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EMF BASED PREDICTIVE CURRENT CONTROL WITH MRAS SENSORLESS SPEED ESTIMATION FOR SVPWM FED INDUCTION MOTOR

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ABSTRACT

In this research work, the speed control of a Sensorless three phase induction motor is carried out using Predictive Current control (PCC) technique and space vector Pulse Width Modulation (SVPWM). Implementation of hardware complexity of a motor drive can be minimized by estimating controls rather than measuring using sensors like Model Reference Adaptive Systems (MRAS) based on Back EMF estimation. The corresponding sensors are to be eliminated so that the system becomes more robust. The PCC is dependent on a closed loop observer, which is insensitive to parameter variations. This closed loop observer is used to calculate the required state variables over a wide frequency range. This methodology is more reliable even at very low frequency due to its insensitiveness nature against changes of motor parameters. Further, a SVPWM voltage source inverter is to be used to serve the induction motor rather than SPWM. The drive system with proposed adaptive mechanism is simulated by MATLAB/Simulink for the verification of the performance of the technique.

Keywords*: SVPWM, Predictive Current Control, Closed loop observer, MRAS, Inverter, Induction Motor.*

I. INTRODUCTION

Induction motors are more rugged and since they are inexpensive machines, they are extensively used in industrial applications. Therefore much concentration is given to their controlling aspects for various applications with different control requirements. Traditional open-loop control scheme with variable frequency for the induction machine may suggest a satisfactory solution under limited constraints. When high performance dynamical operation is required, these methods are unsatisfactory. Hence, more sophisticated control schemes are essential to study and enhance the performance induction motor and to be compared with DC motors. Recent advancements in the area of drive control methods, high speed semiconductor power devices, powerful and high performance microcontrollers made induction motors as alternative for DC motors in industrial applications [1],[2],[3]. In last few decades, the vector control theory has been receiving much attention because of its better steady and dynamic performance over conventional control schemes in controlling torque and speed parameters. The most extensively used induction motor drive vector control method has been the field oriented control (FOC) [4]. Out of all the available vector control schemes, the Sensorless speed vector control [5] has been a relevant area of interest for many researchers due to its low operating cost, high reliability and simple maintenance. There are two main parameters which are required in sensorless speed vector control of induction motor [6], which are, the motor flux and speed estimation. The above listed parameters are necessary for establishing the outer speed loop feedback, flux and torque control algorithms.

To get good performance of sensorless vector control [7], different speed estimation strategies have been proposed. Such as direct calculation method, adaptive full order observer [8], model reference adaptive system (MRAS) [9], Luenberger observers, Extended Kalman Filter (EKF) [10], Slot harmonics, Sliding mode observer [11], Estimators using fuzzy logic control [11] and artificial intelligence etc., [4]. Out of the various speed estimation schemes, MRAS-based speed Sensorless estimation [12]-[13] has been commonly used in AC speed regulation systems due to its good performance characteristics, simple, ease of implementation and needs a low computation power and have a high speed adaptation even at zero speeds, where the output of the reference model is compared with the output of an adaptive model until the errors between the above two models is totally minimized.

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In this work, it can be stated that the rotor speed estimation performance of these methodologies is quite satisfactory in the obtained simulation and experimental results. Finally, a combination of renowned open loop observers, voltage model & current model are used to estimate the rotor flux and rotor flux angle, which can be used in direct field orientation. Similarly, several studies have already been done to find an efficient technique for the predictive stator current control (PSCC) [25] of Sensorless induction motor drives [14]-[16].

A. Proposed System

This work emphasizes the real-time operating system of the electric drive. The system has PSCC implemented in the IM sensorless speed system with field oriented control methodology. FOC is employed along with PSCC and space vector PWM [17]-[18] so as to maintain a constant switching frequency. The predictive current control method is altered by employing an observer system instead of a simple load model [19]. The usage of the observer eliminates the problems associated with system start up. To simplify the control system, the proposed PCC is integrated with the speed/flux observer for FOC IM drives [20].

The proposed sensorless control scheme of induction motor drive system is simple, robust and can be operated with a wide speed range, including extremely low and very high speeds.

The simplified block representation of the proposed control scheme is shown in Fig 1. The motor is supplied via a frequency converter, which include diode rectifier and transistorized voltage source inverter. The FOC is incorporated for speed and flux control with PI controllers. The resultant motor current which has been transformed from the d^s,q^s to the d,q coordinates, is controlled by the PSCC. Subsequently, the PSCC collaborates with the space vector PWM, for assuring a constant switching frequency for the inverter. The inverter along with SVPWM and PSCC works as a controlled current source. The developed control system functions without a speed sensor and measures inverter input voltage and output currents. Other essential parameters required by the control scheme are evaluated in a closed-loop observer system.

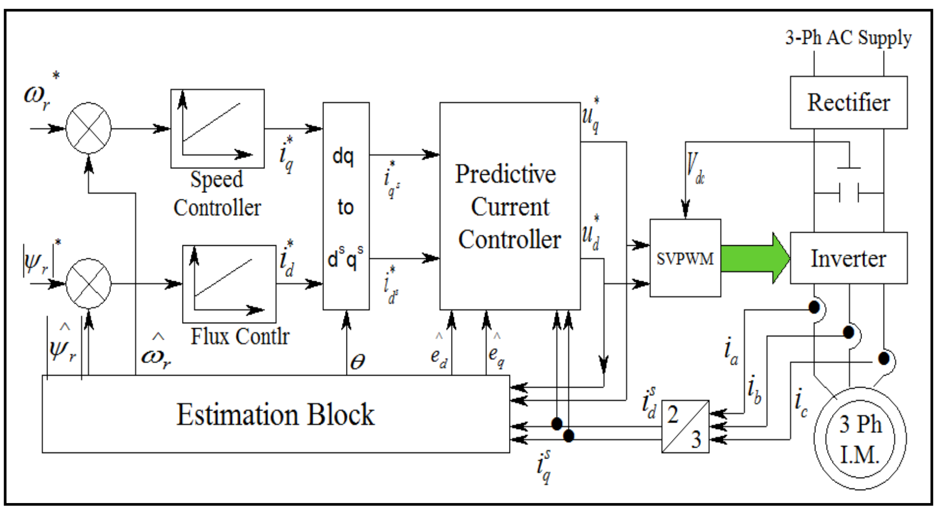

Fig.1. Block representation of the proposed system

II. INDUCTION MOTOR MODEL

The mathematical analysis of an ideal squirrel cage type induction motor in an arbitrary reference frame is shown in fig.2 (a, b) [21].

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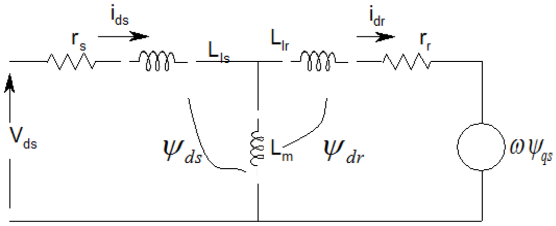

Fig. 2(a)

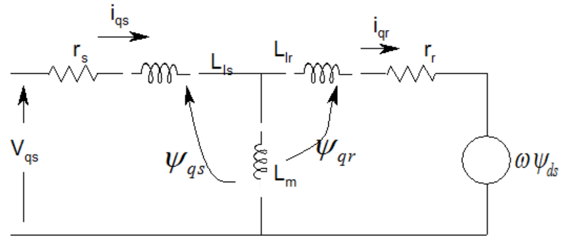

Fig. 2(b)

Fig.2. Equivalent circuit diagram of induction motor (a) d-axis, (b) q-axis

From the above equivalent circuit diagrams we can write the d-q stator and rotor voltage equations, the stator voltage equations can be written as

$$\left.\begin{aligned} v_{ds} &= r_s i_{ds} + \frac{d\Psi_{ds}}{dt} - w\Psi_{qs} \\ v_{qs} &= r_s i_{qs} + \frac{d\Psi_{qs}}{dt} + w\Psi_{ds} \end{aligned}\right\} \qquad (1)$$

Similarly we can write for the rotor voltage eqn's, But it is known that, $V_{dr} = 0$ and $V_{qr} = 0$

Then the rotor voltage equations,

$$\left.\begin{aligned} 0 &= r_r i_{dr} + \frac{d\Psi_{dr}}{dt} - (w - w_m)\Psi_{qr} \\ 0 &= r_r i_{qr} + \frac{d\Psi_{qr}}{dt} + (w - w_m)\Psi_{dr} \end{aligned}\right\} \qquad (2)$$

Then write the stator flux linkage equation's from the above stator voltage equations,

$$\left.\begin{aligned} \Psi_{ds} &= \int (v_{ds} - i_{ds} r_s + w\Psi_{qs}) \\ \Psi_{qs} &= \int (v_{qs} - i_{qs} r_s - w\Psi_{ds}) \end{aligned}\right\} \qquad (3)$$

Similarly for rotor,

$$\left.\begin{aligned} \Psi_{dr} &= \int (v_{dr} - i_{dr} r_r + (w - w_m)\Psi_{qr}) \\ \Psi_{qr} &= \int (v_{qr} - i_{qr} r_r - (w - w_m)\Psi_{dr}) \end{aligned}\right\} \qquad (4)$$

The flux linkages expressed in terms of currents

$$\left.\begin{aligned}
\Psi_{qs} &= L_{ls} i_{qs} + L_m (i_{qs} + i_{qr}) \\
\Psi_{qr} &= L_{lr} i_{qr} + L_m (i_{qs} + i_{qr}) \\
\Psi_{qm} &= L_m (i_{qs} + i_{qr}) \\
\Psi_{ds} &= L_{ls} i_{ds} + L_m (i_{ds} + i_{dr}) \\
\Psi_{dr} &= L_{lr} i_{dr} + L_m (i_{ds} + i_{dr}) \\
\Psi_{dm} &= L_m (i_{ds} + i_{dr})
\end{aligned}\right\} \quad (5)$$

The electromagnetic torque is produced by the coupling of air gap flux and rotor mmf which can be expressed in general vector form as

$$T_{em} = \frac{3}{2} \times \frac{P}{2} \times \Psi_m \times I_r \quad (6)$$

The torque equations can be written in stationary frame with corresponding variables as

$$\left.\begin{aligned}
T_{em} &= \frac{3}{2} \times \frac{p}{2} \times (\Psi^s_{dm} i^s_{qr} - \Psi^s_{qm} i^s_{dr}) \\
T_{em} &= \frac{3}{2} \times \frac{p}{2} \times (\Psi^s_{dm} i^s_{qs} - \Psi^s_{qm} i^s_{ds})
\end{aligned}\right\} \quad (7)$$

The electromagnetic torque in terms of load torque and inertia,

$$T_{em} = T_L + J \frac{dwr}{dt} \quad (8)$$

III. DIRECT OR FEEDBACK VECTOR CONTROL

The basic representation of the direct vector control method for a PWM voltage fed inverter drive is shown in the fig.3 [22]. The primary vector control parameters, i_{ds}^* and i_{qs}^*, which are the DC components used in synchronously rotating frame, are transformed into stationary frame with the help of a standard unit vector ($\cos\theta_e$ & $\sin\theta_e$) generated from flux vector signals ψ_{dr}^s and ψ_{qr}^s. The obtained stationary frame components are then transformed to phase current commands for the inverter. The flux signals ψ_{dr}^s and ψ_{qr}^s are being generated from the machine terminal voltages and currents with the help of voltage model estimator. The torque component of current i_{qs}^* is generated from the speed control loop through a bipolar limiter. The torque proportional to i_{qs} (with constant flux), can be bipolar. It is negative with negative i_{qs} and correspondingly, the phase position of i_{qs} becomes negative. An Additional torque control loop can be added within the speed loop, if desired. Through programming the flux command, fig. 3 can be extended to field- weakening mode so that the inverter remains in PWM mode [23].

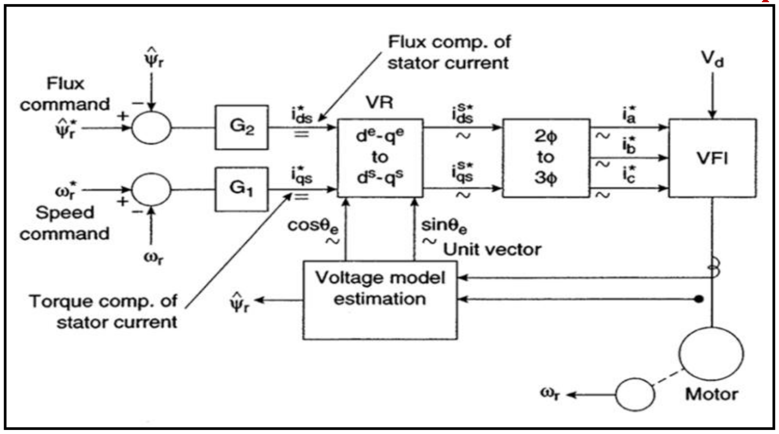

Fig.3. Direct vector control with rotor flux orientation [23]

With the help of stationary frame rotor flux vectors ψ_{dr}^{s} and ψ_{qr}^{s}, the following equations are obtained

$$\left.\begin{aligned} \Psi_{dr}^{s} &= \hat{\Psi}_r \cos\theta_e \\ \Psi_{qr}^{s} &= \hat{\Psi}_r \sin\theta_e \end{aligned}\right\} \quad (9)$$

$$\hat{\Psi}_r = \sqrt{\Psi_{qr}^{s\,2} + \Psi_{dr}^{s\,2}} \quad (10)$$

A. Flux Vector Estimation- Voltage Model

In this method, the machine terminal voltages and currents are sensed and the fluxes are computed from the stationary frame (d^s-q^s) equivalent circuit as shown in the fig. 4. The equations are given as:

$$\left.\begin{aligned} i_{qs}^{s} &= \frac{2}{3} i_a - \frac{1}{3} i_b - \frac{1}{3} i_c = i_a \\ i_{ds}^{s} &= -\frac{1}{\sqrt{3}} i_b + \frac{1}{\sqrt{3}} i_c \end{aligned}\right\} \quad (11)$$

Similarly,

$$\left.\begin{aligned} v_{qs}^{s} &= \frac{2}{3} v_a - \frac{1}{3} v_b - \frac{1}{3} v_c \\ v_{ds}^{s} &= -\frac{1}{\sqrt{3}} v_b + \frac{1}{\sqrt{3}} v_c \end{aligned}\right\} \quad (12)$$

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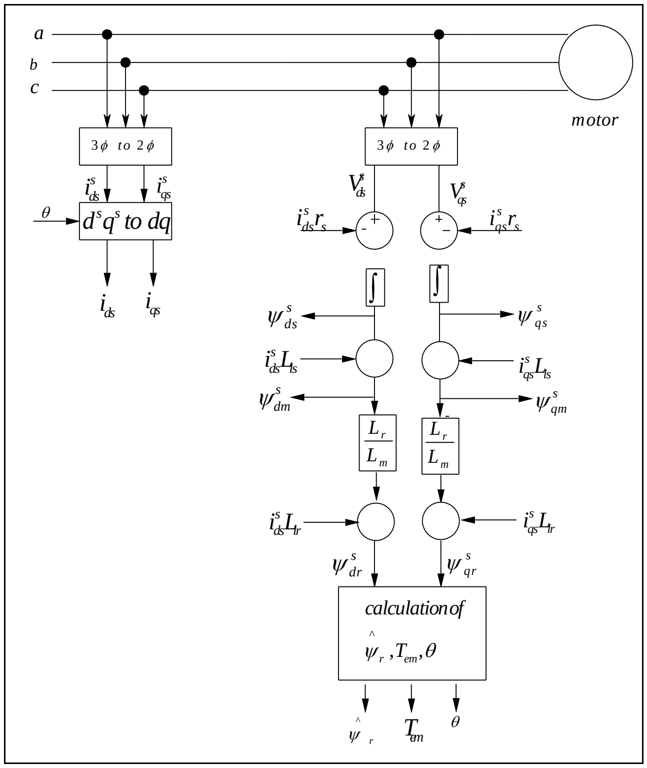

Fig.4 Voltage model feedback signal estimation block diagram

Stator flux linkages,

$$\left.\begin{aligned} \Psi_{ds}{}^{s} &= \int (v_{ds}{}^{s} - i_{ds}{}^{s} r_s)dt \\ \Psi_{qs}{}^{s} &= \int (v_{qs}{}^{s} - i_{qs}{}^{s} r_s)dt \\ \hat{\Psi}_s &= \sqrt{\Psi_{qs}^{s\;2} + \Psi_{ds}^{s\;2}} \end{aligned}\right\} \tag{13}$$

Rotor flux linkages,

$$\left.\begin{aligned} \Psi_{dr}^{s} &= \frac{L_r}{L_m}(\Psi_{ds}^{s} - \sigma L_s i_{ds}{}^{s}) \\ \Psi_{qr}^{s} &= \frac{L_r}{L_m}(\Psi_{qs}^{s} - \sigma L_s i_{qs}{}^{s}) \end{aligned}\right\} \tag{14}$$

Where,

$$\sigma = 1 - \frac{L_m{}^2}{L_r L_s} \tag{15}$$

Then finally we get from the above equation

$$T_e = \frac{3}{2}\left(\frac{p}{2}\right)\frac{L_m}{L_r}(\Psi_{dr}^{s} i_{qs}^{s} - \Psi_{qr}^{s} i_{ds}^{s}) \tag{16}$$

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IV. MODEL REFERENCE ADAPTIVE SYSTEM (MRAS)

This system is one of the most familiar adaptive control scheme employed in motor control applications for monitoring and tracking system parameters and the state estimations. There are multiple model reference adaptive control mechanisms such as parallel modeling approach, series modeling approach, direct model and indirect model etc,. MRAS used in this research work is a parallel model that establishes a comparison between the outputs of the reference model and the adaptive model, and further processes the error between these two which do not deteriorate the stability requirements of the applied system [12],[23].

The speed can be calculated by the model referencing adaptive system where the output of reference model is compared with the output of an adaptive model until the errors between the two models gets minimized. The corresponding block diagram for speed estimation using this technique is shown in fig 5. In this work, MRAS based Sensorless speed control of predictive controller is presented [28].

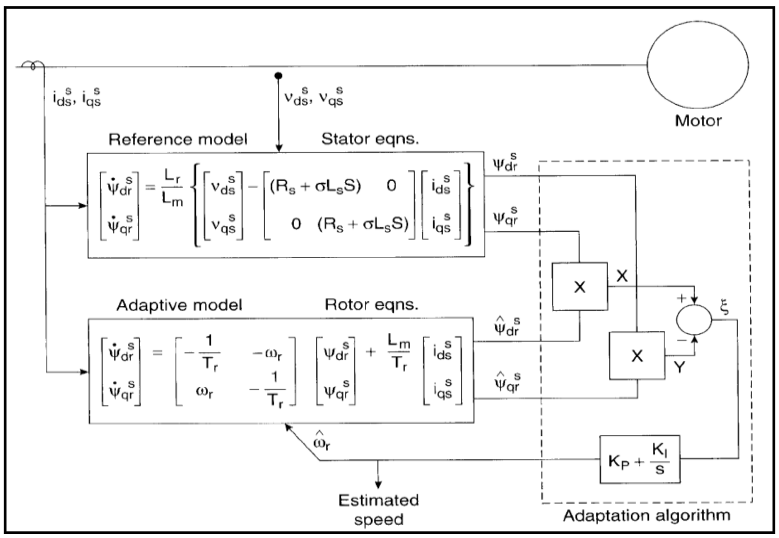

Fig.5 Model reference adaptive system [23]

Based on flux linkage equation's the mechanism for speed can be expressed as

$$\hat{w}_r = \xi(k_p + \frac{k_i}{s}) \qquad (17)$$

Where,

$$\xi = \hat{\Psi}_{\alpha r}\Psi_{\beta r} - \hat{\Psi}_{\beta r}\Psi_{\alpha r} \qquad (18)$$

$$T_r = \frac{L_r}{R_r}$$

$$\sigma = 1 - \frac{L^2{}_m}{L_s L_r}$$

A. MRAS Based on Back-EMF Estimation

The proposed MRAS is using the state observer model and the magnet flux model and these two models for the back-EMF estimation [24]. The rotor speed is obtained from the adaptation mechanism based on the error between

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estimated quantities obtained by the two models. Those are reference model and adaptive model. The equation for estimation of EMF is given by,

For d-axis

$$\hat{e}_d = L_m \frac{di_{md}}{dt}$$

$$\hat{e}_d = \frac{L_m}{L_r} \frac{d\Psi_{rd}}{dt}$$

$$\hat{e}_d = v_{sd} - (r_s i_{ds} + \sigma L_S \frac{di_{ds}}{dt})$$

Similarly for q-axis

$$\hat{e}_q = L_m \frac{di_{mq}}{dt}$$

$$\hat{e}_q = \frac{L_m}{L_r} \frac{d\Psi_{rq}}{dt}$$

$$\hat{e}_q = v_{sd} - (r_s i_{qs} + \sigma L_S \frac{di_{qs}}{dt})$$

B. Predictive Current Control

Predictive control is a very wide class of controllers that have found rather recent application in power converters. The main feature of the predictive control is the use of the future behavior of the controlled variables. This feature is used by the controller to obtain the optimal actuation. The major advantage of the predictive control scheme is that the concepts are very simple and intuitive [14],[16],[26].

C. Deadbeat current control method

A sophisticated deadbeat current control [27] method is shown in Fig.6. It can be observed that instead of PI controller deadbeat controller is employed. Modulator is employed to apply the reference voltage and the load model for a generic RLE load is described by the following space vector equation -

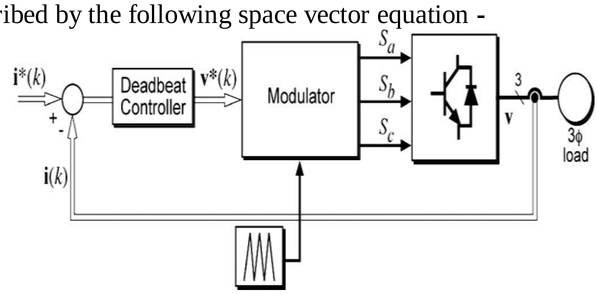

Fig.6 Deadbeat current control [13]

$$v = ir + L\frac{di}{dt} + e$$

Where, ‘v’ & ‘i’ are the voltage and current space vectors and the EMF voltage space vector is represented by ‘e’.

For a sampling time Ts, The following discrete-time equation can be formulated:

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$$\frac{1}{\delta}i(k+1)-\frac{x}{\delta}i(k)=v(k)-e(k)$$

$$where,\ \delta=e^{-\frac{T_S r}{L}}$$

$$and\ x=\frac{1}{r(1-e^{-\frac{T_S r}{L}})}$$

Based on the discrete-time model (2), the reference voltage vector is obtained as

$$v^{*}=\frac{1}{\delta}((i^{*}(k+1)-i(k))+e(k)) \qquad (19)$$

Reference voltage **v*** is applied in the converter using a modulator.

V. SIMULATION RESULTS

For the analysis using MATLAB/Simulink, 5KW, 4-pole three phase induction motor is used with the following Parameters:

Stator Resistance, Rs = 2.65 ohms
Rotor resistance, Rr = 2 ohms
Stator inductance, Ls = 301.4 mH
Rotor inductance, Lr = 306.5 mH
Mutual inductance, Lm = 291.1 mH
Number of poles, P = 4
Inertia constant, J = 0.0055 kgm^2/sec
Frequency, F =50 Hz
Rated voltage, V= 600 V
Rated power, P =5KW

Simulation is worked out by using MATLAB/Simulink. The obtained results are shown in the following figures. (Fig 7(a) through 7(h)).

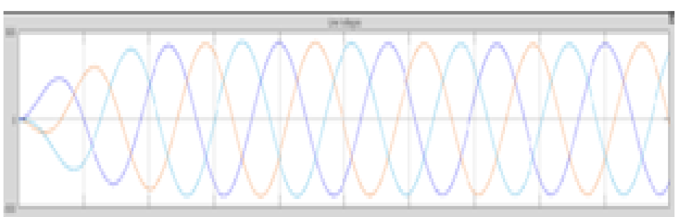

Fig7 a).Stator voltage waveforms

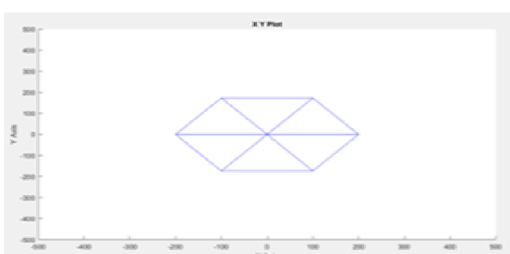

Fig7 b). Switching vector for the SVPWM inverter

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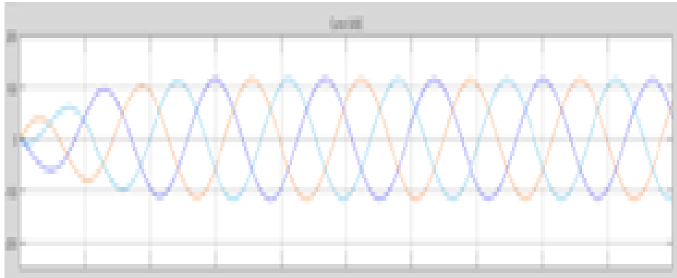

Fig7 c). Stator current waveforms

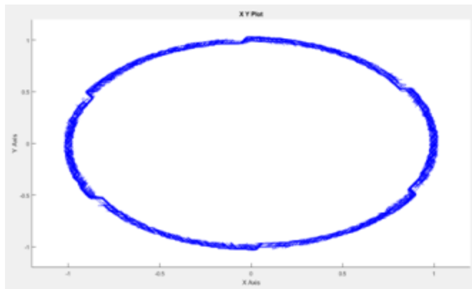

Fig7 d). d-q axis graph for rotor currents

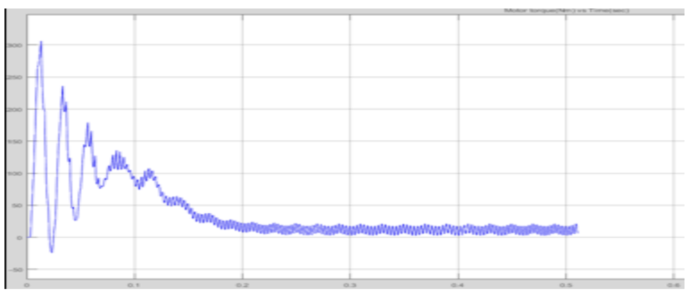

Fig7 e). Electromagnetic torque without load

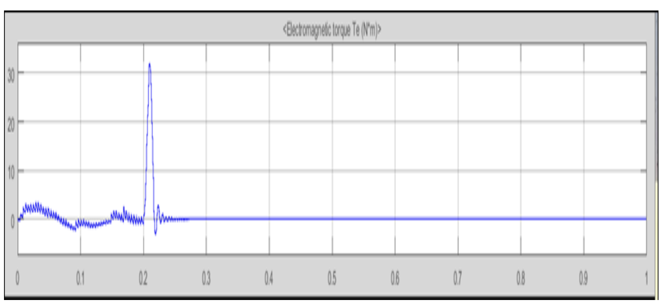

Fig7 f). Electromagnetic torque with load torque of 5NM with 0.2 sample time

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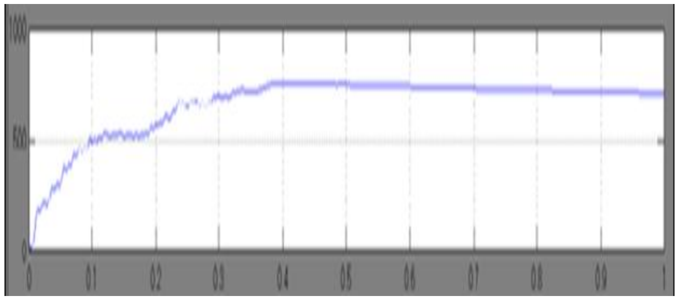

Fig7 g). Speed of the IM w.r.t MRAS back emf (rad/sec)

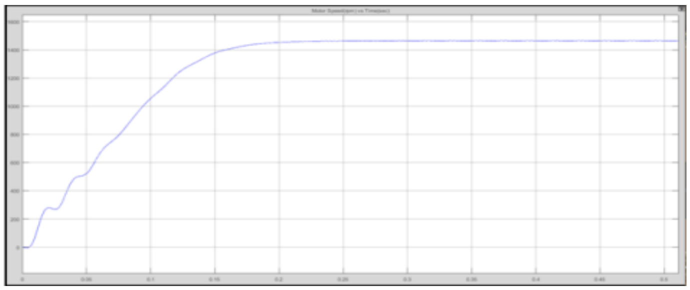

Fig7 h). rated speed (N=1500 rpm) of the induction motor

VI. CONCLUSIONS

By Integrating multiple model blocks in the form of subsystems, an elaborate Simulink schematic of a sensorless induction motor drive system that is capable of simulating typical problems, connected with a high performance digital controlled drive has been proposed. The simulation results validate the assumptions and equations used for modeling the related electromagnetic and electromechanical phenomena and the control system and also reveal the significant features of the closed loop drive system. The attainment and retention of field orientation and realization of independent control of air gap flux and shaft torque through predictive control of the stator currents are important results of this investigation. It is found that the drive system is capable of quickly responding to required changes in speed and torque on the load side and the controller is versatile for fast variation of inverter frequency and voltage; while at the same time make the induction motor to run as a virtual current fed machine. Comparing with the previous methodologies, the proposed MRAS based method is more reliable with high dynamic performance and more accurate in terms of sustaining speed and flux. For inverter switching operation, SVPWM technique is presented and on comparison with SPWM, this technique exhibit less harmonics, 15% THD reduction in SVPWM. The proposed predictive controller is a predefined optimization criterion

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