Design and Simulation of Three Level Neutral Point Clamped Inverter Fed Induction Motor Drive

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Abstract—The objective of the project is to obtain some proficiency in building converters to control speed of medium voltage induction motors. The thesis covers summary of literature surveyed, selection of a topology of converter for design, design of a 3-Level NPC inverter power circuit and FOC parameters. The entire system has been simulated using PSIM software for verification of design. Results obtained are presented. Results prove that the understanding of the student (author) is satisfactory. A low voltage 10kW squirrel cage induction motor has been chosen as the load to converter for design and simulation. Future work may be design of complete control circuits implementing FOC, building complete hardware and testing the small motor chosen.

Index Terms—3-Level NPC Inverter; Dynamic Modeling of IM; Losses Calculation; Scalar Control; Vector Control and IFOC.

I. INTRODUCTION

Variable speed drives have a wide variety of applications such as pumps, fans, compressors, transportation, cement mills, rolling mills, paper mills, machine tools, robotics drives, automation etc. A few decades ago DC motors dominated the market as it is much easier to control DC motors than induction motors although induction motors are more robust and cheaper. Induction motors are used in a wide range of sizes. Larger motors are used at medium voltage (>1000V). Medium voltage induction motor drives (MV drives) in general cover voltages from 1.1 kV-13.2 kV and power from 0.5 MW-40 MW. Larger drives (up to 100kW) are usually with synchronous motor. As induction motors are singly fed, it is difficult to control speed and torque independently. Theories to obtain high performance from induction motors were developed in seventies but implementation required high speed computation and switching devices. The basic modeling of induction motor is quite established and simplified equivalent circuit model of the motor gives good performance prediction for steady state operation of the motor with sinusoidal supply voltages. But it fails good model for dynamic performance. The dynamic model of the induction motor is somewhat involved because of rotating magnetic field. The relationship of these magnetic fields depends on speed and load [1, 2].

The idea of multilevel inverter was introduced in 1975 and in 1981, A. Nabae, I. Takahashi, and H. Akagi presented the first three level inverter. In order to achieve

high power, multilevel inverter used an array of semiconductor devices with several lower DC voltage sources for power conversion resulting in stepped voltage waveform. There are two types of inverter, the two level and multilevel. A Two level inverter generates an output voltage with two levels while the minimum numbers of voltage levels in multilevel inverter are three. The increase in number of levels of the inverter will have a good output voltage waveform with reduced harmonic distortion but with increase in control complexity. A multilevel inverter has great advantages compared to two level inverter mentioned as below:

- a) The output waveforms contain very low total harmonic distortion (THD) and lower dv/dt.
- b) Multilevel inverter can draw input current with very little distortion.
- c) They can operate at both high switching frequency as well as low switching frequency.

With several advantages multilevel inverters do have some disadvantages. As with the increase of voltage level the number of power semiconductor switches also increased so overall system becomes more expensive and complex. The NPC multilevel inverter is widely used in various industrial applications among other multilevel inverter topology. This paper presents a brief review on three-level NPC inverter topology and its modulation strategies. The major issue of this inverter topology is neutral point voltage imbalance in the DC-link capacitors is a commonly reported problem for NPC and flying capacitor type multilevel inverters. However, this does not arise when two DC voltage sources (transformer windings with rectifiers) of equal voltage are connected in series [3, 4].

The vector control can be implemented by direct called feedback and indirect called feed forward control method. The direct vector control possesses more complex calculation for estimate rotor flux and unity vector. But indirect vector control possesses the rotor speed is forwarded to find the unity vector and flux is estimated from the linear relationship of speed and rotor flux. The vector control decouples the three phase (a, b, c) currents into two phase (d, q) currents. Here the d-axis is the current producing magnetic field and q-axis is the current producing the torque. So the vector control method provides the performance of induction motor drive like a DC motor. It is difficult to implement complex computations are involved [5].

This paper focused about the Power circuit design of a 3L Neutral point clamped inverter for 10kW motor and as much heat is generated heat sinks are required to keep the junction temperature within limits. It is planned to use 1 air cooled heat sink for all the semiconductor devices namely three IGBT module and two rectifier modules. Losses in all semiconductor devices are calculated and one force air

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cooled heat sink is found to be enough to mount all 5 five modules. Heat sink thermal resistance for good margins in junction temperature is estimated and a standard heat sink with bonded fins has been chosen. Thermal simulations are done by using free software (R-Tools) from Mersen and the temperatures have been found to be within limits [6].

II. DYNAMIC MODELING OF INDUCTION MOTOR

The dynamic modeling of induction motor is of two types. One is based on dq-axis theory and other losses in semiconductor devices. The dq-axis model does not need to use complex numbers or variables. Both models are equally valid for the analysis of transient and steady-state performance of the induction motor.

A. d-q axis motor model

The induction motor dq-axis model can be derived using three-phase circuit theory and then transformed into the two-phase (dq-axis) frame can be performed by using abc/dq Frame Transformation.

The balanced three phase (a,b,c) and two phase (d,q) quantities and The angle between x_d and x_q is θ .

By resolving a, b and c quantities on to d q axes, we get:

$$\begin{aligned} x_d &= \cos(\theta)x_a + \cos(\theta - \frac{2\pi}{3})x_b + \cos(\theta - \frac{4\pi}{3})x_c \\ x_q &= -\sin(\theta)x_a - \sin(\theta - \frac{2\pi}{3})x_b - \sin(\theta - \frac{4\pi}{3})x_c \end{aligned}$$

To maintain equivalence between the two frames, a factor $2/3$ is used.

The transformation of the three-phase (abc-axis) variables of an induction motor to the equivalent two-phase (dq-axis) variables can be performed by

$$\begin{bmatrix} x_d \\ x_q \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos\theta & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta - \frac{4\pi}{3}) \\ -\sin\theta & -\sin(\theta - \frac{2\pi}{3}) & -\sin(\theta - \frac{4\pi}{3}) \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad (1)$$

Where x represents current, voltage, or flux linkage and θ is the angular displacement between the a-axis and d-axis of the three-phase and two-phase reference frames as shown in Fig.1.

Fig. 1. Variables in three-phase (abc) stationary frame and two-phase (dq) synchronous frame.

The three-phase variables, x_a , x_b and x_c are in the stationary

reference frame which does not rotate in space whereas the two-phase variables, x_d and x_q are in the synchronous reference frame whose direct (d) and quadrature (q) axes rotate in space at the synchronous speed ω_e . Note that ω_e is the angular electrical (not mechanical) speed of the rotating magnetic field of the motor, given by:

$$\omega_e = 2\pi f_s \quad (2)$$

where f_s is the frequency of the stator variables. The angle θ can be found from:

$$\theta(t) = \int_0^t \omega_e(t) dt + \theta_o \quad (3)$$

The transformation equation of (1) is valid only for a three-phase balanced system, in which

$$x_a + x_b + x_c = 0 \quad (4)$$

Similarly, the two-phase variables in the synchronous frame can be transformed back to the three-phase stationary frame by

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \begin{bmatrix} \cos\theta & -\sin\theta \\ \cos(\theta - \frac{2\pi}{3}) & -\sin(\theta - \frac{2\pi}{3}) \\ \cos(\theta - \frac{4\pi}{3}) & -\sin(\theta - \frac{4\pi}{3}) \end{bmatrix} \begin{bmatrix} x_d \\ x_q \end{bmatrix} \quad (5)$$

It can also be obtained by decomposing the space vectors in the space vector motor model into the d- and q-axis components that is,

$$\begin{aligned} \overline{v_s} &= v_{ds} + jv_{qs}; \overline{i_s} = i_{ds} + ji_{qs}; \overline{\lambda_s} = \lambda_{ds} + j\lambda_{qs} \\ \overline{v_r} &= v_{dr} + jv_{qr}; \overline{i_r} = i_{dr} + ji_{qr}; \overline{\lambda_r} = \lambda_{dr} + j\lambda_{qr} \end{aligned} \quad (6)$$

Substituting (1) to (6), the dq-axis voltage equations for the induction motor can be obtained:

$$\begin{aligned} v_{ds} &= R_s i_{ds} + p\lambda_{ds} - \omega \lambda_{qs} \\ v_{qs} &= R_s i_{qs} + p\lambda_{qs} - \omega \lambda_{ds} \\ v_{dr} &= R_r i_{dr} + p\lambda_{dr} - \omega \lambda_{qr} \\ v_{qr} &= R_r i_{qr} + p\lambda_{qr} - \omega \lambda_{dr} \end{aligned} \quad (7)$$

where the stator and rotor flux-linkages can be calculated by

$$\begin{aligned} \lambda_{ds} &= L_{ls} i_{ds} + L_m (i_{ds} + i_{dr}) \\ \lambda_{qs} &= L_{ls} i_{qs} + L_m (i_{qs} + i_{qr}) \\ \lambda_{dr} &= L_{lr} i_{dr} + L_m (i_{ds} + i_{dr}) \\ \lambda_{qr} &= L_{lr} i_{qr} + L_m (i_{qs} + i_{qr}) \end{aligned} \quad (8)$$

The electromagnetic torque can be expressed in a number of ways. Some of the commonly used expressions are

$$T_e = \left\{ \frac{3p}{2} (i_{qs} \lambda_{ds} - i_{ds} \lambda_{qs}) \right. \\ \left. \frac{3pL_m}{2} (i_{qs} i_{dr} - i_{ds} i_{qr}) \right. \\ \left. \frac{3pL_m}{2L_r} (i_{qs} \lambda_{dr} - i_{ds} \lambda_{qr}) \right\} \quad (9)$$

Equations (7) to (9) together with the motion equation of (1) represent the d_q-axis model of the induction motor

B. Mathematical Modeling Losses in semiconductor devices

The power losses in a 3L-Neutral point clamped inverter topology can be calculated according to

Losses in IGBTs: Modulation Index M is assumed to be 1.

Outer IGBTs:

Conduction losses in each IGBT

$$p_{cond} = \frac{M\bar{i}}{12\pi} \{3v_{ceo} [[(\varphi - \pi) \cos(\varphi) + \sin(\varphi)]] + 2r_{ce} \bar{i} [1 + \cos^2(\varphi)]\}$$

Switching losses in each IGBT

$$p_{sw} = f_{sw} E_{sw} \left(\frac{\bar{i}}{I_{ref}} \right)^{k_i} \left(\frac{v_{cc}}{v_{ref}} \right)^{K_p} \left(\frac{1}{2\pi} [1 + \cos(\varphi)] \right) G_i$$

Total losses in outer IGBTs= 2(conduction losses + switching losses)

Inner IGBTs: Conduction losses is given by

$$p_{cond} = \frac{\bar{i}}{12\pi} \{v_{ceo} [12 + 6M[\varphi \cos(\varphi) - \sin(\varphi)]] \\ + r_{ce} \bar{i} [3\pi - 2M(1 - \cos^2(\varphi))]\}$$

Switching losses in each IGBT is given by

$$p_{sw} = f_{sw} E_{sw} \left(\frac{\bar{i}}{I_{ref}} \right)^{K_i} \left(\frac{v_{cc}}{v_{ref}} \right)^{K_p} \left(\frac{1}{2\pi} [1 - \cos(\varphi)] \right) G_i$$

Total losses in inner IGBTs=2 *conduction losses

Losses in clamping diodes:

Conduction losses is given by

$$p_{cond} = \frac{\bar{i}}{12\pi} \{v_{fo} [12 + 3M[(2\varphi - \pi) \cos(\varphi) - 2\sin(\varphi)]] \\ + r_f \bar{i} [3\pi - 4M(1 + \cos^2(\varphi))]\}$$

Switching losses is given by

$$p_{sw} = f_{sw} E_{sw} \left(\frac{\bar{i}}{I_{ref}} \right)^{k_i} \left(\frac{v_{cc}}{v_{ref}} \right)^{K_p} \left(\frac{1}{2\pi} [1 + \cos(\varphi)] \right) G_i$$

Total losses in each clamping diode= Conduction losses+ Switching losses.

C. Selection of heat sink

Let there be a single common heat sink for rectifier diode modules (2 no's) and IGBT module (3 no's).

Forced cooling with a fan is desired to reduce the size of heat sink.

Total losses of all devices on heat sink =2*Total IGBT Losses+3*Losses per module

Maximum allowable junction temperature of all IGBTs and all diodes as per data sheet = 150° C

To be safe, let us limit the junction temperature to 125° C

Maximum ambient temperature = 50° C

Rectifier module junction to sink drop = 0.28° C .

Hence allowable heat sink temperature =125-0.28x125 =120-35≈90° C

IGBT junction to sink drop = 0.6° C / W .

Maximum losses occur in inner IGBTs

Hence allowable heat sink temperature=125-0.6x43 =125-25.8≈99° C

Clamping diode junction to sink drop = 0.8° C / W .

Hence allowable heat sink temperature=125-0.8x24≈ 105° C

Thus the allowable common sink temperature =the lowest of all the above = 90° C

Maximum allowable thermal resistance of heat sink =(90-50)/780 =0.051° C / W

Select a heat sink with thermal resistance of 0.045° C / W or less to be safe.

Heat sink with fabricated fins is preferred for its efficacy. Mounting area chosen is 125mm x 300mm.This allows a 125 sq fan.

R-Tools 3D Heatsink Thermal Modeling

R-Tools Simulation - Design Output Summary

User: DVRK Rao (dvrk@ecil.co.in)
Project: AC drive-mtech
Simulation Date: 2018-06-06 01:31:14

Thermal Design Details

Heatsink type: Swaged fin
Part number: Fablin: DF300T18A17AS118A
Weight: 5.73204 kg
Heatsink dimensions: 125.0 mm wide x 300.0 mm long x 136.3 mm high
Material: Aluminum
Finish: A

Environment

Ambient Temperature: 50.0 C
Altitude: 300.0 m

Thermal Design Details

Source Names	%sc	Power	Temperature	
			Tsink-avg C	Tsink-max C
source2	3.2%	120.0 W	86.8	88.0
source3	7.3%	180.0 W	86.3	87.8
source4	7.3%	180.0 W	86.8	88.3
source5	7.3%	180.0 W	86.8	88.2
source1	3.2%	120.0 W	85.8	87.3

Fig. 2. Base plate Temperature Profile.

Fig. 3. Base plate Temperature Profile.

Fig. 4. Temperature distributions on heat sink.

As heat is generated heat sinks are required to keep the junction temperature within limits. It is planned to use 1 air cooled heat sink for all the semiconductor devices namely three IGBT module and two bridge rectifier modules. As we observed the temperature ranges the highest maximum temperature on heat sink is 88.3°C and power is 180W at 7.3 % our allowable temperature is 90°C as shown in Fig.4.

III. SPEED CONTROL METHODS

There are various possible ways of controlling speed of an Induction Motor. They are:

- Pole Changing
- Variable Supply Voltage Control
- Variable Rotor Resistance Control
- Slip Recovery
- Variable Supply Frequency Control

(i) Scalar Control

(ii) Vector Control

First three methods are almost obsolete. Method be is used for small motors like in domestic fans.

- Slip recovery is used in large drives where the speed control range can be low. This method requires smaller converter as only the slip power has to be controlled.
- Variable supply frequency is the most popular method and in this method both magnitude and frequency of supply voltage to motor are simultaneously controlled using Pulse Width Modulation (PWM).
- There are essentially two schemes viz., scalar control and vector control. A variation of vector control is Direct Torque Control (DTC).

(i) Scalar control

Scalar control, as the name indicates, is due to magnitude variation of the control variables only. For example, the voltage of the machine can be controlled to control the flux, and the frequency and slip can be controlled to control the torque. However, flux and torque are also functions of frequency and voltage, respectively. Scalar control in contrast to the vector control or field oriented control, where both the magnitude and phase are controlled. Scalar controlled drives give somewhat inferior performance, but they are easily implemented.

Scalar controlled drives have been widely used in industry. However, their importance has diminished recently because of the superior performance of vector controlled drives, which is demanded in many applications.

(ii) Vector control

Scalar control is simple to implement, but the inherent coupling effect i.e. both torque and flux are functions of voltage or current and frequency gives the sluggish response and the system is prone to instability when rapid changes are made.

These problems are solved in Vector Control also called (Field Oriented Control). The invention of vector control in the beginning of 1970 as made it possible to control an induction motor like a separately excited DC motor and brought a revolution in the high performance of ac drives.

Indirect Field Oriented Control

The rotor flux angle θ_f for field orientation is obtained from the measured rotor speed and calculated slip angle based on motor parameters. As shown in Fig.5.

Fig. 5. Indirect field-oriented controls with rotor flux orientation.

The rotor speed w_r is directly measured, the rotor flux angle θ_f can be found from

$$\theta_f = \int (w_r + w_{sl}) dt \quad (10)$$

where w_{sl} is the angular slip frequency.

The slip frequency w_{sl} can be derived from the synchronous frame motor model of Fig.5 from which

$$p\lambda_r = -R_r \bar{i}_r - jw_{sl} \bar{\lambda}_r \quad (11)$$

Substituting the rotor current

$$\bar{i}_r = \frac{1}{\lambda_r} (\bar{\lambda}_r - L_m \bar{i}_s) \quad (12)$$

from (11) yields

$$p \bar{\lambda}_r = -\frac{R_r}{L_r} (\bar{\lambda}_r - L_m \bar{i}_s) - j\omega_{sl} \bar{\lambda}_r \quad (13)$$

from which

$$\bar{\lambda}_r (1 + T_r (p + j\omega_{sl})) = L_m \bar{i}_s \quad (14)$$

where T_r is the rotor time constant

$$T_r = \frac{L_r}{R_r} \quad (15)$$

Decomposing (15) into the dq-axis components and taking into account the rotor flux orientation ($\lambda_{qr} = 0$ and $\lambda_{dr} = \lambda_r$) we have

$$\begin{aligned} \lambda_r (1 + pT_r) &= L_m i_{ds} \\ \omega_{sl} T_r \lambda_r &= L_m i_{qs} \end{aligned} \quad (16)$$

from which

$$i_{qs} = \frac{L_m}{T_r \lambda_r} i_{qs} \quad (17)$$

As shown in Fig, the rotor flux and torque are controlled by two feedback loops separately. The relationship between the rotor flux reference λ_r^* and d-axis current reference i_{ds}^* can be expressed as

$$i_{ds}^* = \frac{(1 + pT_r)}{L_m} \lambda_r^* \quad (18)$$

Since λ_r^* is normally kept constant during operation ($p\lambda_r^* = 0$), can be simplified to

$$i_{ds}^* = \frac{1}{L_m} \lambda_r^* \quad (19)$$

The q-axis current reference i_{qs}^* can be obtained from the torque equation of

$$i_{qs}^* = \frac{1}{K_T \lambda_r^*} T_e^* \quad (20)$$

For a given λ_r^* , the torque producing current i_{qs}^* is proportional to T_e^* .

IV. RESULTS AND ANALYSIS

Induction motor ratings:

Output power: 10 kW

Rated frequency: 50Hz

Supply voltage: 230Vrms L-L

No of poles: 4

Slip at rated torque: 4%

Efficiency at rated speed and load: 0.76

Power factor =0.91

A. Calculations of Converter Circuit Parameters

Rated output current = output power/ efficiency/ $\sqrt{3}$ / output line voltage/pf

Max current from converter when input voltage at min=36/0.9=40Arms.

Nominal DC link Voltage per rectifier bridge ($V_{dc}/2$) = $\sqrt{2} \times 230 = 162.1V$

Max. DC link Voltage per rectifier bridge ($V_{dc}/2$) = $1.1 \times 162.1 = 179V$

Considering 110% input voltage

Nominal $V_{dc} \approx 2 \times 162 = 324V$

Max $V_{dc} = 2 \times 179 = 358V$

Min $V_{dc} = 0.9 \times 324 \approx 292V$

Maximum DC link current = DC link current when motor is loaded fully and DC link voltage is minimum

$\approx$ Rated output power/ inv efficiency/Min DC link voltage
 $\approx 10000/0.76/292A$
 $\approx 45A$

Nominal DC link current = $10000/0.76/324 = 40.3A$

Transformer

Secondary nominal voltage = $179/1.35 \approx 133V$ rms

Secondary nominal current = max DC current/ ($\sqrt{1.5}$)
 $= 45/0.817 = 36.8A$ rms.

(i) Rectifier diodes

Max. Forward average current in each rectifier diode = DC link Current/3
 $= 45/3 = 15A$

Peak reverse voltage across each rectifier diode = $\sqrt{2} \times$ max rms Voltage at secondary
 $= \sqrt{2} \times 1.1 \times 133 \approx 207V$

Maximum ripple current in DC link capacitor as a fraction of load current occurs when $m \approx 0.6$ and it is about 65% of load current. In case of pump loads, the load current is almost proportional to square of voltage and hence the ripple current in capacitor is always less than the current at 100% voltage at output.

Maximum ripple current in DC link capacitor at nominal input voltage

(ii) Inner IGBTs

Max peak current $\approx$ Max peak current in load = $1.414 \times 40 \approx 57A$,

Max average current in each IGBT = Peak load current $\times (1 + \cos \phi) / \pi / 2$

$$=57*1.76/3.14/2 \approx 16.5A$$

(iii) Outer IGBTs

Max average current in each IGBT = Maximum average current in Inner IGBT*2/ π

$$=16.5*2/3.14 \approx 10.6A$$

(iv)Clamping Diodes

Max current values are close to those of inner IGBTs at very low values of modulation index

Max peak current $\leq 57A$

Max average current $\leq 16.5 A$

Max voltage across IGBTs= Max $V_{dc}/2= 179V$

Max voltage across clamping diodes= Max $V_{dc}/2 = 179V$

During regenerative loading, currents in IGBTs are equal to currents in the anti-parallel diodes calculated above. Similarly, anti-parallel diode currents will be like the IGBT currents calculated above.

Thus considering our test schemes, the ratings of anti-parallel diodes should be same as the IGBT currents.

Loss calculations are done in motoring, regeneration and zero pf modes to get the worst case junction temperatures.

B. Simulation of Drive System

Fig.6shows the Simulation of drive system. Block A is the rectifier for three level NPC inverter. Mains voltage is given to a 3 phase transformer with two secondaries (in star and delta configurations). The secondary voltages which are equal in magnitude but have a phase difference of 30 degrees are rectified by two uncontrolled diode bridges. Electrolytic capacitors filter the DC voltages. The equal DC voltages thus produced are connected in series. The junction point is the neutral point.

Block B is the three-level inverter with 12 IGBTs and 6 clamping diodes. AC output is connected to a three phase squirrel cage induction motor IM. Block C is the load to motor. For simulating sudden application of mechanical

load, the software (PSIM) uses a block that converts electrical load (current) to mechanical torque. The electrical load can be easily manipulated by the software. In this case, the electrical load (current source) is switched into circuit 500ms after the inverter is turned ON. Block D shows the PWM pulse generation for turning IGBTs ON and OFF. A 2 kHz triangular voltage source with unity magnitude is the carrier wave generator. A DC voltage of unity magnitude biases the triangular wave and produces another carrier. Thus we have two carriers. Reference voltages produced in block E are compared with the carrier voltages by 6 comparators and 6 pulses required for the top six IGBTs are generated. Complements of these pulses are used for the bottom IGBTs. Block E is the FOC block. Speed I_d and loops are used to control speed and flux of the motor. References for I_d and speed are provided by DC sources. A speed sensor coupled to motor provides speed feedback in RPM. PI controllers are used in speed and I_d loops. PSIM provides abc/dq and dq/abc transformations by function blocks. Slip obtained from I_d and I_q is added to rotor speed to obtain rotor flux speed. It is integrated to obtain rotor flux angle. Reference voltages generated in this block are sent to block D for PWM pulse generation.

C. Simulation Results

Top plot shows how the motor responds to step input of speed. It may be observed from the figure6 that the Motor reaches 800rpm in less than 40ms at no load. There is a small overshoot and fine tuning of PI controller in speed loop is expected to reduce it. 10Nm load is suddenly applied by a load switch at 0.5 sec as shown in the middle plot. It may be observed from the figure Speed does not fall and the disturbance at switching of load is very little.

Fig.6.Simulation of drive system.

Fig. 7. Top: speed, Middle: load switch and Bottom: output line current.

Bottom plot shows motor current. It may be observed from the figure that is substantially sinusoidal with very little harmonics at switching frequency.

Fig.8. Red trace Id and Blue trace Iq.

I_d (Current producing magnetic field) and I_q (current producing torque) as shown in Fig. 8. As may be observed from the figure, after initial transients lasting about 50 ms, currents stabilize with a little ripple. I_q Responds quickly at load switching and settles down at higher value to meet higher demand of torque.

Fig.9. Output line voltage.

Output Line voltage is shown in the Fig.9. It is observed that the Output voltage wave form changes when load is switched in. At no load less voltage is required as shown in the plot. When load is switched in, the wave form develops into the form of full 3-level operation. Fundamental voltage is extracted from the wave form by obtaining the average of absolute voltage and multiplying it with form factor (1.1)

Fig. 10. Output phase voltage.

Output phase voltage is shown in the Fig.10 as we observed from the figure Phase voltage is obtained from a

star connected dummy load. RMS voltage of fundamental is extracted from the wave form by obtaining the average of absolute voltage and multiplying it with form factor (1.1)

Fig. 11. Top outer IGBT and Bottom: Top inner IGBT.

Gate pulses to outer IGBT (top) and inner IGBT (top) are shown in the Fig.11 It may be observed that during positive half cycle, inner IGBT is on and the outer IGBT is turned on and off for voltage modulation. During negative half cycle, top IGBT is off and inner IGBT is turned on and off for connection to neutral point (0V level).

Fig. 12.DC link voltage.

DC link voltage is shown in Fig. 12. As we observed from figure 7 V_{dc} (Sum of the filtered rectifier outputs) shows that (a) very little ripple voltage and that phase shift (30 deg) in the input transformer windings cancels the 300Hz ripple voltages substantially and (b) the capacitance in the filter is adequate to reduce switching frequency ripple.

Fig. 13.DC link currents.

DC link current is shown in the Fig. 13. As we observed from figure 8 DC link currents appear with expected envelope (similar to 6 pulse controlled rectifier output voltage) and 2 kHz switching. AC current in capacitors = $\sqrt{I_{rms}^2 - I_{dc}^2} = 8.5 \text{ Arms}$.

Fig.14. Top: Delta Current, Middle: star current and Bottom: input current.

Delta current, star current and input current are shown in the Fig. 14. As we observed from Fig.10 Currents in the transformer secondary's (rectifier input) show the expected waveforms with low leakage inductance of transformer (predominantly capacitive filter). They are equal in magnitude and are out of phase. Thus the 5th and 7th harmonics of 6 pulse rectifiers are almost absent in the current of transformer primary (bottom plot).30-degree phase angle between star and delta outputs of transformer help to get 12- pulse rectification.

Fig.15. Frequency spectrum of input current.

Frequency spectrum of input current as shown in Fig.15.As we observed from figure 10 Frequency spectrum shows that harmonics of only $12n \pm 1$ order are present in the current.

V. CONCLUSION

3-Level NPC inverter with minimum components and stepped output voltage is a good candidate topology for our study. Its operation and PWM schemes are well documented in literature. Carrier based modulation is simple to implement. Third harmonic injection results in increased output like in SVM. Literature covers problem of NP fluctuations and solutions. If a transformer with two secondaries (star and delta) is used, NP fluctuation problem is avoided and input current harmonics are reduced. In scalar control, where dynamic performance is poor, simply controlling frequency while keeping v/f ratio constant, is enough. However, if good dynamic performance and performance at very low speeds close to standstill condition are required, a dynamic model of the motor needs to be worked out and a complex control scheme to maintain proper angle between stator and rotor fluxes at all times has to be made.

Field Oriented Control (FOC) works well for steady state and dynamic conditions. Scheme involves many

mathematical operations to compute proper instantaneous stator voltages. To reduce complexity in calculations, first, three phase ac voltages and currents are reduced to two phase ac values and vice versa by CLARK transformations. Then, to convert ac values to DC values, the reference frame is changed from stationary to rotating frame and vice versa by PARK transformations. Rotor flux oriented scheme is very popular. Power circuit for a 10kW motor has been designed and components have been selected.

Since the current and voltage rating of required devices are low it is possible to have six diodes of a bridge rectifier in one module with isolated base. Similarly, 4 IGBTs and 2 clamping diodes of one inverter phase leg are available in one module with isolated base. Semikron modules SK70D 08 and SKiiP 27MLI07E3V1 have been selected for diode bridges and IGBTs with clamping diodes, respectively. Sufficient margins in voltage and current ratings are available in the modules and the converter becomes compact as many devices are integrated in modules. A single air cooled heat sink is adequate for all devices.

Losses in all semiconductor devices have been calculated and one force air cooled heat sink is found to be enough to dissipate heat generated from all the 5 five modules. Heat sink thermal resistance for good margins in junction temperature is estimated and a standard heat sink with bonded fins has been chosen. Thermal simulation has been done using free software (R-Tools) from Mersen and the temperatures have been found to be within limits.

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