resistance change and its Compensation

Performance of the DTC drive can be enhanced if the stator resistance is continuously estimated and updated in the controller during operation of the machine. It is seen that if an error in stator resistance between the controller and the motor is present, there will be errors in the estimated flux linkage and torque. The estimated torque is larger than the actual torque when the actual stator resistance is larger than the resistance in the controller and vice versa. This will reduce the maximum output torque of the drive. Based on the relationship between change of resistance and change of current, a PI stator resistance estimator (SRE) can be constructed in Fig. 2.

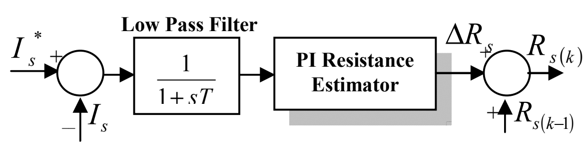

Figure 2. Block diagram of a PI stator resistance estimator.

where change in stator ΔR resistance is given by

$$\Delta R = \left(k_p + \frac{k_s}{s}\right)\Delta I \tag{15}$$

The compensation process can be based on the premise that as stator resistance changes, so will the amplitude of the stator current vector. The error between the amplitude of the measured current vector and that of the reference current vector can be used to compensate for the change in stator resistance until the error in current becomes zero in the steady state. This zero error is forced by the integrator in the estimator.

The reference current vector can be derived from the reference torque and reference flux equations of (5),(8) and (9). [11]

B. Detection of Initional Rotor Position withhout Sensor

The integration process in (14), which runs continuously, requires knowledge of the initial stator flux position $\lambda_s|_{t=0}$ at start. The stator current is assumed to be zero at start, so that the flux linkage due to the rotor only, i.e. λ_S , at t=0 , must be known in order to start the motor without jitter. If the initial position information in the controller is inaccurate, the motor may initially rotate in the wrong direction. This dithering is unacceptable in applications such as disk drives, electric propulsion, and high-performance servos. Since the DTC scheme does not explicitly require the rotor position after start, an initial position estimator without any position sensor is desirable.

A sensorless method to estimate the initial rotor position has been investigated in this section. The method is based on the injection of a high-frequency sinusoidal voltage and considering the effects of saliency on the amplitude of the corresponding stator current component. The magnetic pole is identified using the effect of magnetic saturation. The method is suitable to be combined with a DTC drive. This method is independent of load or on any motor parameters [11-12].

IV. VARIABLE-STRUCTURE CONTROL

The linear and variable structure DTC (LVSC) block diagram is shown in Fig. 3. The controller employs a switching component and a linear one, and has dual behaviour. This is a flexible DTC scheme that takes advantage of the best features of linear control, smooth operation, and of VSC, robustness to perturbations. An SVM unit, which generates the VSI switching signals S_a , S_b and S_c is the output stage.

The sliding surface $s = s_x + js_y$ is selected as

$$s_x = e_{\lambda s} + c_{\lambda s} Se_{\lambda s} \tag{16}$$

$$s_y = e_{Ts} + c_{Ts} Se_{Ts} \tag{17}$$

Where $e_{\lambda s} = \lambda_s^* - \hat{\lambda}_s$ and $e_{Te} = T_e^* - \hat{T}_e$ are the flux and torque control errors, respectively. The symbol " * "denotes reference quantities and " ^ " denotes estimated quantities. Design constants $c_{\lambda s}$ and c_{Te} are selected so as to incorporate the desired dynamics in the sliding mode.

The control law that produces the reference voltage vector $v_s^* = v_{sx}^* + jv_{sy}^*$ in the stator flux reference frame is

$$v_x^* = \left(K_{p\lambda} + K_{I\lambda} \frac{1}{S} \right) (e_{\lambda s} + K_{VSC\lambda} \operatorname{sgn}(S_x)) \qquad \textbf{(18)}$$

$$v_x^* = \left(K_{pT} + K_{IT} \frac{1}{S} \right) (e_{Te} + K_{VCST} \operatorname{sgn}(S_y)) \qquad \textbf{(19)}$$
$$+ \hat{\omega}_{\lambda s} \hat{\lambda}_s$$

Where S is the Laplace operator, $K_{P\lambda}$, $K_{I\lambda}$, K_{PT} ,and K_{IT} are the PI controller gains, and $K_{VSC\lambda}$, K_{VSCT} are the VSC gains. Adequate balance between the linear PI controller and switching behavior of the VSC can be achieved by proper gains selection of both the controllers. For the transient responses, the linear component is dominant, and the PI gains are selected so as the linear control realizes the desired dynamic response.

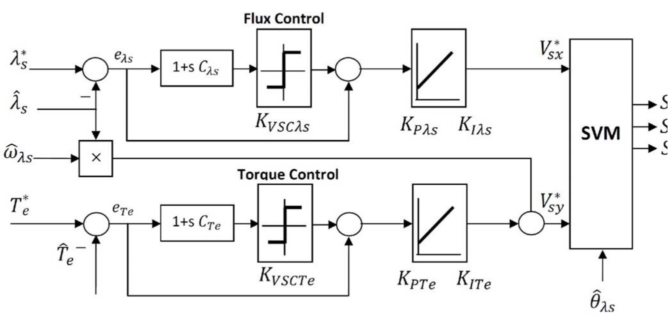

Figure 3. Linear and variable-structure control scheme.

In the steady state, the VSC component is prevailing, and the ripple magnitude depends on the K_{VSC} gains. It can be proved that large enough values for $K_{VSC\lambda}$ and K_{VSCT} will fulfil the stability condition $s\frac{ds}{dt}\langle 0$. Since the linear PI part operates independently of the switching one, selection of the $K_{VSC\lambda}$ and K_{VSCT} gains is not restrictive, and it was determined that any positive values can result in stable operation. These gains are selected just as large as needed to obtain the desired performance in terms of steady-state robustness and chattering free operation.

Assuming accurate state estimation and ideal VSI $\left(v_s = v_s^*\right)$, from equations (10) and (14), the torque dynamics result in ,

$$T_e = \frac{k_T (K_{PT} s + K_{IT})}{(1 + k_T KPT)s + k_T K_{IT}} (T_e^* + K_{VSC} \operatorname{sgn}(S_{Te})) \quad (20)$$

Where $k_T = \dfrac{3P\lambda_s}{2R_s}$, and λ_S was assumed constant.

The PI dynamics has been added in order to ensure zero steady-state torque error. This is an advantage with respect to classic DTC that works with non-zero torque error. Since equation (20) is valid under constant flux assumption, it is important to achieve very fast and robust flux control.

V. SIMULATION FOR PROPOSED METHOD

A complete block diagram of proposed method for IPMSM drive is shown in Fig.4. The D and Q axes stator flux linkages λ_D and λ_Q in the stationary reference frame can be given by

$$\lambda_D(K) = \lambda_D(k-1) + \{v_D(k-1) - Ri_D(k)\}T_s \qquad (21)$$

$$\lambda_Q(K) = \lambda_Q(k-1) + \{v_Q(k-1) - Ri_Q(k)\}T_s \qquad (22)$$

Where T_S is the sampling interval v_D , v_Q are the stator voltages in the D and Q axes, i_D , i_Q are the stator currents in the D and Q axes, k and $k-1$ variables in brackets refer to the k th and $k-1$ th sampling instants, respectively. The stator flux linkage vector is then given by

$$\lambda_s = \sqrt{\lambda_{D(k)}^2 + \lambda_{Q(k)}^2} \qquad (23)$$

The motor developed torque in terms of the stator flux linkages is given by

$$T(k) = \frac{3}{2} P\{\lambda_{D(k)} i_{Q(k)} - \lambda_{Q(k)} i_{D(k)}\} \qquad (24)$$

The estimated stator flux speed $\hat{\omega}_{\lambda s}$ is calculated in a stationary reference frame in each step as follows

$$\hat{\lambda}_{\varphi s} = \frac{d}{dt}(\tan^{-1}\frac{\lambda_{Qs}}{\lambda_{DS}}) = \frac{\lambda_D(v_\varphi - Ri_\varphi) - \lambda_Q(v_D - Ri_D)}{\lambda_s^2} \tag{25}$$

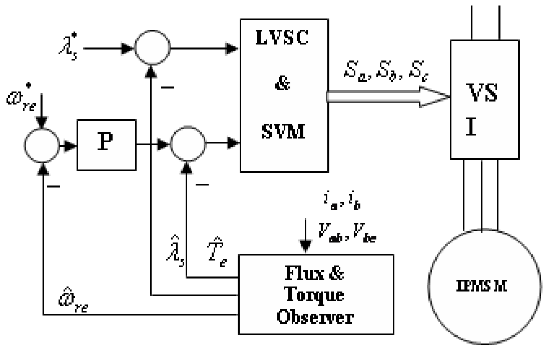

Figure 4. Complete block diagram of IPMSM drive.

VI. SIMULATION RESULT

Modeling studies were performed on an IPM synchronous motor. The pertinent parameters for the test motor are given in table I.

TABLE I. PARAMETER OF AN IPMSM FOR SIMULATION

Number of Pole pairs	Stator resistance	Magnet flux Linkage	d-axis inductance	q-axis inductance	Phase voltage
4	1.9Ω	0.447Wb	0.388H	.475H	230V

Table II shows coefficients of PI and switching surface used for simulation.

TABLE II. COEFFICIENT THAT CONSIDERED IN SIMULATION

$K_{P\lambda}$	$K_{I\lambda}$	K_{PT}	K_{IT}	$K_{VCS\lambda}$	K_{VCST}
320	2780	450	220	10	10

The torque –time response of proposed controller for the IPM motor drive for step command from 0.5 to 1.6 Nm at t = 1.0 sec is in Figure 5. Figure 6 shows the flux-time response t under load condition for step change in torque. There is no major change in the flux when the torque command is applied at t=1.0 sec. Figure 7 shows the speed response of the drive under same torque change command. Figure 8 shows the corresponding trajectories of flux under step changing of torque in λ_D and λ_Q coordinate. In the case when the flux is increased from 70% of nominal value to 100% of nominal value, the responses are shown in fig 9 and fig.10. Under this transient condition, the value of torque is found constant as shown in fig. 11.

It is to be noted that the detailed experimental results will be presented in another paper in the near future.

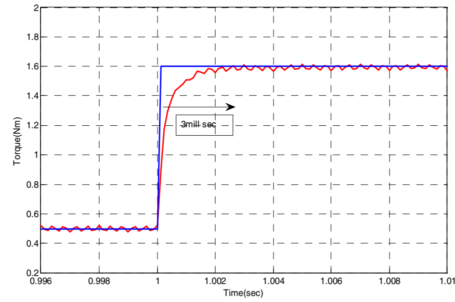

Figure 5. Dynamic torque-time response for proposed controller for step change in torque

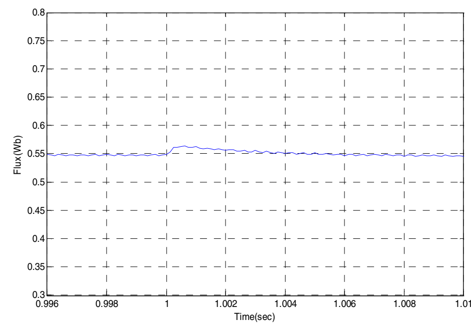

Figure 6. operation of flux under change torque.

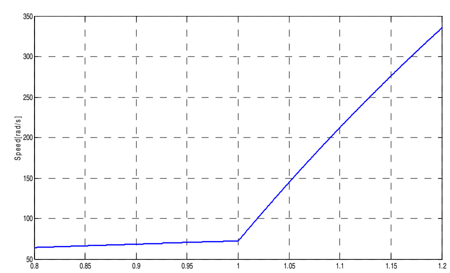

Figure 7. Speed response of IPM motor under same change torque

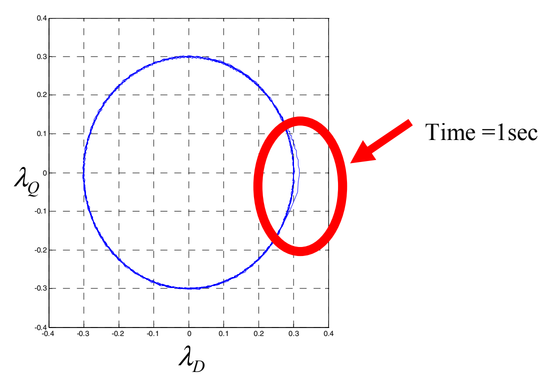

Figure 8. Tragedy of flux in λ_D and λ_Q coordinate .

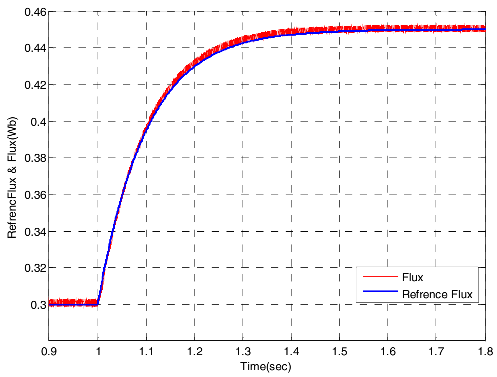

Figure 9. Dynamic opearation of proposed controller under changing flux.

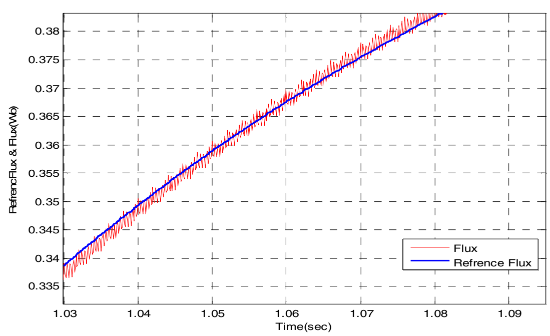

Figure 10. Dynamic performance for proposed controller under changing flux

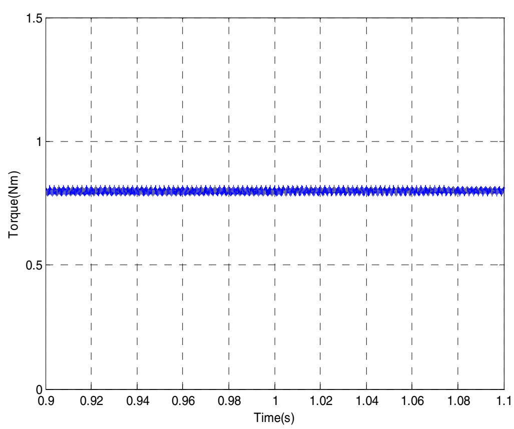

Figure 11. Operation of torque under changing Flux.

VII. Conclusions

In this paper, a variable structure controller has been proposed for direct torque and stator flux control of a sensorless IPMSM drive in stator flux reference frame. Combining the principle of the sliding mode technique with PI controller and space vector modulation technique yields simple but robust performances for the IPMSM drive system. The proposed system has minimum number of depend-parameters. Simulation results establish the efficacy of the proposed controller. Detailed experimental results will be presented in another paper.

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