Design and implementation of A Sliding Mode Controller Technique for DC to DC Boost Converter fed Induction motor drive

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Abstract—in this paper, a PI and sliding mode controller for the control of dc-dc boost converter fed induction motor drive application is designed and analyzed. Dynamic equations describing the boost converter are derived and sliding mode controller is designed. A two-loop control is employed for a boost converter. The sliding mode controlled buck converter system is tested for step load changes and input voltage variations. The theoretical predictions are validated by means of simulations. The performance of conventional PI controller and non-singular terminal sliding mode controller for current control is tested for reference disturbance, line disturbance and load disturbance conditions. The simulation results are presented. The boost converter is tested with operating point changes and parameter uncertainties. Fast dynamic response of the output voltage and load and input voltage variations are obtained.The traditional IM drive which integrates phase controlled boost converter in the facade end upshot tribulations like DC link fluctuations and deprived DC link voltage level. To overcome the problem, the conventional DC-DC converter is replaced with Proportional Integral (PI) and sliding mode controlled in the front end of IM drive that produces the DC link voltage in geometric progression. The Voltage Source Inverter (VSI) of the suggested system renders both open loop and closed loop control scheme for IM by feedback regulated Pulse Width Modulation (PWM) technique.

Keywords- Boost converter, Cascade Control, Sliding Mode Controller, Non-singular Terminal Sliding Mode Controller.

I. INTRODUCTION

Electronic power converters are used as an actuator for electromechanical systems. The boost type dc-dc converters are used in applications where the required output voltage is lower than the source voltage [1]. Different control algorithms are applied to regulate dc-dc converters to achieve an output voltage. As dc-dc converters are nonlinear and time variant systems, the application of linear control techniques for the control of these converters are not suitable. In order to design a linear control system using classical linear control techniques, the small signal model is derived by linearization around a precise operating point from the state space average model [2-5]. The controllers based on these techniques are simple to implement however, it is difficult to account the variation of system parameters, because of the dependence of small signal model parameters on the converter operating point [6-8]. Variations of system parameters and large signal transients such as those produced in the startupor against changes in the load, cannot be dealt with these techniques. Multi loop control techniques, such as current mode control, have greatly improved the dynamic behavior, but the control design remains difficult especially for higher order converter topologies [9].

A control technique suitable for dc-dc converters must cope with their intrinsic nonlinearity and wide input voltage and load variations, ensuring stability in any operating condition while providing fast transient response. Since switching converters constitute a case of variable structure systems, the sliding mode control technique can be a possible option to control this kind of circuits [10]. The use of sliding mode control enables to improve and even overcome the deficiency of the control method based on small signal models. In particular, sliding mode control improves the dynamic behavior of the system, and becomes very useful when the system is required to operate in the presence of significant unknown disturbances and plant uncertainties [11].

Induction Motors are the Workhorse of the Industry due to its economy of procurement, installation and use. Recent development in the field of Power Electronics planted a wide usage of Induction Motor (IM) in the adjustable speed drives where speed control of the motor is highly required and squirrel cage type of IM is very popular in that case. The basic control involved in variable speed control of IM is application of a variable frequency and variable magnitude of AC voltage to the motor for the attainment of variable speed operation.

Many techniques are already introduced to control the IM parameters [12]. However, the method by name constant Voltage/Hertz (V/f) is versatile in use. Further, the V/f technique is classified into open loop and closed loop control. The open loop V/f control of IM presented in [13-14] use the arrangement of Voltage Source Inverter (VSI) fed IM with source input as DC which fails to discuss the influence of DC linkfluctuations in the IM drive due to load disturbance and also the results obtained are ideal. The closed loop V/f control overcomes the problem with the load disturbance that prevails in open loop V/f control of IM.

II. MODELING THE DC TO DC BOOST CONVERTER

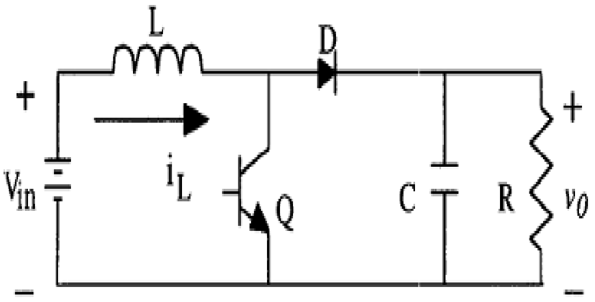

Figure.1 DC to DC boost converter.

Fig.1 shows a DC–DC boost converter. It consists of a DC input voltage source (v_{in}), a controlled switch (S_w), a diode (D), a filter inductor (L), a filter capacitor (C), and a load resistor (R). The equations describing the operation of the converter can be written for the switching conditions ON and OFF, respectively as

$$\frac{di_L}{dt} = \frac{1}{L}(V_{in}) \tag{1}$$

$$\frac{dv_o}{dt} = \frac{1}{C}(-\frac{v_o}{R}) \tag{2}$$

And

$$\frac{di_L}{dt} = \frac{1}{L}(V_{in} - v_o) \tag{3}$$

$$\frac{dv_o}{dt} = \frac{1}{C}(1 - \frac{v_o}{R}) \tag{4}$$

Combining (1)-(4) gives

$$\frac{di_L}{dt} = \frac{1}{L}(V_{in} - uv_o) \tag{5}$$

$$\frac{dv_o}{dt} = \frac{1}{C}(u - \frac{v_o}{R}) \tag{6}$$

Where ‘u’ is the control input which takes 1 for the ON state of the switch and 0 for the OFF state.

III. SLIDING MODE CONTROL METHOD FOR BOOST CONVERTER

A. Need of Cascade Control for DC to DC Boost Converter

The topological modification of locating the inductor before the switching element rather than after it as for the Buck converter enables a higher output voltage than the source voltage. However, from a control point of view, the Boost converter is more difficult for controlling the output voltage. This lies in the fact that the control ‘u’ appears in both the current and voltage equations, as mentioned in equations (5) and (6), and in bilinear way. Such a configuration implies a highly nonlinear system with difficult control design.

In contrast to the Buck converter the equation for the output voltage of the Boost converter depends on control ‘u’ directly. Selecting the control in the form

$$u = \frac{1}{2}(1 + sign(s))$$
$$v_o - v_o^* = 0 \tag{7}$$

The sliding mode can be enforced in s = 0 with the output voltage equal to the reference input v_0^*, if values s and $\dot{s}$ have opposite signs (v_0^* is assumed to be constant). After the sliding mode starts, the equation for the current i_L can be obtained by using the equivalent control method.

$$[1 - u]_{eq} = \left(\frac{v_o^*}{R}\right)(1/i_L) \tag{8}$$

Equation (8) is found from equation $\dot{s} = 0$ and substituted into the equation (5)

$$\frac{di_L}{dt} = -\frac{v_o^{*2}}{RL}\frac{1}{i_L} + \frac{V_{in}}{L} \tag{9}$$

It is evident that the equilibrium point $i_L^* = \frac{v_0^{*2}}{V_{in}R} > 0$ is unstable, and so the voltage control method is not acceptable, though it keeps the output voltage equal to the desired value. It means that the zero dynamics with respect to the output voltage are unstable. Therefore, the voltage tracking problem for a constant reference input should be reformulated in terms of the current one.

So the voltage controller is used as a PI controller, the output of which is treated as the current reference and further another controller is used for current regulation which is the proposed Sliding Mode Controller.

B. Design of Sliding Mode Current Controller

As mentioned above, the output of PI voltage controller is treated as I_{ref} and actual load current is I_0.

The state variables considered in SMC design are the current error and the rate of change of the error. i.e.

$$x_1 = I_o - I_{ref} \tag{10}$$

$$x_2 = \dot{x}_1 = \dot{I}_0 - \dot{I}_{ref} = \dot{I}_0 \tag{11}$$

A linear sliding surface function can be expressed as

$$s = \lambda x_1 + x_2, \qquad \lambda > 0 \tag{12}$$

Where λ is a real sliding co-efficient. The dynamic behavior of (3.12) without external disturbance on the sliding surface is

$$s = \lambda x_1 + \dot{x}_1 = 0 \tag{13}$$

In the phase-plane, S= 0 represents a line, called sliding line, passing through the origin with a slope equal to λ. The sliding mode (S = 0) is described by the following first-order equation

$$\dot{x}_1 = -\lambda x_1 \tag{14}$$

During the sliding mode, the output voltage error is expressed as

$$x_1(t) = x_1(0)e^{-\lambda t} \tag{15}$$

It should be noted that λ must be a positive real constant for achieving system stability. In general, the SMC exhibits two modes: the reaching mode and the sliding mode. While in the reaching mode, a reaching control law is applied to drive the system states to the sliding line rapidly. When the system states are on the sliding line, the system is said to be in the sliding mode in which an equivalent control law is applied to drive the system states, along the sliding line, to the origin. The equation of the control input can be written as

$$u = -\frac{1}{2}[sgn(s) - 1] \tag{16}$$

However, direct implementation of this control input causes the converter to operate at an uncontrollable infinite switching frequency which is not desired in practice. Hence, the following hysteresis modulation (HM) method employing a hysteresis band with the boundary layer is used to solve this very high frequency operation

$$u = \begin{cases} 1 \text{ when } s < -h \\ 0 \quad \text{when } s > h \end{cases} \tag{17}$$

In order to ensure that the movement of the error variables is maintained on the sliding line, the following existence condition must be satisfied

$$s\dot{s} < 0 \tag{18}$$

The time derivative of (12) can be written as

$$\dot{s} = \lambda \dot{x}_1 + \dot{x}_2 \tag{19}$$

Choice of value of λ is discussed thoroughly in [1] and [2]. To summarize the discussion, it can be concluded that in order to ensure that λ is large enough for fast dynamic response and low enough to retain a large existence region; it has been proposed in to set λ as follows:

$$\lambda = \frac{1}{RC} \tag{20}$$

However, despite the dynamic response can be made faster by utilizing a large λ in the sliding surface function, the system states still cannot converge to the equilibrium point in finite time.

C. Terminal Sliding Mode Controller

A non-linear sliding surface function for the buck converter system defined in (9) and (10) can be defined as

$$s_t = \lambda {x_1}^{\gamma} + \dot{x}_1 \tag{21}$$

Where

$$\lambda > 0$$

And

$$0 < \left(\gamma = \left(\frac{q}{p}\right)\right) < 1$$

Where p and q are positive odd integers satisfying p > q. When the system is in the terminal sliding mode (St = 0), its dynamics can be determined by the following non-linear differential equation

$$\dot{x}_1 = -\lambda {x_1}^{\gamma} \tag{22}$$

Note that Equation (22) reduces to$\dot{x}_1 = -\lambda x_1$ for$\gamma = 1$, which is the form of conventional SMC. It has been shown that $x_1 = 0$ is the terminal attractor of the system defined in (22). The term ‘‘terminal’’ is referred to the equilibrium which can be reached in finite time.

D. Non-Singular Terminal Sliding Mode Controller

A non-linear sliding surface function for the boost converter system defined in (9) and (10) can be defined as

$$s_n = x_1 + \lambda' {\dot{x}_1}^{1/\gamma} \tag{23}$$

Where $\lambda > 0,\ 0 < \left(\gamma = \left(\frac{q}{p}\right)\right) < 1$ and $\lambda' = \left(1/\lambda\right)^{(1/\gamma)}$

When the system is in the terminal sliding mode ($s_n = 0$), its dynamics can be determined by the following non-linear differential equation

$$-{\dot{x}_1}^{1/\gamma} = \frac{1}{\lambda'} x_1 \tag{24}$$

In order to ensure that the movement of the error variables is maintained on the sliding line, the following existence condition must be satisfied

$$s_n \dot{s}_n < 0 \tag{25}$$

Taking time derivative of (23) and using it in (25) gives

$$s_n \dot{s_n} = \dot{x}_1 + \frac{\lambda'}{\gamma} \dot{x}_1^{(1/\gamma)-1} \dot{x}_2 \tag{26}$$

From equations of $\dot{x}_1$ and $\dot{x}_2$and equation (26), solution for control input 'u' gives

$$u_{eq}^{NTSMC} = \frac{1}{\omega_0^2 V_{in} \lambda'} \left(\frac{\lambda' x_2}{\gamma RC} + \frac{\lambda' \omega_0^2}{\gamma} (V_{ref} + x_1) - {x_2}^{2-(1/\gamma)} \right) \tag{27}$$

In equation (27), it is obvious that as long as 1 < (1/γ) < 2 (i.e., q < p < 2q), the term $x_2^{2-(1/\gamma)}$ is always non-singular. Therefore, the singularity problem will not occur in the NTSMC.

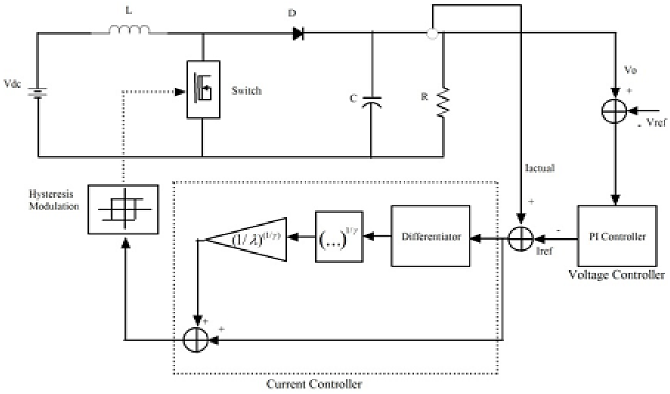

Figure.2.Schematic block diagram of cascade control for DC to DC boost converter.

IV.INDUCTION MOTOR

Induction Motor (1M) An induction motor is an example of asynchronous AC machine, which consists of a stator and a rotor. This motor is widely used because of its strong features and reasonable cost. A sinusoidal voltage is applied to the stator, in the induction motor, which results in an induced electromagnetic field. A current in the rotor is induced due to this field, which creates another field that tries to align with the stator field, causing the rotor to spin. A slip is created between these fields, when a load is applied to the motor.

Compared to the synchronous speed, the rotor speed decreases, at higher slip values. The frequency of the stator voltage controls the synchronous speed. The frequency of the voltage is applied to the stator through power electronic devices, which allows the control of the speed of the motor. The research is using techniques, which implement a constant voltage to frequency ratio. Finally, the torque begins to fall when the motor reaches the synchronous speed. Thus, induction motor synchronous speed is defined by following equation,

$$n_s = \frac{120f}{p}$$

Where f is the frequency of AC supply, n, is the speed of rotor; p is the number of poles per phase of the motor. By varying the frequency of control circuit through AC supply, the rotor speed will change.

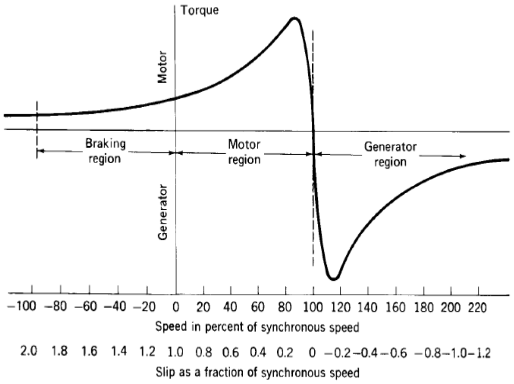

Fig.3.Speed torque characteristics of induction motor.

V.THE THREE-LEVEL NPC VSI

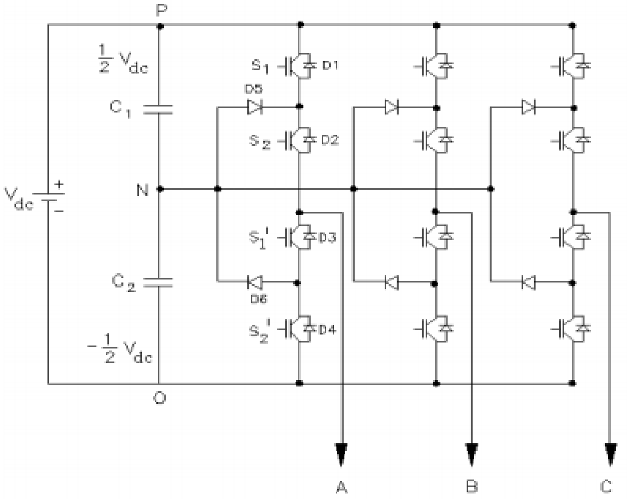

Fig.4. Diode Clamped Three-Level Inverter.

The diode-clamped inverter [5] was also called the neutral-point clamped (NPC) inverter because when it was first used in a three-level inverter the mid-voltage level was defined as the neutral point level.

A three-level diode-clamped inverter is shown in Fig. 4. In this circuit, the dc-bus voltage is split into three levels by two series-connected bulk capacitors, C1 and C2. The middle point of the two capacitors N can be defined as the neutral point. The output voltage VAN has three states: Vdc/2, 0, and -Vdc/2. For voltage level Vdc/2, switches S1 and S2 need to be turned on; for -Vdc/2, switches S'1 And '2s need to be turned on; and for the 0 level, 2s and S'1 need to be turned on. The two diodes D5 and D6 clamp the switch voltage to half the level of the dc-bus voltage. When both 1s and 2s turn on, the voltage across A and O is Vdc, i.e., VAO = Vdc. In this case, D6 balances out the voltage sharing between S'1 and S'2 with S'1 blocking the voltage across C1 and S'2 blocking the voltage across C2.

Some of the important features of diode clamped inverter are given below: Low voltage power semiconductor devices: The m-level diode clamped inverter requires (m+1) active devices (GTO and IGBT's etc) per phase and each active device will see a blocking voltage of (Vdc / (m-1)).

Duty cycle of switching devices: The duty cycle of the power switches is different. So switches of different current rating have to be used for optimal design.

(a) Rating for clamping diodes: For five and higher level inverters, the voltage blocking capability of the diodes are different. So the diodes will have different voltage ratings. Assuming that the characteristics of diodes are identical, then multiple diodes of same voltage rating have to be used to achieve required voltage-blocking capacity. Hence, for a sufficiently large number of levels, the number of diodes required will become too large and will make the circuit less reliable. Also power circuit layout and packaging becomes difficult.

(b) Capacitor voltage unbalance: The midpoint voltage is derived using capacitors and these carry load current.

Unequal loading of the capacitors leads to imbalance in the dc bus capacitor voltages and this will cause the dc mid pint voltage to drift. This is not a serious problem for utility applications such as, static VAR generators (SVG), active power filters, etc., where the inverters need to supply only the reactive power.

(c) High voltage surge: During turn off, the devices will experience a high transient over voltage and also snubbers are required to distribute the voltage across clamping diodes in a uniform fashion. The design of snubbers is complicated, as the current through these snubbers is bi-directional.

When the number of output levels is sufficiently high, the inverter system required a huge number of clamping devices due to the series connection of clamping diodes and capacitors. These not only increase the cost of the system, but also controlling the inverter output and capacitor voltage balance becomes more complex when the number of levels is higher than five. Thus diode clamped inverters are usually limited to three or (maximum).

VI.MATLAB/SIMULATION RESULTS

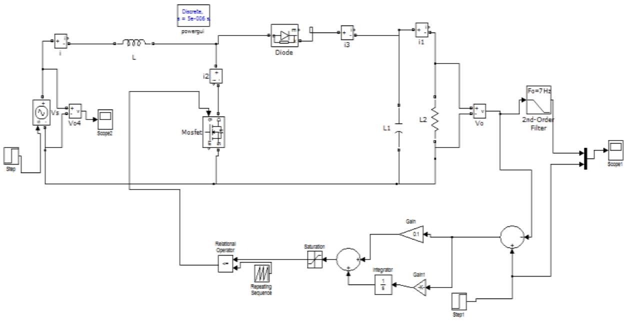

Fig.5.Matlab/simulation model of PI control for DC to DC boost converter.

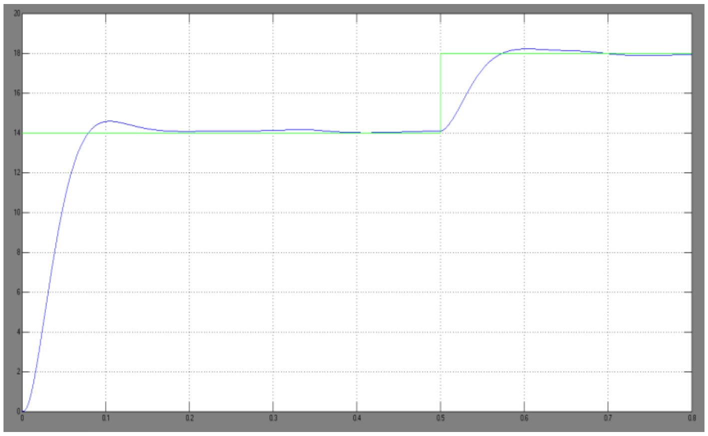

Figure.6. PI Current Controller for DC to DC boost converter under reference disturbance.

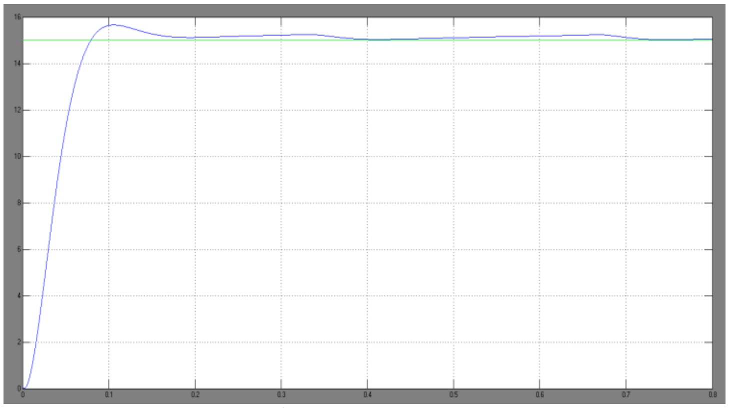

Figure.7. PI Current Controller for DC to DC boost converter under line disturbance.

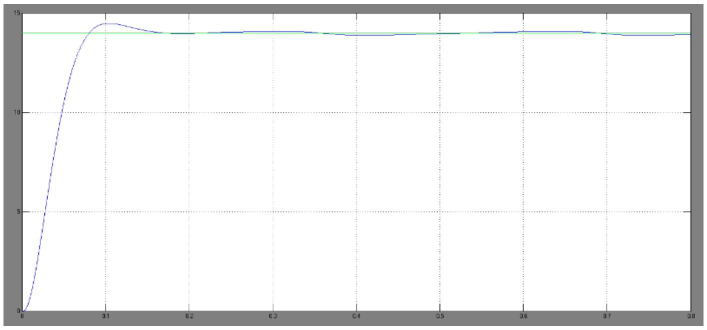

Figure.8. PI Current Controller for DC to DC boost converter under load disturbance.

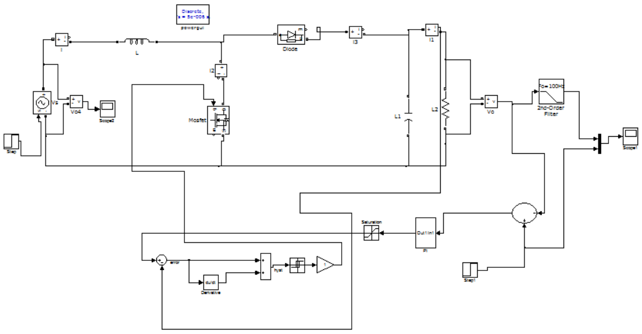

Fig.9.Matlab/simulation model of PI and sliding mode control for DC to DC boost converter.

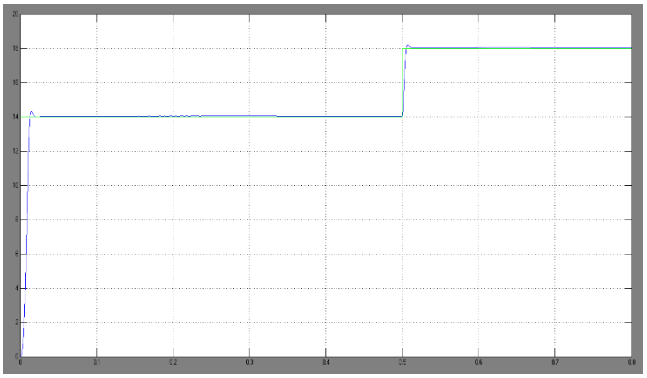

Figure.10.Sliding Mode Current Controller for DC to DC boost converter under reference disturbance.

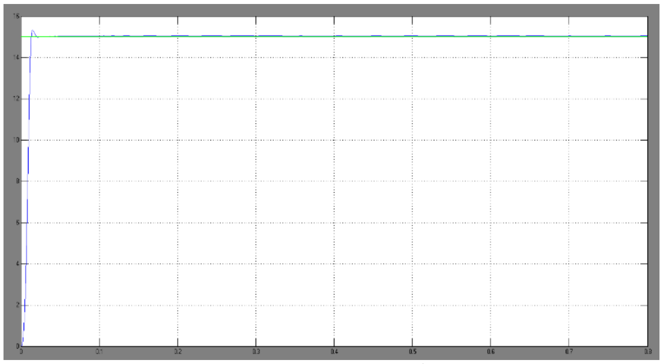

Figure.11.Sliding Mode Current Controller for DC to DC boost converter under line disturbance.

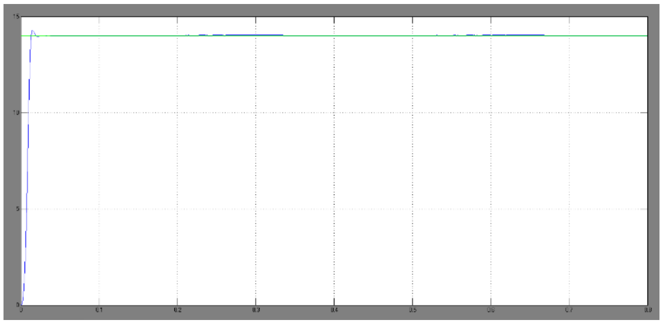

Figure.12.Sliding Mode Current Controller for DC to DC boost converter under load disturbance.

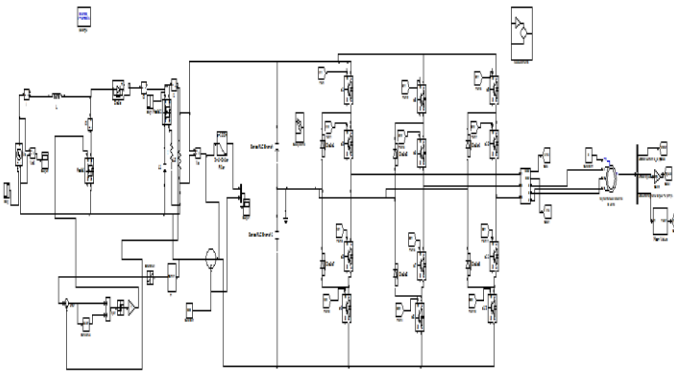

Fig.13.Matlab/simulation model of PI and sliding mode control for DC to DC boost converter fed Induction motor drive applications.

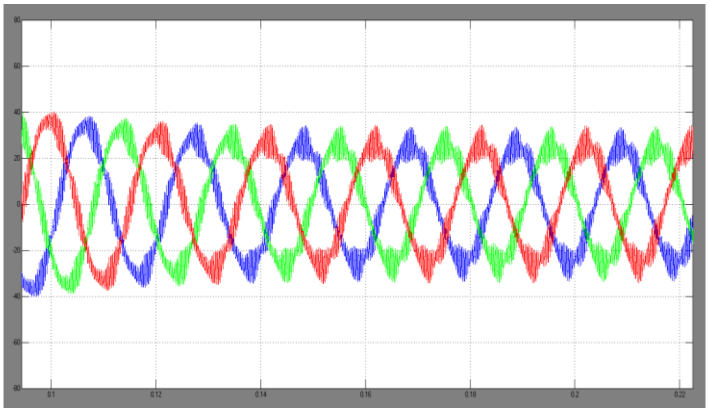

Fig.14.inverter line to line voltage.

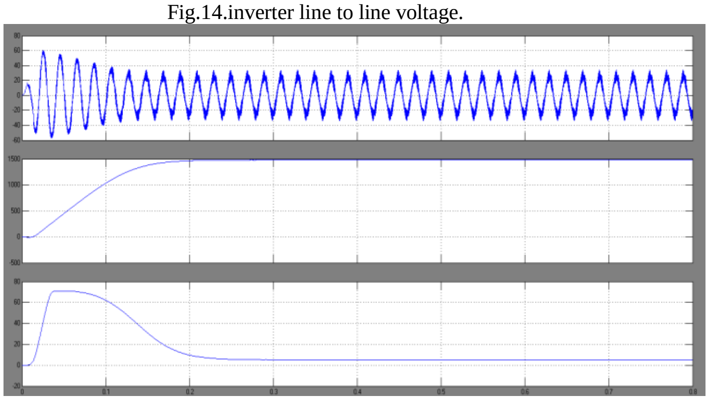

Fig.15.Stator Current, speed and torque for induction motor drive.

VII.CONCLUSION

A dc-dc boost converter with sliding mode control is simulated in this study. It allows evaluation of closed loop performances like tracking the desired output voltage. The simulation results show the validity of the sliding mode controlled boost converter model and the PIof this control technique against changes in the load or variations in the input voltage. Therefore the system achieves a sliding output voltage against load disturbances and input voltage variations to guarantee the output voltage to feed the load without instability.Moreover the proposed SMC technique facilitates easy implementation owing to the simple mathematical model. Also, the system stability is ensured under all parametric disturbance conditions. Nevertheless, the use of PI controller as voltage controller makes the overall system response sluggish. This problem can be solved by adopting a novel way rather than the use of PI voltage controller to reformulate the voltage reference problem in boost converter. The scheme suggested utilize minimum rectified DC voltage for driving a three phase IM and therefore, the design using renewable energy as a source by eliminating the single phase AC supply cascading diode rectifier is possible. Basis of future work has been left for the alteration and implementations of experimental drives under involvement of some other control techniques.

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