A COMPARATIVE STUDY OF VECTOR CONTROLLED INDUCTION MOTOR DRIVE USING DIFFERENT SPEED CONTROLLERS

A DISSERTATION

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requirements for the award of the degree
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By

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CANDIDATE DECLARATION

I hereby declare that the work presented in this Seminar report entitled **"A comparative study of vector controlled induction motor drive using different speed controllers"** submitted in partial fulfilment of the requirements for the award of **Integrated Dual Degree** in Electrical Engineering with specialization in **Power Electronics** at the Indian Institute of Technology Roorkee, is an authentic record of my original work carried out during the period from June 2011 to June 2012 under the guidance of **Dr. S. Ghatak Choudhuri**, Assistant Professor, Department of Electrical Engineering, IIT Roorkee.

I have not submitted the matter embodied in this Seminar Report for the award of any other degree.

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CERTIFICATE

This is to certify that the above statement made by the candidate is correct to the best of my knowledge.

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I must concede that there were lot more who helped me directly or indirectly in the completion of this work, whose name I could not mention.

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RATHI VIVEK VIJAYKUMAR

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CERTIFICATE

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Chapter 1
Introduction

The induction motor is superior to the dc motor with respect to size, weight, rotor inertia, maximum speed capability efficiency, and cost. However, the case of control of dc motor cannot be equated because the induction motor has a nonlinear control structure, whereas the separately excited dc motor has a decoupled control structure with independent control of flux and torque. This report describes the technique of field oriented control or vector control which can be used with both induction and synchronous machines and essentially transforms the dynamic structure of the ac machines into that of a separately excited compensated dc motor. As a result, the induction motor drive can achieve four-quadrant operation with a fast torque response and good performance down to zero speed.

In a dc motor the armature mmf axis is established at 90 degrees electrical to the main field axis. This orthogonal or perpendicular relationship exists between the field flux and armature mmf axis. Hence in a separately excited dc motor with a constant flux value, torque is directly proportional to armature current. Direct control of armature current gives direct control of motor torque and fast response; because motor torque can be altered as rapidly as armature current can be altered.

The principle of field orientation originated in the work of Hasse and Blaschke. A variety of techniques are now developed but these techniques can be broadly classified into two groups: direct control and indirect control. The classification is based on the method used to determine the rotor flux vector. Direct field oriented control, as originally suggested by Blaschke, determines the magnitude and position of the rotor flux vector by direct flux measurement or by computation based on terminal conditions. Indirect flux orientation as suggested by Hasse, requires a high-resolution rotor position sensor to determine the rotor flux position.

The aim of vector control scheme is to resolve stator current into two mutually perpendicular direct and quadrature axis components which are decoupled from each other, in such a way that one component is responsible for production of torque and other for excitation so that the induction motor can be operated in way, similar to a separately excited compensated dc motor.

This implementation of vector control requires information regarding the magnitude and position of the rotor flux vector. The control action takes place in a field coordinate system using rotating rotor flux vector as frame of reference for stator currents and voltages.

A description mathematical model for vector controlled induction motor drive system, using synchronous reference frame, choosing the rotating d-axis to be the angle of the rotor flux linkage is given in this report. This choice offers a lot of advantages of simplifying control and analysis of the motor. Other choices frequently used in direct vector control are stator flux linkage frame (d-axis is aligned to the stator flux linkage) and air-gap flux linkage frame, which are discussed in brief, but does not come under scope of this investigation.

In fully compensated dc motor, the stationary poles are produced either with dc-excited field winding or permanent magnets, established in a magnetic field in which the armature rotates. The field winding is used to excite the field flux. Armature current is supplied to the rotor via brush and commutator for the mechanical work. The armature mmf is established at quadrature with the main field axis.

In an induction motor torque is produced by the electromagnetic interaction of the current in rotor conductor and magnetic flux density in the air-gap to which the rotor conductor is subjected. But in 3-phase induction motor of cage type all electrical inputs are into the stator circuit only. There is no external electrical connection to the rotor circuit. Neither it is practical to sense rotor current. The currents flowing in stator coil are primarily responsible for inducing the currents in the rotor. The primary cause of creation of magnetic field in the air-gap is also the flow of stator current. The technique of vector control is based on a method of separating out these two basic functions of stator current. In vector controlled induction motor, we look upon the current flowing in each stator phase as sum of two components at every instant – one component responsible for producing the magnetic field in the air-gap and the other component responsible for inducing the torque by inducing e corresponding current in the rotor conductors. These components are called field component and torque component respectively.

The torque (T_e) developed by a dc motor is proportional to the product of field flux (φ_f) and armature current (I_a) where K_{fdc} is the torque constant of dc motor.

$$T_e = K_{fdc} * I_a * \varphi_f \quad (1.1)$$

In dc motor, the field is determined by the current in the field coils. The mmf resulting from the armature current has ideally, no effect on the field flux because the spatial direction in which the armature mmf is oriented is in quadrature with direction of the field flux. This spatial angular displacement of π/2 radians is because of the commutator the angular orientation of the armature ampere-turns as well as the angular displacement between them, are not affected by the rotation of the shaft. Therefore changes in armature current, irrespective of whether they are caused by the controller or by changes in load, do not affect the field flux. This 'decoupling' naturally exists in dc motor.

In separately excited dc motor as shown in *figure 1.1(a)*, torque is given by

$$T_e = K_{fdc} * I_a * I_f \quad (1.2)$$

K_{fdc} is torque constant of the dc motor, I_a is the torque component of the current and I_f is the field or flux component of current.

The angle of π/2 radians between two fields created by I_a and I_f is independent of both the speed of the motor and the load on its shaft. These two fields are considered as orthogonal or decoupled vectors and are termed as independent control variables. The current I_f and the corresponding field flux is decoupled from the armature current I_a, thus accomplishing the fast dynamic response through quick control of the developed electromagnetic torque in transient as well as steady state operating conditions of the dc motor. This mode of control can be made applicable to an induction motor by controlling it in a synchronously rotating reference frame where the sinusoidal variables of the motor become dc quantities.

The induction motor with inverter and control is shown in *figure 1.1 (b),* with two control inputs i_{ds}^* and i_{qs}^*. These are direct axis and quadrature axis components respectively of the stator current and are expressed in synchronously rotating reference frame. In vector control i_{ds}^* is analogous to field current I_f and i_{qs}^* is analogous to armature current I_a of a dc motor. Therefore the torque in an induction motor is expressed as,

$$T_e = K_{fim} * i_{qs}^* * i_{ds}^* \quad (1.3)$$

K_{fim} is torque constant of induction motor, i_{qs}^* is torque producing component of stator current vector, i_{ds}^* is flux producing component of stator current vector.

The torque expressed as in *equation (1.3)* is identical to that of *equation (1.2)*. The two components i_{qs}^* and i_{ds}^* are orthogonal and the angle is independent of speed of motor or load on its shaft. For normal operating conditions below base speed the current i_{ds}^* is kept constant and the torque is varied by changing variable i_{qs}^*. Thus the induction motor behaves like a separately excited dc motor.

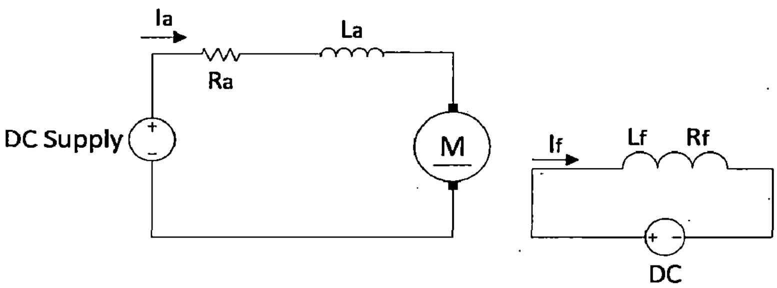


Fig 1.1 (a)

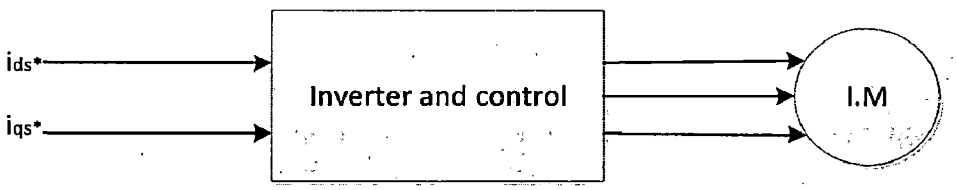


Fig. 1.1 (b)

Fig 1.1: Induction motor analogy with DC Motor in vector control.

As shown in *figure 1.2*, basic model of vector controlled induction drive system consists of two loops, torque control loop and flux control loop. The quantities i_{ds} and i_{qs} are flux and torque components respectively. These two currents are expressed in synchronously rotating reference frame therefore are dc in nature. These two quantities can be controlled independently. The feedback signals obtained from winding currents and voltages, rotor speed and fluxes etc., and these two current quantities i_{ds} and i_{qs} constitute vector controlled structure of induction motor. The vector controlled structure provides quick transient response due to decoupled control also, the conventional stability problem faced by an induction motor (i.e. crossing the breakdown torque point) does not exist in vector control mode.

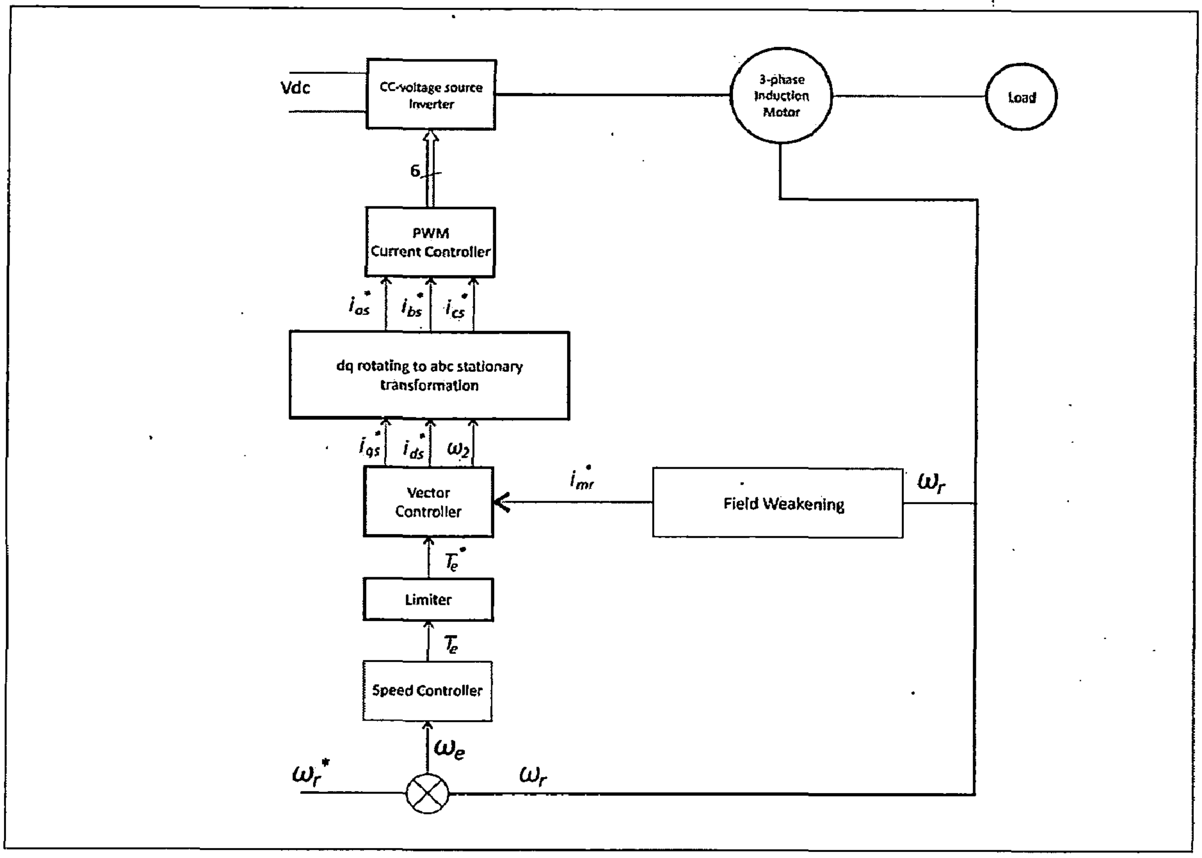


Figure 1.2 Primitive block-diagram model of vector controlled induction motor drive

The brief description of the proposed drive system is hereby stated.

Vector controlled induction motor drive (VCIMD) system comprises of: (a) three phase Voltage Source Inverter, (b) three phase squirrel cage induction motor, (c) the feedback control circuits receiving sensed signals, namely, the speed signal ω_m and current signals $i_{as} and\ i_{bs}$ respectively. The induction motor is star connected with an isolated neutral. Sensing currents corresponding to two phases is sufficient to yield feedback signals for all three-phase currents. The speed of the motor ω_m is compared with its reference value ω_m^* and the error is processed in the speed controller. A limit is kept on the output of the speed controller depending on the maximum permissible winding current. The output of the speed controller after the limiter, is considered as the reference torque (T*) signal and similarly the output of the field weakening block is considered as the reference value of excitation current i_{mr}^* respectively. These two signals are the main command signals for the controller. The controller calculates the d component i_{ds}^* and the q component i_{qs}^* of the reference current signal and also the slip frequency reference signal ω_2 in the synchronously rotating reference frame aligned with the rotor field. These signals are then converted into three-phase reference currents i_{as}^*, i_{bs}^* and i_{cs}^* in the stationary reference frame.

The induction motor is operated in the vector controlled mode with voltage source inverter in current controlled mode so that the winding currents are in the same pattern as that of the reference currents i_{as}^*, i_{bs}^* and i_{cs}^* computed by the controller in the Stationary Reference Frame. For current sensing, two of the three-phase currents are sensed using Hall Effect current sensors and the third current is computed from the two sensed currents (sum of all the three currents for a star connected balanced motor is zero). The current error signals are amplified and used as modulating signals for the Pulse Width Modulated (PWM) current controller. Triangular Carrier Waveform is generated at the required switching frequency. The point of intersection of the Triangular Carrier Waveform and the modulating signals, acts as the point of state change over for the resulting PWM driver signals. Six driver PWM signals emanate from the output of the PWM current controller and these are then fed to the respective gate driver circuits of the devices forming the VSI. Because of the closed loop current controlled operation of VSI, the winding currents follow the same pattern as the reference currents computed by the vector controller.

Further, the report discusses the speed controller design used for VCIMD the control over a wide operating range. It is being applied for a number of applications both in the low and high power ranges. Fast current control is essential for the vector control in order to achieve decoupling and torque control. Usually, a linear current controller of the proportional plus integral (PI) type and a feed-forward pulse-width-modulator are used for current vector control.

In fuzzy control it is not necessary to change the control parameters as the reference speed changes, however with the conventional PI controller this does not happens. With results obtained from the simulation and real time implementation and its comparison, it is clear that for the same operating condition the indirect vector controlled induction motor speed control using fuzzy logic technique is better performance than the conventional PI controller. In addition, the motor speed to be constant when the load varies. This proposed scheme is very useful for adjustable speed drive in industrial applications.

This report gives a comparative story of vector controlled induction motor drive using different speed controllers, viz. Proportional-Integral, Fuzzy logic and pre-compensated fuzzy proportional integral speed controllers. This report evaluates the performance of the drive over various operating conditions, viz. starting, speed reversal, load perturbation (load application and removal), evaluating the dynamic response under these conditions. This report shows how operation of PI speed controller results in occurrence of specific characteristics such as overshoot, undershoot, etc., which gets eliminated on use of Fuzzy logic pre-compensation or fuzzy logic controllers. This report also investigates the effect of pre-compensation over improver tuning of PI speed controller and gives a detailed comparison of above mentioned speed controller.

Chapter 2

Literature Review

R. J. Lee, P. Pillay and R. G. Harley [13] have discussed the equations of three preferable reference frames for use in the simulation of induction machines when using the d-q axis theory. It uses case studies to demonstrate that the choice of the reference frame depends on the problem to be solved and the type of computer available (analog or digital). During start-up and other severe motor operations the induction motor·draws large currents; produces voltage dips oscillatory torque and can even generate harmonics in the power system. It is therefore important to be able to model the induction motor in order to predict these phenomena.

Various models have been developed, and the d,-q or two-axis model for the study of transient behaviour has been well tested and proven to be reliable and accurate. It has been shown by P. C. Krause and C. H. Thomas [6] that the speed of rotation of the d,-q axes can be arbitrary although there are three preferred speeds or reference frames as follows:

(a) The stationary reference frame when the d-q axes do not rotate;
(b) The synchronously rotating reference frame when the d-q axes rotate at synchronous speed;
(c) The rotor reference frame when the d-q axes rotate at rotor speed.

On the basis of comparison between these frames it is concluded that:

(a) When a single induction motor is being studied, anyone of the three preferred frames of reference can be used to predict transient behaviour;
(b) If the stationary reference frame is used, then the stator d-axis variables are identical to those of the stator phase A variables. This eliminates the need to go through the inverse of Park's transform to obtain the actual stator variables, so saving in computer time. This would be useful when interest is confined to stator variables only, as for example in variable speed stator-fed induction motor drives;
(c) If the rotor reference frame is used then the rotor d-axis variables are exactly the same as the actual rotor phase-A variable. This again saves computer time by eliminating

the need to go through the inverse of Park's Transform to obtain the rotor phase-A variables. This would be useful when interest is confined to rotor variables only, as for example in variable speed rotor-fed induction motor drives;

(d) When the synchronously rotating reference frame is used, the steady DC quantities both of the stator and rotor d-q variables make this the preferred frame of reference when employing an analog computer. In the case of a digital computer solution, the integration step length may be lengthened without affecting the accuracy of results;

(e) Should a multi-machine system be studied then the advantages of the synchronously rotating reference frame appear to outweigh those of the other two reference frames.

Dal Y. Ohm [24] explains the fundamental dynamic mechanism of the motor in the synchronous frame is developed and the basic principles of vector control are discussed in general terms. For the alignment of the rotating reference frame with respect to the physical coordinate, one can choose the rotating d-axis to be the angle of the rotor flux linkage. In fact, this choice offers a lot of advantages of simplifying control and analysis of the motor.

Bhim Singh and S. Ghatak Choudhuri [28] have investigated DSP Based Implementation of Vector Controlled Induction Motor Drive using Fuzzy Pre-compensated Proportional Integral Speed controllers. The power circuit consists of a three-phase IGBT based voltage source inverter (VSI) feeding a three-phase squirrel cage induction motor. Test results are presented for drive starting, speed reversal and load perturbation modes using PI and FPPI speed controllers respectively and the advantage of pre-compensation is demonstrated in detail.

S. Masoudi, M. Reza Feyzi, M. B. B. Sharifian [32] have studied the field orientation control of an induction motor and found out that field orientation permitted fast transient response by decoupled torque and flux. The conventional PI controller has been widely used in industrial application due to the simple control algorithm and easy implementation. With help of the Matlab/Simulink, block model of induction motor drives can be constructed. They presented a novel fuzzy controller and posicast controller of an indirect field oriented induction motor drive for high performance. A superiority of the proposed fuzzy and posicast controller over conventional PI controller in handling nonlinear such as an induction motor has been effectively demonstrated by comparing speed controller with conventional PI controller under varying operating conditions like step change in speed reference and torque reference. Their results validate the robustness and effectiveness of the fuzzy logic controller and posicast controller for high performance of induction motor drive system.

The FLC is a knowledge-based control that uses fuzzy set theory and fuzzy logic for knowledge representation [32]. It has following main characteristics [11]:

(a) The FLC is a linguistic controller. It is not necessary to find a precise and accurate mathematical model of the controlled object

(b) The FLC is an ideal flexible nonlinear type controller. It can overcome the influence of only non-linear variations.

(c) The FLC has strong robustness as it is not sensitive to parametric variations of the controlled process [32].

Many FLC methods were presented in [11], [22], [27], [29] that are suitable for speed control of induction motor drives.

Chapter 3

Description of the Drive

Vector controlled induction motor drive and its main parts are briefly described below:

3.1. Speed Sensor

It measures the motor speed, since in the indirect vector control the accurate measurement of position of rotor flux vector is required which may be measured with the help of high resolution and high precision speed sensor Normally for this purpose either speed resolvers or a shaft encoders are used for the closed loop vector control of the cage induction motor.

3.2. Speed Controller

The measured speed (ω_r) is compared with the set reference speed (ω_r^*) in the error detector and the resulting output is known as speed error (ω_{re}) and this becomes the command input to the speed controller. The command input (ω_{re}) may be positive or negative depending upon the values of set reference speed and the motor shaft speed. Speed error (ω_{re}) is processed in the speed controller which may be of different types depending upon the required dynamic performance of the drive. In this report **PI, Fuzzy, pre-compensated fuzzy PI** speed controllers are considered for operation. The speed controller gives its output as command ($T^*_{(n)}$) which is input to the limiter.

3.3. Limiter

The limiter puts a limit on the output of the speed controller $T_{(n)}$ and its output is considered as reference torque. Limit on $T_{(n)}$ is required since during transients (such as starting, speed reversal, loading) experienced by the drive , the output of the speed controller jumps quickly to a very high value which may even go beyond the breakdown torque limit of the drive causing instability and over current. This is because the reference torque T^*affects the reference value of the torque producing component of the stator current i^*_{qs}. Therefore, in order to operate the drive in safe region with safe operating current putting a limit on the command output is necessary. This limit ensures that in no case the inverter output current jumps more than a set value thereby providing the feature of inherent over current protection in the drive.

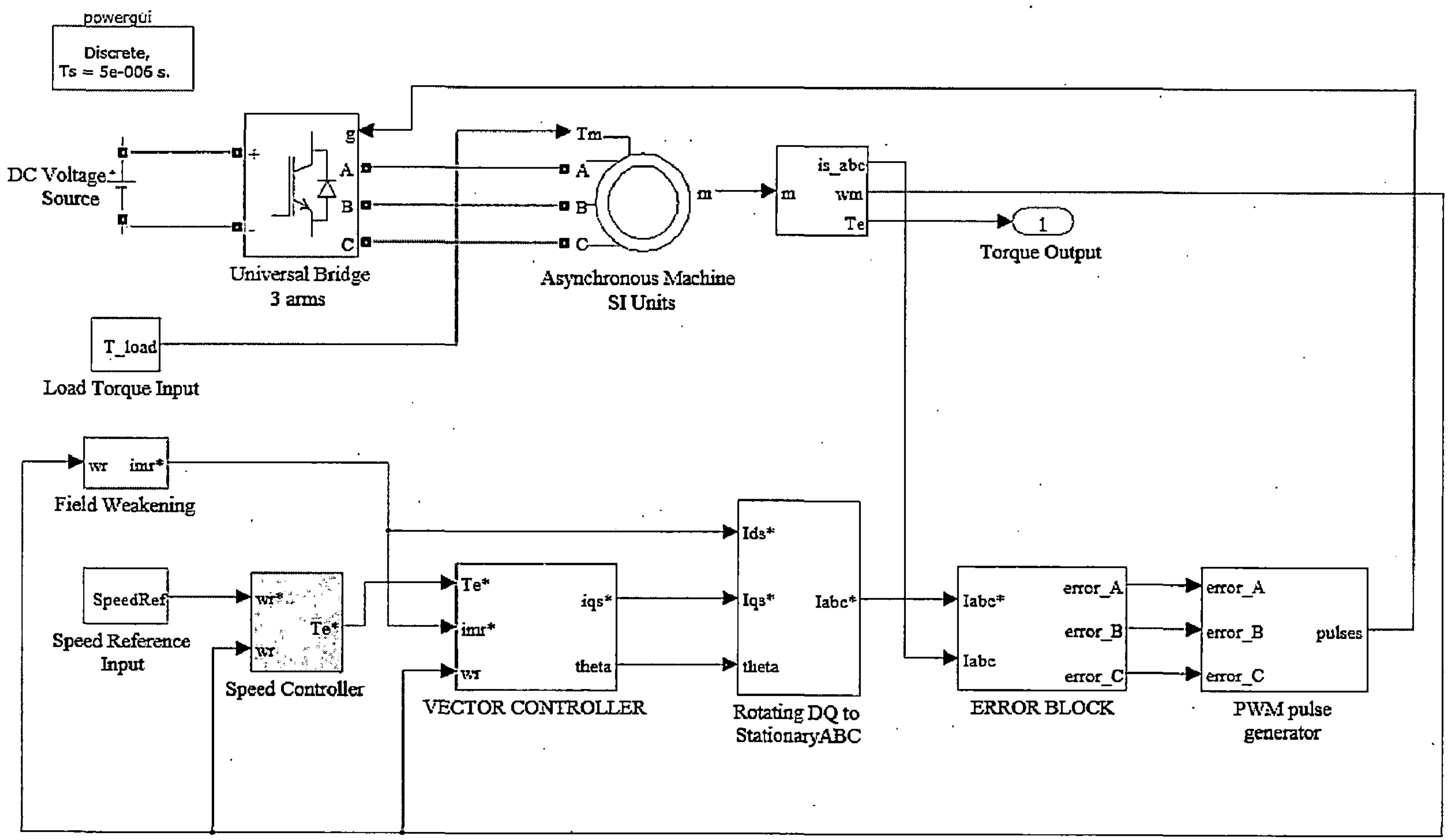


Fig. 3.1. Block diagram of vector controlled induction motor

When an abnormality arises because of the step changes in speed or load on the motor shaft, the reference torque which is command signal at the output of the limiter gets saturated at its maximum value with an appropriate polarity depending upon the desired direction of rotation of the motor. With this limit on the command torque T^*, it is ensured that in no case the current crosses a set limit for breakdown torque. This provides inherent stability to drive.

3.4. Field Weakening Control

Field weakening in the vector controlled drive is analogous to the field control of armature reaction in separately excited dc motor. Field weakening is normally used for operating the motor about its base speed. It is also used for efficiency optimization of the drive by balancing variable (winding) loss against the fixed (core) losses of the drive.

In case of vector controlled drive field weakening is normally not considered for operations below base speed. For operations below base speed a constant flux is maintained while for operation above the base speed the field weakening is considered in order to realize a constant power drive. This therefore exhibits characteristics similar to that of dc motor. Field weakening control incorporated in vector control structure of the drive constitutes its flux control loop.

The reference value of the exciting current i_{mr}^* is a function of the rotor speed. Mathematically, the logic for computation of the exciting current i_{mr}^* may be stated as follows:

$$i_{mr(n)}^* = I_m \; if \; \omega_r < base \; speed \; of \; the \; motor.$$

$$i_{mr(n)}^* = \frac{K_f I_m}{\omega_{m(n)}} \; if \; \omega_r < base \; speed \; of \; the \; motor$$

K_f is flux constant and I_m is rms value of the magnetising current.

3.5. Vector Controller

The output of the speed loop, reference torque (T^*), is processed in the vector control structure along with reference magnetizing current vector ($i_{mr}{}^*$). The vector control structure computes the orthogonal current components ($i_{ds}{}^*$ and $i_{qs}{}^*$) of the stator current vector ($i_s{}^*$). The online command slip speed ($\omega_2{}^*$) is also computed in the vector control structure. The currents $i_{ds}{}^{*}$ *and* $i_{qs}{}^{*}$ being in synchronously rotating reference frame are dc quantities in nature. In order to ensure the vector control of induction motor, its stator currents (i_{as}, i_{bs} *and* i_{cs}) are controlled in such a way that orthogonal currents (i_{ds} *and* i_{qs}) flowing through the stator windings have one to one correspondence with these reference orthogonal current components ($i_{ds}{}^{*}$ *and* $i_{qs}{}^{*}$).

For coordinate transformation the vector controlled structure should have knowledge about rotor flux position which is estimated by integrating the sum of rotor speed (ω_r) and command slip speed ($\omega_2{}^{*}$). The quantity so obtained gives flux angle (ψ) in radians.

Vector control consists of three stages as stated by following equations:

$$i_{ds}^{*}(n) = i_{mr}^{*}(n) + \tau_r \frac{d}{dt} i_{mr}^{*} \quad (3.1)$$

$$i_{qs}^{*}(n) = \frac{T^{*}(n)}{k\, i_{mr}^{*}(n)} \quad (3.2)$$

$$\omega_2^{*}(n) = \frac{i_{qs}^{*}(n)}{\tau_r i_{mr}^{*}(n)} \quad (3.3)$$

$$k = \frac{3}{2}\left(\frac{P}{2}\right)\frac{M}{1+\sigma_r} \quad (3.4)$$

Where $i_{ds}^{*}(n)$ and $i_{qs}^{*}(n)$ refer to the flux and torque components of the stator current at n^{th} instant respectively, $\omega_2^{*}(n)$ refers to the n^{th} instant reference slip speed of rotor. P, M, σ_r are the number of poles, mutual inductance and rotor leakage factor respectively. In addition to the decoupled components of the stator current, the flux angle ψ is defined at n^{th} instant as:

$$\psi(n) = \psi(n-1) + (\omega_2^{*} + \omega_r)\Delta T \quad (3.5)$$

Where $\psi(n)$ is the value of flux angle at the n^{th} instant, $\psi(n-1)$ is the value of flux angle at the $(n-1)^{th}$ instant, ω_2^{*}, ω_r and ΔT are the reference slip, frequency, rotor speed and sampling time respectively.

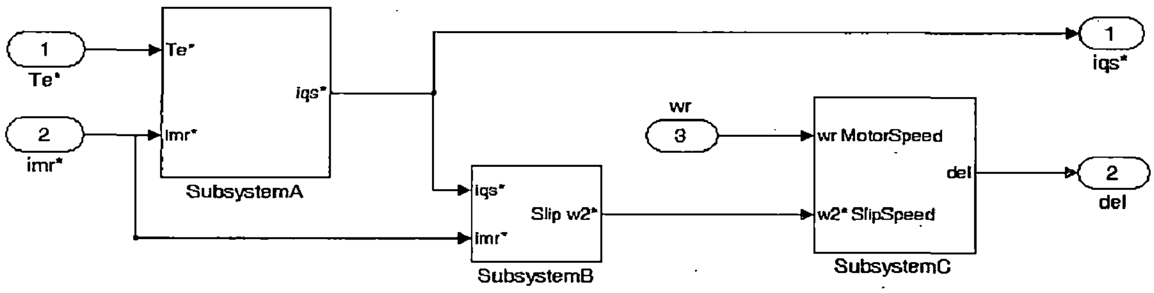


Figure 3.2: Vector Controller

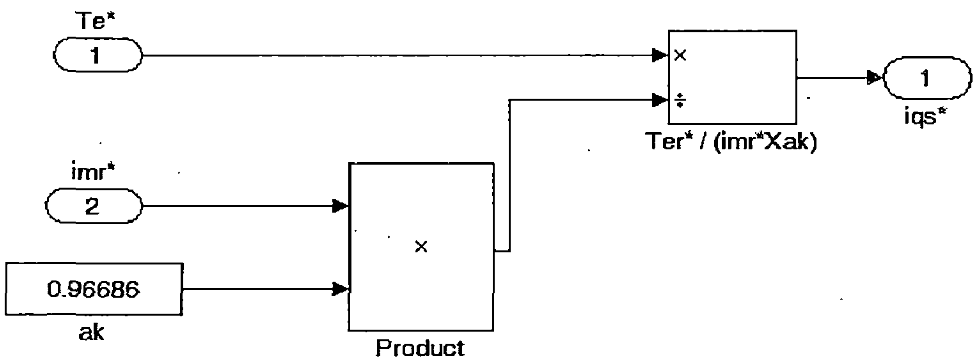


Figure 3.2 (a): Vector Controller- Subsystem (A)

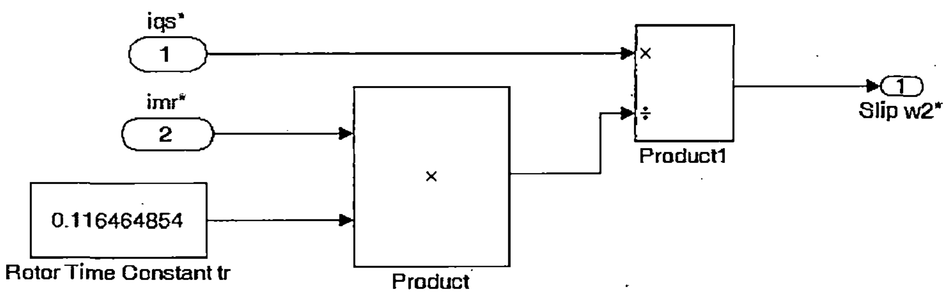


Figure 3.2 (b): Vector Controller- Subsystem (B)

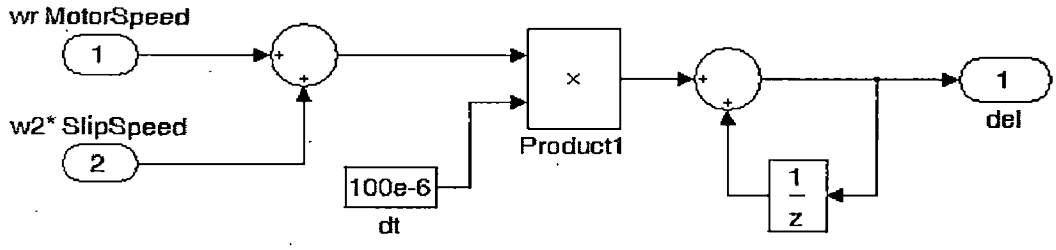


Figure 3.2 (c): Vector Controller- Subsystem C

3.6. Two phase rotating frame to three phase stationary frame converter

The modelling is based on following equations:

$$i_{as}^{*} = -i_{qs}^{*} \sin\psi + i_{ds}^{*} \cos\psi \quad (3.6)$$

$$i_{bs}^{*} = \frac{1}{2}\left[\, i_{ds}^{*}(-\cos\psi + \sqrt{3}\sin\psi) + i_{qs}^{*}(\,\sin\psi + \sqrt{3}\cos\psi)\right] \quad (3.7)$$

$$i_{cs}^{*} = -(\, i_{as}^{*} + i_{bs}^{*}) \quad (3.8)$$

Where i_{ds}^{*} and i_{qs}^{*} refer to the decoupled components of the stator current i_s^{*} in the two phase synchronously rotating reference frame. i_{as}^{*}, i_{bs}^{*} and i_{cs}^{*} refer to the three phase currents in the stationary reference frame.

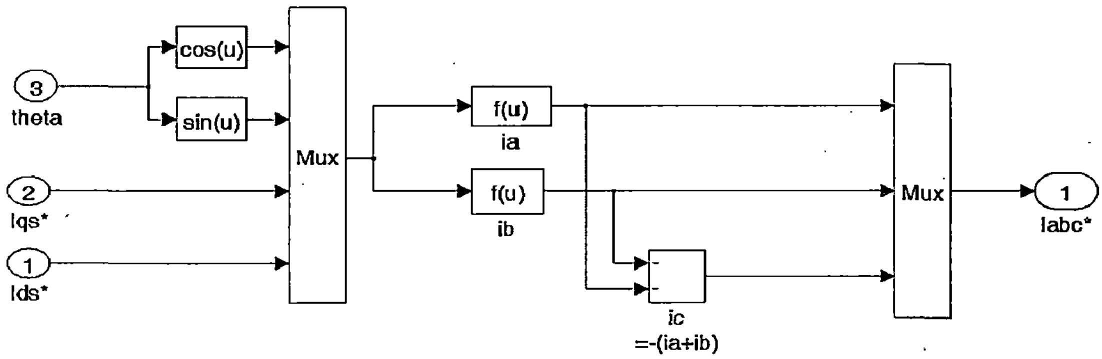


Figure 3.3: Two phase rotating frame to three phase stationary frame converter

3.7. Sinusoidal Pulse Width Modulated (PWM) Current Controller (CC)

In order to ensure that the three phase currents flowing through the motor windings follow the three-phase reference currents a suitable current controller is used. For this purpose a variable frequency inverter is employed along with a current controller. For maintaining the motor currents in such a way that they vary I a definite manner the inverter requires a particular type of ON/OFF (PWM) pattern for its switches (an IGBT with anti-parallel diode). The current controller is responsible for generating desired switching. The current controller compares the winding currents i_{as}, i_{bs}, i_{cs} with the reference currents i_{as}^*, i_{bs}^*, i_{cs}^*. As a result of this comparison a switching pattern is obtained which controls ON/OFF time of inverter switches.

The PWM-CC compares the sensed winding currents i_{as}, i_{bs} and i_{cs} and their reference values i_{as}^*, i_{bs}^* and i_{cs}^* and evaluates the current error. This signal is amplified by a simple gain factor and compared with a high frequency triangular wave carrier signal, with a frequency same as switching frequency f_s of the inverter. Based on the comparison of modulating signal and switching frequency triangular carrier wave, combination of switching functions for three phases are decided. The current is as:

$$i_{ase} = i_{as}^* - i_{as}, \quad i_{bse} = i_{bs}^* - i_{bs}, \quad i_{cse} = i_{cs}^* - i_{cs};$$

$$If\ Ki_{ase} > y(t),\ then\ SF_a = 1, else\ SF_a = 0; \quad (3.9)$$

$$If\ Ki_{bse} > y(t),\ then\ SF_b = 1, else\ SF_b = 0; \quad (3.10)$$

$$If\ Ki_{cse} > y(t),\ then\ SF_c = 1, else\ SF_c = 0; \quad (3.11)$$

Where y(t) refers to the instantaneous value of triangular carrier wave with K being the gain of P controller.

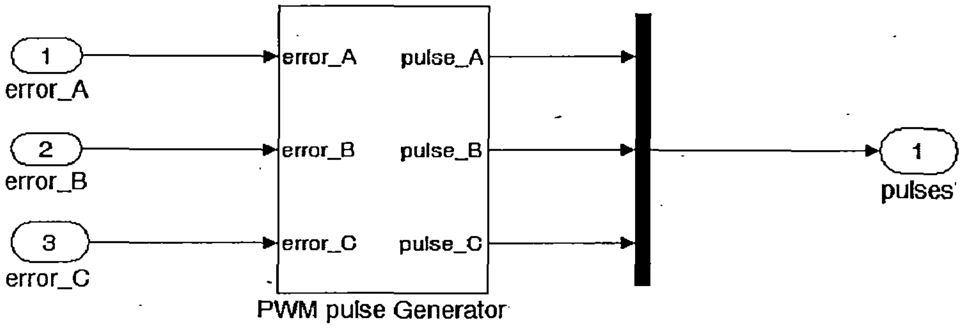
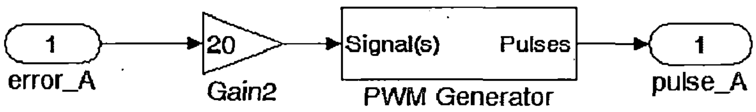


Figure 3.4: PWM Generator



3.8. Current Controlled (CC) Voltage Source Inverter (VSI)

The current controlled voltage source inverter which gets its switching signals from the current controller maintains the motor winding currents in a manner which conforms to the need of vector control. It has a constant voltage source at its input and gives three phase PWM voltage waves at its output. This results in a set of three phase sinusoidal currents of varying voltage and frequency. These in turn depend upon the set speed of the drive and different operating conditions being encountered by the drive such as starting, speed reversal and also smooth running of the drive in any quadrant of torque speed characteristics.

The ac output voltage of VSI is dependent upon the combination of the switching functions SF_a, SF_b and SF_c received from the sinusoidal PWM current controller.

The magnitude of three phase voltages can be mathematically expressed as:

$$v_{as} = \frac{v_{dc}(2SF_a - SF_b - SF_c)}{3} \quad (3.12)$$

$$v_{bs} = \frac{v_{dc}(2SF_b - SF_a - SF_c)}{3} \quad (3.13)$$

$$v_{cs} = \frac{v_{dc}(2SF_c - SF_b - SF_a)}{3} \quad (3.14)$$

Where v_{as}, v_{bs} and v_{cs} are the per phase PWM voltages respectively. SFa, SFb and SFc are the switching functions for phase a, b and c respectively.

3.9. Three phase induction motor

The induction motor is modelled using d-q stationary reference frame. The voltage-current relationship in the stationary frame of the induction motor in terms of d-q variables is expressed as:

$$[v] = [R][i] + [L]p[i] + \omega_r[G][i] \quad (3.15)$$

Where p refers to the differential operator (d/dt) and ω_r is the rotor speed in electrical rad/s.

Current and voltage vectors are given as:

$$[i] = \left[i_{qss} \; i_{dss} \; i_{qrs} \; i_{drs}\right]^t \quad (3.16)$$

$$[v] = \left[v_{qss} \; v_{dss} \; v_{qrs} \; v_{drs}\right]^t \quad (3.17)$$

In the above equations v_{qss}, v_{dss} are the forcing functions impressed across the stator windings. v_{qrs}, v_{rds} are zero in the case of three-phase squirrel cage induction motor.

$$v_{qss} = (v_{bs} - v_{cs})/\sqrt{3} \quad (3.18)$$

$$v_{dss} = v_{as} \quad (3.19)$$

[L], [R] and [G] are referred as the inductance matrix, resistance vector and the rotational inductance matrix respectively. d, q, s and r represent the direct axis, quadrature axis, stator quantities and rotor quantities respectively. $i_{qss}, i_{dss}, i_{qrs}, i_{drs}, v_{qrs}, v_{drs}$ are expressed in stationary reference frame.

The derivative of rotor speed is obtained from the torque balance equation as follows:

$$p\omega_r = \left(\frac{P}{2}\right)\left[\frac{T_e - T_l}{J}\right] \quad (3.20)$$

Tl refers to the load torque on the motor shaft, which includes friction and windage torque opposing the movement of the motor shaft and Te refers to the electromagnetic torque developed by the motor as defined as:

$$T_e = \left(\frac{3}{2}\right)\left(\frac{P}{2}\right) L_m\left(i_{qss}i_{drs} - i_{dss}i_{qrs}\right) \quad (3.21)$$

Where P = number of poles, L_m magnetising inductance.

Chapter 4
Speed Controllers

Each component of VCIMD is modeled by a set of mathematical equations. Such sets of equations when combined together represent the mathematical model of the complete VCIMD. The modeling of various types of speed controllers forms the topic of discussion of this chapter.

Speed Controllers

4.1. Proportional Integral (PI) Controller

The general block diagram of the PI speed controller is as shown in the Fig.4.1: The output of the speed controller at the n^{th} instant is given as:

$$e_{(n)} = \omega_{re(n)} = \omega^*_{r(n)} - \omega_{r(n)} \qquad (4.1)$$

$$\Delta e_{(n)} = \Delta\omega_{re(n)} = \omega_{re(n)} - \omega_{re(n-1)} \qquad (4.2)$$

$$T_{(n)} = T^*_{(n-1)} + K_p\left(\omega_{re(n)} - \omega_{re(n-1)}\right) + K_i\omega_{re(n)} \qquad (4.3)$$

Where K_p and K_i are the proportional and integral gain parameters of the PI speed controller.

$e_{(n)}$ and $\Delta e_{(n)}$ are error and change in error respectively, at n^{th} instant.

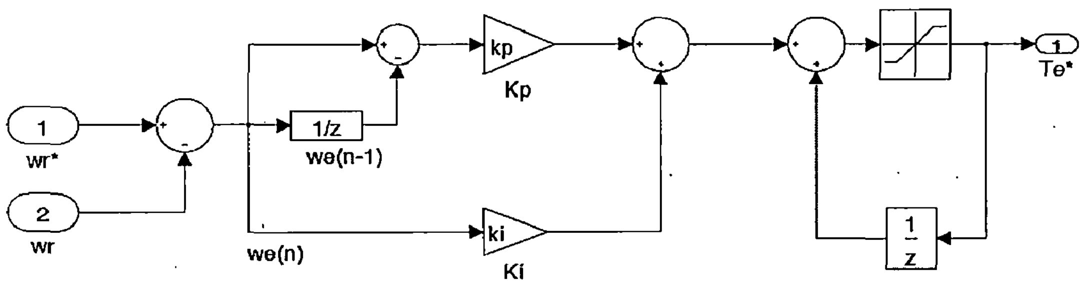


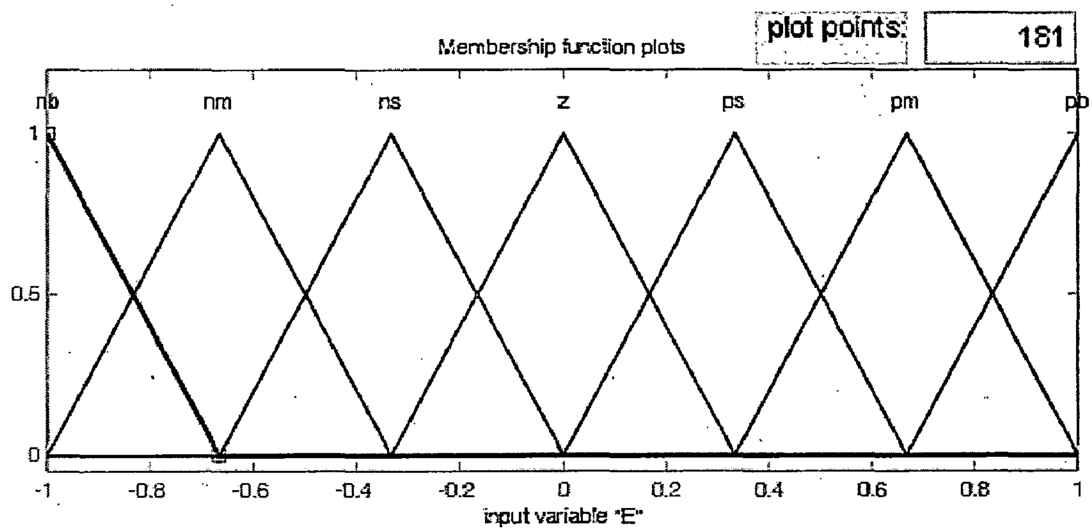
Figure 4.1: PI controller design

4.2. Fuzzy Logic Controller

Different schemes of Fuzzy logic control methods are used widely in induction motor control. The design and synthesis of a fuzzy controller will be presented using fuzzy logic toolbox with MATLAB/Simulink. In this report, fuzzy logic controller employs speed error and change of speed error as inputs, the changes in torque component of current that drives the induction motor is output.

In the first stage, the crisp variables $e_{(n)}$ and $\Delta e_{(n)}$ are converted into fuzzy variables. The fuzzification maps the error, and the error changes to linguistic labels of the fuzzy sets.

Knowledge base involves defining the rules represented as statements governing the relationship between input and output variables in term of membership functions. The control rules are represented as a set if then rules. The fuzzy rules of proposed controller for speed control of induction motor are presented in the following table.



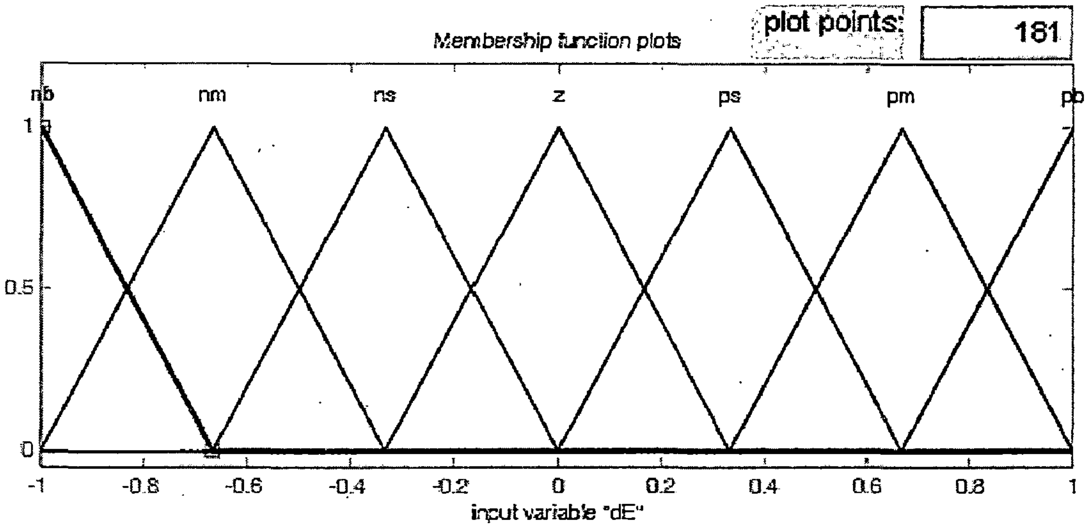


Fig 4.2: Membership Functions for input variables for fuzzy controller

Change in Error	**Error**						
	NB	**NM**	**NS**	**Z**	**PS**	**PM**	**PB**
NB	NB	NB	NB	NB	NM	NS	Z
NM	NB	NB	NM	NM	NS	Z	PS
NS	NB	NM	NS	NS	Z	PS	PM
Z	NB	NM	NS	Z	PS	PM	PB
PS	NM	NS	Z	PS	PS	PM	PB
PM	NS	Z	PS	PM	PM	PB	PB
PB	Z	PS	PM	PB	PB	PB	PB

4.3. Pre-Compensated Fuzzy PI Speed Controller

Scheme:

The basic structure of Fuzzy Pre-compensated PI speed controller is shown in Fig. 4.2 It consists of a pre-compensator block followed by a conventional PI speed controller. The pre-compensator block uses reference speed and the actual speed as inputs. It generates the output signal $\omega^{*\prime}_r$ on the basis of these inputs and fuzzy compensation term as follows:

$$u_{(n)} = F\left[e_{(n)}, \Delta e_{(n)}\right] \qquad (4.4)$$

$$\omega^{*\prime} = \omega^{*} + u_{(n)} \qquad (4.5)$$

where $e_{(n)}$ is speed error, $\Delta e_{(n)}$ is change in error and $u_{(n)}$ is a compensation term which is determined on the basis of fuzzy logic mapping function at an instant n.

The compensation $F\left[e_{(n)}, \Delta e_{(n)}\right]$ term is used to change the reference speed so that the transients are avoided. This term is computed using fuzzy logic control.

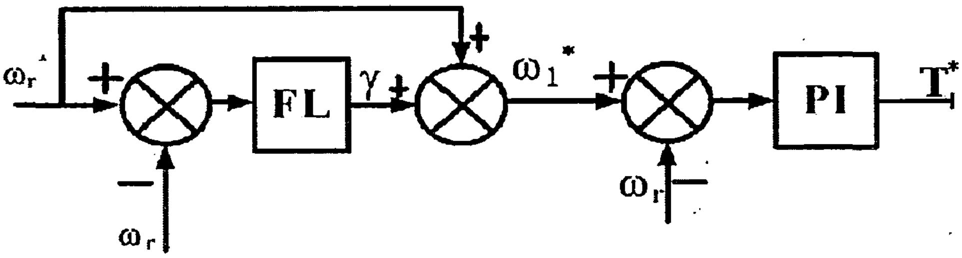


Fig. 4.3: Block schematic of fuzzy pre compensated proportional integral (FPPI) speed controller

The compensated reference signal $\omega^{*\prime}_{r}$ is then applied to the conventional PI controller to obtain the torque reference T^{*} at that instant as follows:

$$e'_{(n)} = \omega^{*\prime}{}_{r(n)} - \omega_{r(n)} \qquad (4.6)$$

$$\Delta e'_{(n)} = e'_{(n)} - e'_{(n-1)} \qquad (4.7)$$

$$T_{(n)} = T^{*}_{(n-1)} + K_p \Delta e'_{(n)} + K_i e'_{(n)} \qquad (4.8)$$

Fuzzy Pre-compensator

The intricate details of the fuzzy logic control used in this work are given below:

The basic structure of fuzzy logic speed controller is given in Fig. The input to the fuzzy logic speed controller is speed error and change in error. These two inputs are normalized to obtain error $e_{(n)}$ and change in error $\Delta e_{(n)}$ in desired range. The compensation term $u_{(n)}$ is then determined by using fuzzy logic control. The process of fuzzification is basically assigning membership functions for various fuzzy subsets on the basis of normalized inputs (error and change in error).

Control Rules

The decision-making in fuzzy logic controller is based on rules written using linguistic terms as follows:

Change in Error	**Error**						
	NB	**NM**	**NS**	**Z**	**PS**	**PM**	**PB**
NB	NB	NB	NM	NM	NS	NS	Z
NM	NB	NM	NM	NS	NS	Z	PS
NS	NM	NM	NS	NS	Z	PS	PS
Z	NM	NS	NS	Z	PS	PS	PM
PS	NS	NS	Z	PS	PS	PS	PM
PM	NS	Z	PS	PS	PM	PM	PB
PB	Z	PS	PS	PM	PM	PM	PB

Following figure gives schematic of Pre-Compensated Fuzzy PI controller, which takes error (E) and change in error (dE) and generates U. This quantity is added to reference speed to generate$\omega_r^{*\prime}$, which then drives the controller in traditional PI fashion.

This concludes the description of the controllers used in this investigation.

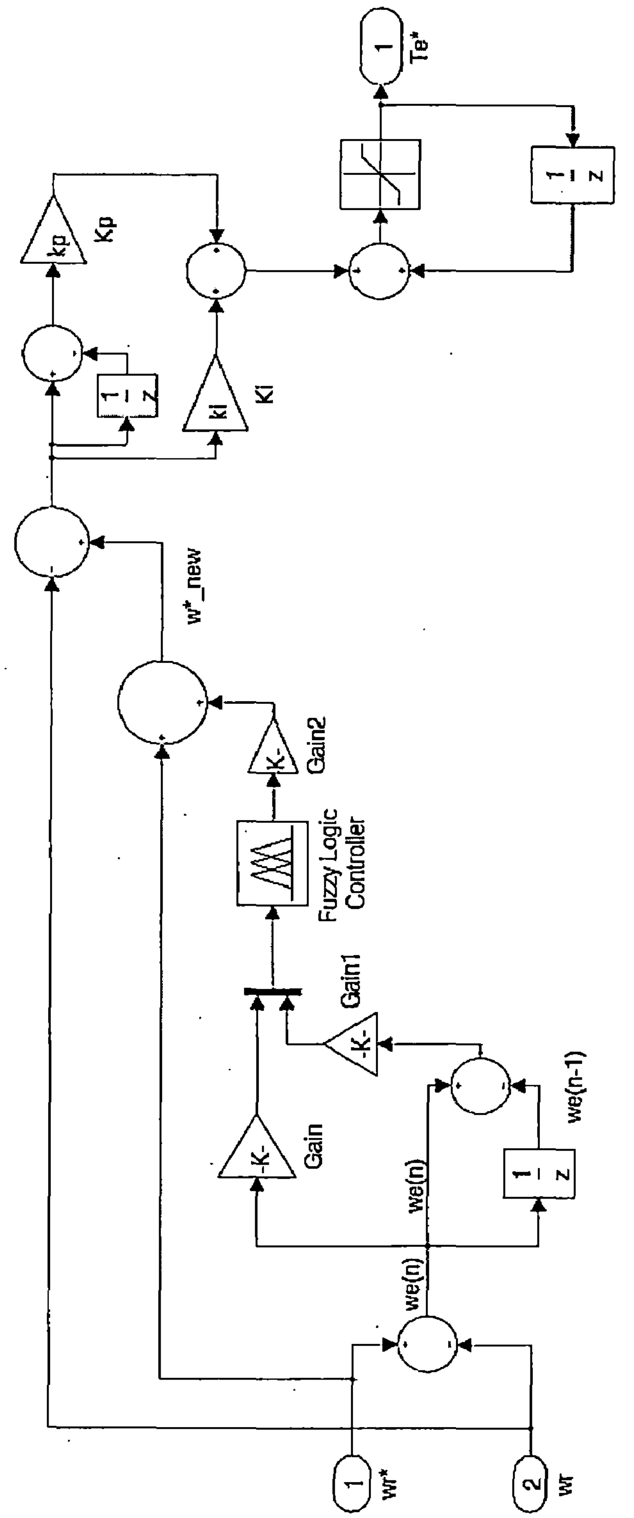


Fig 4.4: Pre-compensated Fuzzy PI Speed Controller

Chapter 5
Results and Discussion

The main objective of the thesis has been aimed towards modeling and simulation of performance of vector controlled induction motor drive under various operating conditions using different speed controllers in MATLAB environment using simulink and power system blockset (PSB) toolboxes. This chapter investigates the performance of 1 HP and 30 HP induction motor drive when operated with Proportional-Integral (PI), pre-compensated fuzzy Proportional Integral and Fuzzy logic controllers.

5.1. Tuning of PI controller

First section of this chapter aims at finding optimal proportional gain, K_p and integral gain, K_i parameters of the PI speed controller for a 1 HP motor by use of trial and error. The procedure adopted is as follows:

Keeping K_i constant, K_p is varied and response of the drive is noted for various operating conditions, viz. starting, speed reversal, load application and load removal or, load perturbation. A value of K_p is thus chosen which gives best dynamic response in all the operating conditions. Table 5.1 compares the results for effect of variation of proportional gain. Fig. 5.1 gives response of the drive for the above mentioned operating conditions and compares the results. After K_p is established, K_i is varied against the fixed K_p obtained earlier and that value of K_i is chosen for which the drive gives best dynamic performance. Table 5.2 compares the results for effect of variation of integral gain. Fig. 5.2 gives response of the drive for the above mentioned operating conditions and compares the results.

Here, the tuning of PI speed controller for 1 HP motor is shown, similar procedure was adopted for tuning 30 HP motor. The parameters of the motors used are listed in the appendix-A

Table 5.1: Effect of Variation of Proportional Gain (Kp)				
Kp	0.09	0.19	0.26	0.35
Ki	0.0018	0.0018	0.0018	0.0018
Starting Dynamics				
Starting time (s)	0.3406	0.1544	0.316	0.3584
% overshoot in speed	226.3055	217.1303	212.6356	0
maximum Starting torque (Nm)	6.8	6.8014	6.75	6.7
Speed Reversal				
reversal time (s)	0.785	0.6605	0.7015	0.6815
% overshoot in speed	226.45	217.0254	212.682	0
% steady state error in speed	0	0	0	0
Load Application				
% dip in speed	199.552	202.1882	203.231	204.525
settling time (s)	1.56	1.305	1.366	1.38
Load Removal				
% rise in speed	220.52	217.899	216.8478	215.525
settling time (s)	1.8548	1.7005	1.7375	1.775

Table 5.2: Effect of Variation of Integral Gain (Ki)				
Kp	0.19	0.19	0.19	0.19
Ki	4.50E-04	1.80E-03	3.00E-03	9.00E-03
Starting Dynamics				
starting time (s)	0.337	0.1544	0.2123	0.2095
% overshoot in speed	209.992	217.1303	218.62	217.749
maximum Starting torque (Nm)	6.3	6.8014	6.8	6.75
% steady state error in speed	0	0	0	0
Speed Reversal				
reversal time (s)	0.7902	0.6605	0.69	0.68
% overshoot in speed	0	217.0254	218.445	217.652
% steady state error in speed	0	0	0	0
Load Application				
% dip in speed	199.52	202.1882	203.025	202.2
settling time (s)	1.36	1.305	1.3172	1.32
Load Removal				
% rise in speed	220.05	217.899	216.689	214.58
settling time (s)	1.78	1.7005	1.7445	1.705

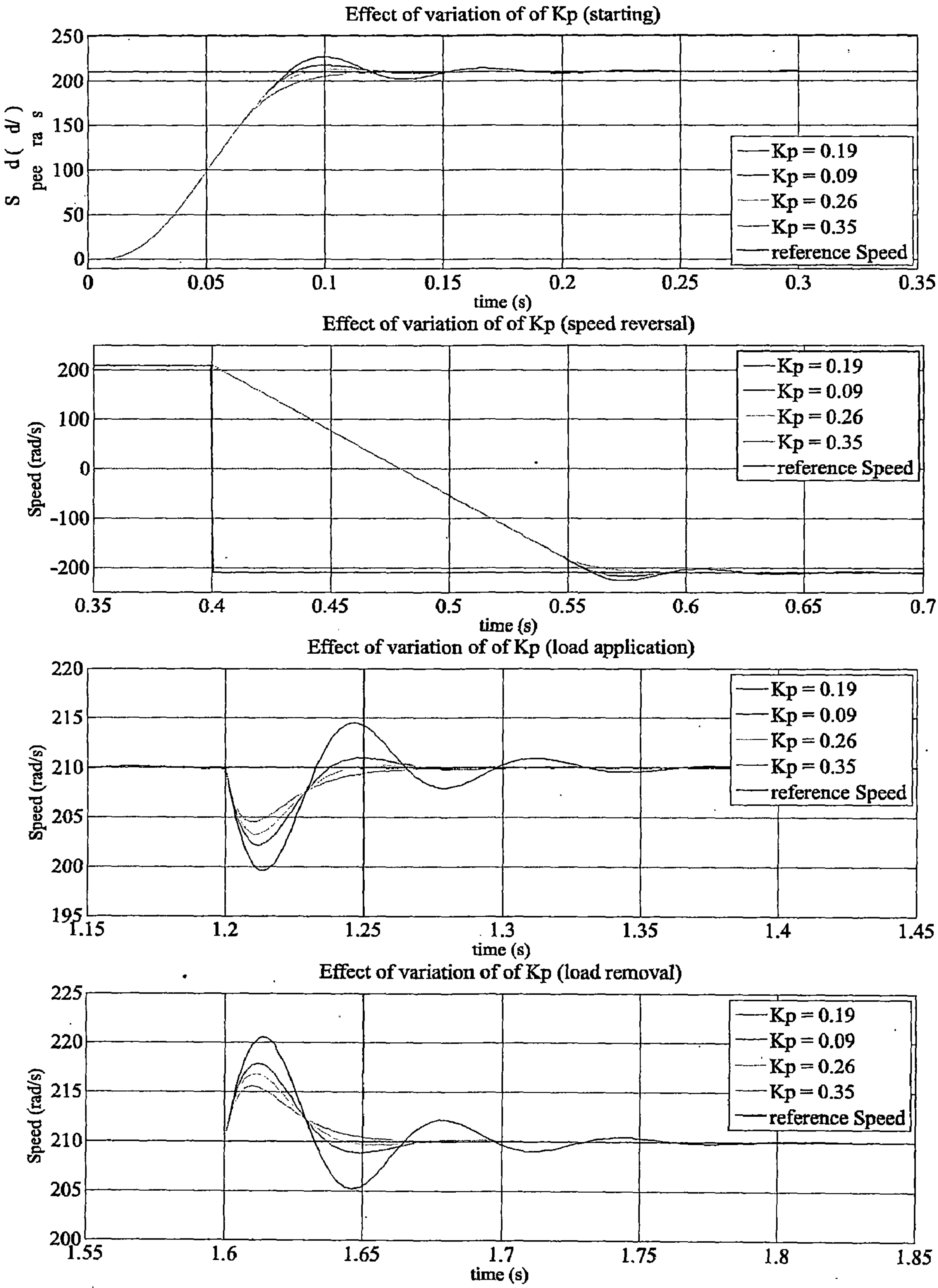


Fig. 5.1 Effect of Proportional Gain (a) Starting (b) Speed Reversal (c) Load Application (d) Load Removal

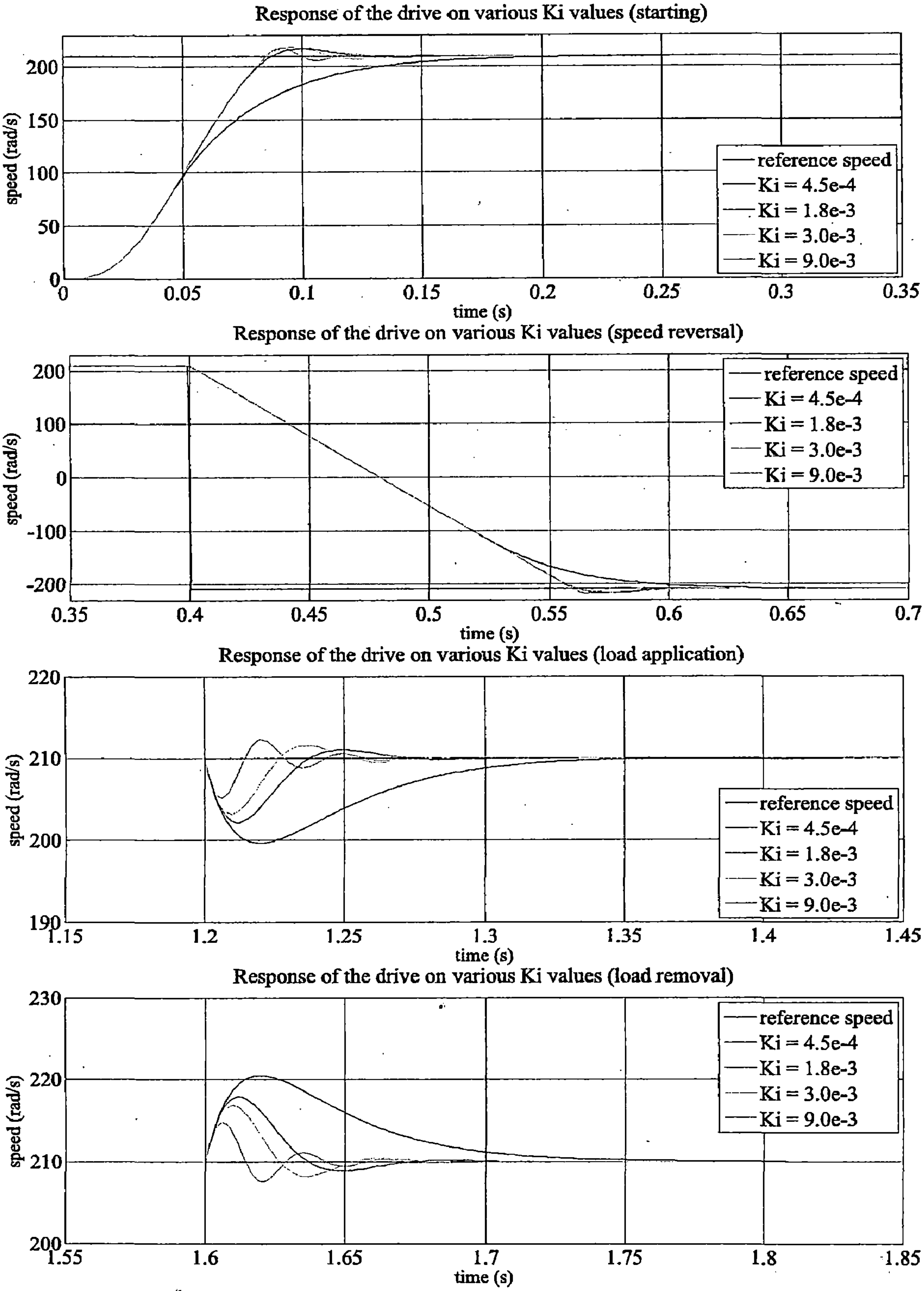


Fig. 5.2 Effect of Integral Gain (a) Starting (b) Speed Reversal (c) Load Application (d) Load Removal.

5.2. Comparison between PI and pre-compensated fuzzy PI speed controller

This section aims at establishing the superiority of pre-compensated fuzzy PI speed controller over conventional PI speed controller. The problem with conventional PI speed controller is with its tuning. It is difficult to tune the controller and many times it becomes cumbersome to use the hit and trial method. Also, the conventional PI speed controller is susceptible to changes in operating conditions. Hence to make the system more reliable and more rugged, we need to have a controller which would take care of tuning mismatches of PI speed controller. As it would be shown in the following discussion, a pre-compensated Fuzzy PI speed controller is the best solution for this purpose.

The mismatch in tuning of conventional PI controller and its removal by the use of pre-compensation technique is illustrated here by taking two extreme cases: low value of proportional gain and large value of integral gain.

Table 5.3 provides comparison between Proportional-Integral (PI) and Pre-compensated Fuzzy PI Controller for low K_p value and Fig. 5.3 plots the same. Table 5.4 provides comparison between Proportional-Integral (PI) and Pre-compensated Fuzzy PI Controller for large K_i value and Fig. 5.4 plots the same.

From these figures it is evident that fuzzy pre-compensation takes care of tuning mismatches of PI controller.

Table 5.3: Comparison between Proportional-Integral (PI) and Pre-compensated Fuzzy PI Controller for low Kp value		
Adjustments to Proportional and Integral Gain	Kp = 0.09	
	Ki = 0.0018	
Controller Type	**Proportional Integral (PI)**	**Precompensated Fuzzy PI**
Starting Dynamics		
Starting time (s)	0.3406	0.21
% overshoot in speed	226.3055	214.024
Maxium Starting torque (Nm)	6.8	6.8
% steady state error in speed	0	0
Speed Reversal		
settling time (s)	0.785	0.703
% overshoot in speed	226.45	213.766
% steady state error in speed	0	0
Load Application		
% dip in speed	199.52	204.82
settling time (s)	1.36	1.329
Load Removal		
% rise in speed	220.05	215.256
settling time (s)	1.78	1.729

Table 5.4: Comparison between Proportional-Integral (PI) and Pre-compensated Fuzzy PI Controller for large Ki value		
Adjustments to Proportional and Integral Gain	Kp = 0.19 Ki = 0.009	
Controller Type	**Proportional Integral (PI)**	**Precompensated Fuzzy PI**
Starting Dynamics		
Starting time (s)	0.2095	0.13
% overshoot in speed	217.749	0
Maxium Starting torque (Nm)	6.75	6.6
% steady state error in speed	0	0
Speed Reversal		
settling time (s)	0.68	0.638
% overshoot in speed	217.652	0
% steady state error in speed	0	0
Load Application		
% dip in speed	202.2	208.45
settling time (s)	1.32	1.28
Load Removal		
% rise in speed	214.58	211.56
settling time (s)	1.705	1.6396

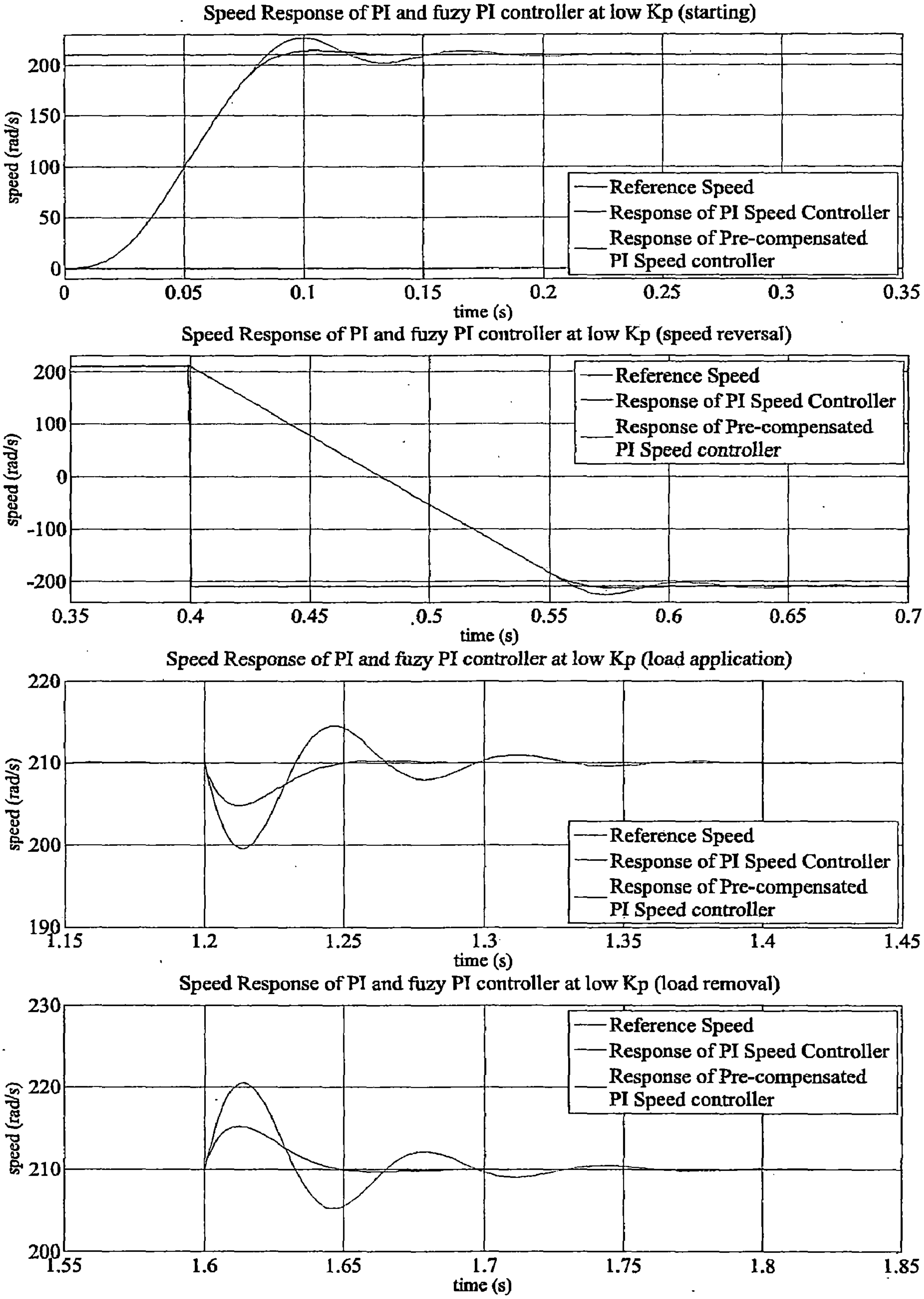


Fig. 5.3 Comparisons between conventional PI and Pre-compensated Fuzzy PI speed controller for various operating conditions (a) Starting (b) Speed Reversal (c) Load Application (d) Load Removal at low value of proportional gain Kp

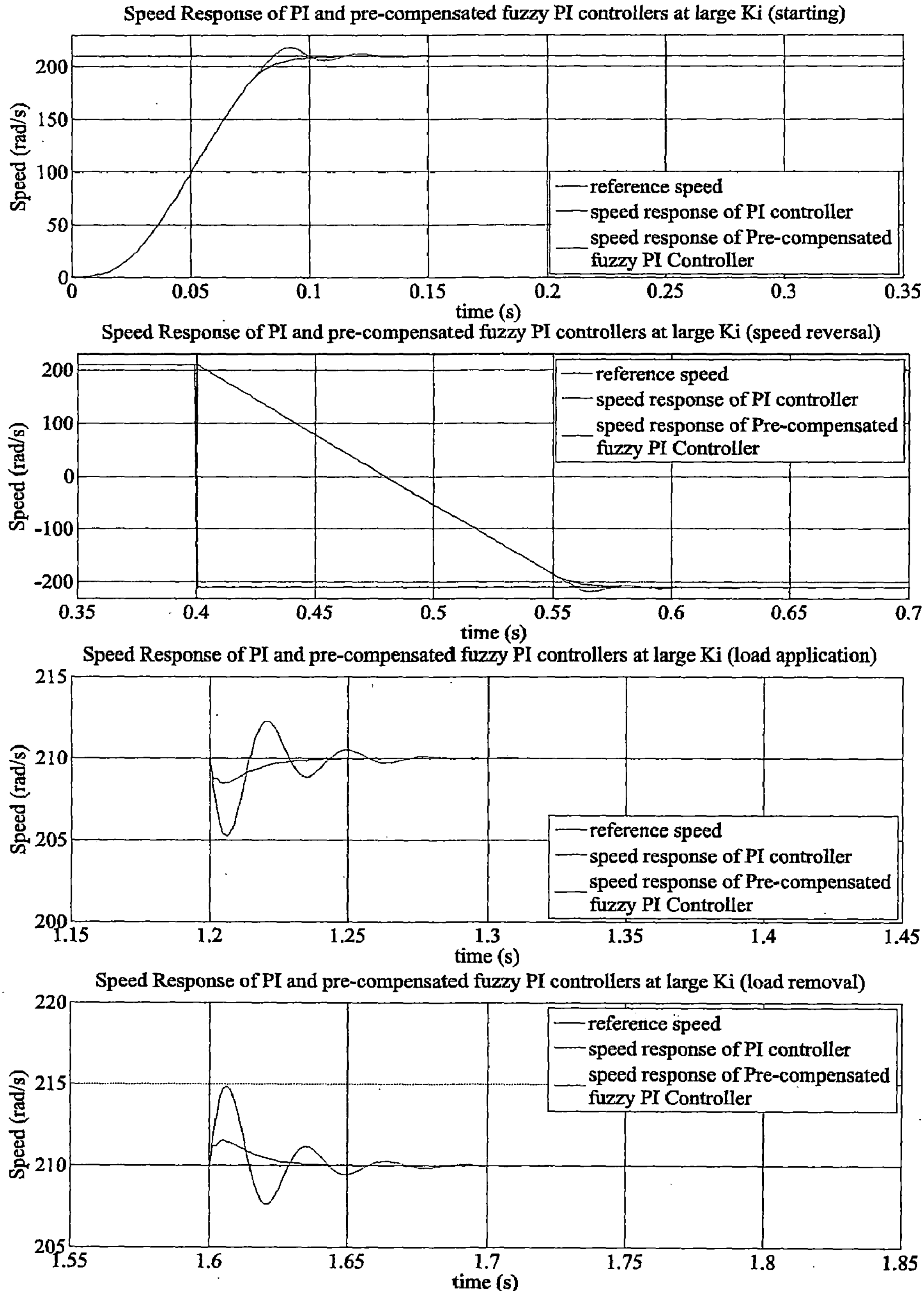


Fig. 5.4 Comparisons between conventional PI and Pre-compensated Fuzzy PI speed controller for various operating conditions (a) Starting (b) Speed Reversal (c) Load Application (d) Load Removal at large value of integral gain Ki

5.3. Comparative study among different speed controllers

5.3.1. Starting Dynamics

Three-phase squirrel cage induction motors are fed from a controlled voltage and frequency source. The motor is started at low frequency decided by the controller and finally runs at the steady state condition at the set reference value of speed. The reference speed is presently set at 210 electrical rad /sec with a torque limit set at twice the rated value. The starting current is also inherently limited when the motor builds up the required starting torque to reach the set reference speed. When the speed error reduces to almost zero rad/secs, the winding current also reduces to the no load value and the developed torque becomes equal to load torque as observed in the starting response shown in the following figures:

with conventional PI speed controller: Fig.5.5 (a) for 1HP drive and Fig. 5.8 (a) for 30 HP drive,
with fuzzy logic speed controller: Fig. 5.6 (a) and Fig. 5.9 (a) for 30 HP drive,
with pre-compensated fuzzy PI speed controller: Fig. 5.7 (a) for 1HP drive and Fig. 5.10 (a) for 30 HP drive.

5.3.2. Reversal Dynamics

While the motor is running at a steady speed of 210 rad/sec, the reference speed is changed from 210.0 rad/sec to -210.0 rad/sec. In response to this change, the controller first reduces the frequency of the stator currents demonstrating braking followed by the phase reversal for starting the motor in the reverse direction. Since just before and after the reversal phenomenon the drive is in the same dynamic state (no load condition), the steady values of the inverter currents are observed to be the same both in magnitude and frequency in the either direction of rotation. However the phase sequence of currents in the two directions is different. Speed reversal response is shown in the following figures:

with conventional PI speed controller: Fig.5.5 (b) for 1HP drive and Fig. 5.8 (b) for 30 HP drive,
with fuzzy logic speed controller: Fig. 5.6 (b) and Fig. 5.9 (b) for 30 HP drive,
with pre-compensated fuzzy PI speed controller: Fig. 5.7 (b) for 1HP drive and Fig. 5.10 (b) for 30 HP drive.

5.3.3. Load Perturbation

When the motor is running at a steady state speed of 210.0 rad/sec, a load torque equal to the rated torque of the motor is applied on its shaft. Sudden application of the load causes an instantaneous fall in the speed of the motor. In response to the drop in speed, the output of the speed controller responds by increasing the reference torque value. Therefore, the developed electromagnetic torque increases, causing the motor speed to settle to the reference value with the increased winding currents.

When the motor is operating at steady state on load, sudden load is removed, which causes aovershoot in the rotor speed. Because of this overshoot the input to the speed controller becomes negative and consequently the output of the PI speed controller (T*) reduces. The control structure results in change in electromagnetic torque developed by the motor. This causes reduction in the rotor speed and hence the speed decreases and settles to the reference value. After the removal of load, the stator current also reduces to the no load value. Load perturbation response is shown in the following figures:

with conventional PI speed controller: Fig.5.5 (c) for 1Hp drive and Fig. 5.8 (c) for 30 HP drive,
with fuzzy logic speed controller: Fig. 5.6 (c) and Fig. 5.9 (c) for 30 HP drive,
with pre-compensated fuzzy PI speed controller: Fig. 5.7 (c) for 1HP drive and Fig. 5.10 (c) for 30 HP drive.

Table 5.5: Starting Dynamics

	Type of Controllers		
	Proportional Integral (PI)	**Fuzzy Logic**	**Precompensated Fuzzy PI**
	1 Hp Motor (Unloaded)		
starting time (s)	0.1665	0.0865	0.1665
% overshoot in speed	3.395380952	0.073809524	0.883571429
Maxium Starting torque (Nm)	6.8014	6.587	6.6
% steady state error in speed	0	0	0
	30Hp Motor (Unloaded)		
starting time (s)	0.52	0.96	0.478
% overshoot in speed	4.978571429	1.70952381	3.227333333
Maxium Starting torque (Nm)	198	194.4	195
% steady state error in speed	0	0	0

Table 5.6: Speed Reversal Dynamics			
	Type of Controllers		
	Proportional Integral (PI)	**Fuzzy Logic**	**Precompensated Fuzzy PI**
	1 Hp Motor (Unloaded)		
settling time (s)	0.235	0.164	0.245
% overshoot in speed	3.345428571	0	0.905904762
% steady state error in speed	0	0	0
	30Hp Motor (Unloaded)		
settling time (s)	0.725	1.048	0.72
% overshoot in speed	4.975285714	0	2.683
% steady state error in speed	0	0	0

Table 5.7: Load Perturbation			
Type of Controller	**Proportional Integral (PI)**	**Fuzzy Logic**	**Precompensated Fuzzy PI**
	1 Hp Motor		
Load Application			
% dip in speed	3.719904762	2.838095238	1.886666667
settling time (s)	0.135	0.052	0.135
% steady state error in speed	0	2.838095238	0
Load Removal			
% overshoot in speed	3.761428571	0	1.924761905
settling time (s)	0.11	0.029	0.11
	30 Hp Motor		
Load Application			
% dip in speed	4.583714286	5.619047619	3.021761905
settling time (s)	0.2604	0.2482	0.215
% steady state error in speed	0	5.619047619	0
Load Removal			
% overshoot in speed	4.674761905	0	3.092857143
settling time (s)	0.268	0.225	0.222

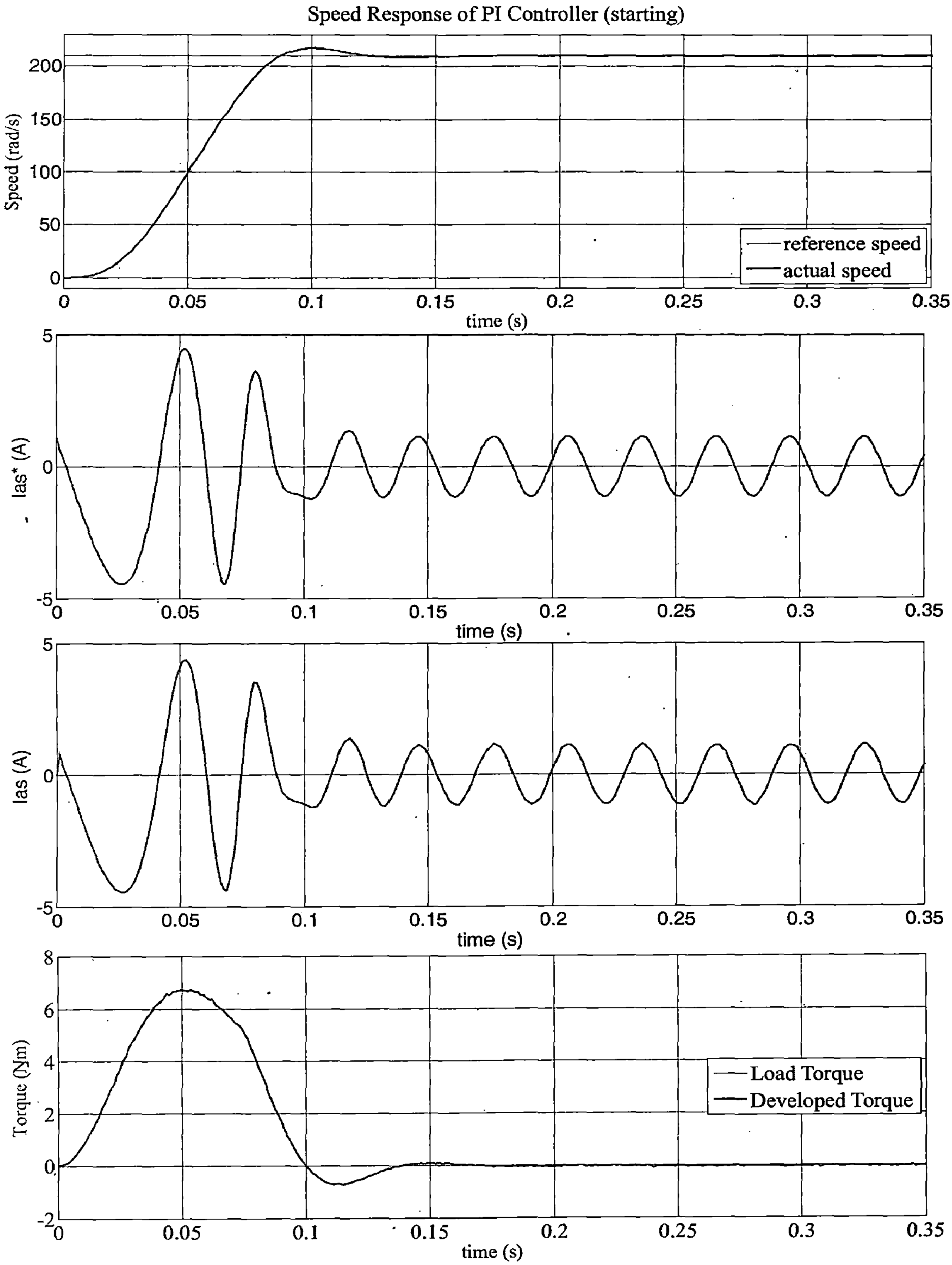


Fig. 5.5 (a) Starting response of 1HP drive with PI Speed controller

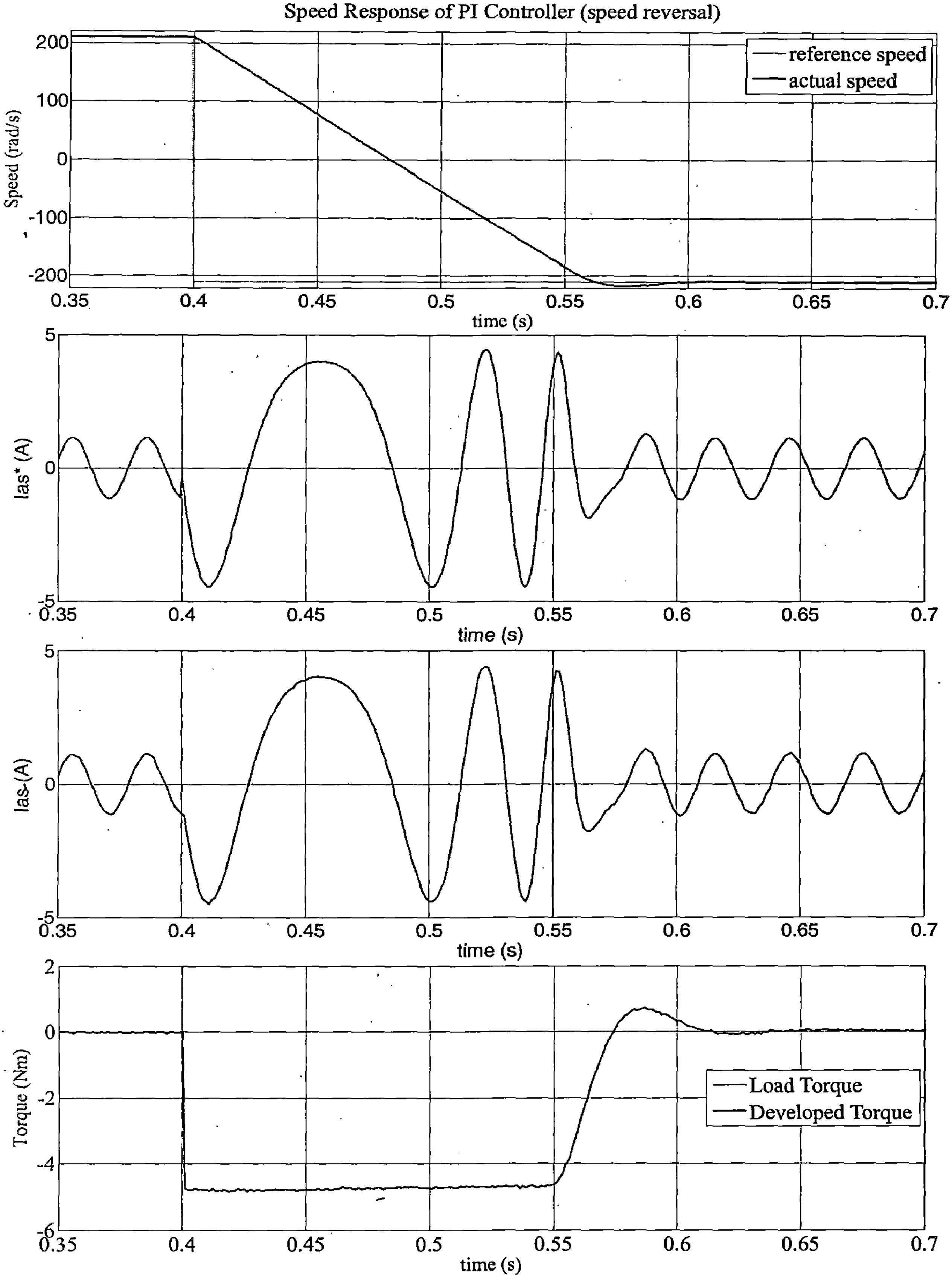


Fig. 5.5 (b) Speed reversal response of 1Hp drive with PI Speed controller

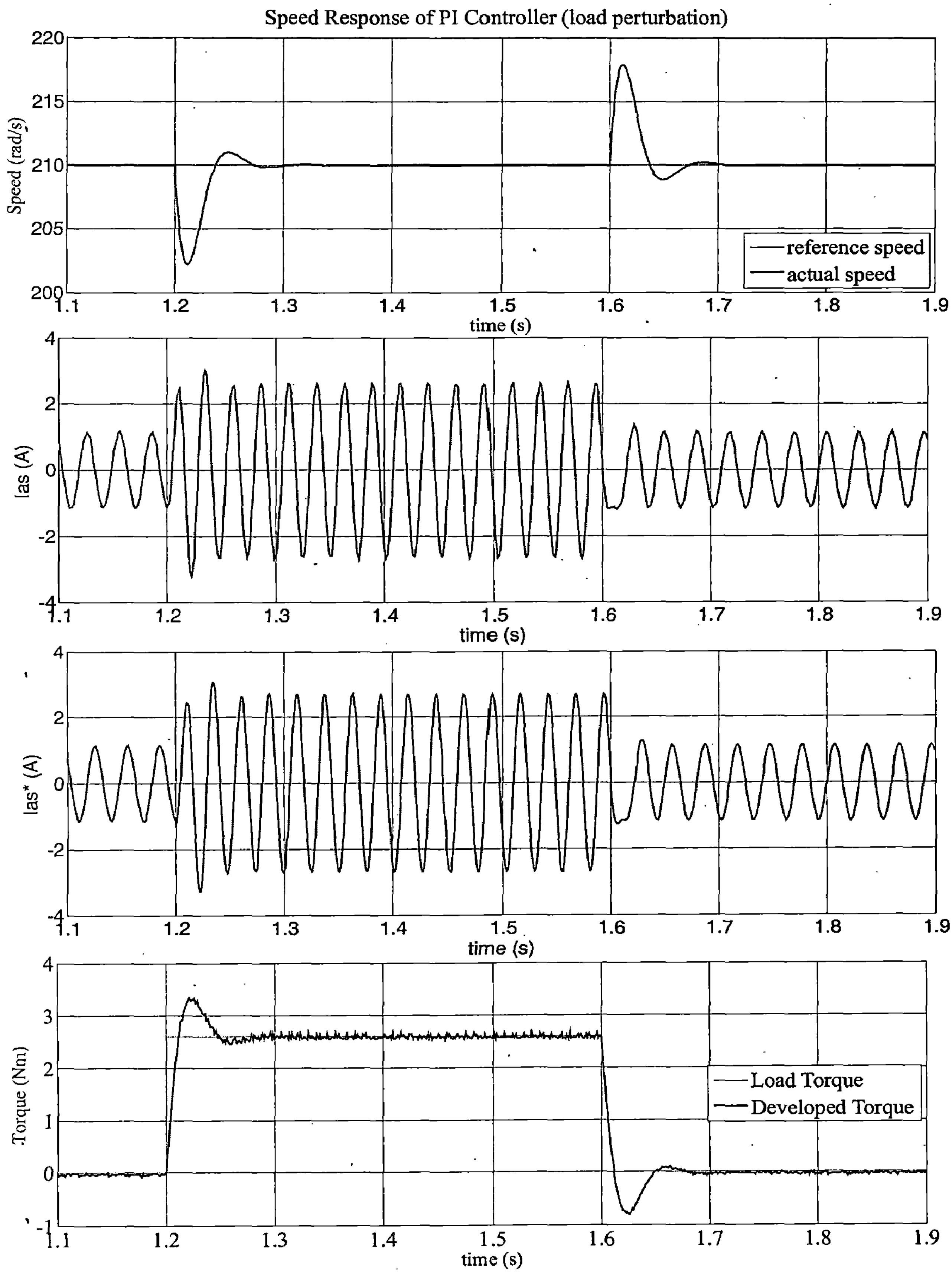


Fig. 5.5 (c) load perturbation response of 1Hp drive with PI Speed controller

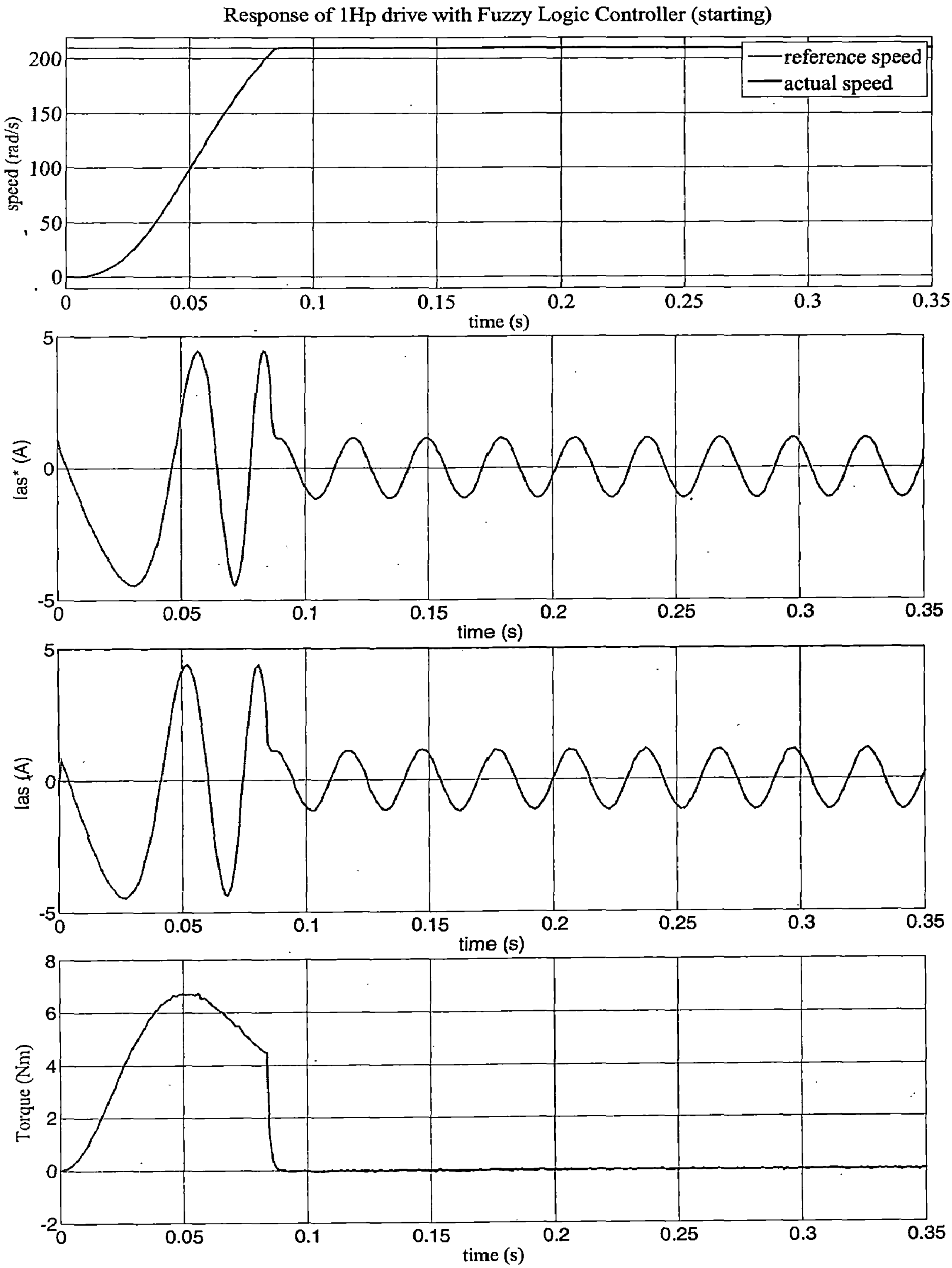


Fig. 5.6 (a) Starting response of 1Hp drive with fuzzy logic speed controller

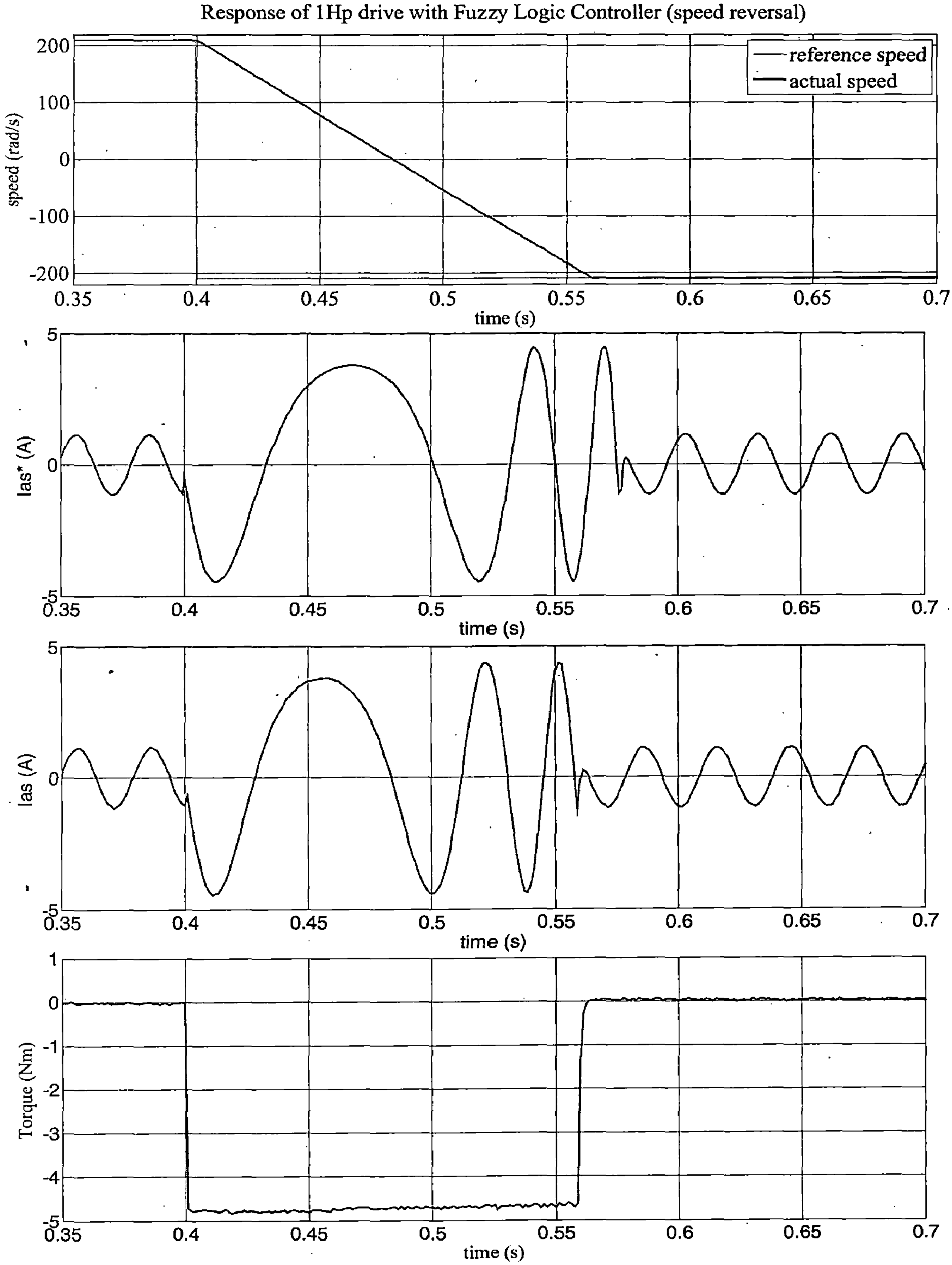


Fig. 5.6 (b) Speed reversal response of 1Hp drive with fuzzy logic speed controller

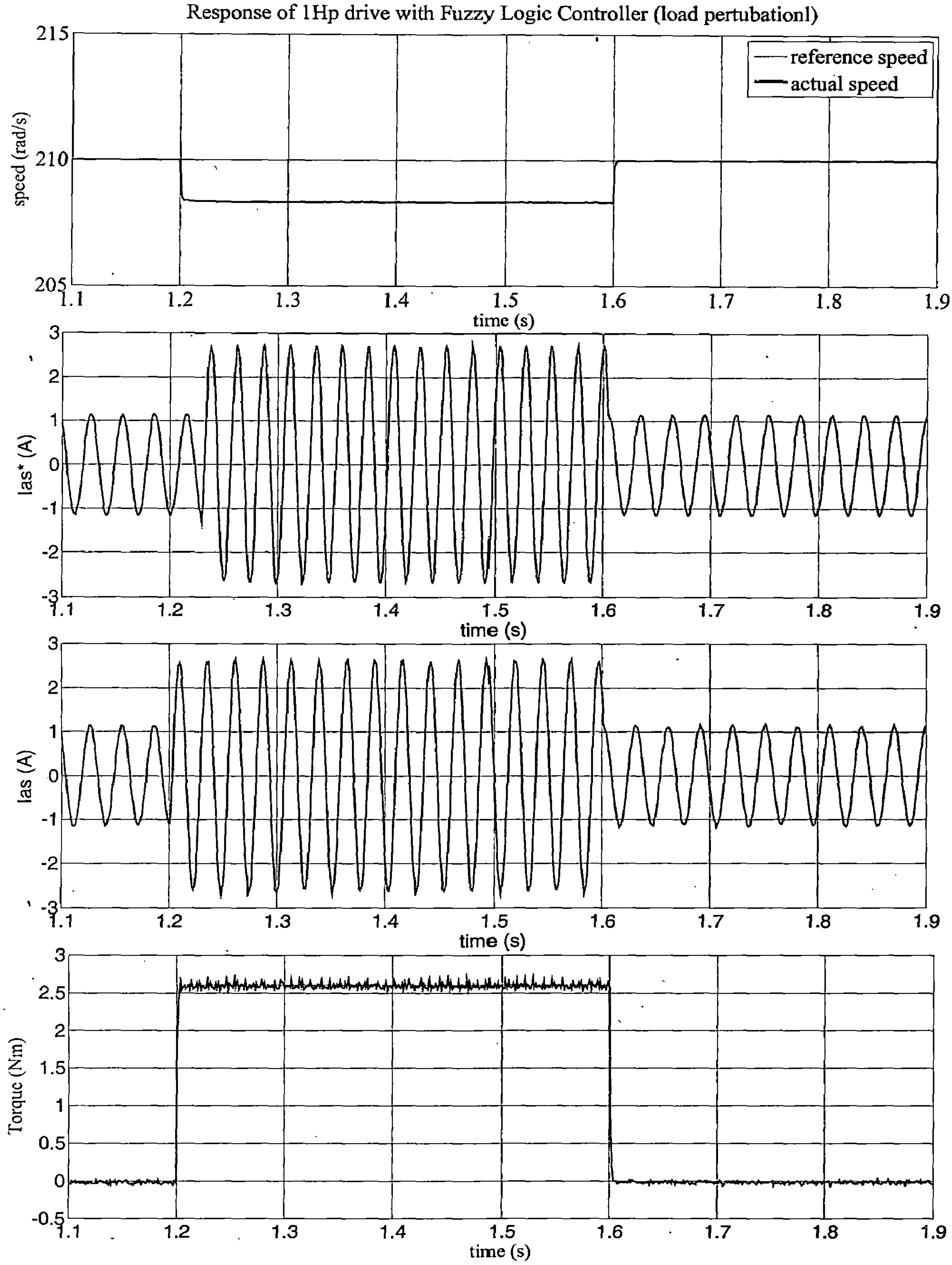


Fig. 5.6 (c) load perturbation response of 1Hp drive with fuzzy logic speed controller

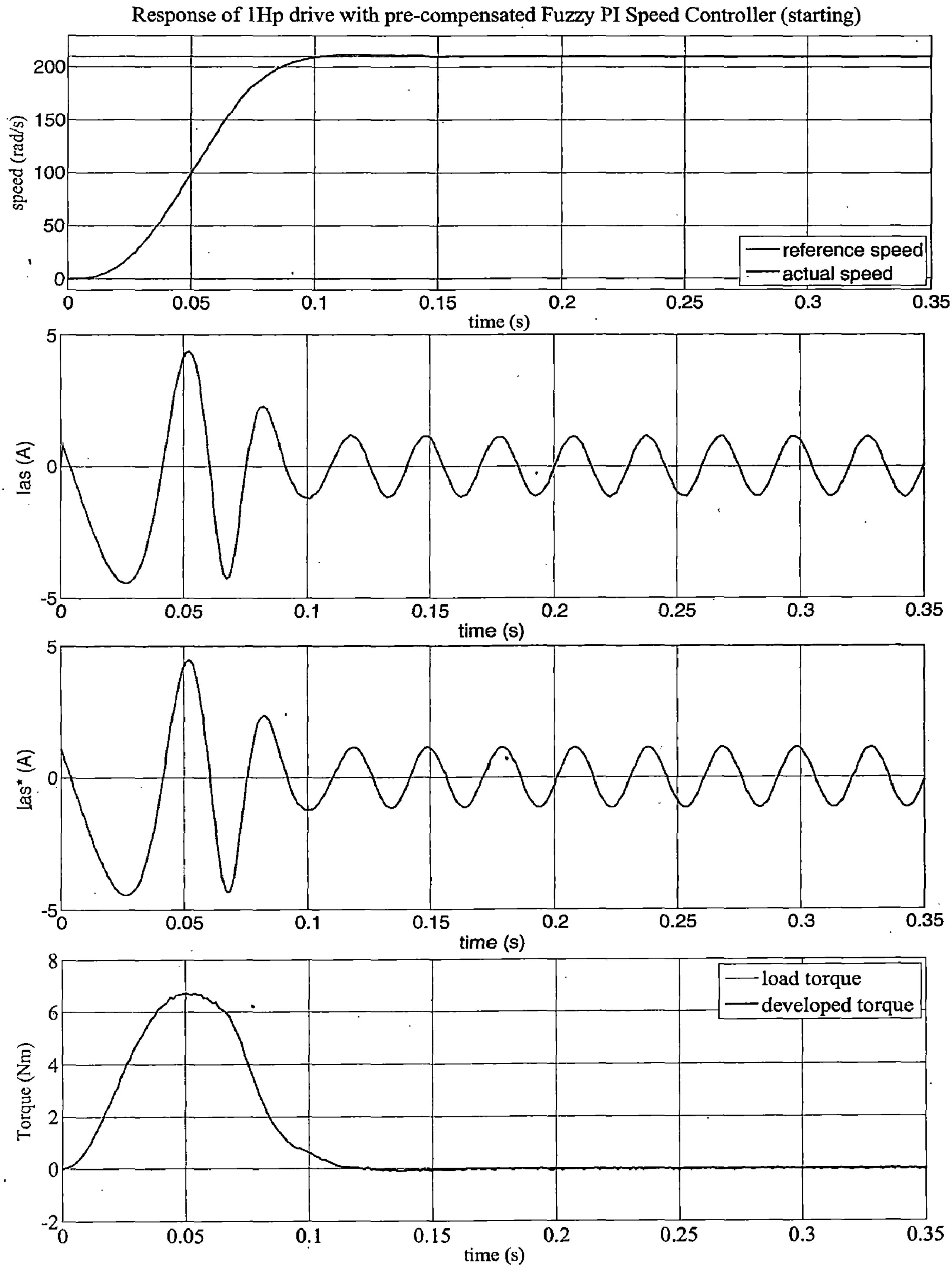


Fig. 5.7 (a) Starting response of 1Hp drive with pre-compensated fuzzy PI speed controller

Response of 1Hp drive with pre-compensated Fuzzy PI Speed Controller (speed reversal)

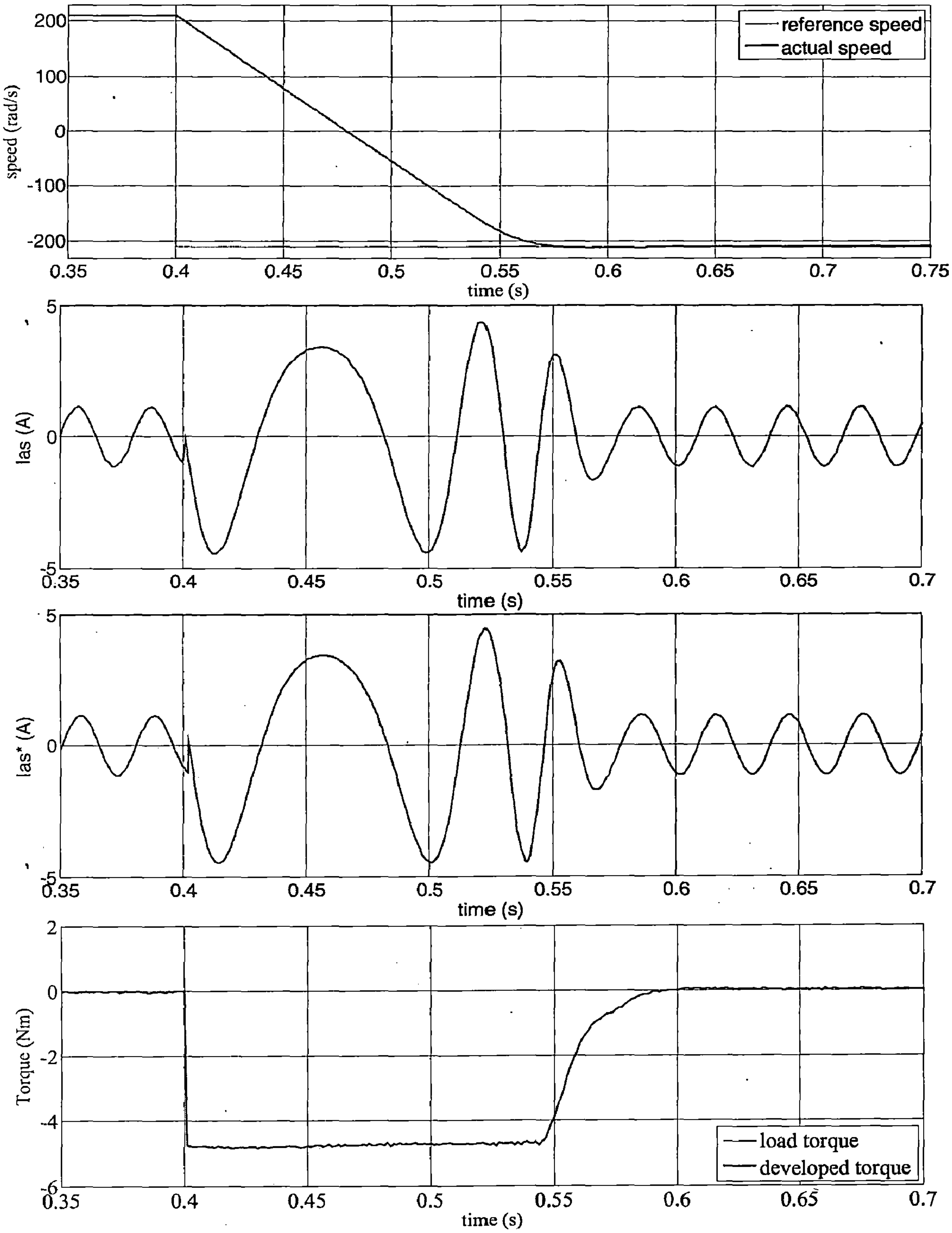


Fig. 5.7 (b) Speed reversal response of 1Hp drive with pre-compensated fuzzy PI speed controller

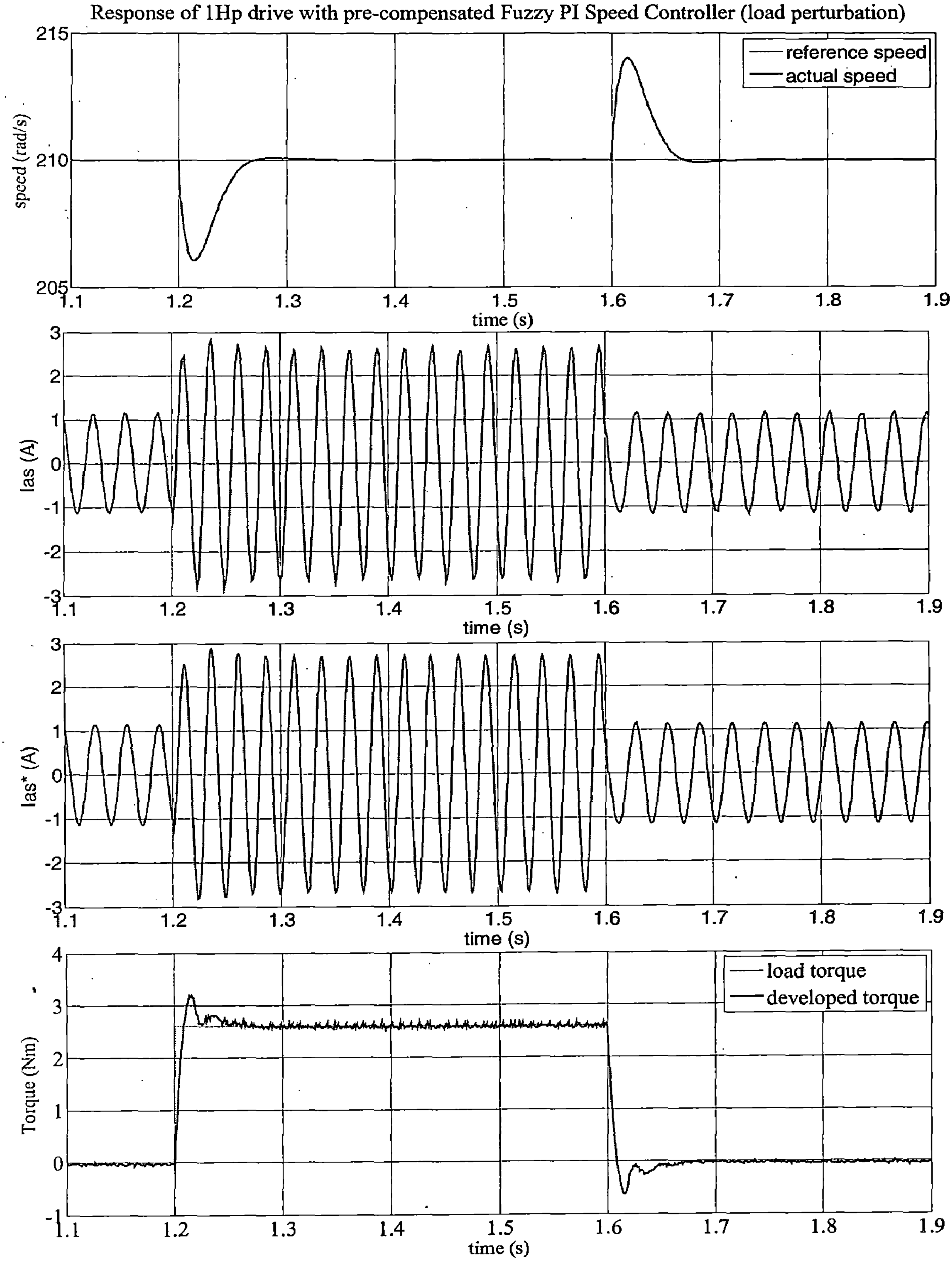


Fig. 5.7 (c) load perturbation response of 1Hp drive with pre-compensated fuzzy PI speed controller

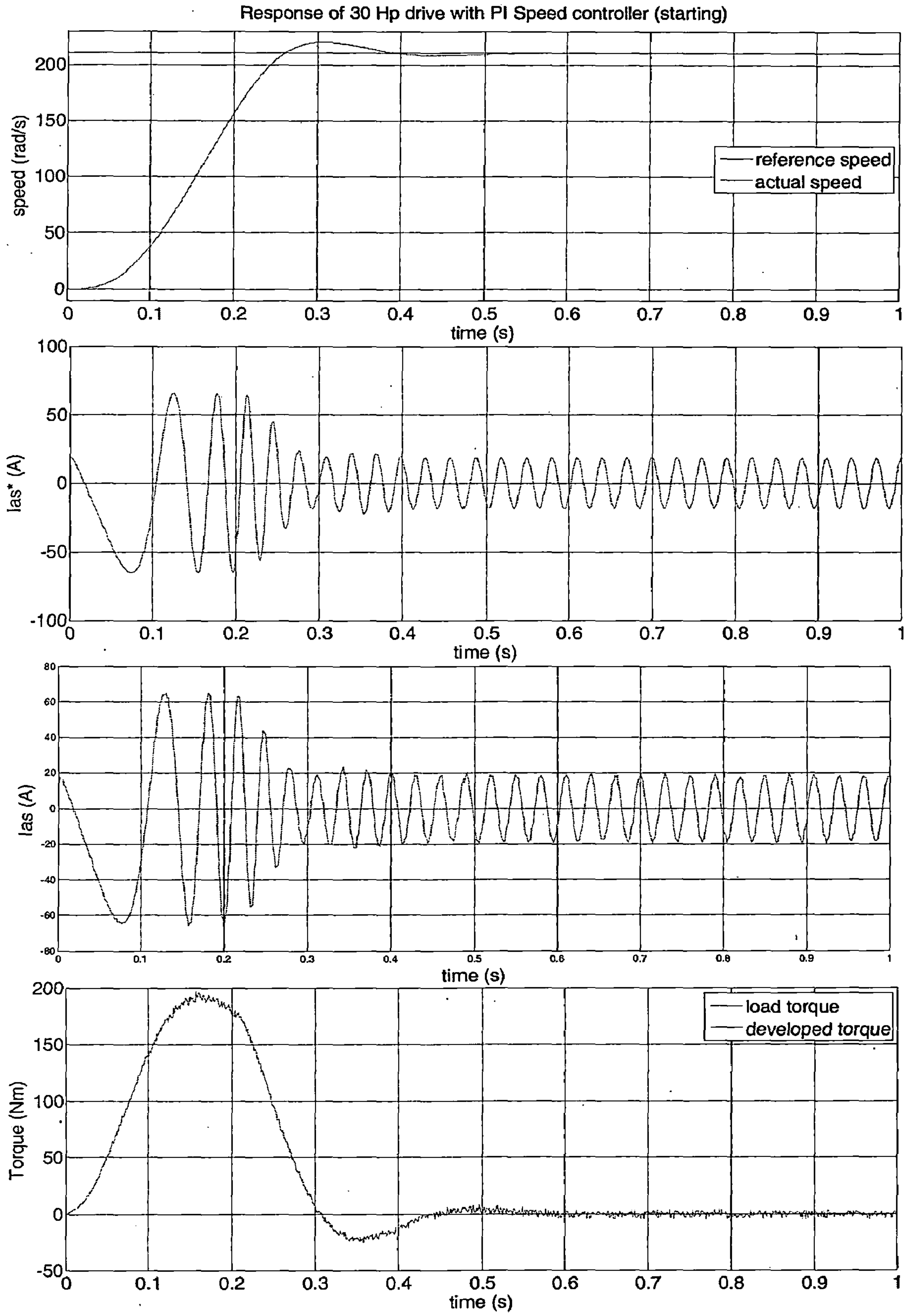


Fig. 5.8 (a) Starting response of 30 Hp drive with PI Speed controller

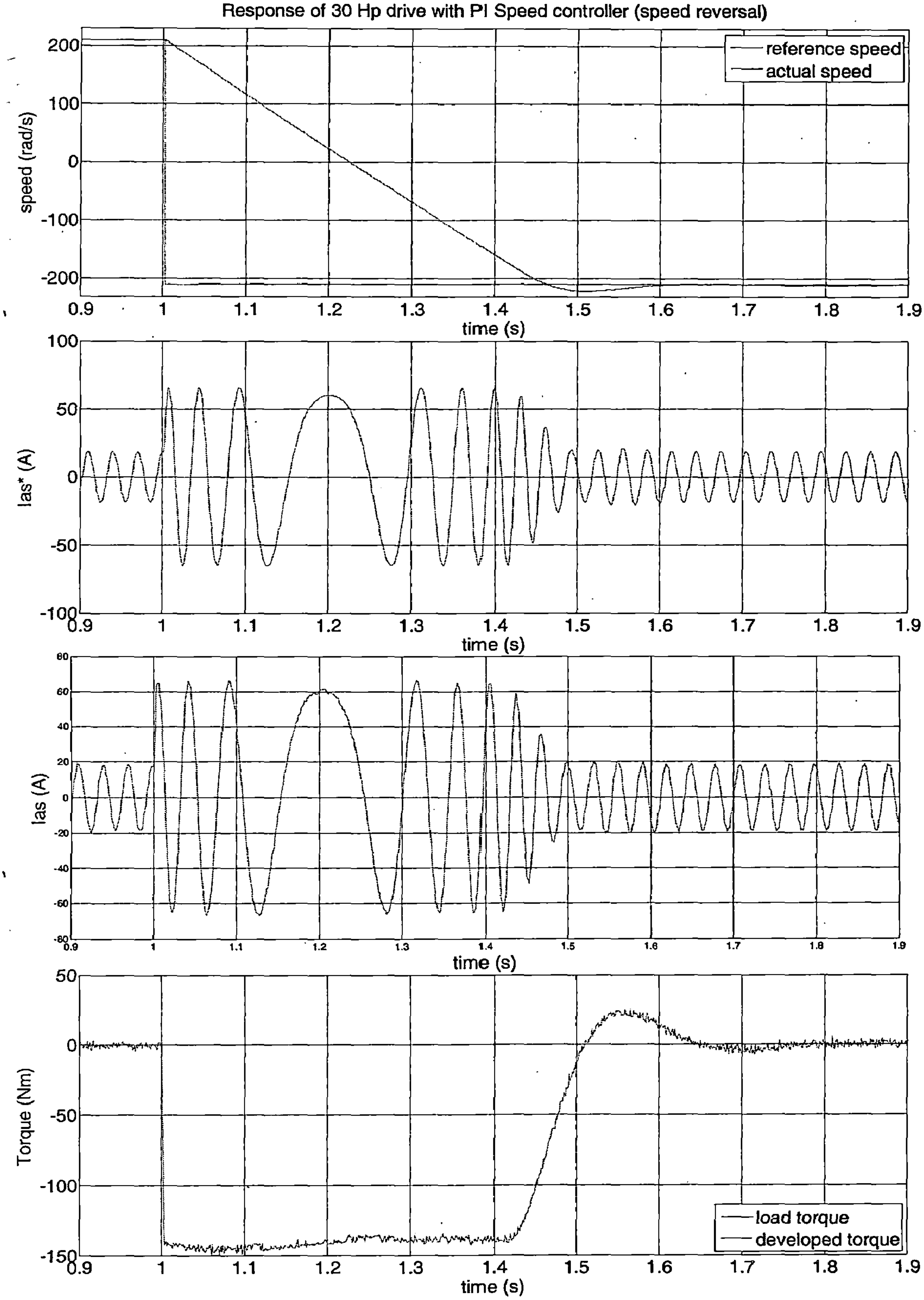


Fig. 5.8 (b) Speed reversal response of 30 Hp drive with PI Speed controller

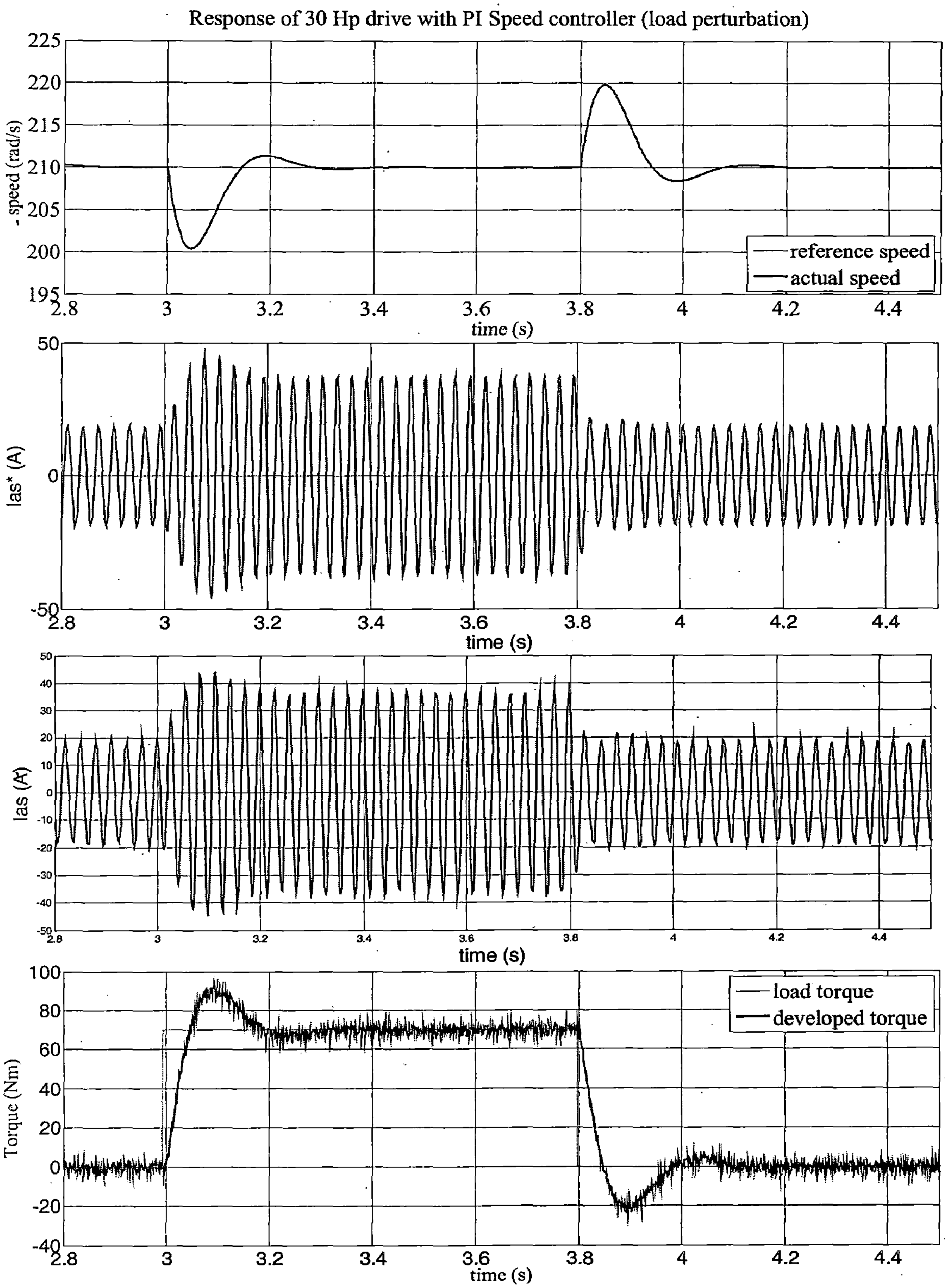


Fig. 5.8 (c) load perturbation response of 30Hp drive with PI Speed controller

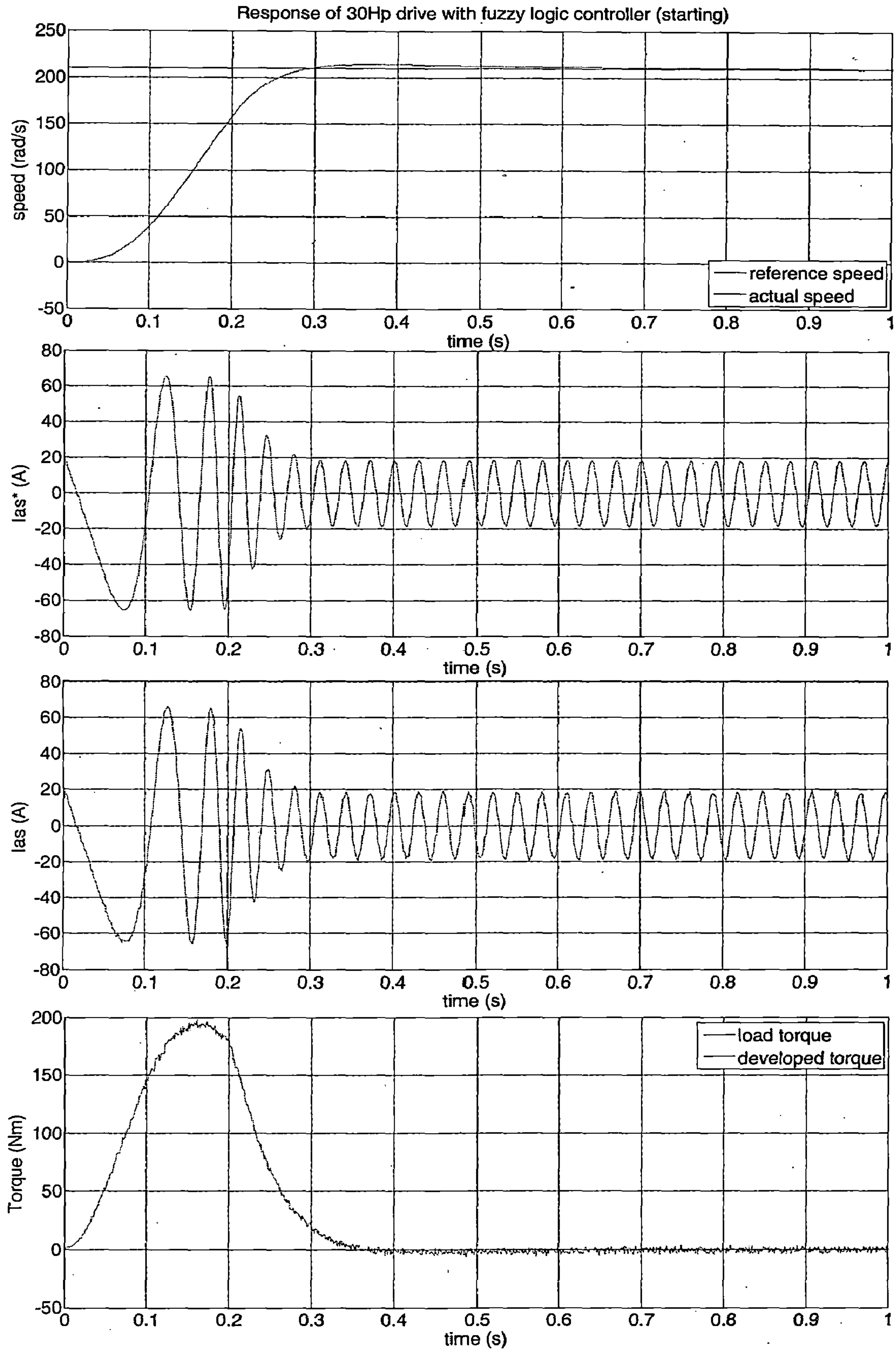


Fig. 5.9 (a) Starting response of 30 Hp drive with fuzzy logic speed controller

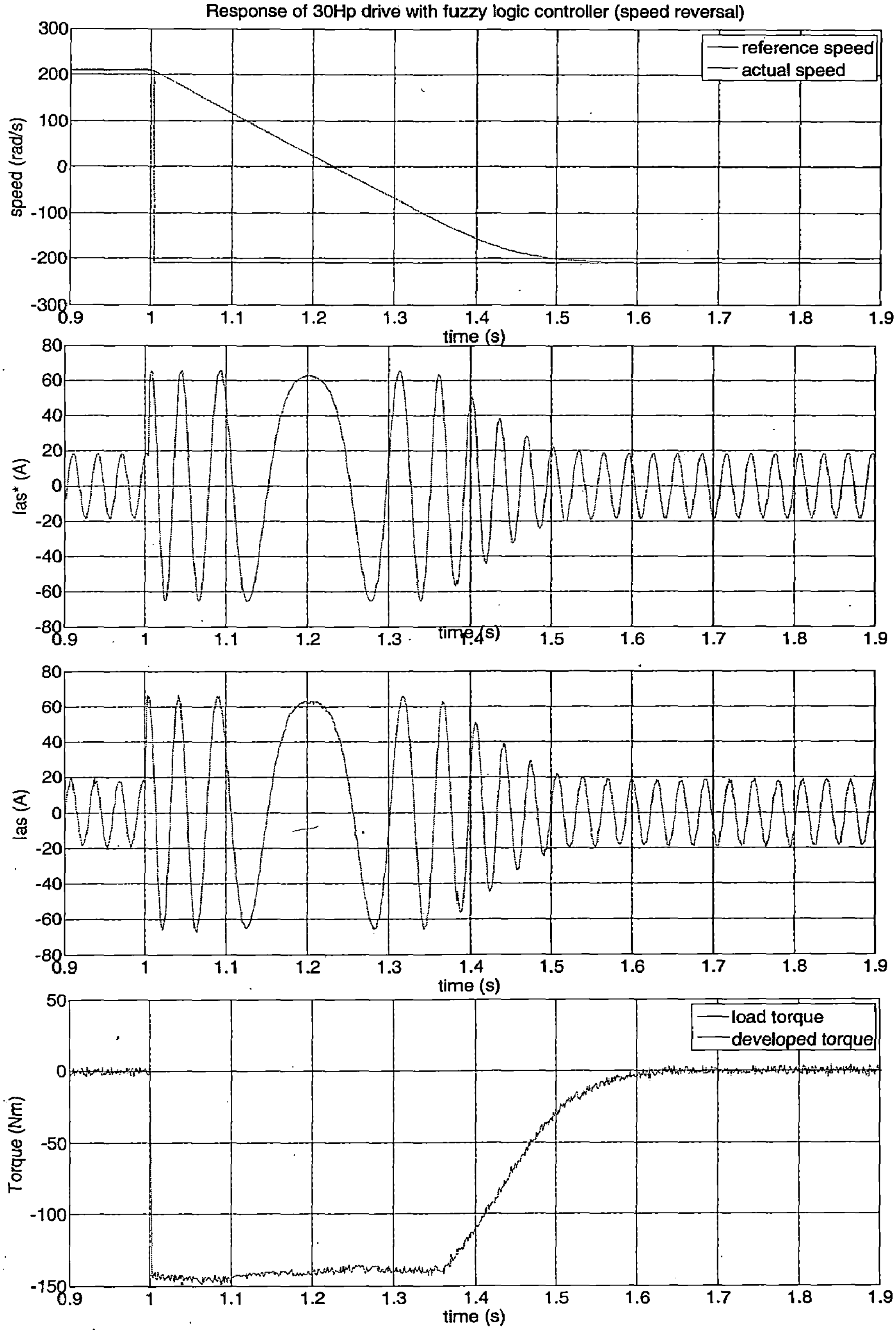


Fig. 5.9 (b) Speed reversal response of 30 Hp drive with fuzzy logic speed controller

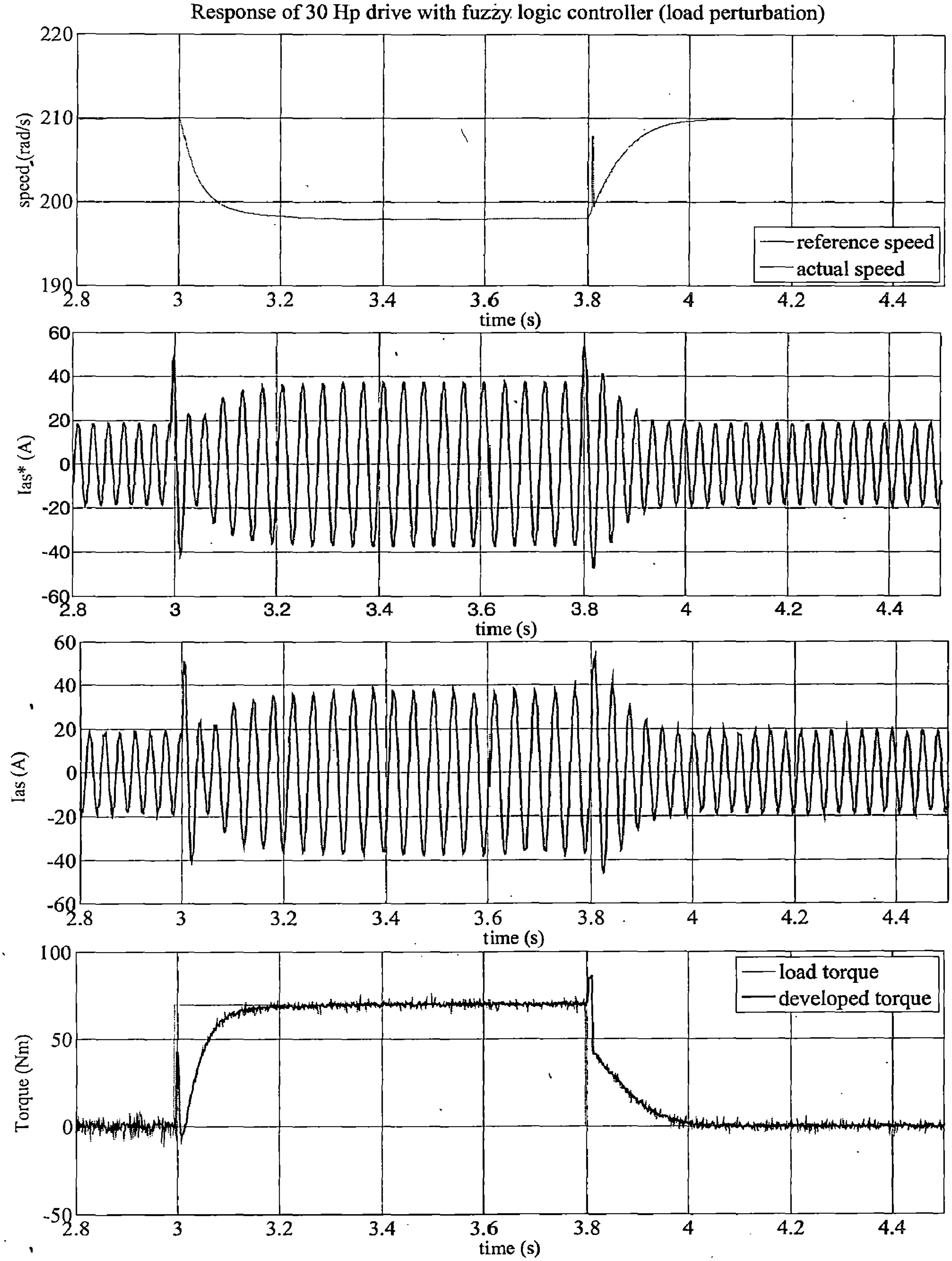


Fig. 5.9 (c) load perturbation response of 30 Hp drive with fuzzy logic speed controller

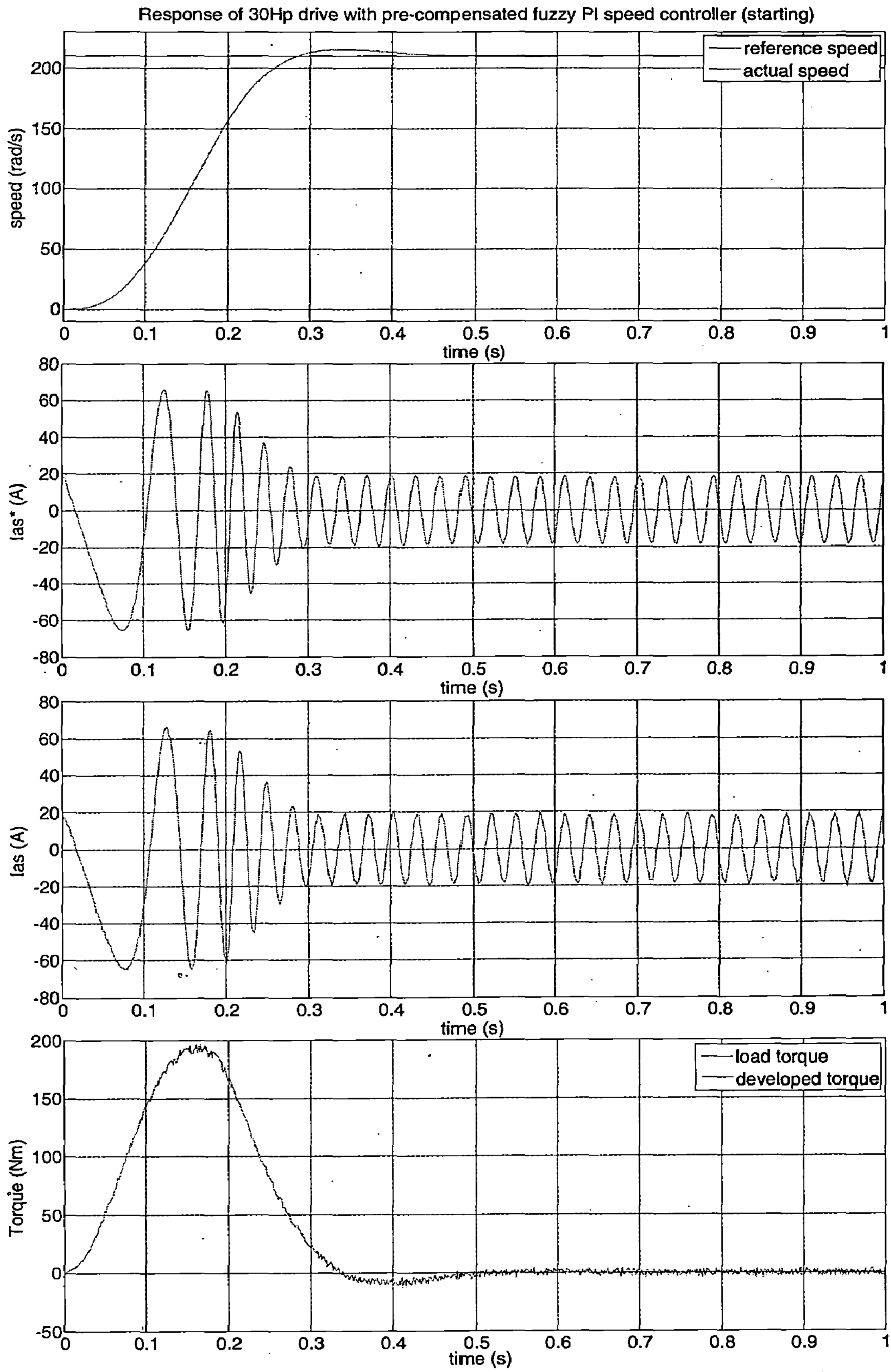


Fig. 5.10 (a) Starting response of 30 Hp drive with pre-compensated fuzzy PI speed controller

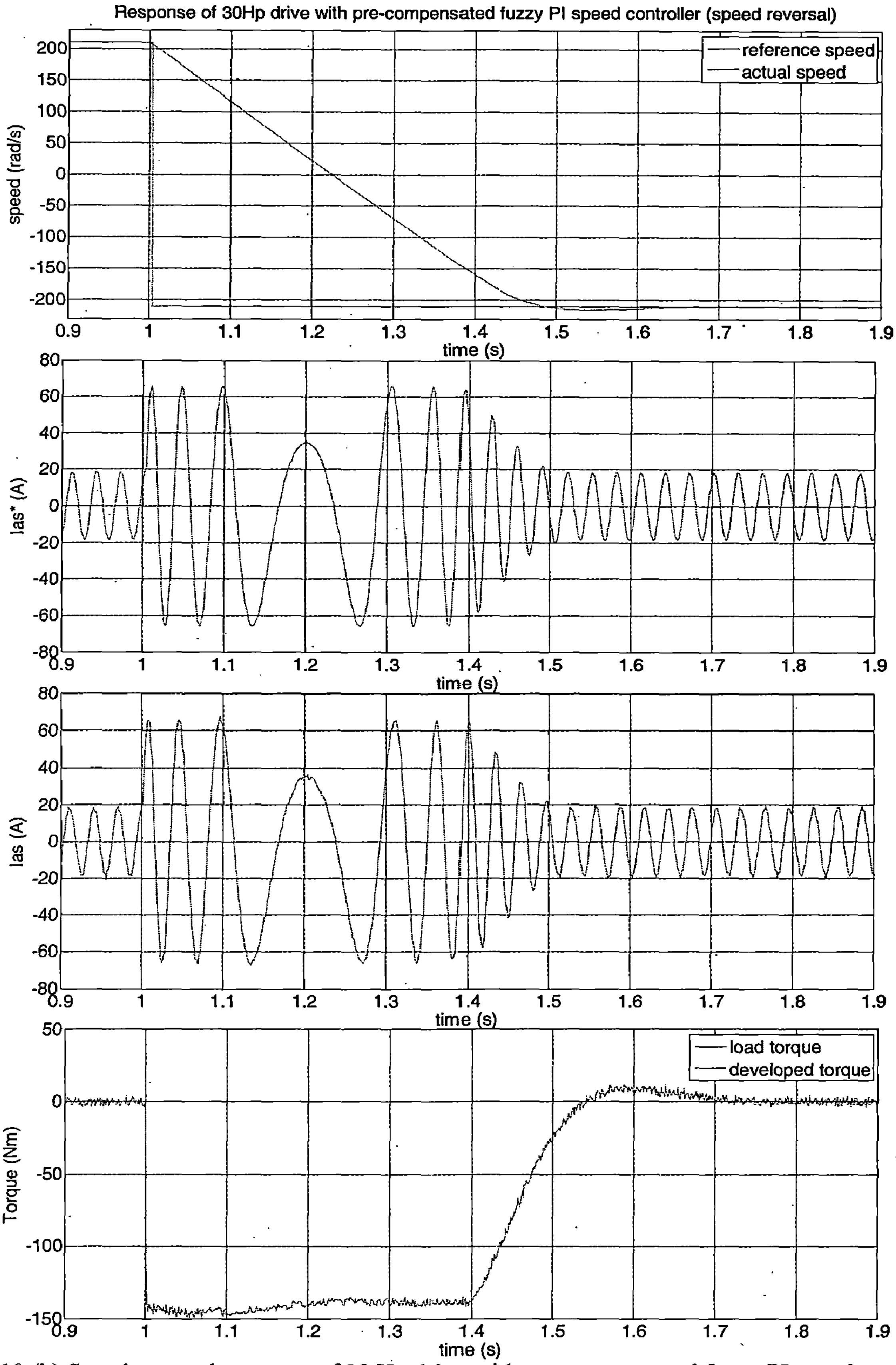


Fig. 5.10 (b) Speed reversal response of 30 Hp drive with pre-compensated fuzzy PI speed controller

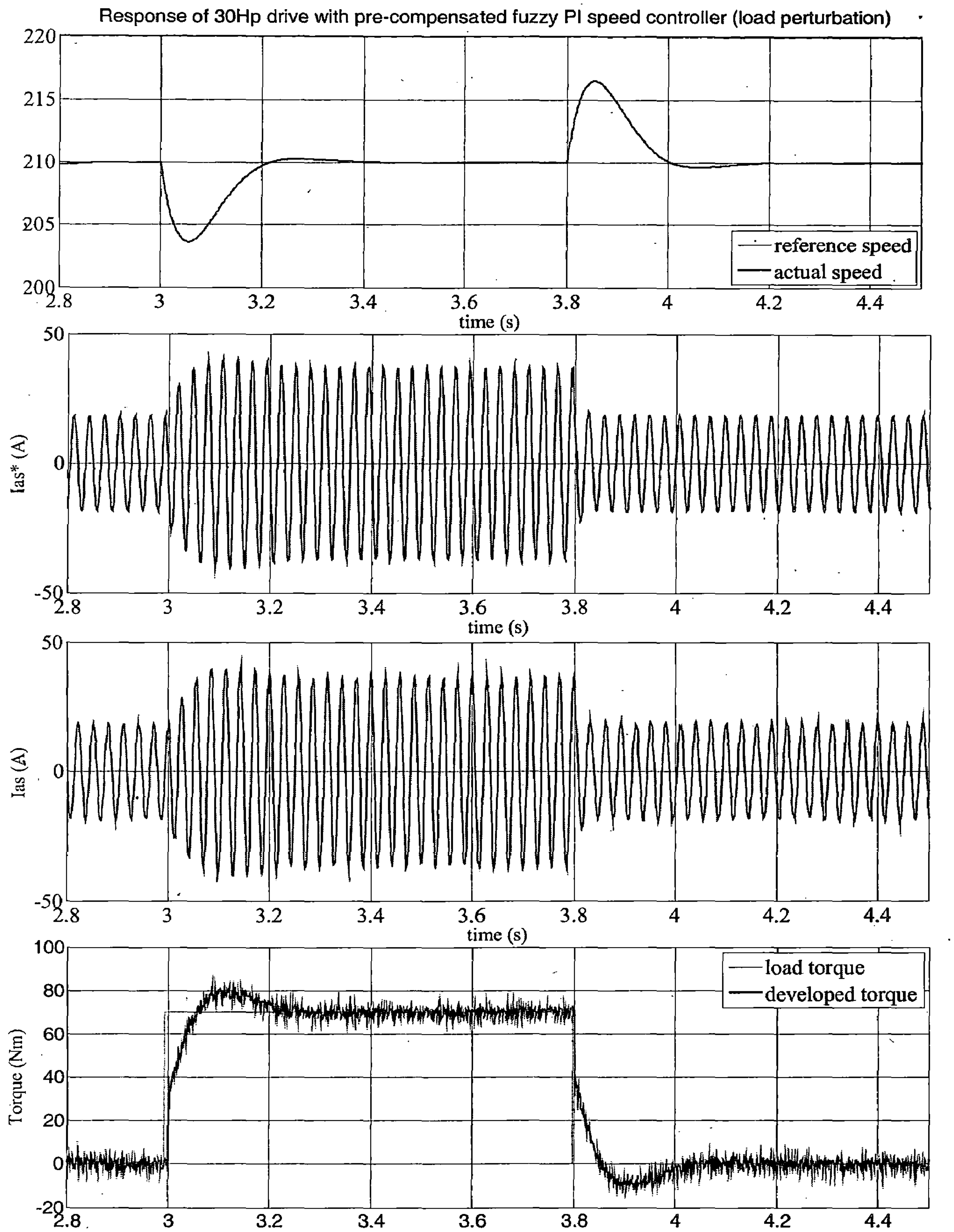


Fig. 5.10 (c) load perturbation response of 30Hp drive with pre-compensated fuzzy PI speed controller

5.4. Conclusion

The dynamic response of VCIMD is studied during various operating conditions such as starting, speed reversal and load perturbation on the VCIMD. The response of VCIMD system has been simulated in MATLAB environment using simulink and PSB toolboxes. The comprehensive study of different speed controllers has shown that the individual speed controllers have their own merits and demerits. The choice of a particular speed controller can be made depending on the application requirement. The proportional integral (PI) speed controller is a simple controller. The fuzzy logic controller is an intelligent and fast controller. Finally, it is observed that the fuzzy pre-compensated proportional integral (FPPI) seed controller is the best choice when requirement is both high level of performance and intelligence.

Tables 5.5, 5.6 and 5.7 show that for the same reference speed, PI, Pre-compensated fuzzy PI, Fuzzy Logic speed controllers show a persistent change in starting and reversal time. The advantages of FL and FPPI controllers over PI is that the tuning efforts are eliminated, the FL controller makes it's decisions based on error and change in error in speed while FPPI controller takes care of tuning mismatches of PI controller. The FL controller provides least of settling time for all the transient conditions. In order to take the advantages of PI and Fuzzy Logic controllers and to eliminate disadvantages of the individual controller, both the control techniques are combined and named as a hybrid speed controller. When the speed error is near to zero, the control is switched over to PI controller therefore eliminating disadvantage of steady state speed error of Fuzzy Logic speed controller. When drive speed is away from the reference value, the control is changed over to Fuzzy Logic controller, which provides improved dynamic performance.

5.5. Suggestions for further work:

Although the objectives set forth in this thesis have been successfully achieved, certain aspects require further investigation. The simulation study has been carried out in MATLAB environment using simulnk and PSB toolboxes, the real time implementation of these converters feeding VCIMD may be carried out to validate the model and designing the control of these converters. Different other controllers like posicast and hybrid of fuzzy logic and PI controller which are left out in this investigation can also be modeled and simulated and even real time design can be implemented for those controllers. The reduction of sensors in VCIMD can be investigated and implemented. The reduction in the number of sensors without any loss in quality of the signal estimated would lead to simplification of the controller making it more robust and reducing overall cost of the drive system.

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APPENDIX-A

Specifications of the motor considered in this investigation:

Cage Induction motor I

1 HP, 3-phase, 2-pole, Y-Connected, 240V, 50Hz

$R_s = 9.45\ \Omega$, $R_r = 11.12\ \Omega$, $X_{ls} = 11.03396\ \Omega$, $X_{lr} = 11.03396\ \Omega$, $X_m = 202.892\ \Omega$,

$J = 0.0018\ Kgm^2$

Cage Induction motor II

30 HP, 3-phase, 4-pole, Y-Connected, 240V, 50Hz

$R_s = 0.251\ \Omega$, $R_r = 0.249\ \Omega$, $X_{ls} = 0.4386\ \Omega$, $X_{lr} = 0.4386\ \Omega$, $X_m = 13.085\ \Omega$,

$J = 0.305\ Kgm^2$