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Robust sliding mode speed controller for hybrid SVPWM based direct torque control of induction motor

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Abstract. The conventional direct torque control (CDTC) is known to produce quick and robust response in induction motor drives. However, during steady state, torque, flux and current occur, which results in acoustical noise and incorrect speed estimations. The main objective of this paper is to present different hybrid space vector PWM (HSVPWM) methods for direct torque controlled induction motor for reduced inverter switching losses and the steady state ripple in torque, flux and current. The proposed PWM techniques are designed based on the notion of stator flux ripple, which is a measure of line current ripple. Expressions for RMS ripple, over a sub-cycle are derived and this analysis together with the total harmonic distortion (THD) performance of the different sequences is used to design HSVPWM techniques, which utilize the zero voltage vector redundancy that leads to clamping sequences. To validate the proposed method, the simulation has been carried out using MATLAB-SIMULINK and the results are presented. The simulation results verify the superiority of the proposed method to the conventional method.

Keywords: direct torque control, hybrid space vector PWM, sliding mode speed control

1 Introduction

In recent years, the development of high-performance control strategies for induction motor drives resulted in a rapid evolution. One of the most popular methods, known as field oriented control (FOC) or vector control, has been proposed by F. Blaschke[3]. In the vector control the motor equations are transformed in a field coordinate system that rotates in synchronism with the rotor flux vector and hence FOC controls the induction motor in a same manner as separately excited DC motor. The disadvantage of this control scheme is inclusion of the pulse encoder, indirect torque control and also it is quite complex due to reference frame transformations. To overcome these disadvantages, in the middle of 1980's, a new quick response technique for the torque control of induction motors was proposed by Takahashi as direct torque control (DTC) [12]. DTC provides very quick response with simple control structure and hence, this technique is gaining popularity in industries[14]. To find suitable solution for any application that requires torque control, FOC and DTC are compared in [5] and it concluded that the DTC give fast response with compare to FOC. Though DTC has high dynamic performance, it has few drawbacks such as high ripple in torque, flux, current and variation in switching frequency of the inverter. The effects of flux and torque hysteresis band amplitudes in the induction motor drive performance has been analyzed in [4]. To improve the performance of conventional DTC in terms of ripple, discrete space vector modulation (DSVM) technique was proposed in [6], which uses a higher number of voltage vectors with respect to basic DTC scheme. But, the conventional DTC and DSVM based DTC give the variation in switching frequency. The problem of variation in switching frequency can be overcome by using space vector pulsewidth modulation (SVPWM)[7, 13]. In this method, for each sampling period, voltage space vectors have been generated and are used for the reduction of flux and torque ripples. In

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[9], both sine-triangle and space vector approaches are compared for different PWM sequences and it is concluded that space vector approach is superior with compared to sine-triangle approach. To reduce the current ripple further at high modulation indices, hybrid SVPWM (HSVPWM) strategies are developed for induction motor drives in [1, 8, 10]. In [11], HSVPWM technique is applied to DTC for reduced ripple. To improve the speed performance, a sliding mode speed controller[15] is used for FOC in [2].

The main objective of this paper is to develop various HSVPWM methods for direct torque controlled induction motor drive to reduce the ripples at all modulation indices. Also to improve the speed performance, under uncertainties, an integral switching surface sliding mode speed controller is developed, which is robust under uncertainties caused by variation in load torque.

2 Principle of conventional DTC

The key idea of the conventional DTC is, the rate of change of torque is proportional to the instantaneous slip between the stator flux and rotor flux under constant stator flux linkage. It has been widely recognized for its fast and robust torque and flux control. As the rotor time constant of a standard squirrel-cage induction machine is very large, the rotor flux linkage changes slowly compared to the stator flux linkage. However, during a short transient, the rotor flux is almost unchanged. Thus rapid changes of the electromagnetic torque can be produced by rotating the stator flux in the required direction, as directed by the torque command. On the other hand the stator flux can instantaneously be accelerated or decelerated by applying proper stator voltage phasors. Thus, the simultaneous and decoupled control of torque and flux is achieved by direct adjustment of the stator voltage, in response to the torque and flux errors. The DTC regularly applies the most appropriate voltage vector in order to maintain the torque and stator flux within two hysteresis bands, which results bang-bang behavior and produces variation in witching frequency and significant ripple in flux, torque and current.

3 Hybrid SVPWM methods

The voltage vectors, produced by a 3-phase, two-level inverter, divide the space vector plane into six sectors as shown in Fig. 1 (a). As the sectors are symmetric, the discussion in this paper is limited to first sector only. In the space vector approach, the desired reference vector is generated by time averaging the suitable discrete voltage vectors in sampling period TS. For a given reference voltage VREF and angle in sector I, the volt-time balance is maintained by applying the active vectors 1, 2 and zero states together for durations T1, T2 and TZ respectively, as given in (1).

$$T_1 = M^{*}T_S{}^{*}\frac{\sin(60^\circ-\alpha)}{\sin(60^\circ)}, \quad T_2 = M^{*}T_S{}^{*}\frac{\sin(\alpha)}{\sin(60^\circ)}, \quad T_Z = T_S - T_1 - T_2 \tag{1}$$

where M is the modulation index, is given by $\frac{3V_{REF}}{2V_{dc}}$.

The conventional SVPWM uses 0127-7210 sequence in the first sector, 0327-7230 sequence in second sector and so on. Different switching sequences are possible by utilizing the freedom of active state vector division, which can be used within the sub-cycle to generate the reference voltage vector as shown in Fig. 1 (b). The conventional sequence, 0127-7210, which divides the zero vector time equally between the two zero states within a sub-cycle. The basic symmetric bus clamping sequences 012-210 and 721-127 use only one zero state. The remaining two asymmetrical bus clamping sequences 0121-1210 and 7212-2127 divide the active vector times into two equal halves within a sub-cycle[1, 8–11].

The states of the inverter are switched at appropriate instants to generate the required fundamental voltage sample, in an average sense, and not in an instantaneous fashion. The difference between the instantaneous applied vector and the instantaneous fundamental voltage vector is the instantaneous voltage ripple vector. The time integral of the voltage ripple vector gives the 'stator flux ripple vector', which is a measure of the ripple in line current[1, 8, 10, 11]. The stator flux ripple vector over a sub-cycle for above sequences for a given reference vector and the corresponding q-axis ripple and d-axis ripple are also shown in [1, 10]. The values of

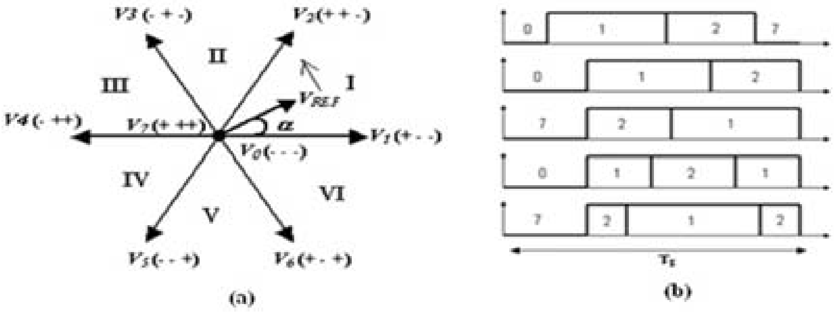

Fig. 1. (a): Voltage space vectors (b): Different possible Switching sequences in sector-I

q-axis and d-axis ripples at the switching instants are given in terms of Q_Z, Q_1, Q_2 and D, which are defined as in (2) [1, 10].

$$Q_Z = -V_{REF}T_Z, \quad Q_1 = (\frac{2}{3}V_{dc}Cos\alpha - V_{REF})T_1 \tag{2}$$
$$Q_2 = (\frac{2}{3}V_{dc}Cos(60^\circ - \alpha) - V_{REF})T_2, \quad D = \frac{2}{3}V_{dc}Sin\alpha T$$

The expressions for mean square value of the stator flux ripple vector over a sub-cycle are also given in [1] as a functions of $Q_Z, Q_1, Q_2, D, T_1, T_2$ and T_s, which gives a measure of the distortion in the line currents.

The proposed hybrid SVPWM techniques, employing the different sequences, result in less ripple current than conventional SVPWM at a given average inverter switching frequency. To develop the proposed HSVPWM methods, first, the stator flux mean square ripple trajectory over a 60° period for every sequence at different modulation indices are plotted as shown in Fig. 2. By comparing the sequences 0127, 012, 721, 0121 and 7212 with respective to each other, the zones of superior performance of each sequence can be identified. The development of hybrid PWM techniques for reduced current ripple involves determination of superior performance for every sequence. The zone of superior performance for a given sequence is the spatial zone within a sector where the given sequence results in less RMS ripple than other sequences considered. The superior performance of sequences in the space vector plane is shown in Fig. 3 for various hybrid SVPWM techniques and the proper selection of sequences within a sector gives reduced current ripple.

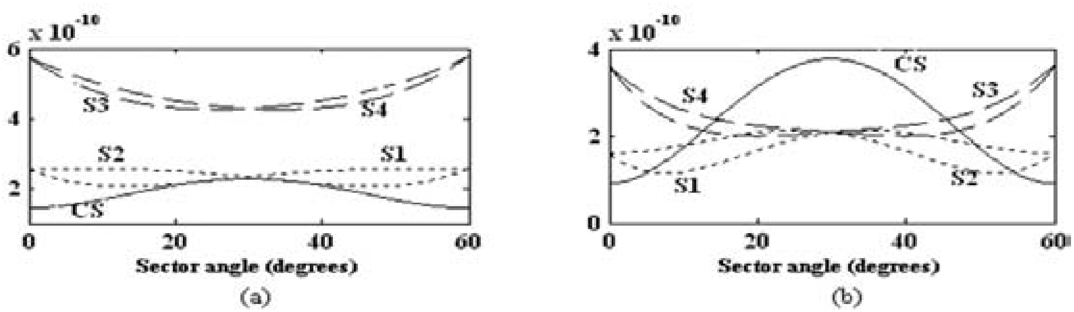

Fig. 2. Mean square stator flux ripple within a sector for (a) M=0.525 (b) M=0.73

4 Proposed DTC

In this paper, various HSVPWM methods for DTC are presented, and the block diagram of the proposed method is as shown in Fig. 4. In the proposed method, speed of the reference stator flux vector $\bar{\psi}_s^*$ is derived by the addition of slip speed and actual rotor speed. The actual synchronous speed of the stator flux vector $\bar{\psi}_s$ is calculated from the adaptive motor model. After each sampling interval, actual stator flux vector $\bar{\psi}_s$ is corrected by the error and it tries to attain the reference flux space vector $\bar{\psi}_s^*$. Thus the flux error is minimized

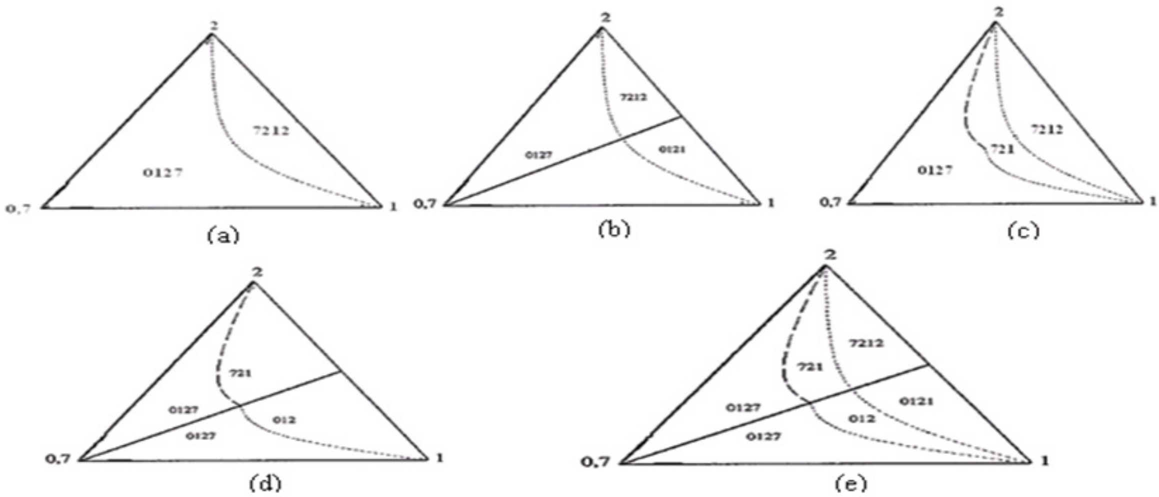

Fig. 3. Proposed hybrid SVPWM methods (a) Type-I (b) Type-II (c) Type-III (d) Type-IV and (e) Type-V

in each sampling interval. The d-axis and q-axis components of the reference voltage vector can be obtained as follows:

Reference value of the d and q- axes stator flux and actual value of the d and q-axes stator fluxes are compared in the reference voltage vector calculator block and hence the error in the d and q-axes stator flux vectors is obtained as in (3).

$$\Delta\psi_{ds} = \psi_{ds}^* - \psi_{ds} \quad \Delta\psi_{qs} = \psi_{qs}^* - \psi_{qs} \tag{3}$$

The knowledge of flux error and stator ohmic drop allows the determination of appropriate reference voltage space vectors along d and q-axes, which are given in (4).

$$V_{ds}^* = R_s^i ds + \frac{\Delta\psi_{ds}}{T_s}, \quad V_{qs}^* = R_s^i qs + \frac{\Delta\psi_{qs}}{T_s} \tag{4}$$

These $d - q$ components of the reference voltage vector are then fed to the HSVPWM block, from where the gating pulses for the inverter are generated.

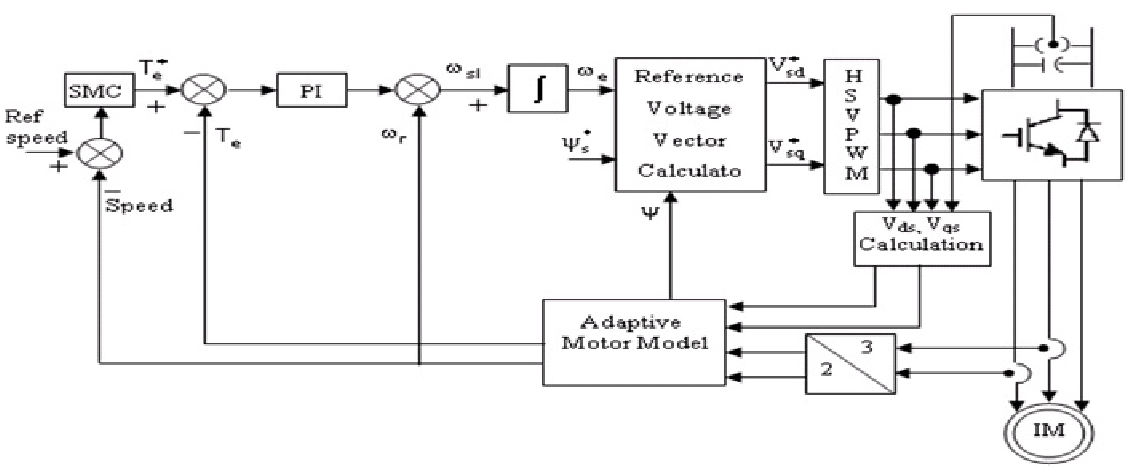

Fig. 4. Block diagram of proposed DTC

4.1 Proposed sliding mode speed controller

To improve the speed performance under uncertainties, an integral sliding mode speed controller is developed, which is robust under uncertainties caused by parameter variation and load torque. In the block diagram it is shown as SMC.

In general, the electromechanical equation of an induction motor is described as

$$J\frac{d\omega_m}{dt} + B\omega_m + T_L = T_e \tag{5}$$

Where B and J denote the viscous friction coefficient and inertia constant of the motor respectively, T_L is the external load torque and ω_m is the rotor mechanical speed in angular frequency. T_e is the electromagnetic torque of induction motor, defined as

$$T_e = \frac{3}{2}\frac{P}{2}(\lambda_{ds}i_{qs} - \lambda_{qs}i_{ds}) \tag{6}$$

The electromechanical equation can be modified as

$$\dot{\omega}_m + a\omega_m + d = bT_e \tag{7}$$

where $a = \frac{B}{J}, b = \frac{1}{J}, d = \frac{T_L}{J}$

Now consider the above electromechanical equation with uncertainties as

$$\dot{\omega}_m = -(a + \Delta a)\omega_m - (d + \Delta d) + (b + \Delta b)T_e \tag{8}$$

$\Delta a, \Delta b and \Delta d$ represents the uncertainties of the terms a, b and d introduced by system parameters J and B. Now we, further define the tracking speed error as

$$e(t) = \omega_m(t) - \omega_m^*(t) \tag{9}$$

where ω_m^* is the rotor reference speed command.

Taking derivative of previous equation with respect to time yields

$$\dot{e}(t) = \dot{\omega}_m(t) - \dot{\omega}_m^*(t) = -ae(t) + f(t) + x(t) \tag{10}$$

where the following terms have been collected in the signal $f(t)$,

$$f(t) = bT_e(t) - a\omega_m^* - d(t) - \dot{\omega}_m^*(t) \tag{11}$$

and the $x(t)$, lumped uncertainty, defined as

$$x(t) = bT_e(t) - \Delta a\omega_m - \Delta d(t) + \Delta bT_e(t) \tag{12}$$

Now, the sliding variable with integral component, is defined as

$$S(t) = e(t) - \int_0^t (h - a)e(\tau)d\tau \tag{13}$$

where h is a constant gain. Also in order to obtain the speed trajectory tracking, the following assumption should be formulated.

Assumtion-1: The h must be chosen so that the term (h-a) is strictly negative and hence $h < 0$. Then the sliding surface is defined as follows:

$$S(t) = e(t) - \int_0^t (h - a)e(\tau)d\tau = 0 \tag{14}$$

Based on the developed switching surface, a switching control that guarantees the existence of sliding mode, a speed controller is defined as

$$f(t) = he(t) - \beta sgn(S(t)) \tag{15}$$

where β is the switching gain, $S(t)$ is the sliding variable defined by (13) and sgn$(\cdot)$ is the sign function defined as

$$\text{sgn}(S(t)) = \begin{cases} +1, & \text{if} S(t) > 0 \\ -1, & \text{if} S(t) < 0 \end{cases} \tag{16}$$

In order to obtain the speed trajectory tracking, the following assumption should be formulated.

Assumtion-2: The gain β must be chosen so that $\beta \geq |x(t)|$ for all time. When the sliding mode occurs on the sliding surface (14), then, $S(t) = \dot{S}(t) = 0$, and the tracking error $e(t)$ converges to zero exponentially. Finally, the reference torque command T_e^*, can be obtained by substituting (15) in (11) as follows.

$$T_e^*(t) = \frac{1}{b}[(h.e) - \beta sgn(S) + a\omega_m^* + \dot{\omega}_m^* + d] \tag{17}$$

5 Simulation results

To verify the proposed scheme, a numerical simulation has been carried out by using Matlab/Simulink. Sampling time of $125\mu s$ and ode 4 (Runge-Kutta) methods are used for a fixed step size of $10\mu s$. For the simulation, the reference flux is taken as 1wb and starting torque is limited to 15 N-m. The induction motor used in this case study is a 1.5 KW, 1440 rpm, 4-pole, 3-phase induction motor having the following parameters: $R_s = 7.83, R_r = 7.55, L_s = 0.4751H, L_r = 0.4751H, L_m = 0.4535H, J = 0.06Kg.m^2$ and $B = 0.01N - m.sec/rad$. The values, which have chosen for sliding mode speed controller, are $h = -200$ and $\beta = 10$. Various conditions such as starting, steady state, step change in load and speed reversal are simulated. The results for CDTC are shown in Fig. 5 ∽ Fig. 7, from which it can be observed that, the ripple in torque, flux and torque is more in steady state. To reduce the steady state ripple, in this paper, hybrid SVPWM methods are proposed. Among the proposed hybrid SVPWM methods, 5-zone PWM method is superior, which gives less harmonic distortion. Hence, this paper describes the simulation results of 5-zone hybrid SVPWM based DTC only, which are given in Fig. 8 ∽ Fig. 13. From Fig. 9 it can be observed that, the steady state ripple in torque, flux and current can be reduced with the proposed method and hence acoustical noise can be reduced. Also to improve the speed performance under uncertainties caused by parameter variation or load changes, an integral sliding speed controller is developed in this paper. From the Fig. 10 and Fig. 11, it can be find that, whether the load disturbance is added or not, the speed response is almost the same with the proposed speed controller. Hence, the proposed controller provides the robustness for the system.

6 Conclusions

The conventional DTC is simple and gives quick response. But, it gives a considerable ripple in torque, flux and current in the steady state. Hence, the acoustical noise is more. To reduce the ripple in steady state, in this paper, a few space vector based hybrid PWM techniques are proposed. From the simulation results, it can be observed that, with the proposed method, the ripple in torque, flux and current in the steady state can be reduced. To develop the hybrid PWM methods, few bus-clamping sequences are used. Hence, the switching losses of the inverter can also be reduced with the proposed method. Also, to improve the speed performance, an integral sliding mode speed controller is developed, which is robust under uncertainties caused by parameter errors or by variation in load torque. From the simulation results, it can be observed that, the proposed method gives good response over the wide speed range.

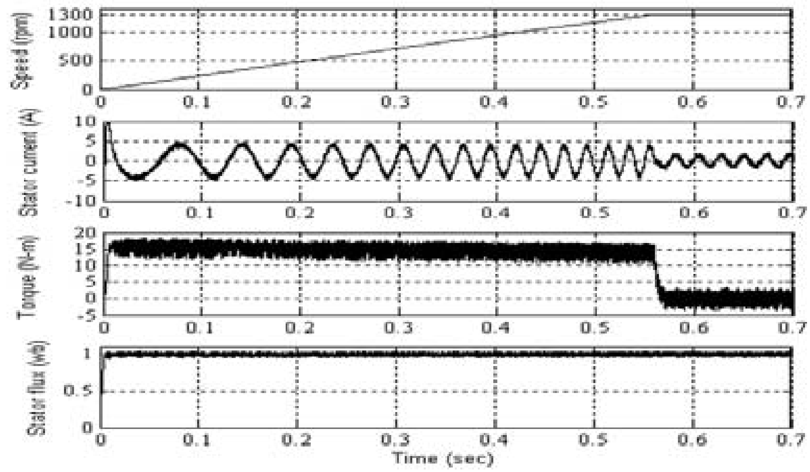

Fig. 5. Simulation results of CDTC: No-load starting transients

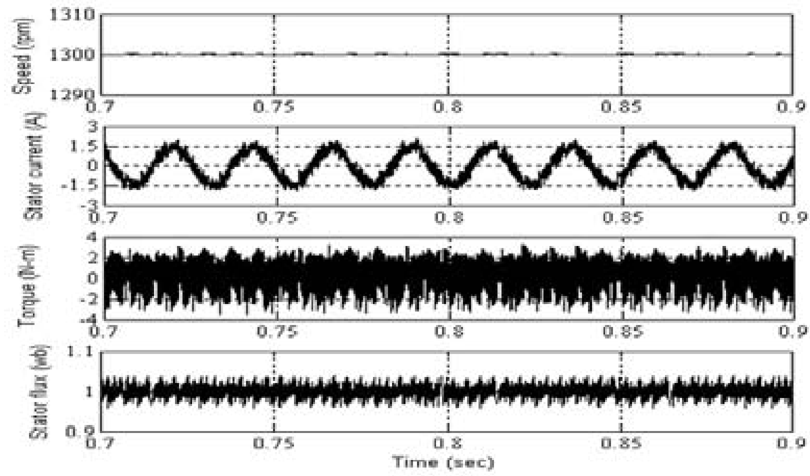

Fig. 6. Simulation results of CDTC: No-load steady state plots

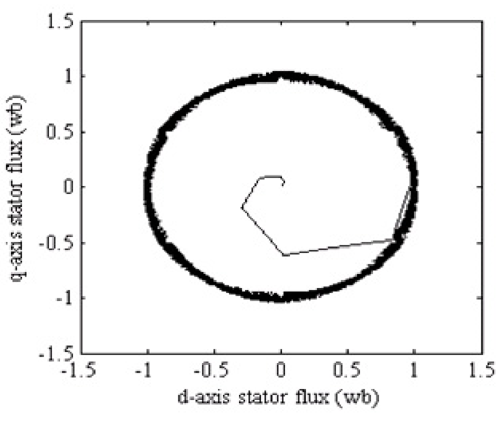

Fig. 7. Simulation results of CDTC: Locus of stator flux

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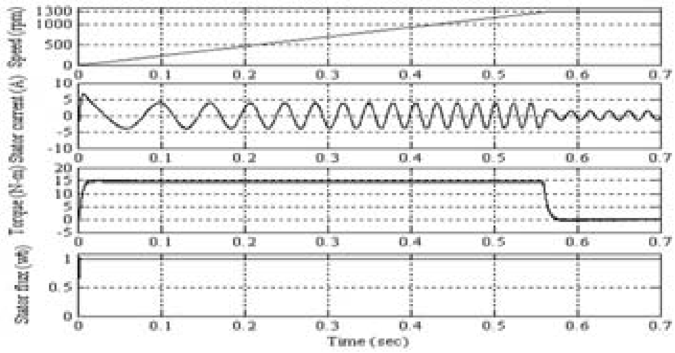

Fig. 8. Simulation results of 5-zone HSVPWM based DTC: No-load starting transients

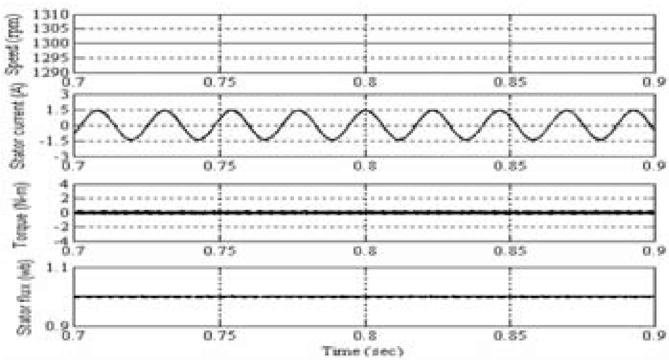

Fig. 9. Simulation results of 5-zone HSVPWM based DTC: No-load steady state plots

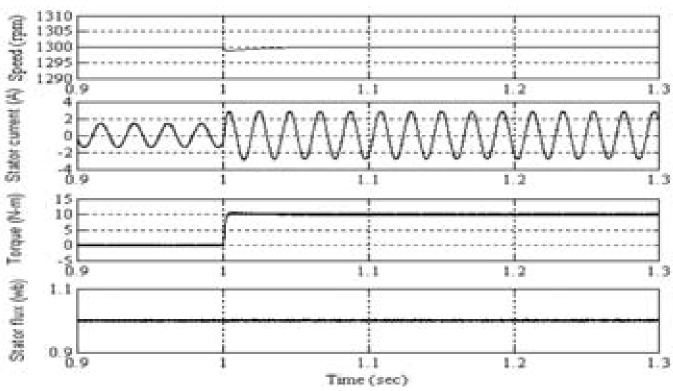

Fig. 10. Simulation results of 5-zone HSVPWM based DTC: A load torque of 10 $N-m$ is applied at $t=1$ sec

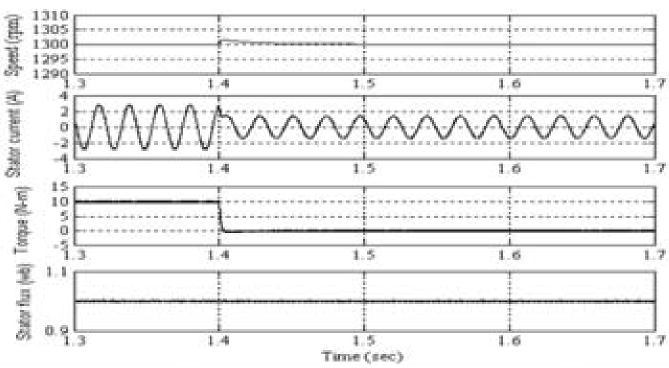

Fig. 11. Simulation results of 5-zone HSVPWM based DTC: A load torque of 10 $N-m$ is removed at $t=1.4$ sec

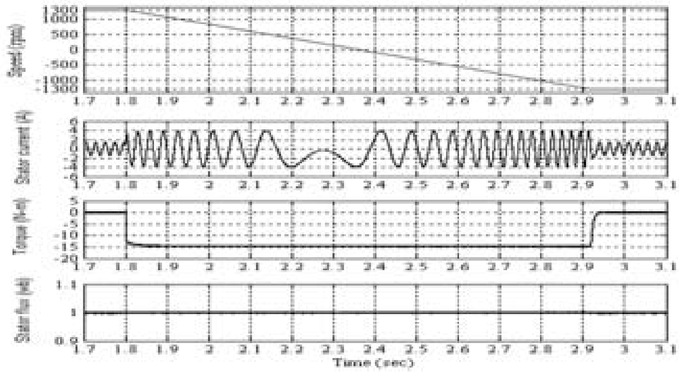

Fig. 12. Simulation results of 5-zone HSVPWM based DTC: Transients during speed reversal

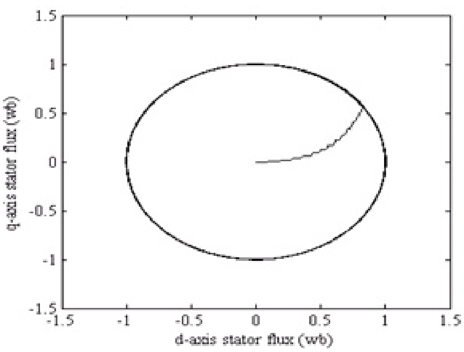

Fig. 13. Simulation results of 5-zone HSVPWM based DTC: Locus of stator flux

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