A New DTC-SVM Based Control of Field Oriented Position Sensorless Induction Motor Drive with Reduced Torque and Flux Ripple

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Abstract—Direct Torque Control (DTC) is widely used for ac drives. Attempts to combine DTC with Space Vector Modulation (SVM) have led to new ways. A new approach to DTC-SVM is presented in this paper. A Correlated Real Time Recurrent Learning (CRTRL) algorithm based Recurrent Neural Network (RNN) is used to estimate stator and rotor flux. Through measurement of the phase flux linkages and phase currents the RNN is able to estimate the rotor position, thereby facilitating elimination of the rotor position sensor. Fast dynamic speed response is obtained through maintaining the rotor flux constant as in the case of field orientation control. The control method proposed in this paper can reduce the torque, stator current, and rotor flux ripples which improve the system dynamic performance and robustness in different operating conditions. The proposed controller is also computationally efficient. The control methodologies and simulation results are given and discussed.

Index Terms—Direct torque control, space vector modulation, induction motor, flux estimation, torque ripple, flux ripple.

I. INTRODUCTION

It is well known that DTC system operates to control the stator flux and the torque directly by selecting the appropriate inverter switching state. DTC is also very simple in its implementation because it needs only two hysteresis comparators and switching vector table for both flux and torque control. Therefore, the only gains to be adjusted are the amplitudes of the hysteresis band. The amplitude of the hysteresis band greatly influences the drive performance such as flux and torque ripples, inverter switching frequency, and current harmonics [1], [2]. Current Researchers are working to maintain constant switching frequency in DTC and to reduce the torque, flux, current, and speed pulsations during steady state [3]-[5]. For minimizing torque and flux

ripple further new DTC-SVM scheme was proposed in [6], [7]. In [6], the DTC-SVM scheme in rotor flux reference frame is presented where the Auto Disturbance Rejection Controller (ADRC) is employed. In [7], a general Space Vector Pulse Width Modulation (SVPWM) algorithm is presented to reduce the torque and flux ripple. Both researchers of [6] and [7] used 3-level inverters for better performance rather than two level inverters.

High performance induction motor drives based on “Field Orientation” have been commercially available for almost three decades. Because of torque flux decoupling, Field Orientation Control (FOC) achieved good dynamic response and accurate motion control. Their performance is such that they can easily replace DC drives without loss of accuracy, stability or speed of response. The most significant barrier to further utilization is at present the rotor position sensor which being external to the controller- power electronics assembly. It is the cause of low reliability, high maintenance and high installation cost. So, demands for fast and accurate rotor position estimator in high performance motor drives are increasing. A number of model-independent solutions for sensorless speed and position control of induction motors have been developed [8]-[11]. However in these studies, the drives were limited to operation by flux saturation in the main path and around the rotor slots of the machine which interferes with that of the position-dependent rotor saliency. Unwanted “saturation harmonics” in the motor currents can dominate over the desired “position harmonic” making position estimation impossible [12].

In this paper, a new DTC-SVM based control of induction motor drive is presented. The inverter switch position is selected combining with the situation of torque error, current error and the position of stator flux angle. The rotor flux is estimated from induction motor currents and speed information. The proposed CRTRL algorithm is used to estimate the rotor flux with its angle and eliminates the rotor position sensor. The main flux saturation effects using a variable magnetizing inductance are also in consideration. The magnetizing inductance is expressed as a polynomial for a more realistic representation of a saturated induction machine. It is demonstrated that the proposed control scheme is robust against the external disturbance in various operating conditions.

II. MATHEMATICAL MODEL

The dynamic model of induction motor can be represented in the synchronous reference frame (d-q) as:

$$v_{ds} = (R_s + pL_s)i_{ds} - L_s\omega_e i_{qs} + pL_m i_{dr} - L_m\omega_e i_{qr} \quad (1)$$

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$$v_{qs} = \omega_e L_s i_{ds} + (R_s + pL_s) i_{qs} + L_m \omega_e i_{dr} + pL_m i_{qr} \quad (2)$$

$$0 = pL_m i_{ds} - \omega_{sl} L_m i_{qs} + (R_r + pL_r) i_{dr} - L_r \omega_{sl} i_{qr} \quad (3)$$

$$0 = L_m \omega_{sl} i_{ds} + pL_m i_{qs} + (R_r + pL_r) i_{qr} + L_r \omega_{sl} i_{dr} \quad (4)$$

$$T_e = Jp\omega_m + B\omega_m + T_L \quad (5)$$

where,

v_{ds} and v_{qs} = the dq-axis stator voltages; i_{ds} and i_{qs} = the dq-axis stator currents; i_{dr} and i_{qr} = the dq-axis rotor currents; R_s and R_r = the stator and rotor resistances; L_s and L_r = the stator and rotor leakage inductances; L_m = the mutual inductance; ω_e = the synchronous reference speed; ω_r = the rotor angular speed; ω_m = the angular speed of the motor; $\omega_{sl} (= \omega_e - \omega_r)$ = the slip angular speed; T_L = the load torque; J = the moment of inertia; B = the rotational damping coefficient; P_p = the number of pole pairs; $p \equiv d/dt$.

The developed electromagnetic torque in terms of d - and q - axes components is given by:

$$T_e = \frac{3}{2} P_p L_m (i_{qs} i_{dr} - i_{ds} i_{qr}) \quad (6)$$

where P_p is the number of pole pairs.

Components of rotor flux are:

$$\psi_{dr} = L_r i_{dr} + L_m i_{ds} \quad (7)$$

$$\psi_{qr} = L_r i_{qr} + L_m i_{qs} \quad (8)$$

From (7) and (8), d - and q - axes rotor currents are:

$$i_{dr} = \frac{1}{L_r} (\psi_{dr} - L_m i_{ds}) \quad (9)$$

$$\omega_m^* \quad i_{qr} = \frac{1}{L_r} (\psi_{qr} - L_m i_{qs}) \quad (10)$$

Substituting (7)-(10) into (3) and (4) yields:

$$\frac{d\psi_{dr}}{dt} + \frac{R_r}{L_r} \psi_{dr} - \frac{L_m}{L_r} R_r i_{ds} - \omega_{sl} \psi_{qr} = 0 \quad (11)$$

$$\frac{d\psi_{qr}}{dt} + \frac{R_r}{L_r} \psi_{qr} - \frac{L_m}{L_r} R_r i_{qs} + \omega_{sl} \psi_{dr} = 0 \quad (12)$$

If the field orientation is established such that q -axis rotor flux is set zero, and d -axis rotor flux is maintained constant

then equations (11), (12), (9), (10) and (6) becomes:

$$\psi_{dr} = L_m i_{ds} \quad (13)$$

$$\omega_{sl} = \frac{1}{\tau_r} \frac{i_{qs}}{i_{ds}} \quad (14)$$

$$i_{dr} = 0 \quad (15)$$

$$i_{qr} = -\frac{L_m}{L_r} i_{qs} \quad (16)$$

$$T_e = \frac{3}{2} P_p \frac{L_m}{L_r} \psi_{dr} i_{qs} \quad (17)$$

Where $\tau_r (= L_r / R_r)$ is the time constant of the rotor. Hence, only q -axis stator current controls the developed electromagnetic torque.

Substituting (13) and rearrangement of (1) and (2), yields:

$$v_{ds} = R_s i_{ds} + \frac{L_s}{L_m} \frac{d\psi_{dr}}{dt} + L_m \sigma \frac{di_{dr}}{dt} - \omega_e \frac{L_m}{L_r} \psi_{qr} - \omega_e \sigma L_s i_{qs} \quad (18)$$

$$v_{qs} = R_s i_{qs} + L_s \sigma \frac{di_{qs}}{dt} + \omega_e L_s i_{ds} + \frac{L_m^2}{L_r} \frac{di_{qs}}{dt} + \omega_e L_m i_{dr} + L_m \frac{di_{qr}}{dt} \quad (19)$$

where $\sigma = 1 - \frac{L_m^2}{L_s L_r}$ and $\omega_e = P_p \omega_m + \omega_{sl}$.

III. PROPOSED CONTROL SCHEME

In the proposed scheme shown in Fig. 1, the reference torque T_e^* is generated from speed error maintaining minimum time control [13] to get the fast speed response. The two errors in i_{ds} and i_{qs} are processed through two PI controllers to generate reference voltages v_{ds}^* and v_{qs}^* . The stator command currents and command voltages are generated as follows:

$$i_{ds}^* = \Psi_r^* / L_m \quad (20)$$

$$i_{qs}^* = k_{p1} \tau + k_{i1} \int \tau dt \quad (21)$$

$$v_{ds}^* = k_{p2} (i_{ds}^* - i_{ds}) + k_{i2} \int (i_{ds}^* - i_{ds}) dt \quad (22)$$

$$v_{qs}^* = k_{p3} (i_{qs}^* - i_{qs}) + k_{i3} \int (i_{qs}^* - i_{qs}) dt \quad (23)$$

Here, τ is the torque error.

The angle of the input stator voltage needed with respect to the reference stator α -axis as:

$$\hat{\theta}_s = \hat{\theta}_v + \hat{\theta}_r \quad \text{where, } \hat{\theta}_v = \tan^{-1}(v_{qs}^* / v_{ds}^*).$$

The magnetizing inductance representing the saturation in the magnetic circuit is given in the appendix and was found from the laboratory test of the motor.

Fig. 1. Proposed control scheme

IV. CRTRL BASED FLUX ESTIMATOR

The CRTRL algorithm [14] is used to train the RNN based flux estimator. The fully connected RNN consists of q neurons with l external inputs, as shown in Fig. 2. Let the q -by-1 vector $\mathbf{x}(k)$ denotes the state of the network in the form of a nonlinear discrete-time system, the $(l+1)$ -by-1 vector $\mathbf{u}(k)$ denotes the input (including bias) applied the network, and the p -by-1 vector $\mathbf{y}(k)$ denotes the output of the network.

The process equation in the state-space description of the network is written in the following form:

$$\mathbf{x}(k+1) = \begin{bmatrix} \varphi(W_1^T(k)z(k)) \\ \varphi(W_2^T(k)z(k)) \\ \vdots \\ \varphi(W_q^T(k)z(k)) \end{bmatrix} \quad (24)$$

where $\varphi(\cdot)$ is an activation function, and the $(q+l+1)$ -by-1 weight vector $W_i(k)$, which is connected to the i th neuron in the recurrent network, corresponds to the i th column of the transposed weight matrix $W^T(k)$.

The adjustment for the weight vector of the i th neuron, ΔW_i , is given by:

$$\Delta W_i = \eta \frac{\partial J(k)}{\partial W_i(k)} = \eta C \Lambda_i(k) e(k), \quad i = 1, 2, 3, \dots, q \quad (25)$$

where $J(k)$ is the cost function, $e(k)$ is the p -by-1 error vector.

The recursive equation Λ_i for the neuron i can be obtained by:

$$\Lambda_i(k+1) = \varphi(k+1)[W_x(k)\Lambda_i(k) + Z_i(k)] \quad (26)$$

Where $Z_i(k)$ is a q -by- $(q+l+1)$ matrix and $\varphi(k)$ is a q -by- q diagonal matrix defined by:

$$\varphi(k+1) = \text{diag} [\varphi'(W_1^T(k)z(k)), \varphi'(W_2^T(k)z(k)), \dots, \varphi'(W_q^T(k)z(k))] \quad (27)$$

Stator flux synthesis by CRTRL algorithm is shown in Fig. 2. After getting the stator fluxes, Ψ_{as} and $\Psi_{\beta s}$, the rotor fluxes, Ψ_{ar} and $\Psi_{\beta r}$ are obtained by the following equations:

$$\Psi_{ar} = \frac{L_r \Psi_{as} - (L_s L_r - L_m^2) i_{as}}{L_m} \quad (28)$$

$$\Psi_{\beta r} = \frac{L_r \Psi_{\beta s} - (L_s L_r - L_m^2) i_{\beta s}}{L_m} \quad (29)$$

Fig. 2. Stator flux synthesis by CRTRL algorithm based RNN

V. SIMULATION RESULTS

The proposed control structure was implemented in the environment software C++, and tested with various operating conditions, to illustrate the performance of the proposed control scheme. The flux estimator was trained off-line so that it estimates the flux components accurately. The numerical method for solving the equations is Runge-Kutta method (order 4). The parameters of the induction motor used are given in the appendix.

A. Stator flux and rotor position estimation

Figs. 3 & 4 show the actual and estimated α -axis stator flux, β -axis stator flux, and rotor flux angle under both in transient and steady state condition. It is apparent that the estimated responses are matching accurately with the actual responses both in transient and steady-state conditions. Thus the acceptability of the proposed CRTRL has been confirmed.

Fig. 3. Actual and estimated responses of (a) α -axis stator flux, (b) β -axis stator flux, and (c) rotor angle for the induction motor drive under transient condition.

Fig. 4. Actual and estimated responses of (a) α -axis stator flux, (b) β -axis stator flux, and (c) rotor angle for the induction motor drive under steady-state condition.

B. Starting Performance of the Induction Motor Drive

The motor was started with a command speed of 1432 rpm (150 rad/sec) with no load from standstill condition. At $t=0.24$ second the motor reaches to the command speed as shown in Fig. 5(a). The motor follows the command speed accurately without steady-state error and oscillations. It can also be seen from Fig. 5(b) that, the developed electromagnetic torque is almost ripple free. In Fig. 5(c), it is observed that the rotor flux also follows the reference flux accurately. The stator phase currents are shown in Fig. 5(d). So the drive operates perfectly with negligible torque, flux, and current ripples.

(b)

Time in sec

(c)

Time in sec

(d)

(a)

Fig. 5. (a) Simulated speed response, (b) developed electromagnetic torque, (c) rotor flux, and (d) stator current under transient and steady-state condition.

C. Performance Under Different Operating Conditions

The performance of the induction motor drive under different operating conditions was also investigated in order to verify the robustness of the proposed control scheme. The performances of the drive system for various changes in speed with no load are shown in Figs 6(a), 6(b), and 6(c). From Fig. 6(a) it is observed that the proposed control scheme follows both the forward and reverse direction reference track quickly with negligible torque and flux ripple. The control scheme also follows the step change in speed and ramp change in speed reference track accurately with negligible torque and flux ripple and shown in Fig. 6(b) & 6(c) respectively. Effectiveness of the proposed control scheme was also tested by applying and reducing load torque in step and shown in Fig. 6(d). The load torque of the motor was suddenly increased from 0 N-m to 4.0 N-m at $t=0.5$ second, then decreased to 2.0 N-m at $t=1.0$ second and again to 1.0 N-m at $t=1.5$ second. No fall and oscillation in speed is noticed due to this load torque disturbance. Also the developed electromagnetic torque and the rotor flux are found almost ripple free.

Time in sec

Time in sec

(b)

(d)

(c)

Fig. 6. Performances of the induction motor drive with (a) speed reverse, (b) step change in speed, (c) ramp change in speed, and (d) step change of load.

In order to obtain a high performance induction motor drive, its controller should be insensitive with the machine parameter deviation. The insensitivity of the proposed controller due to increase of both stator and rotor resistance of the machine up to 100% from its nominal value at $t=1.0$ second can be visualized from Fig. 7(a) and 7(b) respectively.

(a)

Time in sec

Time in sec

(b)

 Fig. 7. Performances of the induction motor drive for change in stator and rotor resistances with (a) $R_s = 2R_s^*$, and (b) $R_r = 2R_r^*$ at $t=1$ second.

There is some instrumental error always present in physical system. For the robustness test 10% instrumental error was introduced to reference voltage vectors. The performance of the motor speed, developed electromagnetic torque, and the rotor flux are shown in Fig. 8. It can be observed that the performances are similar to the motor starting performance. So depending on the performance, it can be concluded that the proposed control scheme is computationally efficient.

Time in sec

Time in sec

Time in sec

Fig. 8. Performances of the induction motor drive under computational errors present in the physical system.

VI. CONCLUSIONS

This paper presents a new high performance control methodology for position sensorless induction motor drive with reduced torque and flux ripple. The results obtained and presented in this work indicate that the proposed control strategy produces very fast response of the induction motor drive and realizes almost ripple free operation in torque, stator current, and rotor flux. It is also observed that the proposed drive with CTRL based flux estimator is capable to estimate stator flux and rotor position both in steady-state and transient conditions accurately. The drive is also robust to load disturbances, parameter variations, and different speed tracking conditions. Due to its simple structure it can be easily implemented on available embedded DSP's, thus will find many practical applications.

APPENDIX

Parameters of induction motor for simulation

Parameter	Value
Rated power (KW)	0.75
Rated voltage (V)	415
Rated current (A)	1.8
Rated speed (rad/s)	150
Stator resistance (Ω)	13.25

Rotor resistance (Ω)	16.818
Mutual inductance (H)	0.7114
Stator self inductance (H)	0.7359
Rotor self inductance (H)	0.7359
Moment of inertia (Kg-m^2)	0.0075
Damping coefficient (Nm-sec/rad)	0.00107
Number of pole pairs	2

Saturation characteristic for induction machine:

Mutual inductance,

$$L_m = f(i_{mag}) =$$

$$b_0 + b_1 i_{mag} + b_2 i_{mag}^2 + b_3 i_{mag}^3 + b_4 i_{mag}^4 + b_5 i_{mag}^5$$

$$b_0 = 0.9328; b_1 = 0.27058; b_2 = -0.95898; b_3 = 0.54123;$$

$$b_4 = -0.5479; b_5 = -0.01859$$

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