Robust Control of a Feedback Linearized Induction Motor through Sliding Mode

K. B. Mohanty, *Member, IEEE* and Madhu Singh

Abstract—**This paper presents a feedback linearization control scheme with sliding mode controller (SMC) for enhancing performance of the induction motor drive. The feedback decoupling controller is designed in a stationary (α-β) reference frame with rotor flux and stator current components as state variables. Since the induction motor drive system is sensitive to parameter variation and load disturbances, a robust control strategy based on sliding mode is implemented. Initially feedback linearization technique is simulated with conventional PI controller. Then the proposed scheme comprising of a sliding mode flux controller and a sliding mode speed controller is simulated. Both control schemes are simulated in MATLAB environment. Simulation results demonstrates that performance of sliding mode based scheme is more robust than fixed gain P-I controller based scheme.**

Index Terms-- ***Feedback linearization, Decoupling control, Stationary reference frame, Sensorless flux estimator, P-I controller, Sliding Mode Controller.***

I. INTRODUCTION

Recently the subject of nonlinear control is occupying an increasingly important place in automatic control engineering and has become a necessary part of the fundamental background to control engineering [1]-[2]. Its potential application in the area of induction motor control is emerging as the thrust area for research work. The induction motors are the most preferred for industrial application because of its simplicity, reliability, low cost, ruggedness, and suitability to work in volatile environment. It does not require maintenance and is pollution free. So it is also acceptable in automation industries. But it requires complex control strategy, because it posses three inherent drawbacks as follows. (i) It is a higher order nonlinear dynamic system with internal coupling. (ii) Some state variables like rotor flux and currents, are not directly measurable. (iii) Variation in parameters like rotor resistance due to temperature and magnetizing inductance due to saturation have significant impact on the system dynamics.

Many attempts have been made in past to optimize the performance and simplify the control strategy of the induction motor. Out of these Field Oriented Control or Vector Control proposed by Blaschke [3] and Hasse [4] has emerged successfully to achieve the high performance requirement. As a result it has been aggressively accepted by the automation industries by replacing bulky, costly dc motor drive which has commutation problem.

The vector control methods are complex to implement, because in vector control method, the decoupling relationship is obtained by means of proper selection of state coordinates, under the hypothesis that the rotor flux being kept constant. The torque is only asymptotically decoupled from the flux i.e., decoupling is obtained only in steady state, when the flux amplitude remains constant. Coupling is still present, when flux is weakened in order to operate the motor at higher speed within the input voltage saturation limit or when flux is adjusted in order to maximize power efficiency [5].

This has further led to introduction of nonlinear differential geometric control theory, particularly feedback linearization, which can achieve completely decoupled torque and flux amplitude of the induction motor [9]-[12]. In the past decade, a good number of research works has been reported incorporating various feedback linearization control techniques to simplify and to enhance the performance of the induction motor [7]-[16].

But these techniques need mathematical models and in practice it is difficult to obtain precise nonlinear model. As a result the control performance is still influenced by the uncertainties of the plant. To alleviate the need of for accurate mathematical model, sliding mode technique is introduced [1]. The sliding mode technique is developed from variable structure control theory. Since long time it has been successfully used for induction motor drive [13]-[17].

The purpose of this work is to show how the combination of feedback linearization and sliding mode controller (SMC) provides an effective tool for the robust control of induction motor drive. At first, a feedback linearization approach for induction motor drive is developed. Then the control scheme is implemented with two methods. In one method two P-I controllers are used for controlling speed and secondary rotor flux. Simulation results show the complete decoupling of the rotor flux and rotor speed. However, the control performance is still influenced by the parameter variation and plant uncertainties and remarkable oscillation appears in the torque response and motor speed is affected at the time of load changes. The uncertainties in the plant can well be taken care by sliding mode controller. Therefore, in another scheme sliding mode speed and flux controllers are incorporated, which reduce ripples in the torque and also compensate it. The testing of robustness sliding mode controllers is confirmed by imposing stator resistance and machine inertia variation.

II. SYSTEM DESCRIPTION

The schematic block diagram of the proposed system is shown in Fig 1. The scheme consists of three controllers, one flux estimator, a current controlled PWM voltage source inverter, and an induction motor. Two controllers are

Kanungo Barada Mohanty is with Electrical Engineering Department, National Institute of Technology, Rourkela-769008, India (e-mail: kbmohanty@nitrkl.ac.in)

Madhu Singh is presently pursuing Ph.D. programme at NIT Rourkela and is on QIP study leave from National Institute of Technology, Jamshedpur-831014, India (e-mail: madhu_nitjsr@rediffmail .com)

regulating flux and speed loop. Voltage model [6] is used for flux estimation. Output of flux and speed regulator and also estimated flux are the inputs to the decoupling controller and its output goes to the current controller. Output of current controller is utilized to generate gate drive signal for PWM voltage source inverter (VSI), which forces reference current in the motor to develop required torque. The control scheme is implemented with two methods. In one method two P-I controllers are used for controlling speed and secondary flux. In another scheme two sliding mode controllers for speed and flux are used, which are designed to overcome the problem of uncertainties and parameter variations.

III. MODELING OF INDUCTION MOTOR

The dynamic equations representing induction motor in the α-β stator fixed frame are as

$$\dot{i}_{\alpha s} = -\frac{1}{\sigma L_s}\left(R_s + \frac{L_m^2}{L_r^2}R_r\right)i_{\alpha s} + \frac{1}{\sigma L_s}\frac{L_m R_r}{L_r^2}\psi_{\alpha r} + \frac{pL_m}{\sigma L_s L_r}\omega_r\psi_{\beta r} + \frac{V_{\alpha s}}{\sigma L_s} \quad (1)$$

$$\dot{i}_{\beta s} = -\frac{1}{\sigma L_s}\left(R_s + \frac{L_m^2}{L_r^2}R_r\right)i_{\beta s} + \frac{1}{\sigma L_s}\frac{L_m R_r}{L_r^2}\psi_{\beta r} - \frac{pL_m}{\sigma L_s L_r}\omega_r\psi_{\alpha r} + \frac{V_{\beta s}}{\sigma L_s} \quad (2)$$

$$\dot{\psi}_{\alpha r} = -\frac{R_r}{L_r}\psi_{\alpha r} - p\omega_r\psi_{\beta r} + \frac{L_m R_r}{L_r}i_{\alpha s} \quad (3)$$

$$\dot{\psi}_{\beta r} = -\frac{R_r}{L_r}\psi_{\beta r} + p\omega_r\psi_{\alpha r} + \frac{L_m R_r}{L_r}i_{\beta s} \quad (4)$$

$$\dot{\omega}_r = -\frac{B}{J}\omega_r + \frac{1}{J}(T_e - T_l) \quad (5)$$

where, $\sigma = (1 - \frac{L_m^2}{L_s L_r})$ is the leakage coefficient, ($i_{\alpha s}$, $i_{\beta s}$), ($\psi_{\alpha r}, \psi_{\beta r}$), ($V_{\alpha s}$, $V_{\beta s}$) are respectively the α-β component of the stator current, rotor flux and stator voltage, (R_s, L_s) and (R_r, L_r) are stator and rotor parameters (resistance and inductance), L_m is magnetizing inductance, ω_r is the motor speed, and p is the number of pole pairs.
The electromagnetic torque developed is given by

$$T_e = K_T(\psi_{\alpha r}i_{\beta s} - \psi_{\beta r}i_{\alpha s}),\ K_T = \frac{3pL_m}{2L_r} \quad (6)$$

IV. FEEDBACK LINEARIZATION

Feedback linearization is an approach to nonlinear control design which has attracted great deal of research interest in recent years. The central idea of the approach is to algebraically transform a nonlinear system dynamics into a fully or partially linear one so that linear control technique can be applied. This differs entirely from conventional linearization techniques. Feedback linearization is achieved by exact state transformation. Therefore, it uses a nonlinear transformation on system variables, expressing them in a new suitable coordinate system which enables the introduction of a feedback, so that an input-output or state linearization in new coordinates is achieved. The theoretical foundation and a systematic approach can be found in [1].

In order to control the induction motor in field orientation schemes to get a dc motor like performance, the rotor speed and rotor flux must be decoupled. Therefore, output to be controlled is chosen as

$$Y^T = [\omega_r, \psi_r] \quad (7)$$

where, ω_r is the rotor speed and ψ_r is the rotor flux. The rotor flux calculated as

$$\psi_r^2 = \psi_{\alpha r}^2 + \psi_{\beta r}^2 \quad (8)$$

The time derivative of ψ_r is

$$\dot{\psi}_r = \frac{1}{\psi_r}\left[\psi_{\alpha r}\dot{\psi}_{\alpha r} + \psi_{\beta r}\dot{\psi}_{\beta r}\right] \quad (9)$$

Substituting $\dot{\psi}_{\alpha r}$ and $\dot{\psi}_{\beta r}$ from equation (3) and (4) into equation (9)

$$\dot{\psi}_r = \frac{1}{\psi_r}\left[\psi_{\alpha r}\left(-\frac{R_r}{L_r}\psi_{\alpha r} - p\omega_r\psi_{\beta r} + \frac{L_m R_r}{L_r}i_{\alpha s}\right) + \psi_{\beta r}\left(-\frac{R_r}{L_r}\psi_{\beta r} + p\omega_r\psi_{\alpha r} + \frac{L_m R_r}{L_r}i_{\beta s}\right)\right] \quad (10)$$

Simplifying equation (10) and we have state equation of the rotor flux as

$$\dot{\psi}_r = -\frac{R_r}{L_r}\psi_r + \frac{L_m R_r}{L_r\psi_r}(i_{\alpha s}\psi_{\alpha r} + i_{\beta s}\psi_{\beta r}) \quad (11)$$

From equation (5) and (6) state equation of rotor speed is

$$\dot{\omega}_r = -\frac{B}{J}\omega_r + \frac{1}{J}K_T(\psi_{\alpha r}i_{\beta s} - \psi_{\beta r}i_{\alpha s}) - \frac{T_L}{J} \quad (12)$$

The state equations (11) and (12) describe flux and mechanical system dynamics, which have $i_{\alpha s}$, and $i_{\beta s}$ as two control inputs; ψ_r and ω_r as the two outputs. Thus, it represents a coupled system. Therefore, the nonlinear feedback theory [1] is used to eliminate this coupling relationship between the control inputs $i_{\alpha s}$, $i_{\beta s}$ and the system outputs ψ_r and ω_r.

Let $u1$ and $u2$ be taken as two new control input which converts coupled system into uncoupled one [15]. Eqns. (11) and (12) with new control input can be rewritten as:

$$\dot{\psi}_r = -\frac{R_r}{L_r}\psi_r + \frac{L_m R_r}{L_r}u1 \quad (13)$$

$$\dot{\omega}_r = -\frac{B}{J}\omega_r + \frac{1}{J}K_T u2 - \frac{T_L}{J} \quad (14)$$

Thus, the new inputs $u1$ and $u2$ can be used to control the rotor flux and motor speed via following PI controllers [8].

$$u1 = K_{p1}(\psi_r^* - \psi_r) + K_{i1}\int_0^t(\psi_r^* - \psi_r)dt \quad (15)$$

$$u2 = K_{p2}(\omega_r^* - \omega_r) + K_{i2}\int_0^t(\omega_r^* - \omega_r)dt \quad (16)$$

From equations (11), (12), (13) and (14) the expression for control inputs can be written as equations (17) and (18) [15].

$$u1 = \frac{1}{\psi_r}(\psi_{\alpha r}i_{\alpha s} + \psi_{\beta r}i_{\beta s}) \quad (17)$$

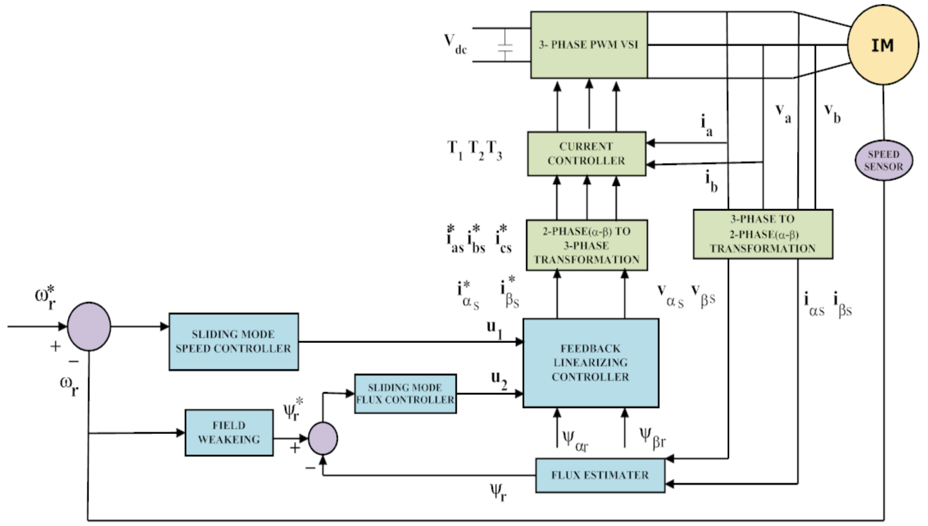


Fig. 1. The proposed drive system with sliding mode speed and flux controller

$$u2 = (\psi_{\alpha r} i_{\beta s} - \psi_{\beta r} i_{\alpha s}) \tag{18}$$

Rewriting the equations (17) and (18) for derivation of $i_{\alpha s}$ and $i_{\beta s}$ in terms of $u1$ and $u2$

$$i_{\alpha s} = \frac{\psi_{\alpha r}}{\psi_r} u1 - \frac{\psi_{\beta r}}{\psi_r^2} u2 \tag{19}$$

$$i_{\beta s} = \frac{\psi_{\beta r}}{\psi_r} u1 + \frac{\psi_{\alpha r}}{\psi_r^2} u2 \tag{20}$$

The above equations (19) and (20) represent a feedback linearization decoupling controller. The schematic diagram is shown in Fig.1.

V. SLIDING MODE CONTROLLER

In sliding mode entire system dynamics is governed by sliding surface parameters. System response is insensitive to parameter variation and external disturbances. The conventional PID controllers are unable to tackle such problem. In past decade, a lot of research work has been reported using SMC in place of PID controller in many applications. Because its control strategy does not require mathematical model and it can sustain external disturbances up to the limited range, robustness is its inherent characteristics. As induction motor is a multivariable, coupled and nonlinear system and it also faces lots of internal and external disturbances during practical operation. SMC based controllers have been successfully implemented for performance improvement of induction motor.

In SMC, the system response in the phase plane is forced to follow a sliding line [1]. The dynamics of error e(t) and its derivatives $\dot{e}(t)$ need to be driven to zero along the sliding line $s(t) = \dot{e}(t) + \lambda e(t)$ [1]. In time domain, the corresponding response is exponentially decaying. Its time constant (λ) depends on the slope of the sliding line and control signal forces the response to slide on slide-line and system state error always remains on zero state. This process can be easily implemented by switching process back and forth between negative and positive controller gain. The system error can not only be made zero, but its response can be made independent of the plant parameters [6].

To design a sliding mode speed and flux controller for the feedback linearized induction motor system designed in section 4.

The steps are as follows:

$$e_1(t) = \omega_r^* - \omega_r \tag{21}$$

$$e_2(t) = \psi_r^* - \psi_r \tag{22}$$

Taking the derivative of error

$$\dot{e}_1(t) = \dot{\omega}_r^* - \dot{\omega}_r \tag{23}$$

$$\dot{e}_2(t) = \dot{\psi}_r^* - \dot{\psi}_r \tag{24}$$

Substituting the expression of $\dot{\omega}_r$ and $\dot{\psi}_r$ from equations (13) and (14) in equation (23) and (24) respectively

$$\dot{e}_1(t) = \dot{\omega}_r^* + \frac{B}{J}\omega_r - \frac{K_T}{J}u2 + \frac{T_L}{J} + d_1(t) \tag{25}$$

$$\dot{e}_2(t) = \dot{\psi}_r^* + \frac{R_r}{L_r}\psi_r - \frac{L_m}{L_r}R_r u1 + d_2(t) \tag{26}$$

Where, extra-terms $d_1(t)$ and $d_2(t)$ are the external disturbances and uncertainty appears in speed and flux loop during practical application [17].

The sliding surface s(t) in integral form is defined as [1]:

$$s_1(t) = e_1(t) + \lambda_1 \int e_1(t)dt \tag{27}$$

$$s_2(t) = e_2(t) + \lambda_2 \int e_2(t)dt \tag{28}$$

So that we have

$$\dot{s}_1(t) = \dot{e}_1(t) + \lambda_1 e_1(t) \tag{29}$$

$$\dot{s}_2(t) = \dot{e}_2(t) + \lambda_2 e_2(t) \tag{30}$$

Substituting the value of $\dot{e}_1(t)$ and $\dot{e}_2(t)$ from equations (25) and (26), we obtain variable surface equation as follows:

$$\dot{s}_1(t) = \dot{\omega}_r^* + \frac{B}{J}\omega_r - \frac{K_T}{J}u2 + \frac{T_L}{J} + d_1(t) + \lambda_1 e_1 \tag{31}$$

$$\dot{s}_2(t) = \dot{\psi}_r^* + \frac{R_r}{L_r}\psi_r - \frac{L_m}{L_r}R_r u1 + d_2(t) + \lambda_2 e_2 \tag{32}$$

The best approximation of $u1 \,\&\, u2$ of a continuous control law that would satisfy the precedent condition, i.e., $\dot{s} = 0$, can be considered in the following form[1]:

$$u2 = u2^{eq} + u2^{n} \tag{33}$$

$$u1 = u1^{eq} + u1^{n} \tag{34}$$

Where u1 and u2 are the control vector, $u1^{eq}$ and $u2^{eq}$ are the equivalent control vector and $u1^{n}$ and $u2^{n}$ are the switching part of the control (the correction factor).
During the permanent sliding mode condition

$$s_1(t) = 0 \quad \dot{s}_1(t) = 0 \quad u2^{n} = 0 \tag{35}$$

$$s_2(t) = 0 \quad \dot{s}_2(t) = 0 \quad u1^{n} = 0 \tag{36}$$

Equivalent control for satisfying above condition is

$$u2^{eq} = \frac{J}{K}(\dot{\omega}_r^* + \frac{B}{J}\omega_r + \frac{T_L}{J} + \lambda_1 e_1) \tag{37}$$

$$u1^{eq} = \frac{L_r}{L_m R_r}(\dot{\psi}_r^* + \frac{R_r}{L_r}\psi_r + \lambda_2 e_2) \tag{38}$$

The switching control law is defined as

$$u1^{n} = -\beta_1 \operatorname{sgn}(s_1) \tag{39}$$

$$u2^{n} = -\beta_2 \operatorname{sgn}(s_2) \tag{40}$$

where, sgn() is signum function. β_1, β_2 are the respective switching gain and must be greater than total uncertainties present in corresponding circuit guaranteed by the Lyapunov stability criterion [1]. In order to reduce chattering J. J. E. Slotine proposed an approach, by introducing boundary layer of with Φ on either side of the switching surface [1].

Then switching control law redefines by

$$u1^{n} = -\beta_1 \operatorname{sgn}(s_1/\Phi_1) \tag{41}$$

$$u2^{n} = -\beta_2 \operatorname{sgn}(s_2/\Phi_2) \tag{42}$$

Finally the speed command and flux command variable *u2* and *u1* can be found as

$$u2 = \frac{J}{K_T}\left[\lambda_1 e_1 + \dot{\omega}_r^* + \frac{B}{J}\omega_r^* + \frac{T_L}{J}\right] - \beta_1 \operatorname{sgn}(s_1/\Phi_1) \tag{43}$$

$$u1 = \frac{L_r}{L_m R_r}\left[\lambda_2 e_2 + \dot{\psi}_r^* + \frac{R_r}{L_r}\psi_r^*\right] - \beta_2 \operatorname{sgn}(s_2/\Phi_2) \tag{44}$$

With the above speed and flux command for the set value of speed and flux are obtained.

VI. ROTOR FLUX ESTIMATION

The knowledge of rotor flux is essential for the feedback linearization. The direct measurement of flux is practically difficult. Therefore, the voltage model of flux estimation has been preferred [6], where measurement of terminal voltage and stator current are required.

The α-β axis stator voltage vector component voltages are estimated as

$$V_{\alpha s} = -\frac{V_b}{\sqrt{3}} + \frac{V_c}{\sqrt{3}} \tag{45}$$

$$V_{\beta s} = \frac{2}{3}(V_a - \frac{V_b}{2} - \frac{V_c}{2}) \tag{46}$$

Similarly the α-β axis components of stator currents are estimated as

$$i_{\alpha s} = \frac{2}{3}(-\frac{\sqrt{3}}{2}i_b + \frac{\sqrt{3}}{2}i_c) \tag{47}$$

$$i_{\beta s} = \frac{2}{3}(i_a - \frac{i_b}{2} - \frac{i_c}{2}) \tag{48}$$

Where (V_a, V_b, V_c) and (i_a, i_b) are the sensed stator voltage and stator currents. The α-β axis component of stator flux can be estimated by using the α-β component of stator voltages and currents.

$$\psi_{\alpha s} = \int (V_{\alpha s} - R_s i_{\alpha s})dt \tag{49}$$

$$\psi_{\beta s} = \int (V_{\beta s} - R_s i_{\beta s})dt \tag{50}$$

The α-β component of rotor current referred to stator reference frame is obtained from the α-β component of stator fluxes and currents.

$$i_{\alpha r} = \frac{\psi_{\alpha s}}{L_m} - \frac{i_{\alpha s}}{L_m}L_s \tag{51}$$

$$i_{\beta r} = \frac{\psi_{\beta s}}{L_m} - \frac{i_{\beta s}}{L_m}L_s \tag{51}$$

Finally the α-β component of rotor flux is estimated by using the α-β component of rotor currents, stator currents and machine parameters.

$$\psi_{\alpha r} = L_r i_{\alpha r} + L_m i_{\alpha s} \tag{53}$$

$$\psi_{\alpha r} = L_r i_{\alpha r} + L_m i_{\alpha s} \tag{54}$$

Where, L_s and L_m are rotor self and magnetizing inductance respectively.

VII. DESIGN OF CURRENT CONTROLLER

By using park transformation three phase reference currents i_{as}^*, i_{bs}^* and i_{cs}^* are obtained from $i_{\alpha s}^*$ and $i_{\beta s}^*$. The switching signal for the inverter devices is obtained by comparison of the motor currents i_{as}, i_{bs}, and i_{cs} with their reference counter parts i_{as}^*, i_{bs}^* and i_{cs}^*. Hysteresis-band with limits is properly selected by better performance of the current controller. The states of the devices are obtained in the following manner.

If $i_{as} \leq i_{as}^{*}$ – band limit then T_1 is on and $V_{an} = V_{dc}$

If $i_{as} \geq i_{as}^{*}$ + band limit then T_4 is on and $V_{an} = 0$

VIII. SIMULATION RESULT AND DISCUSSION

The proposed control scheme is simulated for two methods in MATLAB simulink environment using power system block set. First one uses two P-I controllers for controlling speed and flux and second scheme uses sliding mode (SM) based controller in place of corresponding PI controllers.

Simulation results of speed response, torque response and reference torque, stator current α-β components, estimated rotor flux linkage, estimated rotor flux α-β components, are presented for PI controller based system in Fig. 2, and sliding mode controller based system in Fig. 3. The results of both the schemes are compared and shown in Table 1.

I. Step change of speed and speed reversal response along with addition and removal of external load with PI controller:

The aim behind designing this scheme is to implement feedback linearizing controller successfully in close loop for tracking set speed. The response shows (Fig. 2) that the right from beginning motor picks up speed at constant rate and reaches its set point (from zero to 500 rpm) in 0.39 s. The initial torque response picks-up very fast and it achieves maximum torque limit within 0.1 s. But the huge ripple is investigated in torque response. Its effect is also appearing in the stator α-β axis current components. At t = 1.00 s, 10 Nm external load is subjected to the running drive system (500 rpm). Speed deep is almost low. Further, at t = 1.5 s, load torque is removed. The motor attains the set speed. This shows that feedback controller is inherently robust in characteristics. It has been observed that the α-β axis current and rotor flux components are complete decoupled during transient condition. Therefore, the simulated scheme claim to achieve complete decupling as stated in the reported papers reported using the same feedback linearizing controller [15], [16] and [18].The drive system is also tested for speed reversal. The drive system is also tested for speed reversal. The speed reversal command (from 500 rpm to -500 rpm) is given at t = 2.00 s. The motor takes 0.85 s as total reversal time.

II. Step change of speed and speed reversal response along with addition and removal of external load with sliding mode (SM) controller

The same set-up is tested with SM controller. For starting time, reversal time forward pick up time, the propose scheme with SM controller results in a better performance than the P-I controller shown in Table 1. Beside it, there is negligible variation in speed about 0.035 rpm at the time of application of the load Fig.3, whereas in case of P-I based scheme it is 0.5 rpm (Fig.2). This indicates the inherent robustness of sliding mode. Another significant thing observed in Fig. 2 and Fig. 3 is that the motor speed oscillates around the set speed having some magnitude and frequency. As frequency depends on the switching frequency of the inverter switch, it is another aspect of control but magnitude can be controlled by SM controller. The peak to peak difference in case of SM controller is 0.04 rpm where as in PI controller it is found to be 0.13 rpm. Reduction in magnitude of torque ripple in SM controlled scheme leads to reduced speed oscillation. The sliding mode speed controller based system gives better result, as seen from Fig.3. The robustness test of the SMC based controller is implemented for the following two conditions

a. When the rotor speed is 500 rpm, at 0.5 sec the value of stator resistance R_s is increased by 25% of its rated value.

b. For the same operating condition motor inertia is increased to twice its original value, at time t =0, i.e., just when the motor starts.

The corresponding simulation results for speed, torque and rotor flux response are shown in Fig.4. Robustness of sliding mode controller based system for stator resistance variation is clearly seen from the simulation results. The moment of inertia is increased to twice the original value at the time of motor starting. So starting time is increased to 0.7 s, i.e., nearly doubled.

IX. CONCLUSION

The control for induction motor for two schemes with feedback linearization P-I controller and sliding mode controller are presented. The complete scheme is simulated in MATLAB simulink environment. The performance of the system with both controllers is compared in terms of speed, torque, motor current and flux response. It is found that sliding mode controller gives decoupling at all stages, reduced torque ripple and better rotor flux response. For starting time, reversal time, forward peak-up time, the proposed scheme with sliding mode controller results in a better performance than the PI controller. Robustness of sliding mode controlled system is also verified for stator resistance and moment of inertia variations. Thus sliding mode controller improves overall performance of the drive system.

X. APPENDIX

Motor Specifications

Three Phase Squirrel Cage Induction Motor– 5 HP (3.7 kW), 4 pole, Δ-connected, 415 V, 1445 rpm, R_s = 7.34 Ω, L_{ls} =0.021 H, L_m = 0.5H, R_r = 5.64 Ω, L_{lr} = 0.021 H, J=0.16 kg-m^2, B=0.035 kg-m^2/s.

Controller Specifications

P-I Flux controller : K_p = 1000, K_i= 500;

P-I speed controller: K_p =10,.K_i=0;

SM speed controller : λ_1 =120,β_1 =110

SM flux controller: λ_2 =200, β_1 =.5

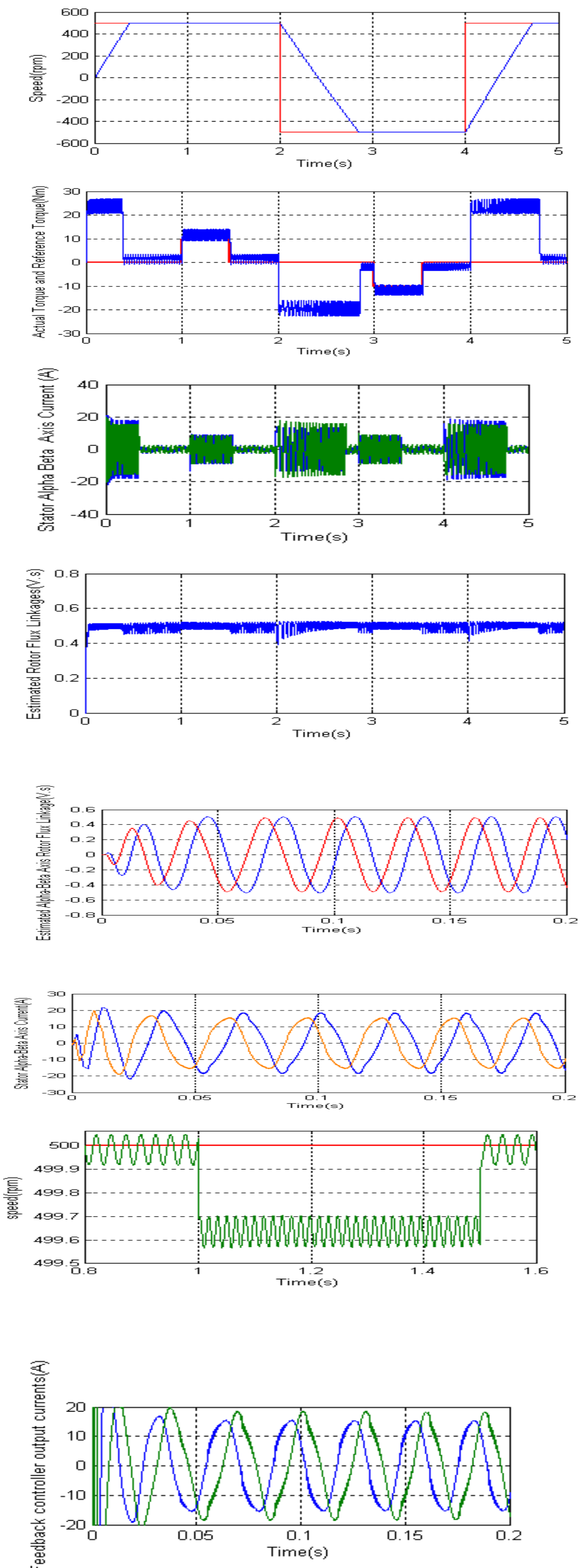


Fig. 2. Simulation response with PI controllers

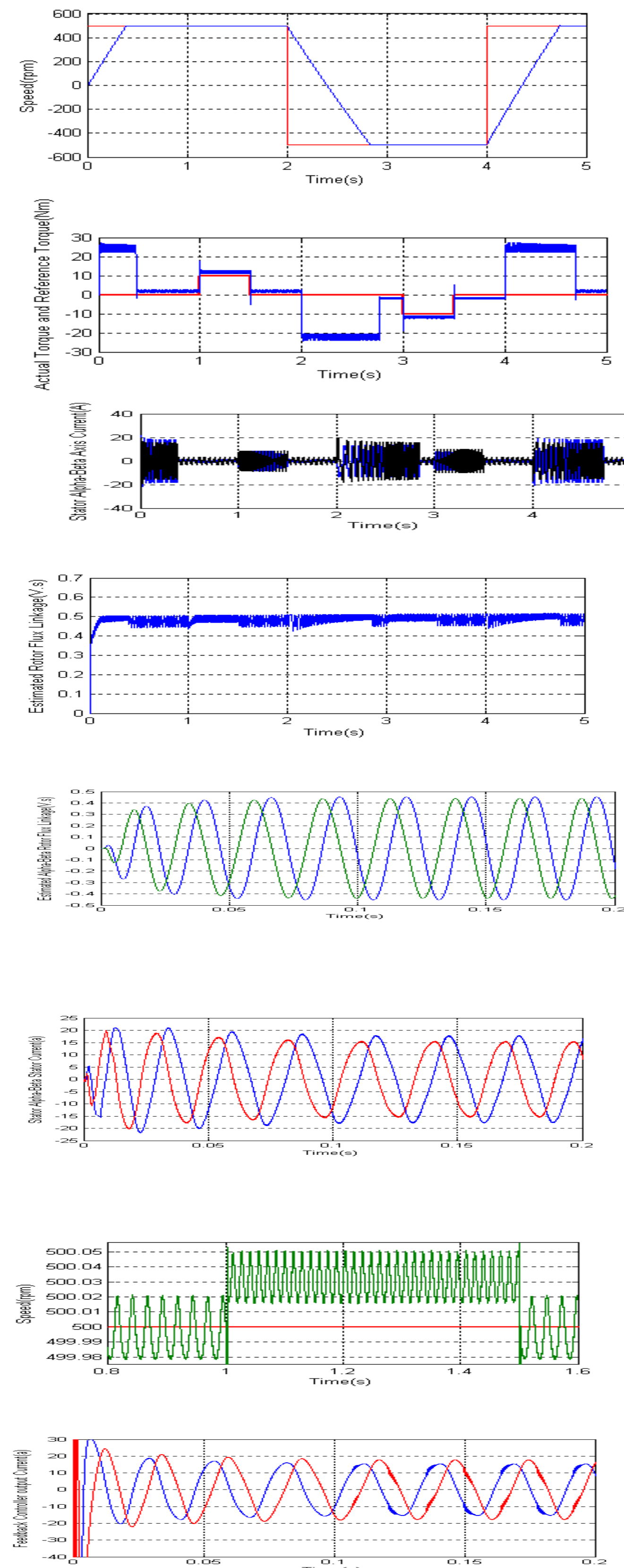
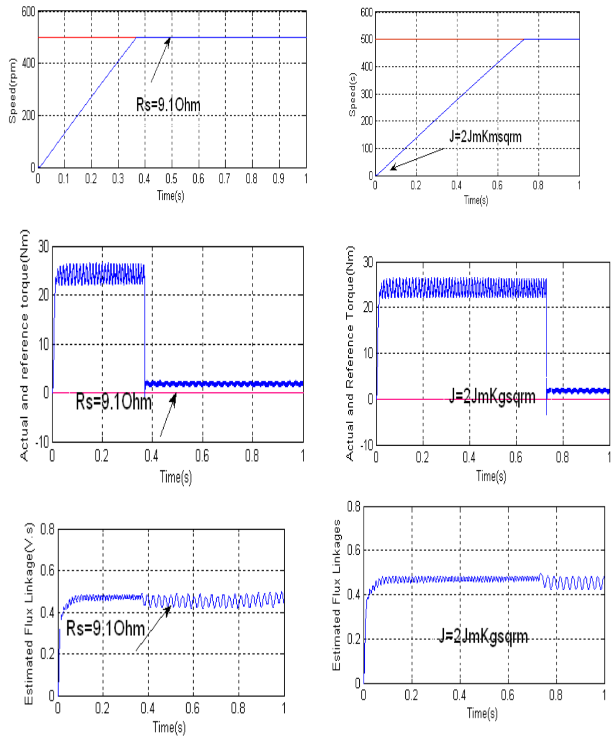


Fig. 3. Simulation response with SM controllers

TABLE 1: Comparative performance of proposed scheme with only PI and with SM controller

Input Signal	PI Controller	SMController
1. At t=0.00s, ω_r^* =500rpm, T_l= 0 N.m	Starting time= 0.395 s	Starting time=0.36 s
2. At t=1.00 s, T_l = 10 N.m	Speed deeps by 0.5 rpm	Speed rises by 0.03 rpm
3. At t=1.50 s, T_l = 0 N.m	Speed rises to original speed of 500rpm	Speed deeps to original speed of 500rpm
4. At t=2.00 s, ω_r^* = -500rpm, T_l= 0 N.m	Reversal Time=0.85 s	Reversal Time=0.77s
5. At t=3.00 s, T_l = -10 N.m	Speed deeps by 0.5 rpm	Speed rises by 0.03 rpm
6. At t=3.5 s , T_l= 0 N.m	Speed rises to original value -500 rpm	Speed deeps to original value of -500 rpm
7. At t=4.00 s, ω_r^* = 500rpm, T_l =0N.m	Forward peak-up time =0.74s	Forward peak-up time =0.68 s



(a) (b)

Fig. 4. Simulation response for sliding mode controller based system with (a) stator resistance increase and (b) motor inertia increase

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Kanungo Barada Mohanty has received B.E. degree from Sambalpur Univ., M.Tech. and Ph.D. degrees from Indian Institute of Technology (IIT), Kharagpur in the years 1989, 1991 and 2002 respectively, in Electrical Engineering. He is a faculty member of Electrical Engg. Dept., National Institute of Technology, Rourkela since 1991, and currently serving as Associate Professor. He has published 14 journal papers and more than 40 conference papers. His research interests include control and estimation in induction machine specifically vector control, DTC and wind energy conversion systems. Dr. Mohanty is a Member of the IEEE, Fellow of the IE (India), Fellow of IETE, and Life Member of Solar Energy Society of India, and System Society of India.

Madhu Singh has received B.Sc. Engg.(Electrical) degree from N.I.T (formally B.C.E),Patna MBA(Personnel Management) degree from L N M Institute of Economic Development and Social Change, Magadh Univ, M.Sc.Engg.(Power Electronics) degree from N.I.T(formally R.I.T), Jamshedpur in the years 1990,1993, and 2007 respectively. She is a faculty member of Electrical Engg. Dept. National Institute of Technology, Jamshedpur since 1999. Her research interest area includes power electronics and drives, nonlinear control and artificial intelligent control application in motor servo drives. She is Member of the IE (India).