Feedback Linearizing Control of Induction Motor Drive by P-I Controllers in RTDS Environment

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***Abstract*—This paper reviews the model of the induction motor in stationary reference frame, with stator current and rotor flux components as state variables. The state feedback linearizing and decoupling control as applied to the induction motor drive is presented. With this control, the drive system is decoupled into two linear subsystems: electrical and mechanical. Comprehensive and systematic procedures are developed to determine the gains of the Proportional–Integral (P-I) controllers for electrical and mechanical subsystems. The controlled drive system is simulated in MATLAB SIMULINK and experimented in RTDS environment, and results are presented.**

***Index Terms*—feedback linearization; flux estimator; stationary reference frame; proportional-integral controller; real time digital simulator (RTDS)**

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 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 well acceptable in automation industries. But it requires complex control strategy, because it possesses three inherent drawbacks as follows. It is a higher order nonlinear dynamic system with internal coupling of states. Some state variables like rotor currents and flux are not directly measurable. Variations 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 a proper selection of state coordinates, under the hypothesis that the rotor flux is kept constant. The torque is only asymptotically decoupled from the flux i.e., decoupling is obtained only in steady state, when the flux amplitude is 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], [6].

This has further led to introduction nonlinear geometric control theory particularly feedback linearization, which can achieve completely decoupled torque and flux amplitude of the induction motor [7]-[18]. But for the satisfactory performance, the motor parameters of the controlled plant must be precisely known and accurate knowledge of the flux is required. In the last decade, a good number of research works has been reported incorporating various control schemes to simplify and to enhance the performance. The work includes several methods for accurate estimation of flux. But the control performance is still influenced by the uncertainties of the plant. Therefore, the motivation behind this work is to design a suitable robust control scheme to combat the uncertainties arising in practical application.

In this work, decoupling of flux and speed in induction motor drive with state feedback linearization technique, and design of the proportional-cum-integral (P-I) controllers for flux and speed control loop are presented. The nonlinear dynamics of the induction motor in the stationary reference frame is presented in section II. The nonlinear model of induction motor is linearized by state feedback linearization and decoupling technique in section III. The induction motor is thus decoupled into two linear subsystems: electrical subsystem and mechanical subsystem. In section IV, systematic procedures are used to select the gains of the P-I controllers for each subsystem. In section V, closed loop control technique based on state feedback linearization algorithm is simulated using P-I controllers and results are discussed. The control scheme is simulated in MATLAB SIMULINK environment and verified in RTDS environment.

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II. Modeling of Induction Motor

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

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

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

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

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

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

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), (R_r, L_r) are stator and rotor parameters (resistance and inductance), L_m is magnetizing inductance and ω_r is the motor speed.

The electromagnetic torque developed is given by

$$T_e = K_T(\psi_{\alpha r}i_{\beta s} - \psi_{\beta r}i_{\alpha s}) \quad (6)$$

where,

$$K_T = \frac{3pL_m}{2L_r}$$

p is the number of pole pairs.

III. 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 systematic approach can be found in [1], [7]-[18].

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, outputs to be controlled are chosen as the rotor speed, ωr and rotor flux, ψr. The magnitude of rotor flux is given as

$$\psi_r^2 = \psi_{\alpha r}^2 + \psi_{\beta r}^2 \quad (7)$$

The time derivative of rotor flux linkage is

$$\dot{\psi}_r = \frac{1}{\psi_r}\left[\psi_{\alpha r}\dot{\psi}_{\alpha r} + \psi_{\beta r}\dot{\psi}_{\beta r}\right] \quad (8)$$

Substituting $\dot{\psi}_{\alpha r}$ $\dot{\psi}_{\beta r}$ from (3) and (4) into (8)

$$\dot{\psi}_r = \frac{1}{\psi_r}\left[\psi_{\alpha r}\left(-\frac{R_r}{L_r}\psi_{\alpha r} - p\omega_r\psi_{\beta r} + \frac{L_m R_r}{L_r}i_{\alpha s}\right) + \psi_{\beta r}\left(-\frac{R_r}{L_r}\psi_{\beta r} + p\omega_r\psi_{\alpha r} + \frac{L_m R_r}{L_r}i_{\beta s}\right)\right] \quad (9)$$

Simplifying (9), the rotor flux dynamic equation is

$$\dot{\psi}_r = -\frac{R_r}{L_r}\psi_r + \frac{L_m R_r}{L_r\psi_r}(i_{\alpha s}\psi_{\alpha r} + i_{\beta s}\psi_{\beta r}) \quad (10)$$

From (5) and (6), rotor speed dynamic equation is

$$\dot{\omega}_r = -\frac{B}{J}\omega_r + \frac{1}{J}K_T(\psi_{\alpha r}i_{\beta s} - \psi_{\beta r}i_{\alpha s}) - \frac{1}{J}T_l \quad (11)$$

Equations (10) and (11) describe flux and mechanical system, which has $i_{\alpha s,}$ and $i_{\beta s}$ as two control inputs. 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 decoupled one [7]. Equations (10) and (11) with new control input can be rewritten as:

$$\dot{\psi}_r = -\frac{R_r}{L_r}\psi_r + \frac{L_m R_r}{L_r}u1 \quad (12)$$

$$\dot{\omega}_r = -\frac{B}{J}\omega_r + \frac{1}{J}K_T u2 - \frac{T_l}{J} \quad (13)$$

From (10), (11), (12) and (13) the expression for control inputs can be written as [7].

$$u1 = \frac{1}{\psi_r}\left(\psi_{\alpha r}i_{\alpha s} + \psi_{\beta s}i_{\beta s}\right) \quad (14)$$

$$u2 = \left(\psi_{\alpha r}i_{\beta s} - \psi_{\beta r}i_{\alpha s}\right) \quad (15)$$

Above equations are rewritten in (16) and (17) for derivation of $i_{\alpha s}$ and $i_{\beta s}$ in terms of $u1$ and $u2$

$$i_{\alpha s} = \frac{\psi_{\alpha r}}{\psi_r}u1 - \frac{\psi_{\beta r}}{\psi_r^2}u2 \quad (16)$$

$$i_{\beta s} = \frac{\psi_{\beta r}}{\psi_r}u1 + \frac{\psi_{\alpha r}}{\psi_r^2}u2 \quad (17)$$

Equations (16) and (17) represent a feedback linearization decoupling controller. The block diagram of feedback linearizing controller is shown in Fig.1. The transformed model of induction motor as given in (12) and (13) is linear and decoupled. The developed torque and the rotor flux are independently controlled. The induction motor model is now decoupled into two linear subsystems: Electrical subsystem and Mechanical subsystem, shown in Fig.2 and Fig. 3, respectively. The electrical subsystem is given by (12) and mechanical subsystem is given by (13). The rotor flux and motor speed can be controlled by P-I controllers [8]. P-I Controllers are developed using linear control theory in [8], to obtained desired steady state and transient performance. The performance of the drive system largely depends upon the choice of the controller. Design of controllers is discussed in section IV.

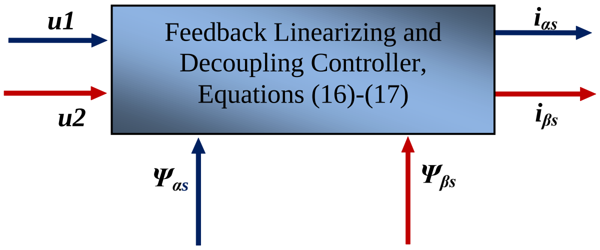


Figure 1. Feedback linearizing and decoupling controller

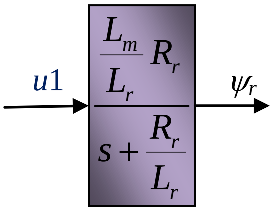


Figure 2. Block diagram of the open loop electrical subsystem

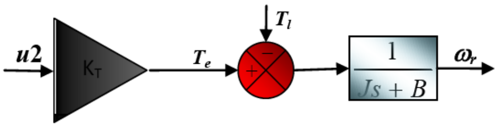


Figure 3. Block diagram of the open loop mechanical subsystem

IV. Design of P-I Controllers

The electrical and mechanical subsystems obtained above are type zero systems as the integral term is absent in both. So, a step input leads to a steady state error. Hence, controllers with integrating action are required for both the subsystems. In this scheme two Proportional-cum-Integral (P-I) controllers are used, one for the electrical subsystem and another for the mechanical subsystem. To have a feel of the influence of controller gains on the characteristic of the drive system in general and the factors affecting the gains of the P-I controllers in particular, the transfer function of the electrical and mechanical subsystem are analyzed. Drive system with designed controllers is simulated and it is verified that simulation response with the designed controllers is satisfactory. P-I controllers are designed using Modulus optimum method. The controller is designed so as to make the modulus of the closed loop transfer function, unity over a wide frequency range, starting from zero.

A. P-I Controller for Electrical Subsystem

Using the induction motor parameters given in Section V, L_r=0.52H, L_m=0.5H and R_r=0.64Ω, the open loop transfer function for the electrical subsystem shown in Fig.2 can be expressed as

$$\frac{\psi_r(s)}{u1(s)} = \frac{\frac{L_m R_r}{L_r}}{s+\frac{R_r}{L_r}} = \frac{5.42}{s+10.846}$$

This is a first order transfer function. Time constant of flux is:

$$\tau_{\psi r} = \frac{1}{10.846} = 0.092\,\text{sec}.$$

The time constant of the open system response is 0.092 s. For a unit step input, 90% rise time of flux is 2.3 times the electrical time constant, i.e., 2.3x 0.092= 0.216 s, thereby giving sluggish response. In order to track the reference flux ψ_r^* and to improve the system response one P-I controller is used as shown in Fig. 4. The transfer function $G_0(s)$ of the forward path including that of the controller is

$$G_0(s) = G_1(s).\ G_2(s)$$

$$= \frac{K_{p1}s+K_{i1}}{s}\cdot\frac{5.42}{s+10.846} = \frac{5.42(R_{p1}s+K_{i1})}{s(s+10.846)}$$

Then, the overall closed loop transfer function of the unity feedback flux control loop

$$\frac{\psi_r(s)}{\psi_r^*(s)} = \frac{G_0(s)}{1+G_0(s)} = \frac{\frac{5.42(K_{p1}s+K_{i1})}{s(s+10.846)}}{1+\frac{5.42(K_{p1}s+K_{i1})}{s(s+10.846)}}$$

$$= \frac{5.42(K_{p1}s+K_{i1})}{s^2+(10.846+5.42K_{p1})s+5.42K_{i1}}$$

The characteristic polynomial of the closed loop transfer function: $s^2+(10.846+5.42K_{p1})s+5.42K_{i1}$ is of second order and compared with the standard form: $(s^2+2\xi\omega_n s+\omega_n{}^2)$, where, ω_n is the natural frequency of oscillation and ξ is the damping factor. Assuming the natural frequency of oscillation, ω_n to be 75 rad/s,

$$5.42K_{i1} = \omega_n{}^2 = 5625$$

$$K_{i1} = \frac{5625}{5.42} = 1037.8$$

In order to make the system critically damped (i.e.,ξ=1), equating coefficients of *s*, proportional gain K_{p1} is obtained.

$$10.846 + 5.42K_{p1} = 2\xi\omega_n = 150$$

$$K_{p1} = \frac{139.154}{5.42} = 25.67$$

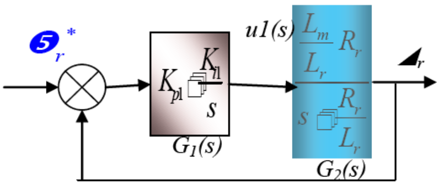


Figure 4. Block diagram of the closed loop electrical subsystem with P-I controller

With these values of K_{p1} and K_{i1}, the control law for the electrical subsystem is:

$$u1 = 25.67(\psi_r^* - \psi_r) + 1037.8\int_0^t(\psi_r^* - \psi_r)dt$$

Thus closed loop poles are at -75, farther away from the origin of complex s-plane, than the open loop pole at -10.846. Rise time for unit step speed input is $2.3\times\frac{1}{75}=0.03$ s. This makes the closed loop system's dynamic response faster and more stable. The drive system is simulated taking different value of ω_n and it is observed from the simulation studies that ω_n = 75 rad/s is more acceptable.

B. P-I Controller for Mechanical Subsystem

Using the induction motor parameters given in Section V, L_r=0.52H, L_m=0.5H and R_r=0.64Ω, J=0.16 kg.m^2 and B=0.035 kg.m^2/s, the open loop speed transfer function is:

$$\frac{\omega_r(s)}{u2(s)}=\frac{K_T}{Js+B}=\frac{18}{s+0.22}$$

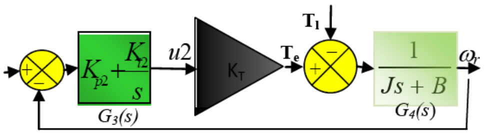


Figure 5. Block diagram of the closed loop mechanical subsystem with a P-I controller

This is a fist order transfer function. Time constant of speed is $\tau_{\omega_r}=\frac{1}{0.22}=4.55\,\text{sec}$ It's rise time is $2.3\times4.55=10.465$ s and pole at -0.22 gives sluggish response. In order to track the reference speed ω_r^* and to improve the transient performance a P-I controller is used as shown in Fig. 5.

The transfer function of the forward path including that of the controller is

$$G_{01}(s)=G_3(s).G_4(s)$$

$$=\frac{K_{p2}s+K_{i2}}{s}.\frac{K_T}{(Js+B)}=\frac{18(K_{p2}s+K_{i2})}{s(s+0.22)}$$

Then, the overall closed loop transfer of using feedback speed control loop can be expressed as:

$$\frac{\omega_r(s)}{\omega_r^*(s)}=\frac{18(K_{p2}s+K_{i2})}{s^2+(0.22+18K_{p2})s+18K_{i2}}$$

The closed loop transfer function is of second order. The characteristic polynomial is in the form $s^2+2\xi\omega_n s+\omega_n^2$. Assuming ω_n to be 4 rad/s, and comparing the terms:

18 K_{i2} = 16; $\qquad K_{i2}=\frac{16}{18}=0.88$

Assuming the system to critically damped (i.e., ξ=1)

0.22 + 18 K_{p2} = 2ξω_n = 2 x4

$$K_{p2}=\frac{8-0.22}{18}=.432$$

With these gain values the speed control law is given by:

$$u2=0.432(\omega_r^*-\omega_r)+0.88\int(\omega_r^*-\omega_r)$$

The closed loop pole is at -4, farther to the left of the origin of complex s-plane than the open loop pole -0.22. Rise time for unit step speed input is $=2.3\times\frac{1}{7.5}=0.306$ s. The dynamic response is faster and more stable. Therefore, above gains are acceptable. The drive system is simulated by taking different value of ω_n and it is observed from the simulation studies that ω_n = 4 rad/s is a good choice.

C. System Description

The schematic block diagram of the proposed system is shown in Fig. 6. The scheme consists of two P-I controllers with feedback linearizing algorithm, one flux estimator, and one current controlled PWM voltage source inverter. Two P-I controllers are 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 the 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.

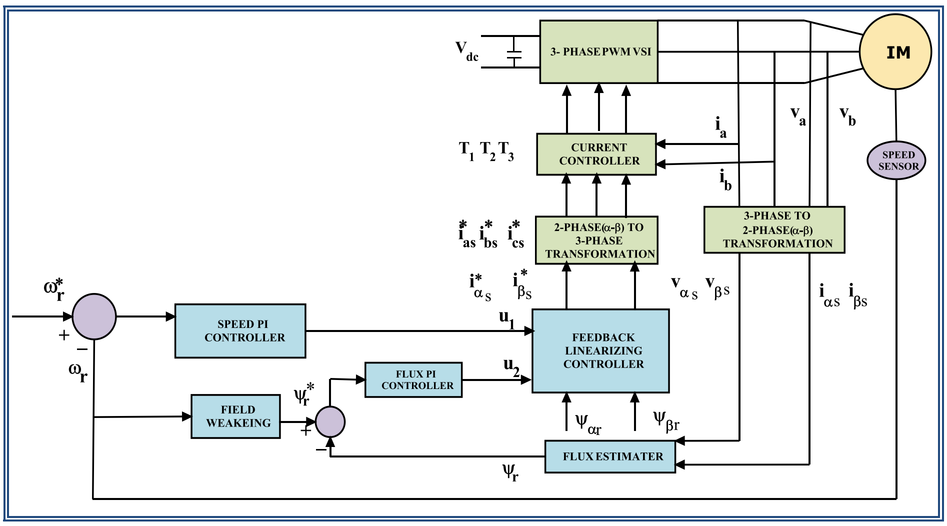


Figure 6. Schematic diagram of a linearized induction motor with P-I speed and flux controller

V. Simulation Results and Discussions

The proposed control scheme is simulated in SIMULINK and experimented with the real-time digital simulator (RTDS). The specifications and parameters of the induction motor are as follows. Three Phase Squirrel Cage Induction Motor –:5 HP (3.7 kW), 4 pole, Δ-connected, 415 V, 1445rpm, R_s = 7.34Ω, L_{ls} =0.021H, L_m =0.5H, Rr = 5.64Ω, L_{lr} = 0.021H, J=0.16 kg-m^2, B=0.035 kg-m^2/s.

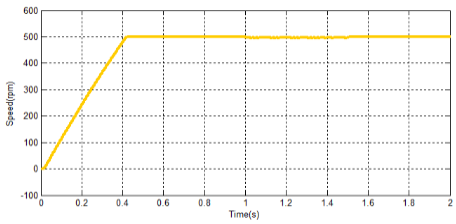
Simulation results corresponding to speed response, electromagnetic torque response, stator current and rotor flux are presented in Fig. 7 for SIMULINK simulation and in Fig. 8 for RTDS. Rotor speed, developed electromagnetic torque, stator phase-a current, α-β components of rotor flux are shown under different dynamic conditions such as starting acceleration and load perturbation (load application and load removal). From the obtained result the following salient features are observed.

Starting Dynamics: The three phase squirrel cage induction motor is fed from a controlled voltage and frequency source. The reference speed is set at 500 rpm with a current limit set at the rated value. Therefore, the starting current is limited to the rated current when the motor builds up the required starting torque to reach the set speed. The motor reaches its set speed in 0.43s. The theoretical value is obtained in section IV as 0.306s. When the speed error becomes zero rpm the winding current also reduces to no load value and the developed torque becomes equal to no load torque as observed in the stator current response shown in Fig. 7. The rotor flux amplitude remains constant throughout.

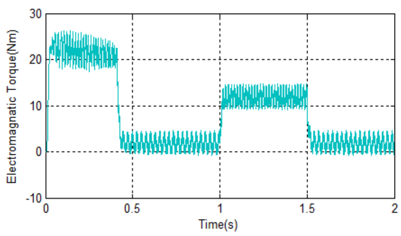
Load Perturbation: As shown in Fig. 7, when the motor is running at a steady state speed of 500 rpm, a load torque equal to the 10 N.m is applied at t = 1s and removed at t = 1.5s. Application of the load results in increase of stator current to corresponding value.

VI. Conclusions

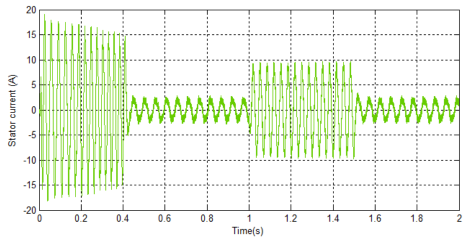
The feedback linearization control for induction motor drive has been presented to decouple speed and flux. Systematic procedure is adopted to design P-I controllers for electrical and mechanical subsystems. The complete scheme is simulated in MATLAB SIMULINK environment. The performance of the system is observed in terms of speed response, torque response, motor current and flux. The results obtained establish that decoupling of flux and speed is obtained at all stages through the proposed control algorithm. The scheme is implemented in RTDS. There is reduction in the torque ripples and better rotor flux response.



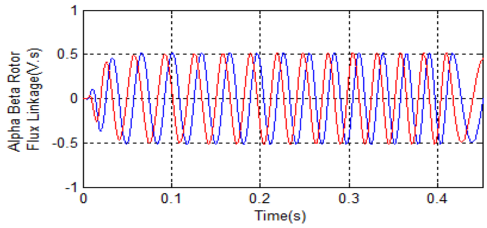
(a) Speed response



(b) Electromagnetic torque response

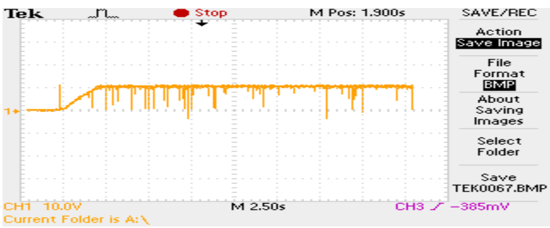


(c) Stator current response

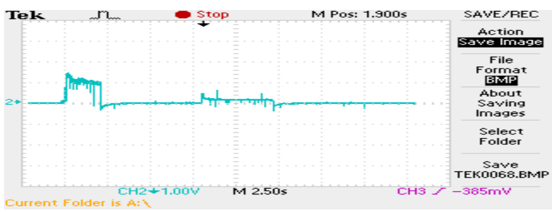


(d) Alpha-Beta components of rotor flux linkage

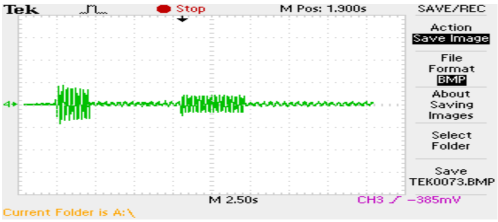
Figure 7. Simulation results of drive system with P-I controllers



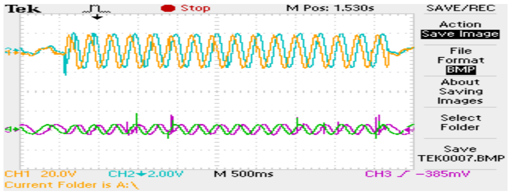
(a) Speed response



(b) Electromagnetic torque response



(c) Stator current response



(d) Alpha-Beta components of rotor flux linkage

Figure 8. Real time simulation results of the system with RTDS

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