JIMMA UNIVERSITY

JIMMA INSTITUTE OF TECHNOLOGY

FACULTY OF ELECTRICAL AND COMPUTER ENGINEERING

Model Reference Adaptive Control Based Adaptive-PID Controller Design for Speed Control of Separately Excited Direct Current Motor

A Thesis Submitted to the School of Post Graduate Studies of Jimma Institute of Technology in Partial Fulfillment of the Requirements for the Degree of Master of Science in Electrical and Computer Engineering (Control and Instrumentation Engineering)

By

Nafbek Begna

June, 2020

JIMMA, ETHIOPIA

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Declaration

I, the undersigned, announce that this thesis is my original work, was not presented at this and any other university for a degree, and all sources of materials used for the thesis have been thoroughly acknowledged.

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Place: Jimma Institute of Technology, Jimma University, Jimma

Date of submission: June, 2020

This thesis has been submitted for examination with my approval as a university advisor.

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Abstract

Separately excited direct current (SEDC) motor is the most used drive in a machine that can be adjusted inside a wide range with the goal that this provides easy controllability and high performance. The primary objective of this thesis is to control the speed of a separately excited direct current motor using the Model Reference Adaptive control (MRAC) based proportional integral derivative (PID) controller approach. The constant gain of the PID controller operation is not effective at the point where the parameter of any system changes regarding time. If MRAC based design should occur, the adjustable PID gain parameters corresponding to changes in the plant will be determined by referring to the reference model which specifies the property of the desired control system. This thesis presents the method for designing a MRAC based PID controller for speed control of armature controlled separately excited DC motor. Simulation results showed that the speed response is less affected under increased reference input, load torque and uncertain output disturbance signal for modified MIT rule than MIT rule. Likewise, at steady state the error is zero, the speed response has less maximum overshoot and fast settling time under modified MIT rule than MIT rule of model reference adaptive control. Under the Lyapunov adaptive control based, the system is stable, and the system response has less or equal to 0.505% maximum overshoot and 3.110 second settling time for all the condition we had considered under Modified MIT rule.

Keywords: Separately Excited DC Motor, Model Reference Adaptive Control (MRAC), MIT rule, Modified MIT rule, PID controller.

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List of Acronyms

DC	Direct Current
EMF	Electromotive force
I/O	Input/output
KVL	Kirchhoff Voltage Law
MIT	Massachusetts Institute of Technology
MRAC	Model Reference Adaptive Control
P	Proportional
PD	Proportional Derivative
PI	Proportional Integral
PID	Proportional Integral Derivative
SEDC	Separately Excited Direct Current

Chapter One

Introduction

1.1. Background

The speed of Direct Current (DC) drives can be adjusted for a wide ranges with the goal that this provides easy controllability and high performance. DC motors are broadly utilized in numerous applications, for example, steel rolling mills, electric trains, electric vehicles, electric cranes, latches, weaving machines, robotic manipulators, …etc. These can be because of their low cost, inexpensive maintenance, simple construction, and simple controlling methodologies [1, 2].

The most flexible control is obtained through separately excited direct current (SEDC) motor in which the armature and field circuits are energized with separate sources. Armature voltage control strategy is used to change the speed up to the rated speed and the motor operates in the constant torque region and field current control method is used to vary the speed above the rated speed by weakening the flux in the constant power region [1].

To control the speed of a separately excited DC motor, there are two comprehensively utilized strategies, in particular open-loop and closed-loop control. The closed-loop control is more attractive than open-loop control in accuracy, performance, and reliability aspects when the parameter variations and external disturbances emerge. In any case, it requests the speed signal that input from the sensor [2].

Conventional controllers such as PID have been applied to control the speed of separately excited DC motors. The drawback of using conventional controllers is that they are sensitive to changes in the drive parameters, load disturbances. Also, it is difficult to tune PID gains to eliminate and reduce overshoot and load disturbance. To maintain a strategic gap from the weaknesses of conventional controllers, applying adaptive control techniques for speed control of separately excited DC motor to achieve parameter insensitivity and fast speed response. The model reference adaptive controller was proposed to solve a problem in which the details are given in terms of a reference model that tells how the process output ideally should respond to the command signal [1].

Model reference adaptive control techniques are more sophisticated than the classic techniques. Usually, unexpected unsettling influence and changes in driving motor exist that since we couldn't get a total and stable design for the system. The adaptive controller's aim would be an output that causes motor output speed to follow the desired speed, and error between output and reference model limits to zero [3].

Adaptive control used to change the control algorithm coefficients in real-time to make up for variations in the environment or in the system itself [4]. In this thesis, we had deal with the modified (normalized) algorithm model reference adaptive control based PID controller to control the speed of separately excited DC motor in which the output response is forced to follow the response of a reference model irrespective of plant parameter variations.

There is no guarantee that an adaptive controller based on the MIT rule will give a stable closed-loop system. It is desirable to check whether there are different methods for designing adaptive controllers that can guarantee the stability of the system. Model reference adaptive control with the Lyapunov stability method is commonly used for first and second-order systems. The Lyapunov techniques used to find the Lyapunov function and an adaptation mechanism in the way that the error between plant and model goes to zero. Likewise, this technique ensures stability for the system [5, 6]. At that point, we have additionally considered the Lyapunov rule because it guaranteed the stability of the system.

Generally, the model reference adaptive control based PID controller is preferred to control the speed of separately excited DC motor. Consequently, the modified MIT rule with model reference adaptive control is less sensitive even for small and large amplitude of reference input.

1.2. Statement of the problem

The separately excited DC motors are frequently utilized in different industry. Different speed can be acquired by changing the armature voltage and the field voltage. Conventional PID technique is commonly used in separately excited DC motor speed and position control, it is not appropriate for high-performance cases, as a result of the conventional controller has low robustness. Although, under fuzzy controller it is difficult to obtain optimal fuzzy rules in the system design. Also robust control guarantees that if the changes are within a given bounds the control law need not be changed and it needs only a priori information about the bounds on the time varying parameters. But adaptive control which adapt to a controlled system with parameters which vary. In spite of the fact that the simple MIT rule gives satisfactory results but it is very sensitive to the changes in the amplitude of the reference input and plant parameter variations. Also when reference input changed to large values the system may become unstable.

1.3. Objectives of the study

1.3.1. General objective

The main objective of this thesis is to design a model reference adaptive control based PID controller for the speed control of separately excited DC motor

1.3.2. Specific objectives

- To drive the mathematical model of the armature controlled separately excited DC motor.
- To design MIT rule and Modified MIT rule of MRAC based PID controller.
- To investigate the performance of MIT rule as compared to modified MIT rule of MRAC based PID controller under different operating conditions.
- To test and validate the overall system through simulation using MATLAB/Simulink software.

1.4. Scope of the study

The scope of this thesis is to study, design, and simulate the speed control of the armature controlled separately excited DC motor using MRAC based PID control scheme only using MATLAB/Simulink software by changing different parameters. The dynamic behavior of this motor is also modeled under a non-flexible shaft for all conditions we have considered.

1.5. Outline of the thesis

This thesis include six chapters. In chapter two, different kinds of literature related to speed control of separately excited DC motor, have been reviewed. System modeling is presented in chapter three. Controller designing for the speed control of separately excited DC motor is designed in chapter four. Simulation results are presented and discussed in chapter five. The contributions of the thesis work are also discussed in the same chapter. Finally, chapter six presents the conclusion and recommendation for future works

Chapter Two

Literature Review

Under this chapter we have discussed the related literature to the thesis title. The techniques of controlling the parameters and the provided result of the system have been considered. As well the differences of their work and this thesis work have been also explained.

(Hameed and Mohamad, 2012) [1] Presented on Speed control of separately excited DC motor using fuzzy neural model reference controller. Under the paper of Speed control of separately excited DC motor using fuzzy neural model reference controller, they explained a technique which preferred then conventional controller because for a high load applied it also gave good performance and high robustness than the conventional controller. In the technique, the learning behavior of the reference model was due to the rule base and in this thesis, the plant output track the reference model by using the parameter adaptation techniques. It is known that, the rule base learning behavior limited on the range that an operator or system designer design for the controller parameters. But in the parameter adaptation techniques the controller parameter update itself depending on the parameter variations and adjust the controller to control the system to the desired operation.

(Pimkumwong and Wang, 2018) [2] Presented on An Online Artificial Neural Network Speed Estimator for Sensorless Speed Control of Separately Excited DC Motor. The paper focused on the armature voltage controlling method and to estimate the speed, the armature current estimation equation coefficient was adjusted according to an artificial neural network learning principle. The speed estimation of the separately excited DC motor that presented on the paper was without considering the disturbance loads and the simulation result shows, the difference between the estimated and the actual speed of the system high. But under this thesis work we have also focused on the armature controlling techniques of separately excited DC motor speed control and the reference model is selected as the desired speed that can be tracked by the response of the system by the help of the adaptation parameters. Also, different load disturbances have been considered and we have seen that under modified MIT rule the speed follows the reference speed with small error.

(SHAHGHOLIAN and MAGHSOODI, 2016) [3] Presented on Analysis and simulation of speed control in separately excited DC Motor drive by using fuzzy control based on model reference adaptive control. Unfortunately, systems designers had difficulty obtaining optimum fuzzy rules because the fuzzy logic controller allows the control of systems whose parameters are obscure, and unknown. In fuzzy controller if-then rules were provided and its input was selected as the error and change of error. In this thesis model reference adaptive control based PID controller has been used and the controller parameter updates itself to the system parameter variations and adjust the common controller to control the system as it perform to the desirable reference.

(Ali et al., 2012) [4] Presented on the adaptive PID controller for DC motor speed control. In this work of simple MIT based adaptive PID controller for the speed control of DC motor, when the adaptation gains increased the performance of the system to adapt the reference model would be higher and the tracking error also reduced continuously to zero. From the presented result part of the paper, the reference model output had higher overshoot than the system output. Also the work does not considered a proper selection of reference model, the loads condition, and the changes of reference input effect. Incorrect choice of the reference model and increasing the reference input signals in simple MIT rule of the MRAC makes the system unstable, and the controller would be unable to control the system. Therefore, in this thesis modified MIT rule is preferred to overcome the problem and we have considered selection of reference model, different load disturbances and the effect of reference input increment.

(Rajpoot et al., 2016) [7] Presented on Design and Simulation of Neuro Fuzzy Controller for Speed Control of a Separately Excited DC Motor. The speed control stated under this technique, rather than using conventional controller and fuzzy logic controller Neuro Fuzzy Controller used to improve the performance of the system where the operating condition was far from the nominal operation. From the result part of the paper, we have seen that Neuro Fuzzy controller had a better response in case of peak overshoot, reduced the settling time and fast rising time than fuzzy and PID controller. But in this thesis, modified MIT rule of MRAC tuned PID controller is used in different operating condition and the result shows that the response of the system track the reference model output with fast rising time, reducing the settling time, less peak overshoot than Neuro Fuzzy controller.

(Khanke and Jain, 2015) [8] Presented on speed control of separately excited DC motor using various conventional controllers. Using conventional controllers to control the speed of separately excited DC motor is common. In the paper, different conventional controllers such as proportional (P), proportional-derivative (PD), proportional-integral (PI), and proportional integral derivative (PID). From the result part of the paper even only linear load was applied to all the controllers their response had high overshoot and took more time to settle. In this thesis, load torque, uncertain output disturbance signal and increment of reference input can be considered in the speed control of separately excited DC motor. Then rather than using the conventional controller model reference adaptive control is better to rise the system response in fast, reduce the overshoot, reduce the steady-state error to zero in both no-load and load disturbance and parameter variations.

(Sar and Dewan, 2014) [9] Presented on MRAC based PI controller for speed control of separately excited DC motor. In this case, the simple MIT rule was developed to track the speed of the system as it follows the reference input. But simple MIT rule has its drawback because it is very sensitive to the changes in the amplitude of reference input and parameter variations. Also the controller parameters updated with the parameter variations and the Proportional Integral (PI) controller was used to tune the system response to the desired value as it became a stable system by eliminating the steady-state error rather than reducing the overshoot. In this thesis the modified MIT rule with MRAC based PID is preferred in case it is less sensitive to the changes of the amplitude of reference input and parameter variations. As well, the PID controller additionally used for reducing the overshoot because of the derivative term exist.

(Yadav et al., 2013) [10] Presented speed control of separately excited DC motor using adaptive PID controller. Using an adaptive PID controller for the speed control of separately excited DC motor had been also considered when the system was at no-load condition only. The simulated result of the paper showed that it had fast rising time, small settling time and high overshoot when the adaptation gain of the controller parameters increased continuously. But the paper did not considered another parameter variations and also it simply used the simple MIT rule with an adaptive PID controller. In this thesis the modified MIT rule with MRAC based PID is preferred to reduce the overshoot and steady-state error under different operating conditions because of the technique is less sensitive to the parameter variations.

(Kumar et al., 2012) [11] Presented on self-tuning of PID controller for DC motor using MRAC technique. In the paper of self-tuning PID controller based MRAC with simple MIT technique the first-order linear system modelling was used by ignoring the armature inductance and any disturbances cannot be considered. But the purpose of using an adaptive control method is when nonlinearity situations can happen it used adaptation gains that make the system adapt our reference model which could be the desired output (speed). The simulation part showed that the performance of the controller for rectangular and sine wave inputs and the response of the system had been follow the reference model output with high overshoot and high steady-state error. But in this thesis, we have used MRAC based PID controller with modified MIT rule to control the speed of the system without ignoring the effect of the armature inductance in the modelling and we have also considered the different operating conditions.

(Jain and Nigam, 2013) [12] Presented on Design of a Model Reference Adaptive Controller Using Modified MIT Rule for a Second Order System. In the Design of a Model Reference Adaptive Controller Using Modified MIT Rule for a Second-Order System, the paper used the normalized algorithm technique to control unknown plants. The simulation was simulated without designing and considering the situation that can disturb the system response and the controller performance. In the work, the effect of adaptation gain and reference input had been discussed and the result showed that when the reference input increased the system response had fast rising time, high overshoot and reduced steady-state error. But in in this thesis, we have considered the selection of the reference model, the effect of load disturbances and to show the performance of the modified MIT rule with MRAC tuned PID controller is better than the MIT rule with MRAC tuned PID controller in different operating conditions.

Generally, many controller methods applied in order to improve performance and efficiency of speed control of separately excited DC motor and to cope with the changes in different operating conditions and to ensure controller performance. In addition to this to ensure the best controller performance at no-load, loaded torque and uncertain output disturbance signal characteristics will be considered when we design the controller.

Chapter Three

Mathematical Modeling of Separately Excited DC Motor

3.1. Introduction

Different kinds of literature that were related to this thesis had studied to have an understanding of speed control of separately excited DC motor. Although the parameters that could affect the speed of the system that have been controlled were defined and the mathematical model of the system had established under different conditions.

3.2. Speed control of separately excited DC motor

Separately excited DC motor comprises two sections i.e. rotor and stator. The Stator consists of field winding while the rotor (also known as armature) consists of armature winding. When both armature and field are energized by DC supply, current flows through windings and magnetic flux proportional to the current is produced. When the field flux interacts with the armature flux, this results in the motor the rotor rotation.

Due to separate field and armature circuit of separately excited DC motors, we have develop the mathematically model of the system in separate form as it become an appropriate for control applications. Two ways to control separately excited DC motors are:

1. Field control and
2. Armature control

3.2.1. Field control

This method requires for the field circuit a variable voltage supply which is separated from the main power supply to which the armature is connected. In this method, the voltage source supplying the field current is different from that which supplies the armature [13, 14].

The speed of separately excited DC motor can be controlled by changing field flux. This method can be called field control. In this method, the speed control of the system is applicable for speed above the rated speed.

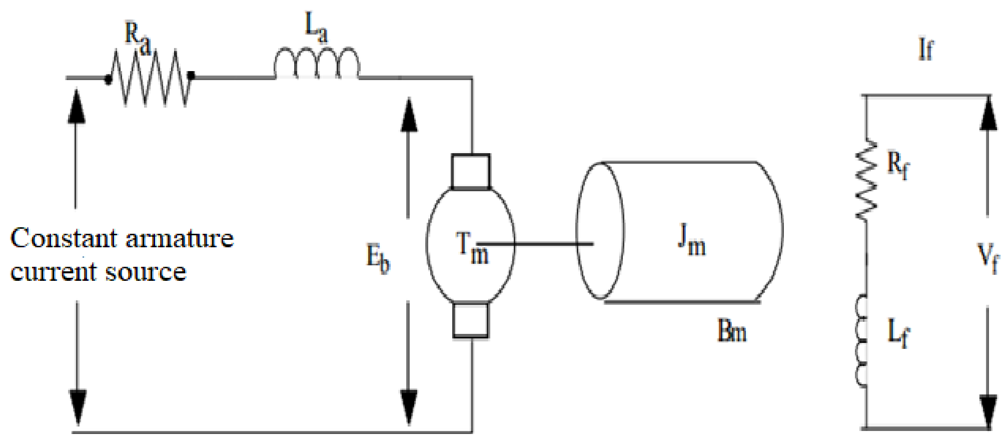

Figure 3.1: Circuit diagram of field controlled separately excited DC motor [8].

In the field control technique, a constant current I_a is fed to the armature and it is known the flux is proportional to the field current. Mathematically it can be expressed in equation (3.1) below.

$$\Phi = k_f I_f \tag{3.1}$$

where k_f = constant due to field current and by using KVL for the circuit diagram in Figure 3.1 we can get the potential difference that can be expressed in equation (3.2).

$$V_f = I_f R_f + L_f \frac{dI_f}{dt} \tag{3.2}$$

Also some torque is produced which is directly proportional to flux (Φ) and armature current (I_a). We can express mathematically as equation (3.3) shown below:

$$T = k' \Phi I_a \tag{3.3}$$

By substituting equation (3.1) into (3.3) we can get:

$$T = k' k_f I_f I_a \tag{3.4}$$

Where both k' and I_a are constant we can take as $k = k' I_a$. Then, equation (3.4) is rewritten as follow:

$$T = k k_f I_f \tag{3.5}$$

As well we introduce the dynamic torque of the system as follows:

$$T = J\frac{dw(t)}{dt} + Bw(t) \tag{3.6}$$

where $J =$ moment of inertia, B coefficient of friction and w angular speed.

Taking the Laplace transform of all equation (3.1), (3.2), (3.5) and (3.6) we can get the equation (3.7), (3.8), (3.9) and (3.10) as follows:

$$\Phi(s) = k_f I_f(s) \tag{3.7}$$

$$V_f(s) = R_f I_f(s) + sL_f I_f(\mathrm{s}) \tag{3.8}$$

$$T(s) = kk_f I_f(\mathrm{s}) \tag{3.9}$$

$$T(s) = sJw(s) + Bw(s) \tag{3.10}$$

From equation (3.8) we have

$$I_f(s) = \frac{V_f(s)}{sL_f + R_f} \tag{3.11}$$

By substituting equation (3.11) into (3.9) we get

$$T(s) = kk_f \frac{V_f(s)}{sL_f + R_f} \tag{3.12}$$

Also by equating equation (3.10) and (3.12) we have

$$\frac{w(s)}{V_f(s)} = \frac{kk_f}{(sJ + B)(sL_f + R_f)} \tag{3.13}$$

For separately excited type, we usually use field control if there is a field supply voltage equal to the rating of the motor and when we need to run the motor in fixed rated speed. Here field control helps adjust the speed for a small range [15]. These are open-loop systems and we require someone to manually adjust field current. As setting it for closed-loop probably lead to an unstable system as the range is very small.

On the other hand control of armature offers a wide range of speed control and this configuration is used for closed-loop systems. To conclude, we usually apply armature control as we can control the input current and voltage to the motor also provide a large range of speed in which motor can be operated. In this thesis, separately excited DC motor speed can be controlled in the armature control techniques by changing the armature voltage.

3.2.2. Armature control

Armature control is the most common control technique for DC motors due to that the speed can be controlled up to the rated speed. To implement this control, the flux is required to be kept constant [14, 16, 17].

Design of the speed controller and different components which are used for the control design such as armature voltage (V_t), armature resistance (R_a), armature inductance (L_a), armature current (i_a), back emf of the motor (e_a), mechanical torque developed (T), moment of inertia (J), friction coefficient (B), load torque (T_L) and etc…, which are components of the diagram shown below.

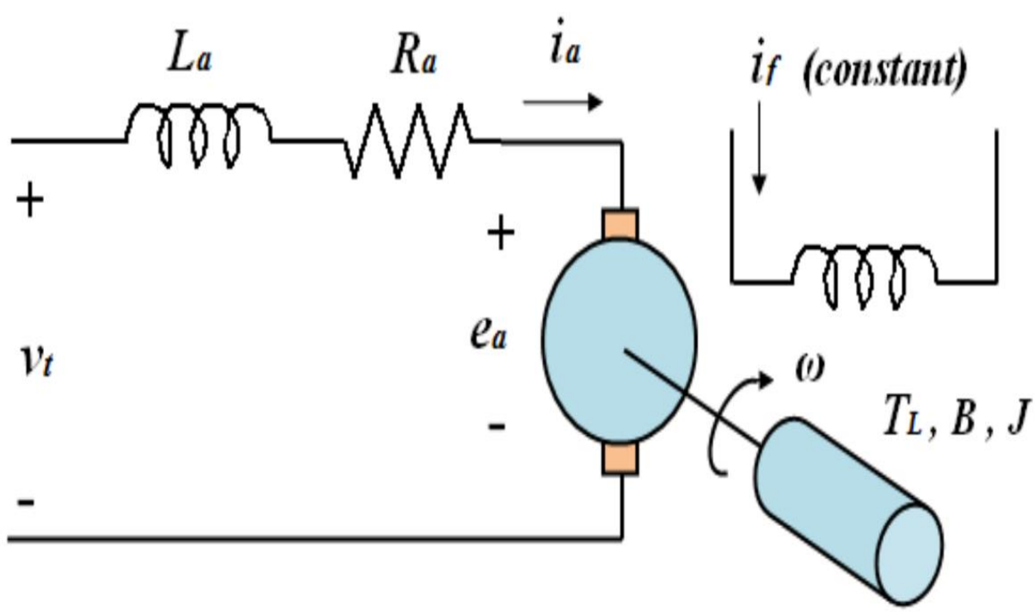

Figure 3.2: Circuit diagram of armature controlled separately excited DC motor [1].

There are three inputs to the plant, namely, the control signal to the plant or adaptive controller output V_t, load torque T_L and output disturbances due to uncertainties d_u [18, 19, 20]. The resultant or ultimate speed of the plant is given by:

$$w_d = w_0 + d_l + d_u \qquad (3.14)$$

In another ways we can express:

$$w_d = w_0 + d \tag{3.15}$$

where w_0 is the speed without disturbance and $d = d_l + d_u$, it is speed disturbance due to load torque and uncertainties.

From circuit diagram of Figure 3.2 above by using KVL we have;

$$\mathrm{V(t)} = \mathrm{e_a(t)} + L_a \frac{di_a(t)}{dt} + R_a i_a(t) \tag{3.16}$$

The dynamic torque produced can be expressed in equation (3.17) as:

$$T_e(t) = J\frac{dw(t)}{dt} + Bw(t) + T_L(t) \tag{3.17}$$

It is clear that, the back electromotive force (emf) of the motor is directly proportional to the angular speed and the torque produced, moreover it is directly proportional to the armature current.

Mathematically, we can express as:

$$e_a(t) = k_b w(t) \tag{3.18}$$

$$T_e(t) = k_t i_a(t) \tag{3.19}$$

where k_b is the back emf constant and k_t is the motor torque constant

By taking the Laplace transform of all equation (3.16), (3.17). (3.18) and (3.19) we get:

$$\mathrm{V(s)} = \mathrm{E_a(s)} + sL_a I_a(s) + R_a I_a(s) \tag{3.20}$$

$$T_e(s) = sJw(s) + Bw(s) + T_L(s) \tag{3.21}$$

$$E_a(s) = k_b w(s) \tag{3.22}$$

$$T_e(s) = k_t I_a(s) \tag{3.23}$$

By substituting equation (3.22) into (3.20) we get:

$$\mathrm{V(s)} = k_b w(s) + (sL_a + R_a)I_a(s) \tag{3.24}$$

Also by equating equation (3.21) with (3.23) we get:

$$sJw(s) + Bw(s) + T_L(s) = k_t I_a(s) \tag{3.25}$$

When the system is with no load torque and uncertainties ($V_t \neq 0, T_L = 0$, and $d_u = 0$) and equation (3.25) rewritten as follows:

$$sJw(s) + Bw(s) = k_t I_a(s) \tag{3.26}$$

From equation (3.24) and (3.26) above we have a block diagram of separately excited DC motor under no loaded condition as shown in Figure 3.3.

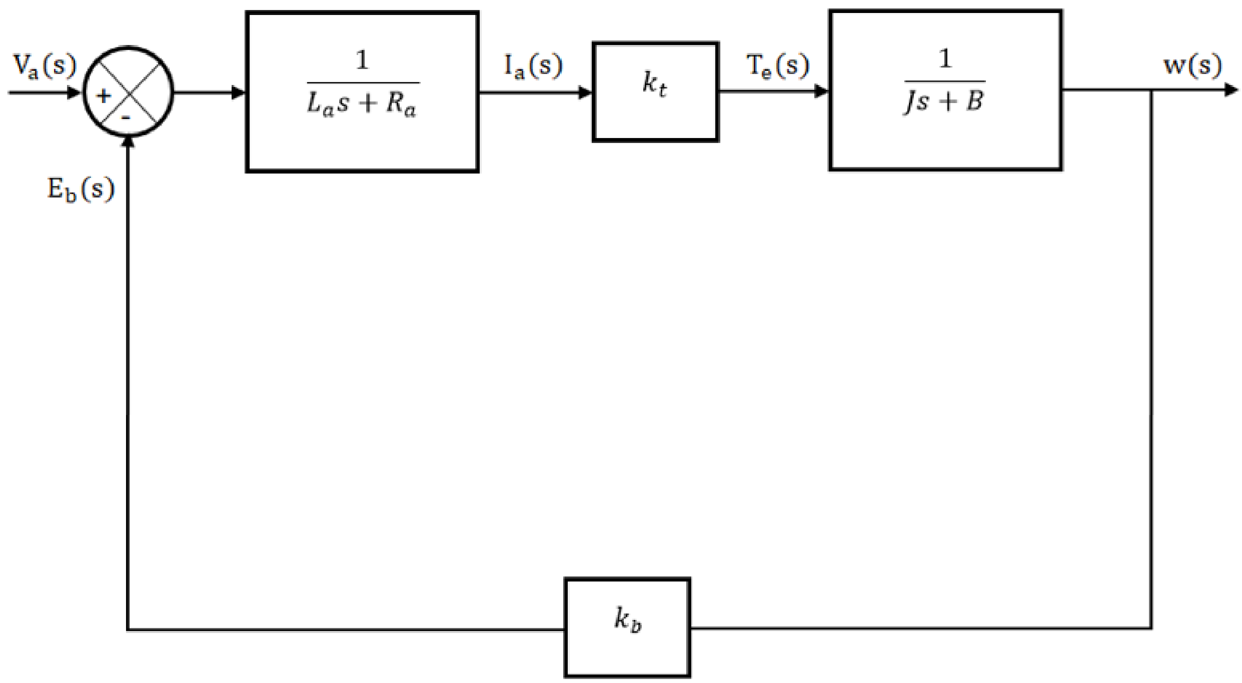

Figure 3.3: Block model of armature controlled separately excited DC motor without load

Then, from equation (3.26) the armature current can be;

$$I_a(s) = \frac{(sJ + B)w(s)}{k_t} \tag{3.27}$$

By substituting equation (3.27) into (3.24) we get:

$$\mathrm{V(s)} = k_b w(s) + (sL_a + R_a)\frac{(sJ + B)w(s)}{k_t} \tag{3.28}$$

The transfer function of armature controlled separately excited DC motor at no load torque and uncertainties condition can be as follow:

$$\frac{w(s)}{V(s)} = G(s) = \frac{k_t}{JL_a s^2 + (JR_a + BL_a)s + (BR_a + k_b k_t)} \tag{3.29}$$

Let's take the constant $k_b = k_t = K$, then equation (3.29) can be rewritten as follows:

$$G(s) = \frac{K}{JL_a s^2 + (JR_a + BL_a)s + (BR_a + K^2)} \tag{3.30}$$

$$G(s) = \frac{\frac{K}{JL_a}}{s^2 + \left(\frac{JR_a + BL_a}{JL_a}\right)s + \left(\frac{BR_a + K^2}{JL_a}\right)} \tag{3.31}$$

By rearranging equation (3.24) and (3.25) at loaded condition (case for $\mathrm{T_L} \neq 0$, and $\mathrm{d_u} = 0$) we get;

$$\mathrm{V(s)} = k_b w(s) + (sL_a + R_a)I_a(s) \tag{3.32}$$

$$T_e(s) = sJw(s) + Bw(s) + T_L(s) = k_t I_a(s) \tag{3.33}$$

From equation (3.33) the armature current can be written in (3.34) as follows:

$$I_a(s) = \frac{sJw(s) + Bw(s) + T_L(s)}{k_t} \tag{3.34}$$

Then, by substituting equation (3.34) in to (3.32) we get equation (3.35) as follows:

$$\mathrm{V(s)} = k_b w(s) + (sL_a + R_a)\left(\frac{sJw(s) + Bw(s) + T_L(s)}{k_t}\right) \tag{3.35}$$

By using equations (3.32) and (3.33) we have a block diagram of separately excited DC motor under loaded condition as shown in Figure 3.4.

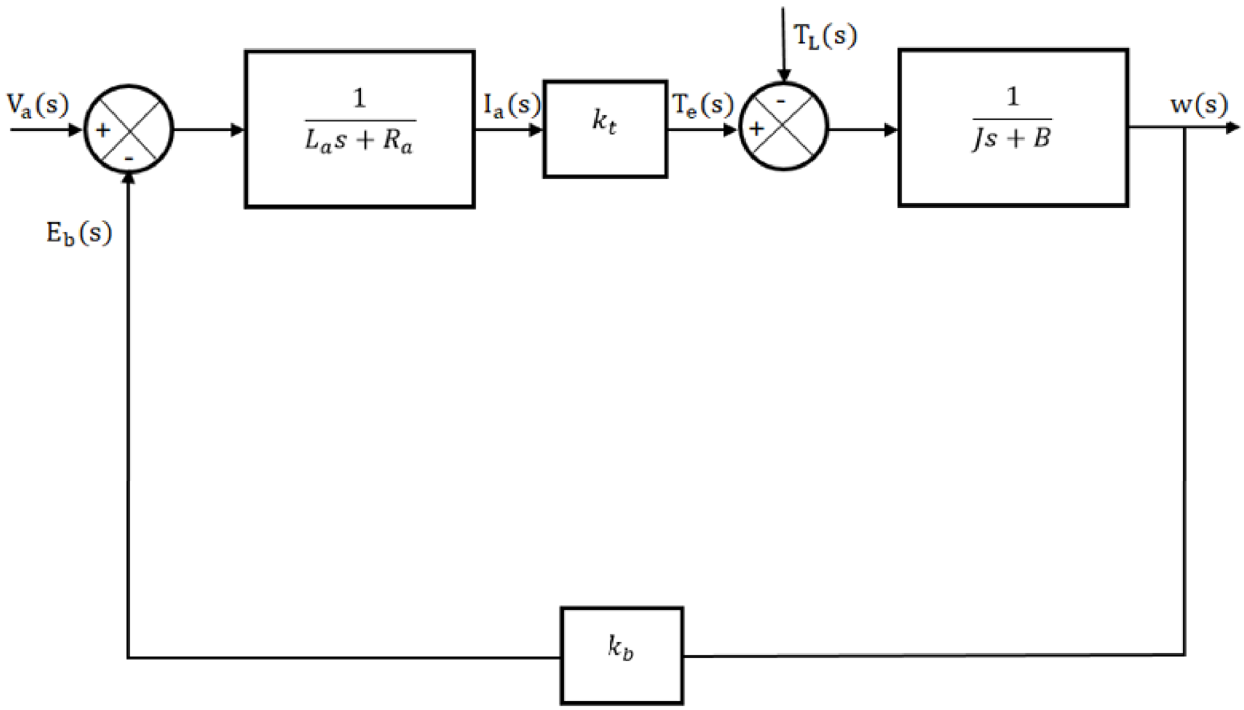

Figure 3.4: Block model of armature controlled separately excited DC motor with load torque

By using superposition principle to make the voltage source zero ($V(s) = 0$) and we can get the transfer function at loaded condition as follows,

$$[JL_a s^2 + (JR_a + BL_a)s + (BR_a + k_b k_t)]w(s) + T_L[L_a s + R_a] = 0 \tag{3.36}$$

where, the constant $k_b = k_t = K$ then the equation (3.36) can be rewritten in (3.37) as follows:

$$[JL_a s^2 + (JR_a + BL_a)s + (BR_a + K^2)]w(s) + T_L[L_a s + R_a] = 0 \tag{3.37}$$

Then, the speed due to load torque can be;

$$d_L(s) = \frac{-T_L(\frac{1}{J}s + \frac{R_a}{JL_a})}{s^2 + \frac{(JR_a + BL_a)s}{JL_a} + \frac{(BR_a + K^2)}{JL_a}} \tag{3.38}$$

where $w(s) = d_L(s)$ is the speed due to load torque

Similarly, when load torque and uncertainties in the input supply are present, the resultant speed is obtained from equation (3.14), (3.31) and (3.38) as;

$$w_d(s) = \frac{\frac{K}{JL_a}V(s) - T_L\left(\frac{1}{J}s + \frac{R_a}{JL_a}\right)}{s^2 + \left(\frac{JR_a + BL_a}{JL_a}\right)s + \left(\frac{BR_a + K^2}{JL_a}\right)} + d_u(s) \tag{3.39}$$

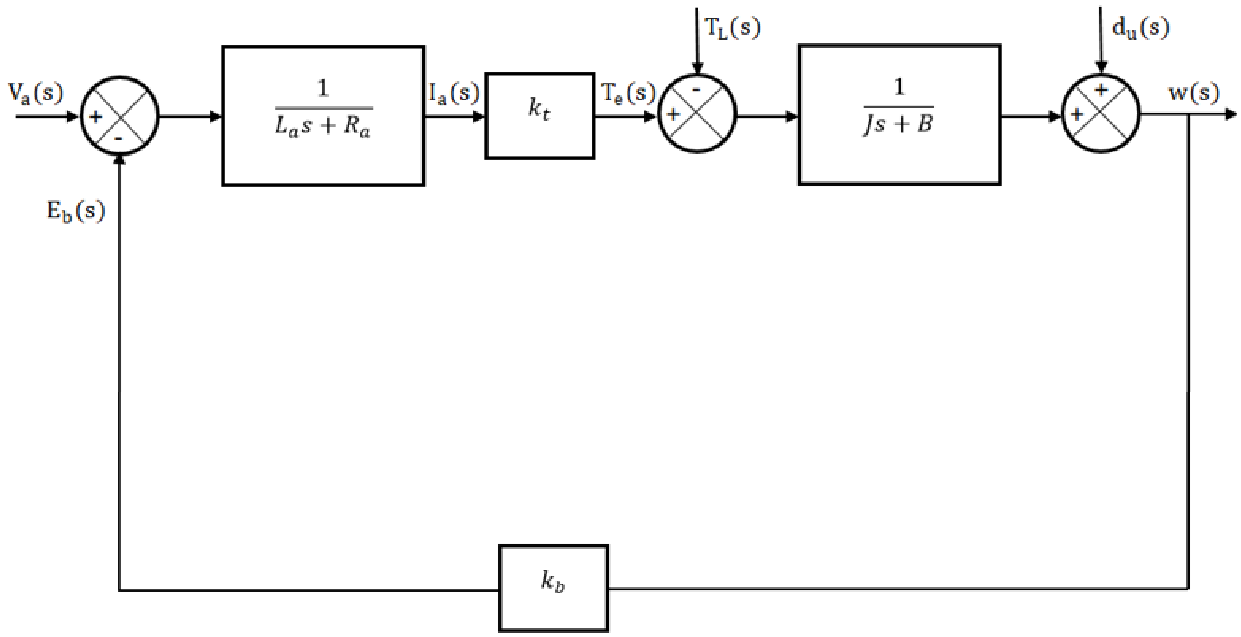

Figure 3.5: Block model of armature controlled separately excited DC motor with load torque and disturbance signal

Chapter Four

Controller Design for Separately Excited DC Motor

4.1. Introduction

In this chapter, we have designed a different controller for speed control of armature controlled separately excited DC motor. The design of the controller for this system had contain PID controller and MRAC based PID controller. This section has been basically focused on the adaptive control which is called model reference adaptive control. In model reference adaptive control, the desired behavior of the system is specified by a reference model. The parameter adaptation mechanism in MRAC can be obtained using the gradient method which is MIT rule and modified MIT rule and stability method which is the Lyapunov rule of stability theory.

4.2. Proportional Integral Derivative (PID) controller

A Proportional Integral Derivative controller is a conventional controller generally utilized in the industrial control system. The Proportional expression reacts momentarily to the present error. The integral term reacts to error accumulation providing a slow response which drives the steady-state error to zero. Also, the derivative term responds to the rate at which the error is changing. This controller can be used in continuous or discrete form. Likewise for tuning of such a controller different strategies are suggested which are given below.

1) Manual tuning method
2) Auto tuning method.

In this thesis, the Auto tuning method is preferred because of it overcomes the shortcomings of manual tuning which is time consuming. In case of speed controlling of separately excited DC motor, we do not require to calculate the value of constants such as k_p, k_i, k_d, as we do in Ziegler – Nichols method. The feedback signal easily calculated and set to the constant value with the help of PID controller. Then, we get the smooth response of a system by using the Auto tuning method.

To improve the performance of the armature controlled separately excited DC motor, a PID controller is applied. A simple feedback control theory is utilized to represent the overall PID controlled system [21, 22]. Therefore the PID controller has to be tuned to have the desired motor response.

The transfer function of this PID controller is:

$$G_c(s) = k_p + \frac{k_i}{s} + k_d s \qquad (4.1)$$

$$G_c(s) = \frac{k_d s^2 + k_i s + k_p}{s} \qquad (4.2)$$

Where k_p, k_i and k_d are proportional, integral and derivative gain respectively.

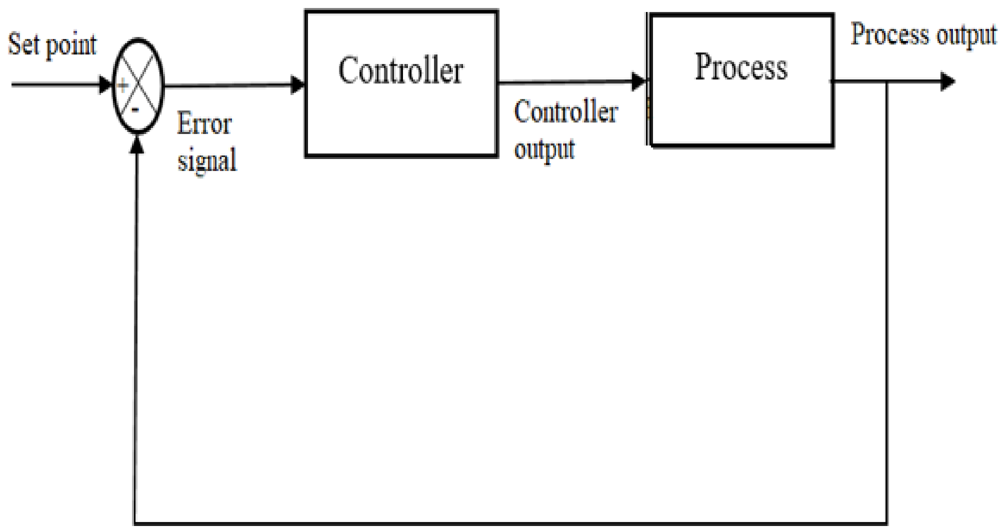

Figure 4.1: Block diagram of process control using conventional PID controller [20]

4.3. Model reference adaptive controller

An approach to designing uncertain systems is given by the adaptive control theory. The adaptive controller changes its actions to the changing property of the controlled processes, unlike the fixed parameter controller. The main difference between conventional and adaptive control is the existence of the adaptation mechanism. The main issue with adaptation design is to synthesize an adaptation mechanism that ensures that the control system remains stable and tracking errors converge to zero even when the parameters changed [9].

The concept behind model reference adaptive control is to construct a closed-loop controller with parameters that can be changed to adjust the system response to the desired model. A good understanding of the plant and performance specifications to be met in model reference adaptive control enables the designer to come up with a model, referred to as the reference model, which defines the required I / O properties of the closed-loop plant. When the plant parameters and the disturbances are slowly or slower than the dynamic behavior of the plant, then a MRAC control is used [10, 23]. The MRAS based PID controller scheme is shown in figure 4.2. The error and the adaption law for the controller parameters are determined by using MIT Rule. The MIT (Massachusetts Institute of Technology) rule states that the time rate of change of the controller parameter (θ) is proportional to negative gradient of the cost function (J).

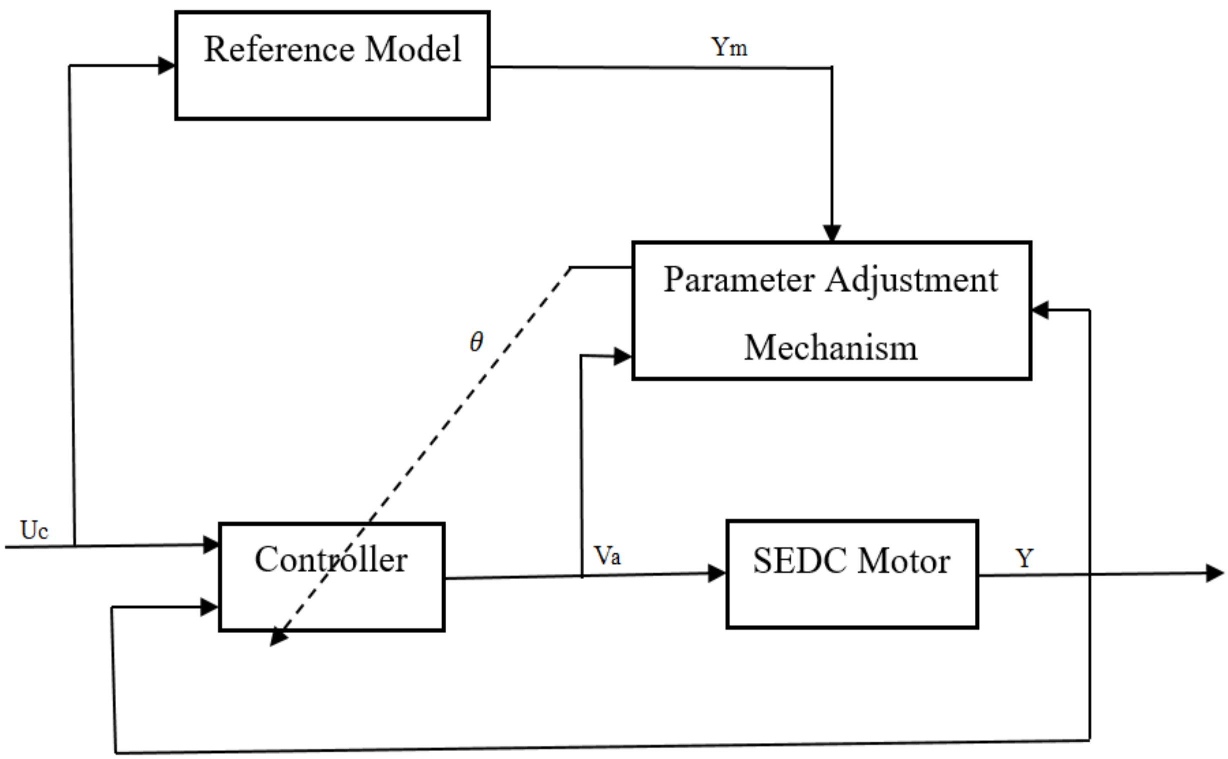

Figure 4.2: Complete proposed system design [10]

The reference model for the MRAC generates the desired trajectory y_m, which the separately excited DC motor speed y, has to follow.

Selection of reference model

A stable first or second-order reference model is chosen whose pole position decides the stability of the whole system with minimum (near to zero) overshoot, fast rise time, small settling time, unity damping ratio and zero steady-state error because it decides the response of the actual plant

[18]. If the preferred reference model has a good response and the adaptation rule is well designed, then the actual plant follows the reference model and if the reference model has a poor response, the actual plant response is also poor.

The standard second order differential equation was chosen as the reference model is given by:

$$G_m(s) = \frac{b_m}{s^2 + a_{m1}s + a_{m0}} \tag{4.3}$$

4.3.1. The MIT rule

This concept is formulated in the Massachusetts Institute of Technology and is extended to the MRAC approach. The time rate of change of controller parameter vector θ is proportional to negative gradient of cost function (J). The MIT approach to the rules aims to minimize the cost function of the squared model. Because as the error function becomes minimum there will be perfect tracking between actual plant output (y) and reference model output (y_m). In model reference adaptive control the structure of the plant is supposed to be known though the parameters are not specified. i.e. the number of poles and zeros are assumed to be known but their locations are not known [9, 24].

Mathematically, the output error can be expressed in (4.4) as follows:

$$e = y - y_m \tag{4.4}$$

Under this rule, the cost function or loss function which is used minimize the error is expressed as:

$$J(\theta) = \frac{1}{2} e^{2}(\theta) \tag{4.5}$$

Also, according to the MIT rule for updating the controller parameter in different operating condition can be determined by using equation (4.6) in the following manner [4].

$$\frac{d\theta}{dt} = -\gamma \frac{\partial J}{\partial \theta} = -\gamma e \frac{\partial e}{\partial \theta} \tag{4.6}$$

where $J =$ Cost function or objective function

$\theta =$ Controller parameter vector

e = Output error (error between actual plant and reference model)

y = Actual plant output

y_m = Reference model output

γ = Adaptation gain

$\frac{\partial e}{\partial \theta}$ = Sensitive derivative of error with respect to controller parameter

4.3.2. Design of MRAC based PID using MIT rule

The output of the PID controller in accordance to the error as the input the adaptive control law of MRAC structure taken in (4.9) as follows.

$$u(t) = k_p e(t) + k_i \int e(t)\,dt + k_d e^*(t) \tag{4.7}$$

where, $e(t) = \mathcal{E}(t) = u_c(t) - y(t)$ is the error input to PID controller, k_p is proportional gain, k_i is integral gain, k_d is derivative gain and u_c is a unit step input. In the Laplace domain, (4.7) can be transformed to:

$$U(s) = k_p \mathcal{E}(s) + \frac{k_i}{s} \mathcal{E}(s) + s k_d \mathcal{E}(s) \tag{4.8}$$

Sometimes in the basic PID algorithm, the system involves impulse function due to direct action of the derivative term when the reference input is a step signal. Such a phenomenon is called a set-point kick. To hold off a set-point kick, we wish to operate a modified PID in which derivative action is directly operated on feedback signal rather than actuating signal [4]. Then, equation (4.8) can be modified and we can get:

$$U(s) = k_p\big(U_c(s) - Y(s)\big) + \frac{k_i}{s}\big(U_c(s) - Y(s)\big) - s k_d Y(s) \tag{4.9}$$

Since it is assumed that the structure of the plant is known even though exact parameters are not known. Now in this case the motor transfer function is of second order, where $b=K/JL_a$, $a_1=(JR_a+BL_a)/JL_a$, $a_2=(BR_a+K^2)/JL_a$.

$$\frac{Y(s)}{U(s)} = \frac{b}{s^2 + a_1 s + a_2} = G_p(s) \tag{4.10}$$

Applying the control law to the system can give the following closed loop transfer function:

$$Y(s) = G_p(s)\left(\left(k_p + \frac{k_i}{s}\right)(U_c(s) - Y(s)) - sk_d Y(S)\right) \quad (4.11)$$

$$Y(s) = \frac{(G_p(s)k_p + \frac{G_p(s)k_i}{s})U_c(s)}{\left(1 + G_p(s)k_p\right) + \frac{G_p(s)k_i}{s} + G_p(s)k_d s} \quad (4.12)$$

The output error of the system the difference between actual plant output and reference model output. Then, we have the following mathematical expression for the output error:

$$e = y - y_m \quad (4.13)$$

$$e = \frac{(G_p(s)k_p + \frac{G_p(s)k_i}{s})U_c(s)}{\left(1 + G_p(s)k_p\right) + \frac{G_p(s)k_i}{s} + G_p(s)k_d s} - \frac{b_m U_c(s)}{s^2 + a_{m1}s + a_{m2}} \quad (4.14)$$

So, the transfer function of reference model can also be assumed as the form:

$$\frac{Y_m(s)}{U_c(s)} = \frac{b_m}{s^2 + a_{m1}s + a_{m2}} \quad (4.15)$$

Now, we can apply the MIT rule to the motor to obtain the controller parameters [4]. Since the controller parameter vector θ has k_p, k_i and k_d, the MIT rule can be splitted up into three parts written in equation (4.16), (4.17) and (4.18).

$$\frac{dk_p}{dt} = -\gamma_p \frac{\partial J}{\partial k_p} = -\gamma_p \left(\frac{\partial J}{\partial e}\right)\left(\frac{\partial e}{\partial y}\right)\left(\frac{\partial y}{\partial k_p}\right) \quad (4.16)$$

$$\frac{dk_i}{dt} = -\gamma_i \frac{\partial J}{\partial k_i} = -\gamma_i \left(\frac{\partial J}{\partial e}\right)\left(\frac{\partial e}{\partial y}\right)\left(\frac{\partial y}{\partial k_i}\right) \quad (4.17)$$

$$\frac{dk_d}{dt} = -\gamma_d \frac{\partial J}{\partial k_d} = -\gamma_d \left(\frac{\partial J}{\partial e}\right)\left(\frac{\partial e}{\partial y}\right)\left(\frac{\partial y}{\partial k_d}\right) \quad (4.18)$$

where $e = y - y_m$, $\frac{\partial e}{\partial y} = 1$ and $\frac{\partial J}{\partial e} = e$

Using the above relationship equation (4.16), (4.17) and (4.18) rewritten in equation (4.29), (4.20) and (4.21) respectively.

$$\frac{dk_p}{dt} = -\gamma_p \frac{\partial J}{\partial k_p} = -\gamma_p e \left(\frac{\partial y}{\partial k_p}\right) \tag{4.19}$$

$$\frac{dk_i}{dt} = -\gamma_i \frac{\partial J}{\partial k_i} = -\gamma_i e \left(\frac{\partial y}{\partial k_i}\right) \tag{4.20}$$

$$\frac{dk_d}{dt} = -\gamma_d \frac{\partial J}{\partial k_d} = -\gamma_d e \left(\frac{\partial y}{\partial k_d}\right) \tag{4.21}$$

By rearranging equation (4.12) we have:

$$\frac{Y(s)}{U_c(s)} = \frac{(G_p(s)k_p + \frac{G_p(s)k_i}{s})}{\left(1 + G_p(s)k_p\right) + \frac{G_p(s)k_i}{s} + G_p(s)k_d s} \tag{4.22}$$

$$Y(s)(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s) = (G_p(s)k_p + \frac{G_p(s)k_i}{s})U_c(s) \tag{4.23}$$

Now by differentiating equation (4.23) with respect to the controller parameters we can get $\frac{\partial y}{\partial k_p}, \frac{\partial y}{\partial k_i}$ and $\frac{\partial y}{\partial k_d}$, as follows:

Differentiating equation (4.23) with respect to k_p

$$[Y(s)\frac{\partial}{\partial k_p}(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s)] + [(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s)]\frac{\partial y}{\partial k_p} = \frac{\partial}{\partial k_p}(G_p(s)k_p + \frac{G_p(s)k_i}{s})U_c(s) \tag{4.24}$$

$$\frac{\partial y}{\partial k_p} = \frac{G_p(s)}{\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(U_c(s) - Y(s)) \tag{4.25}$$

Also differentiating equation (4.23) with respect to k_i

$$[Y(s)\frac{\partial}{\partial k_p}(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s)] + [(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s)]\frac{\partial y}{\partial k_i} = \frac{\partial}{\partial k_p}(G_p(s)k_p + \frac{G_p(s)k_i}{s})U_c(s) \tag{4.26}$$

$$\frac{\partial y}{\partial k_i} = \frac{\frac{G_p(s)}{s}}{\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(U_c((s) - Y(s)) \quad (4.27)$$

Similarly differentiating equation (4.23) with respect to k_d

$$[Y(s)\frac{\partial}{\partial k_p}(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s)] + [(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s)]\frac{\partial y}{\partial k_d} = \frac{\partial}{\partial k_p}(G_p(s)k_p + \frac{G_p(s)k_i}{s})U_c(s) \quad (4.28)$$

$$\frac{\partial y}{\partial k_d} = \frac{G_p(s)s}{\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(-Y) \quad (4.29)$$

It is difficult to use the exact formulas which are derived using the MIT rule. Instead some approximations are required [4]. An approximation made which valid when parameters are closed to ideal value is expressed in equation (4.30) as follows:

$$\frac{G_p(s)}{1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s} \approx G_m(s) \quad (4.30)$$

Then we can apply the MIT gradient rule for determining the value of PID controller parameters (k_p^*, k_i^* and k_d^*). So the adjustment parameters are:

$$\frac{dk_p}{dt} = -\gamma_p e\left(\frac{\partial y}{\partial k_p}\right) = (-\gamma_p)e\frac{G_p(s)}{\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(U_c(s) - Y(s)) \quad (4.31)$$

$$\frac{dk_i}{dt} = -\gamma_i e\left(\frac{\partial y}{\partial k_i}\right) = (-\gamma_i)e\frac{G_p(s)}{s\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(U_c(s) - Y(s)) \quad (4.32)$$

$$\frac{dk_d}{dt} = -\gamma_d e\left(\frac{\partial y}{\partial k_d}\right) = (-\gamma_d)e\frac{sG_p(s)}{\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(-Y(s)) \quad (4.33)$$

Also when we substitute the approximated parameter adaptation in all equations (4.31), (4.32) and (4.33) above, it can be simplified and rewritten in equations (4.34), (4.35) and (4.36) respectively.

$$K_p{}^* = -\gamma_p e G_m \mathcal{E} \tag{4.34}$$

$$K_i{}^* = \frac{-\gamma_i}{s} e G_m \mathcal{E} \tag{4.35}$$

$$K_d{}^* = s\gamma_d e G_m y \tag{4.36}$$

So, by removing the derivative term of the above equation (4.34), (4.35) and (4.36) we get as written in equation (4.37), (4.38) and (4.39) below respectively.

$$K_p = \frac{1}{s}[-\gamma_p e G_m \mathcal{E}] \tag{4.37}$$

$$K_i = \frac{1}{s}[\frac{-\gamma_i}{s} e G_m \mathcal{E}] \tag{4.38}$$

$$K_d = \frac{1}{s}[s\gamma_d e G_m y] \tag{4.39}$$

4.3.3. Design of MRAC based PID controller using modified MIT rule

To drive adaptation rules for the controller parameter vector $\theta = (k_p, k_i, k_d)$ of control law equation (4.9) using modified MIT rule.

Plant transfer function from equation (4.12) written as:

$$\frac{Y(s)}{U_c(s)} = \frac{(G_p(s)k_p + \frac{G_p(s)k_i}{s})}{\left(1 + G_p(s)k_p\right) + \frac{G_p(s)k_i}{s} + G_p(s)k_d s} \tag{4.40}$$

The model transfer function from equation (4.1) can be written as follow:

$$\frac{Y_m(s)}{U_c(s)} = \frac{b_m}{s^2 + a_{m1}s + a_{m0}} \tag{4.41}$$

Model of tracking error is given by:

$$e = y - y_m \tag{4.42}$$

$$e = \frac{(G_p(s)k_p + \frac{G_p(s)k_i}{s})U_c(s)}{\left(1 + G_p(s)k_p\right) + \frac{G_p(s)k_i}{s} + G_p(s)k_d s} - \frac{b_m U_c(s)}{s^2 + a_{m1}s + a_{m0}} \tag{4.43}$$

It is know that, the normalized algorithm modifies the adaptation law in the following manner [12].

$$\frac{\partial e}{\partial \theta} = -\varphi \tag{4.44}$$

$$\frac{d\theta}{dt} = \gamma \frac{e\varphi}{\alpha + \varphi^T \varphi} \tag{4.45}$$

where $\theta =$ Controller parameter vector

$e =$ Output error (error between output of the plant and reference model)

$\gamma =$ Adaptation gain

$\frac{\partial e}{\partial \theta} =$ Sensitive derivative of error with respect to θ

$\alpha =$ Constant

Then, by differentiating the output error (e) in (4.43) with respect to controller parameters (k_p, k_i, k_d) and it can be written in (4.46), (4.47) and (4.48) respectively.

$$\frac{\partial y}{\partial k_p} = \frac{-G_p(s)}{\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(U_c(s) - Y(s)) \tag{4.46}$$

$$\frac{\partial y}{\partial k_i} = \frac{\frac{-G_p(s)}{s}}{\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(U_c(s) - Y(s)) \tag{4.47}$$

$$\frac{\partial y}{\partial k_d} = \frac{G_p(s)s}{\left(1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s\right)}(Y(s)) \tag{4.48}$$

An approximation made which valid when parameters are closed to ideal value is expressed in equation (4.49) as follows:

$$\frac{G_p(s)}{1 + G_p(s)k_p + \frac{G_p(s)k_i}{s} + G_p(s)k_d s} \approx \; G_m(s) \tag{4.49}$$

Then after substituting equation (4.49) in to equation (4.46), (4.47) and (4.48) we can get the simplified form of the equations in equation (4.50), (4.51) and (4.52) respectively.

$$\frac{\partial y}{\partial k_p} = -G_m \mathcal{E} \tag{4.50}$$

$$\frac{\partial y}{\partial k_i} = \frac{-1}{s} G_m \mathcal{E} \tag{4.51}$$

$$\frac{\partial y}{\partial k_d} = sG_m(y) \tag{4.52}$$

Actually, we know that from the general equation (4.45) the controller parameters of the normalized algorithm can be determined as follows:

$$\frac{dk_p}{dt} = -\gamma_p \frac{eG_m \mathcal{E}}{\alpha + (G_m \mathcal{E})^2} \tag{4.53}$$

$$\frac{dk_i}{dt} = \frac{-\gamma_i}{s} \frac{eG_m \mathcal{E}}{\alpha + (G_m \mathcal{E})^2} \tag{4.54}$$

$$\frac{dk_d}{dt} = s\gamma_d \frac{eG_m y}{\alpha + (G_m y)^2} \tag{4.55}$$

Also, by removing the derivative term of the above equation (4.53), (4.54) and (4.55) we get as written in equation (4.56), (4.57) and (4.58) respectively.

$$k_p = \frac{1}{s} [-\gamma_p \frac{eG_m \mathcal{E}}{\alpha + (G_m \mathcal{E})^2}] \tag{4.56}$$

$$k_i = \frac{1}{s} [\frac{-\gamma_i}{s} \frac{eG_m \mathcal{E}}{\alpha + (G_m \mathcal{E})^2}] \tag{4.57}$$

$$k_d = \frac{1}{s} [s\gamma_d \frac{eG_m y}{\alpha + (G_m y)^2}] \tag{4.58}$$

4.3.4. Design of MRAC based PID controller using Lyapunov rule

To drive adaptation rules for the controller parameters vector θ (k_p, k_i, k_d) using the control law of equation (4.9) in a Lyapunov rule.

Plant transfer function from equation (4.10) written as:

$$d^2 y = -a_1 dy - a_2 y + bu \tag{4.59}$$

The reference model transfer function from equation (4.1)

$$d^2y_m = -a_{m1}dy_m - a_{m2}y_m + b_m u_c \tag{4.60}$$

The output error is given by:

$$e = y - y_m \tag{4.61}$$

Since we are trying to make the error small, it is natural to drive a differential equation for the tracking error [25, 26, 27]. We get

$$\frac{d^2e}{dt^2} = \frac{d^2}{dt^2}(y - y_m) \tag{4.62}$$

$$\frac{d^2e}{dt^2} = -a_{m1}\frac{de}{dt} - a_{m2}e - (bk_d + a_1 - a_{m1})y' - \left(bk_p + bk_i + a_2 - a_{m2}\right)y + \left(bk_p + bk_i - b_m\right)u_c \tag{4.63}$$

We will attempt to construct a parameter adjustment mechanism that will derive the parameters k_p, k_i and $, k_d$ to their desired values. For this purpose, assume that $b\gamma > 0$ and introduce the following quadratic function.

Then the candidate Lyapunov function

$$V\left(e, e', k_p, k_i, k_d\right) = \frac{1}{2}[e'^2 + a_{m2}e^2 + \frac{1}{b\gamma}(bk_d + a_1 - a_{m1})^2 + \frac{1}{b\gamma}\left(bk_p + bk_i + a_2 - a_{m2}\right)^2 + \frac{1}{b\gamma}\left(bk_p + bk_i - b_m\right)^2] \tag{4.64}$$

Where $a_1, a_2, b_1$and a_{m1}, a_{m2}, b_m are constants of plant and model parameters.

For this function to qualify as Lyapunov function, the derivative must be negative.

$$\frac{dV}{dt} = e'\frac{d^2e}{dt^2} + a_{m2}e\frac{de}{dt} + \frac{1}{\gamma}\left(bk_p + bk_i + a_2 - a_{m2}\right)\frac{dk_p}{dt} + \left(bk_p + bk_i - b_m\right)\frac{dk_p}{dt} + \frac{1}{\gamma}\left(bk_p + bk_i + a_2 - a_{m2}\right)\frac{dk_i}{dt} + \left(bk_p + bk_i - b_m\right)\frac{dk_i}{dt} + \frac{1}{\gamma}(bk_d + a_1 - a_{m1})\frac{dk_d}{dt} \tag{4.65}$$

The parameters are updated as:

$$\frac{dk_p}{dt} = -\gamma_p e' \mathcal{E} \tag{4.66}$$

$$\frac{dk_i}{dt} = -\gamma_i e' \mathcal{E} \tag{4.67}$$

$$\frac{dk_d}{dt} = \gamma_d e' y' \tag{4.68}$$

When we remove the derivative term from equations (4.66), (4.67) and (4.68) above we get:

$$k_p = -\frac{\gamma_p}{s} e\mathcal{E} \tag{4.69}$$

$$k_i = -\frac{\gamma_i}{s} e\mathcal{E} \tag{4.70}$$

$$k_d = \frac{\gamma_d}{s} ey \tag{4.71}$$

Where Ɛ is the error input to PID controller such that; $\mathcal{E} = u_c - y$ and $e = y - y_m$ is output error between plant and model reference. The derivative of V with respect to time is thus negative semidefinite but not negative definite.

$$\frac{dV}{dt} = -a_{m1} {e'}^2 \tag{4.72}$$

$$\frac{d^2V}{dt^2} = -2a_{m1} e' \frac{de'}{dt} \tag{4.73}$$

Where a_{m1} coefficient of reference model which is positive number. This implies V (t) ≤ V (0) and thus that e , k_p, k_i, k_d must be bounded and this implies that $y = e + y_m$ also bounded.

Chapter Five

Result and Discussion

5.1. Introduction

In the above two chapters (3 and 4) mathematical models of the system and controller design were developed. In this chapter, the numerical parameters that had been used for evaluating the performance of the designed controller were introduced and the simulations were simulated using Matlab/Simulink R2019a. Also, the Performance of the speed response had been investigated by varying the plant parameters such as reference input, load torque, and uncertain output disturbance signal. After the formation of the sets of the initial conditions, the results of the performed computer simulations were presented in the form of tables and figures. Finally, the results were discussed in terms of the selected parameters for comparison.

5.2. Simulation Parameters

During the simulation, the first step is to select armature controlled separately excited DC motor parameters and load opposition torque parameters. Selection of armature controlled separately excited DC motor parameters are taken from previously done journal [4] as given in Table 5.1.

Table 5.1: Parameters of armature controlled separately excited DC motor [4].

No	Description	Parameter	Value	Unit
1	Rated voltage	V_a	240	V
2	Armature resistance	R_a	1	Ω
3	Armature inductance	L_a	0.046	H
4	Moment of inertia	J	0.093	$Kg.m^2$
5	Friction coefficient	B	0.08	$m/s/rad$
6	Back emf constant	k_b	0.55	
7	Motor torque constant	k_t	0.55	
8	Rated speed	N	1500	rpm

As we have been stated under chapter four in (4.3), depending on the selection of reference model criteria we had choose the following second-order reference model transfer function. Then, the reference model which expressed in (5.1) has been selected.

$$G_m(s) = \frac{127.667}{s^2 + 22.599s + 127.667} \qquad (5.2)$$

Its response is shown as Figure 5.1.

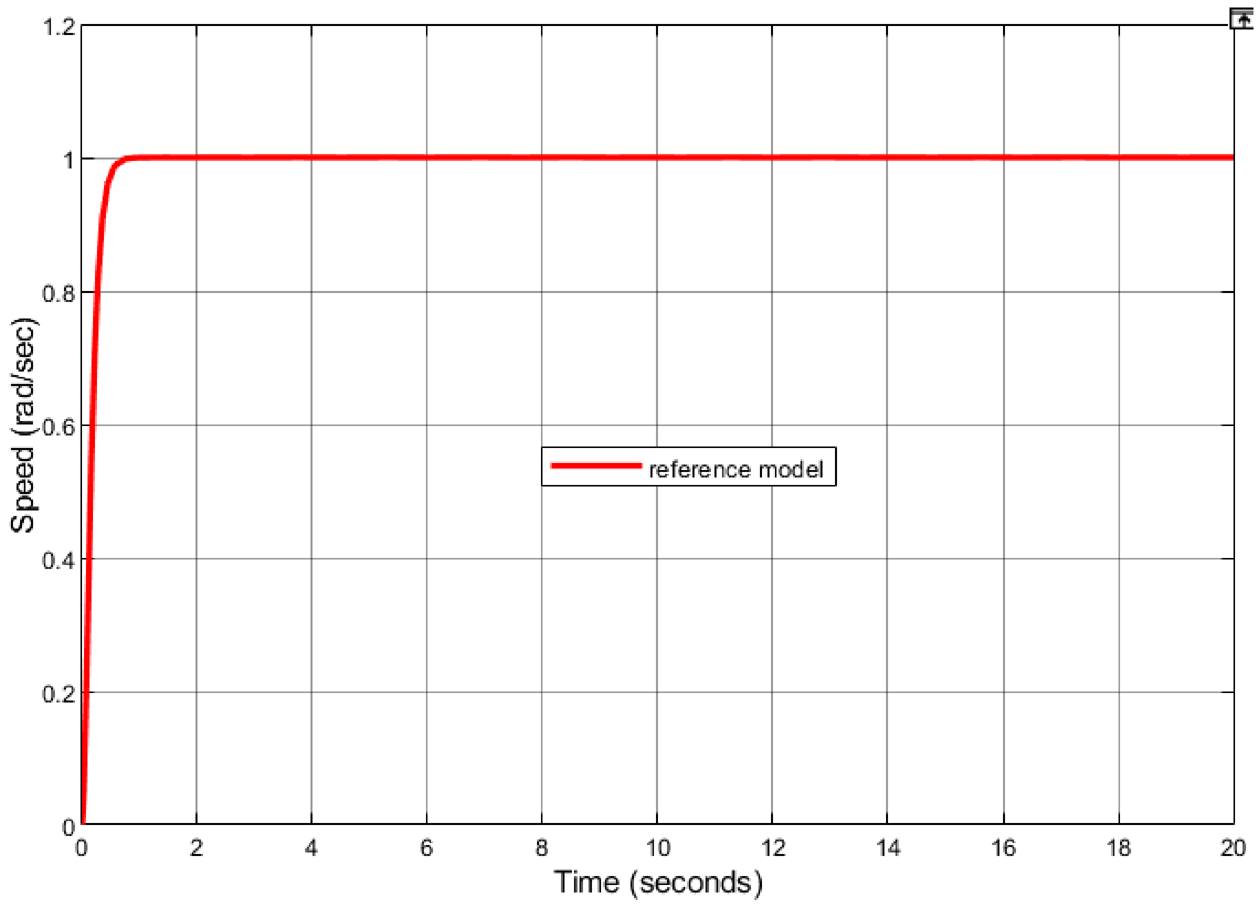

Figure 5.1: Unit step response of reference model

This system has a rise time of 0.297 second, settling time of 1.169 second, 0.505% maximum overshoot and zero steady-state error.

5.3. Response of the system with no load under conventional PID controller

In chapter four different PID tuning methods have been discussed, Matlab automatic PID tuning system block diagram was developed in Simulink library.

The system with conventional PID controller under no-load condition block diagram was developed using Matlab/ Simulink library as shown in Figure 5.2.

For no load condition with the parameters listed in Table 5.1 under conventional PID controller, the block diagram is given by:

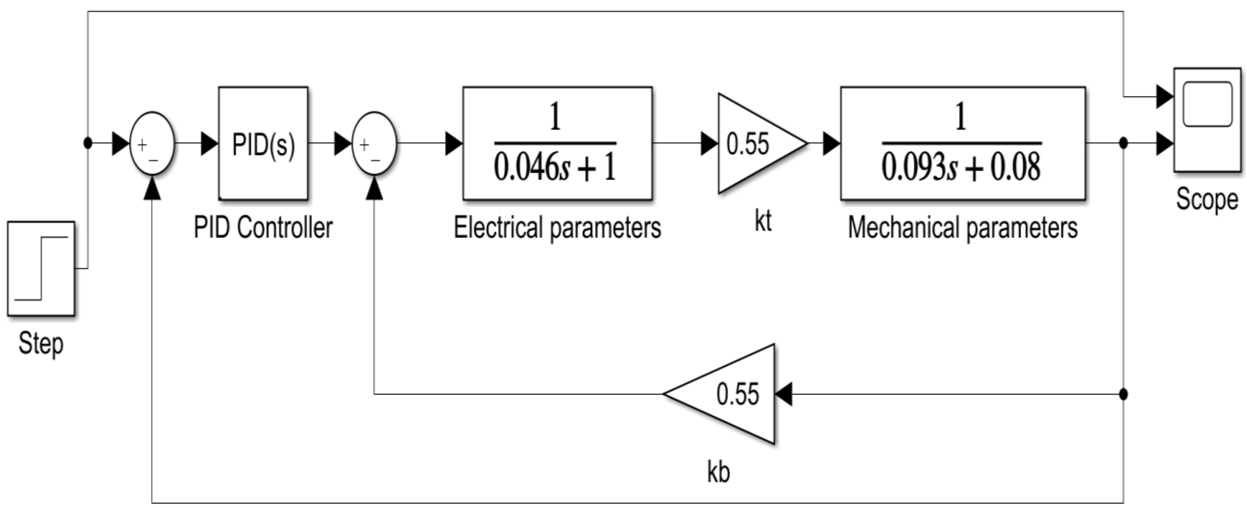

Figure 5.2: Matlab/Simulink model of the system with no load under conventional PID controller

The response of the system under step input and PID gains of $k_p = 0.9328$, $k_i = 0.7164$, and $k_d = -0.07411$ is shown in Figure 5.3.

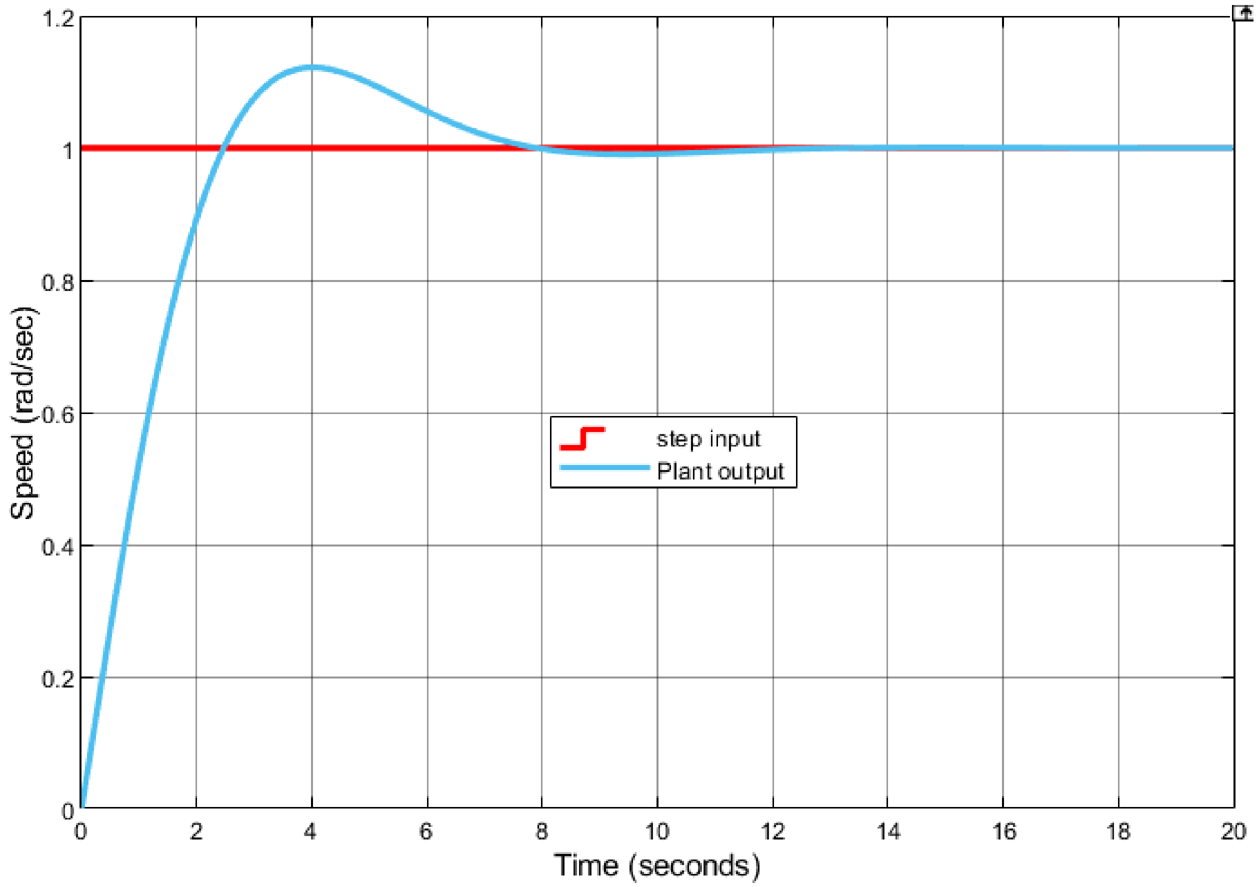

Figure 5.3: Step response of the system with no load under conventional PID controller

This system has a rise time of 2.259 second, settling time of 16.736 second, 8.152% maximum overshoot, and zero steady-state error. From the response of the system, we have seen that it is slowly rising, it takes more time to settle and it has considerable overshoot. This can be improved when we apply a model reference adaptive control tuned PID controller.

As we have seen from Table 5.1 the rated seed is 1500 rpm, which equals 157 rad/sec and when reference input is commanded with this signal the response is as shown in Figure 5.4.

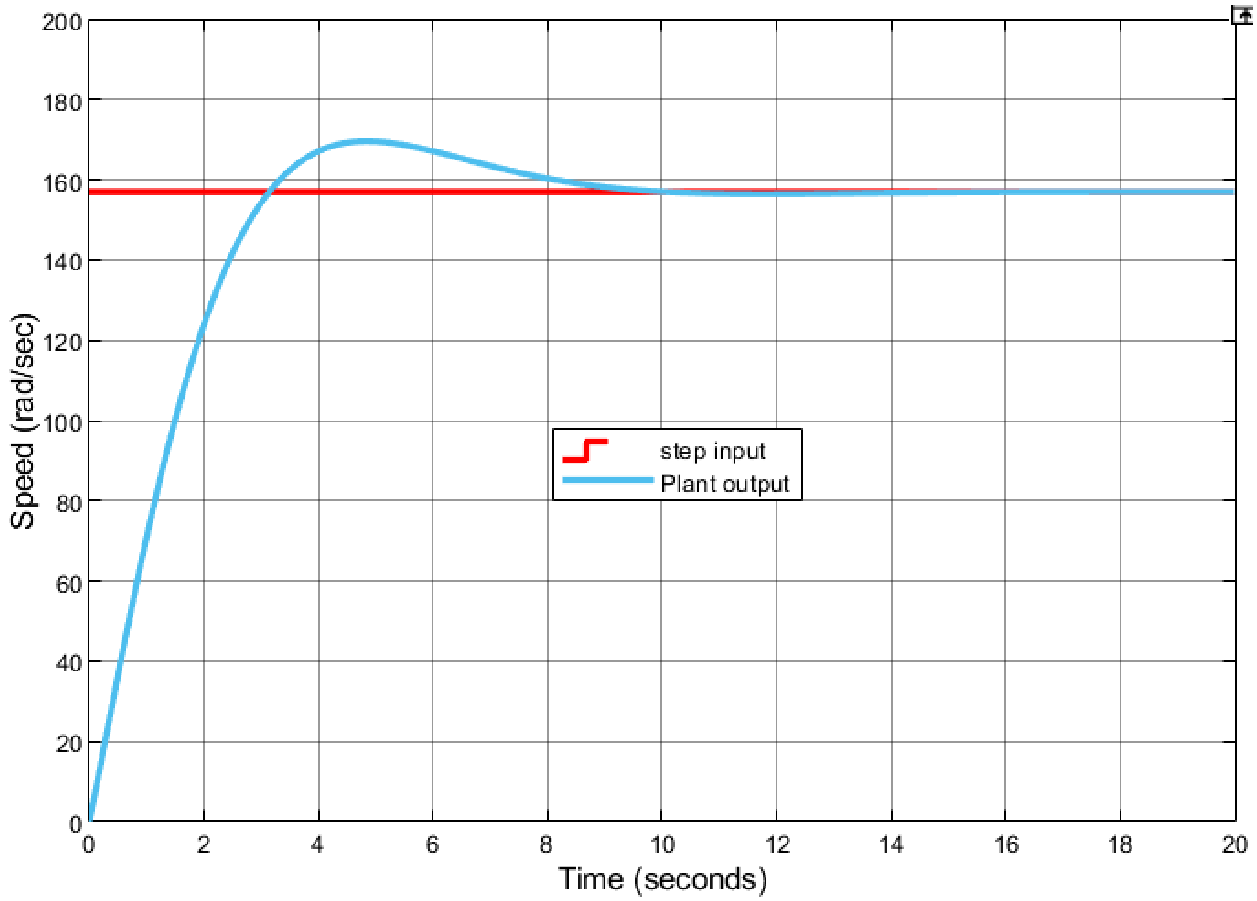

Figure 5.4: Rated speed response of the system with no load under conventional PID controller

This system has a rise time of 1.847 second, settling time of 17.064 second, 11.798% maximum overshoot and zero steady-state error.

5.4. Response of the system with no load under MRAC based PID controller using MIT rule

The Matlab/Simulink block diagram of subsystem contains a system under no load condition for both MIT rule and modified MIT rule was shown in Figure 5.5.

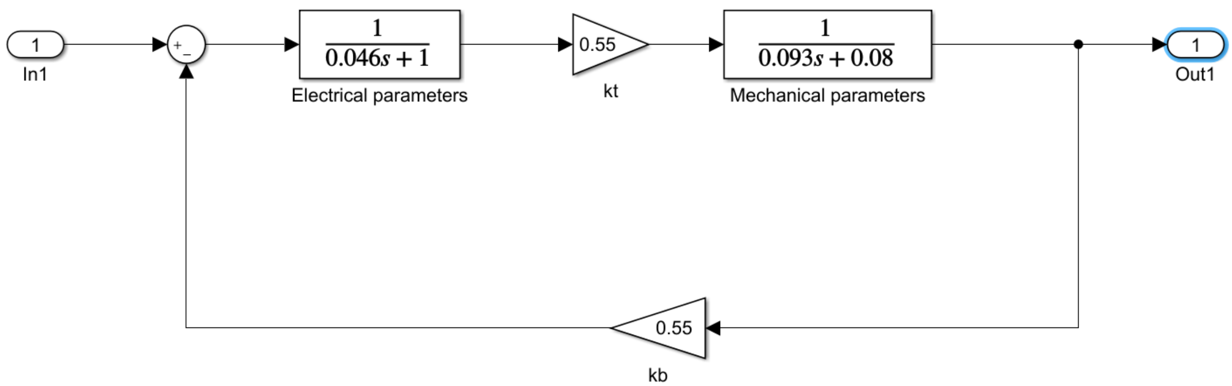

Figure 5.5: Matlab/Simulink model of the system without load

In chapter four, the structure of the controller and parameter adaptation rules have been selected and designed.

The controller parameter adaptation rules were:

$$K_p = \frac{1}{s}[-\gamma_p e G_m \mathcal{E}]$$

$$K_i = \frac{1}{s}[\frac{-\gamma_i}{s} e G_m \mathcal{E}]$$

$$K_d = \frac{1}{s}[s\gamma_d e G_m y]$$

Now the block diagram of the system in combination with the controller in Matlab/Simulink was shown in Figure 5.6.

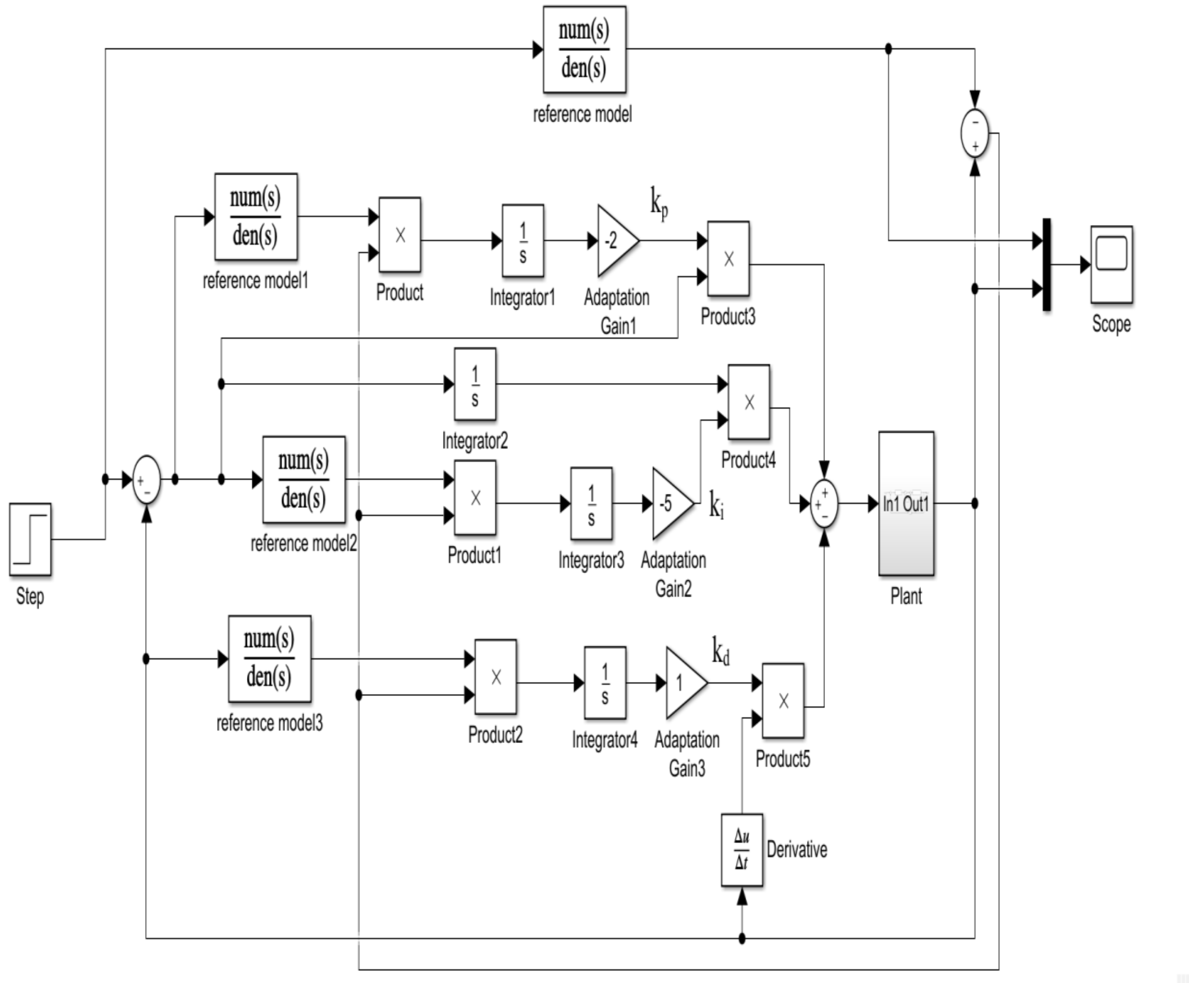

Figure 5.6: Matlab/Simulink model of system with no load under MRAC based PID controller using MIT rule

Unit step response of the system is as shown in Figure 5.7.

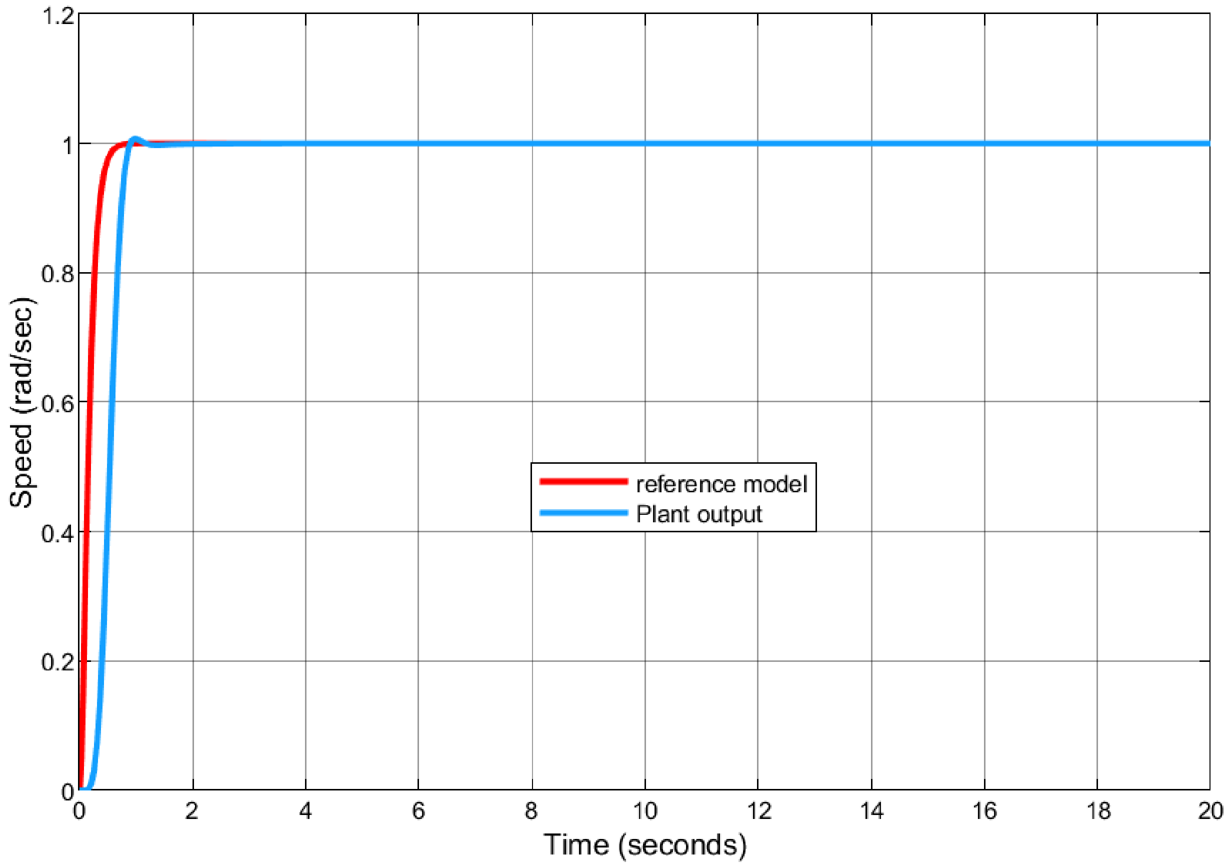

Figure 5.7: Step response of the system with no load under MRAC based PID controller using MIT rule

We can observe that the system has a rise time of 0.407 second, settling time of 4.380 second, 0.508% maximum overshoot and zero steady-state error.

The question comes up when the reference input is increased to a certain value. As it is already stated in chapter one under the introduction part, increasing reference input or other signals to a large value can produce an unstable system unless we use a modified MIT rule.

To see the effect, when the command input is set to 2 rad/sec, the response is shown in Figure 5.8.

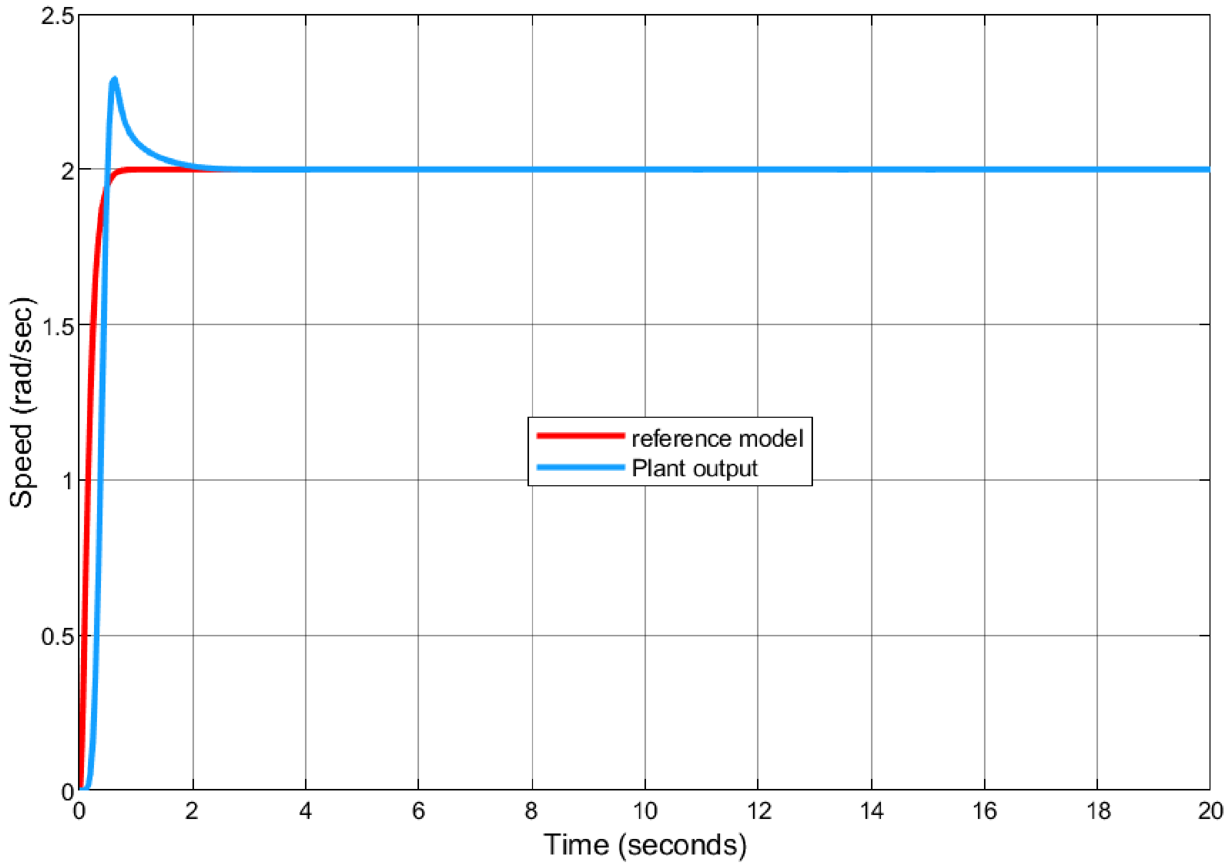

Figure 5.8: Step response of the system with no load under MRAC based PID controller using MIT rule when reference input is increased

As we have seen from the above response that the system has a rise time of 0.217 second, settling time of 3.791 second, 14.368% maximum overshoot and zero steady-state error.

This shows that MRAC using MIT rule can produce an unstable response of the system if the reference input signal is increased. In this case, due to increasing the input reference from 1 rad/sec to 2 rad/sec, the response produces high overshoot with a fast rise time. This can be further improved by using MRAC with modified MIT rule adaptation techniques.

5.5. Response of the system with no load under MRAC based PID controller using modified MIT rule

In chapter four, the controller parameter adjustment mechanism was designed using modified MIT rule. It can be overcome the instability of the system due to an increase in the reference input signal or plant parameter variations.

The modified controller parameter adjustment rules are:

$$k_p = \frac{1}{s}[-\gamma_p \frac{eG_m\mathcal{E}}{\alpha + (G_m\mathcal{E})^2}]$$

$$k_i = \frac{1}{s}[\frac{-\gamma_i}{s}\frac{eG_m\mathcal{E}}{\alpha + (G_m\mathcal{E})^2}]$$

$$k_d = \frac{1}{s}[s\gamma_d \frac{eG_m y}{\alpha + (G_m y)^2}]$$

The block diagram is given as shown in Figure 5.9.

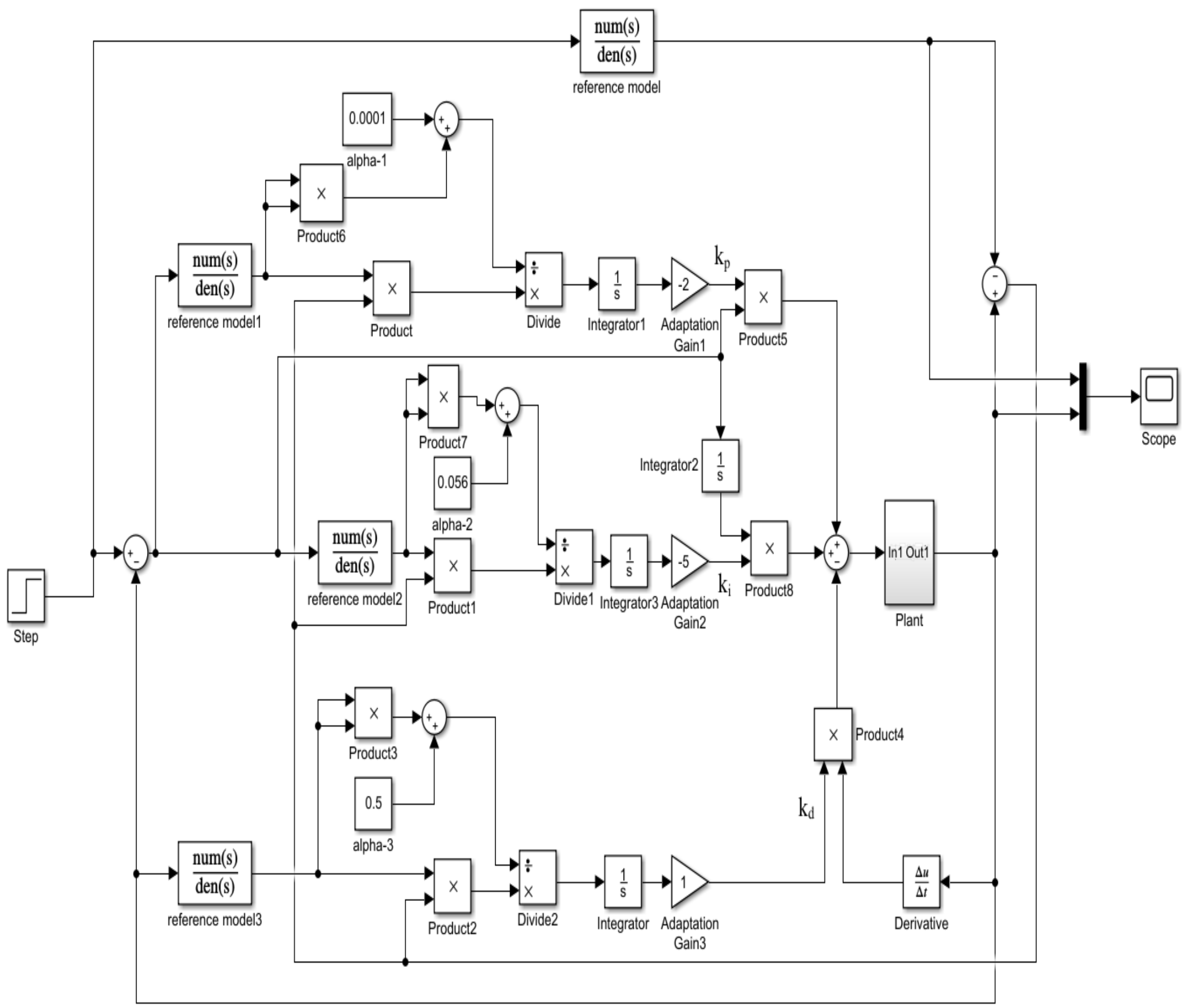

Figure 5.9: Matlab/Simulink model of the system with no load under MRAC based PID controller using modified MIT rule

The response is as shown in Figure 5.10.

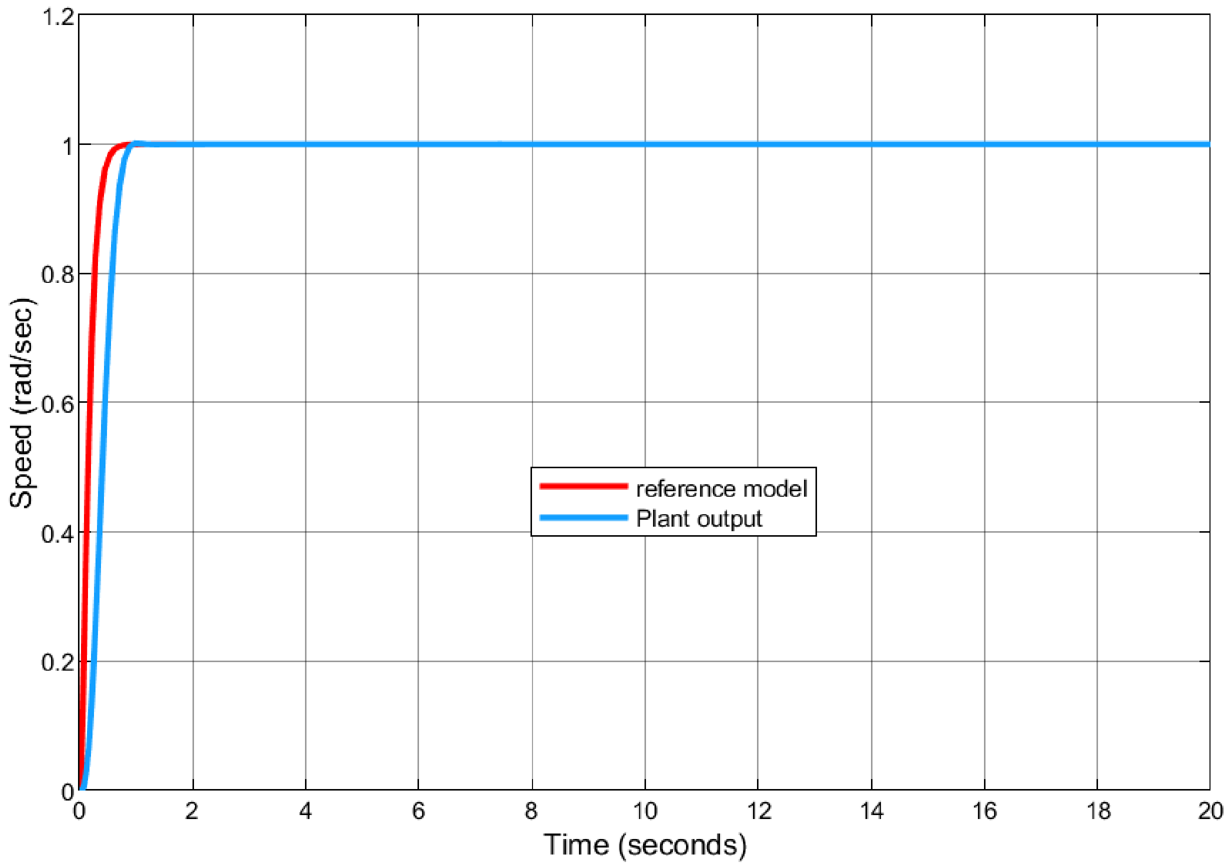

Figure 5.10: Step response of the system with no load under MRAC based PID controller using modified MIT rule

The system has a rise time of 0.473 second, settling time of 3.110 second, 0.487% maximum overshoot and zero steady-state error.

This can be indicates that the modified MIT rule is preferred than MIT rule because of the system response track the reference model output with small settling time and reduced overshoot under the same adaptation gains. As well when an increment of an input signal with the same adaptation gain is taken the modified MIT rule gives the stable system than the MIT rule.

Also to see the effect, when the command input is set to 2 rad/sec, the response is shown in Figure 5.11.

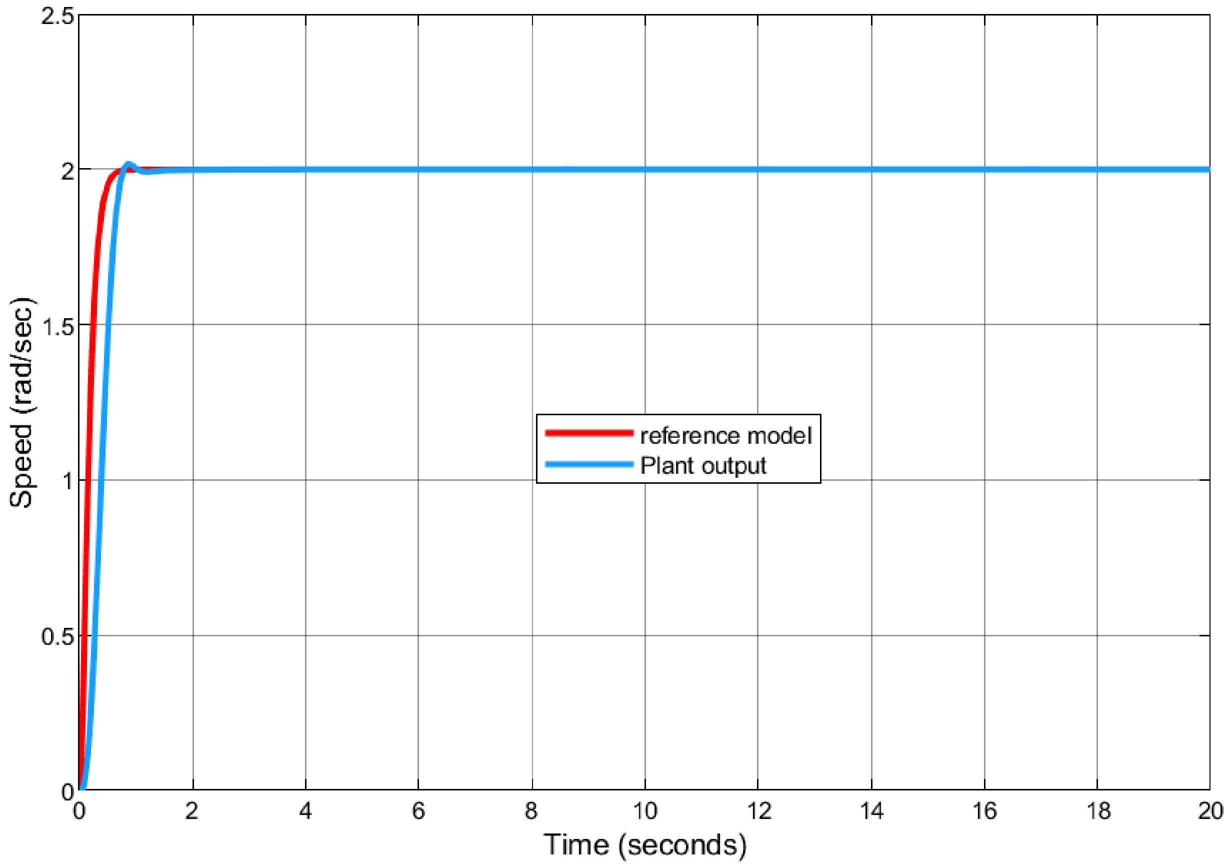

Figure 5.11: Step response of the system with no load under MRAC based PID controller using modified MIT rule when reference input is increased

This system has a rise time of 0.438 second, settling time of 3.100 second, 0.505% maximum overshoot and zero steady-state error.

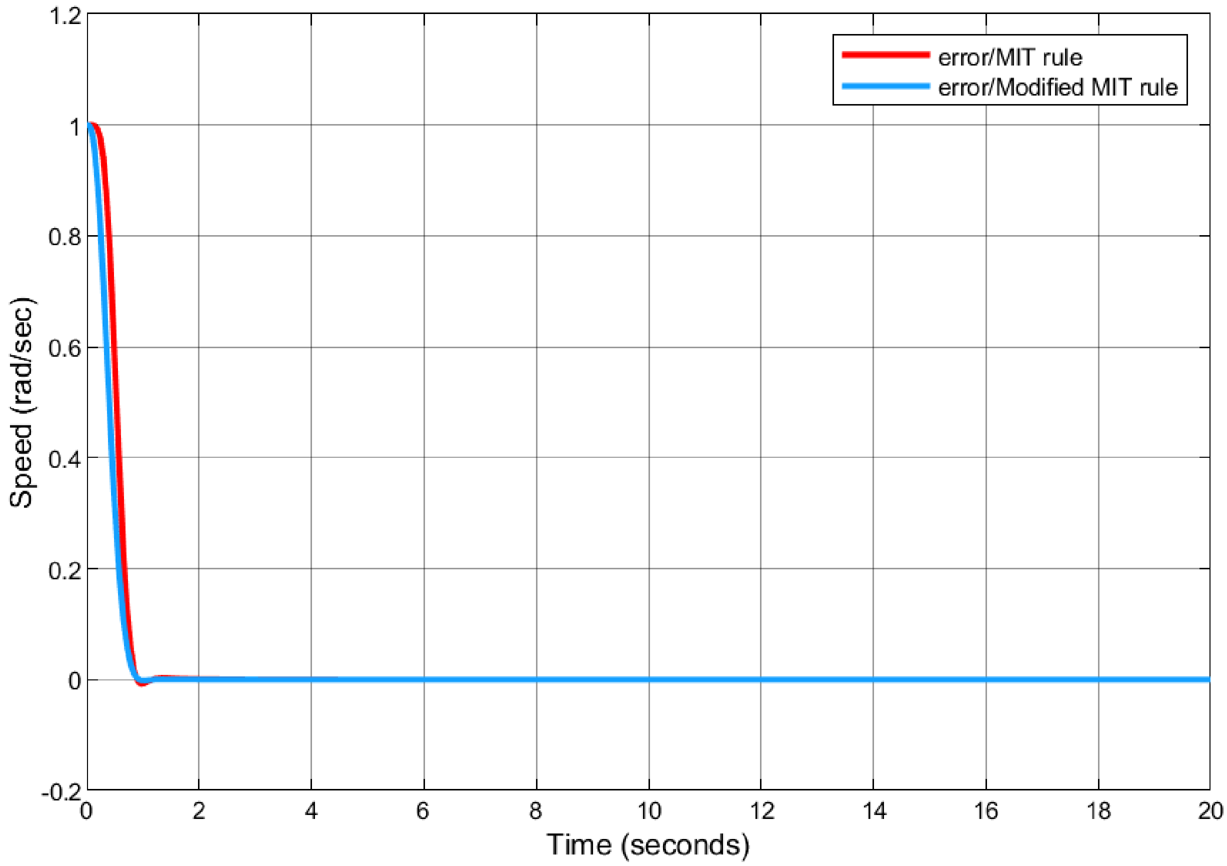

Figure 5.12: Step response error under the same axis with no load

As we have seen from Figure 5.12 of the tracking error response, the modified MIT rule converges the error to zero in fast. While the MIT rule is delayed to converges the tracking error to zero with more disturbance.

From Table 5.1 the rated speed is 1500 rpm, which equals 157 rad/sec and when reference input is commanded with this signal the response is as shown in Figure 5.13.

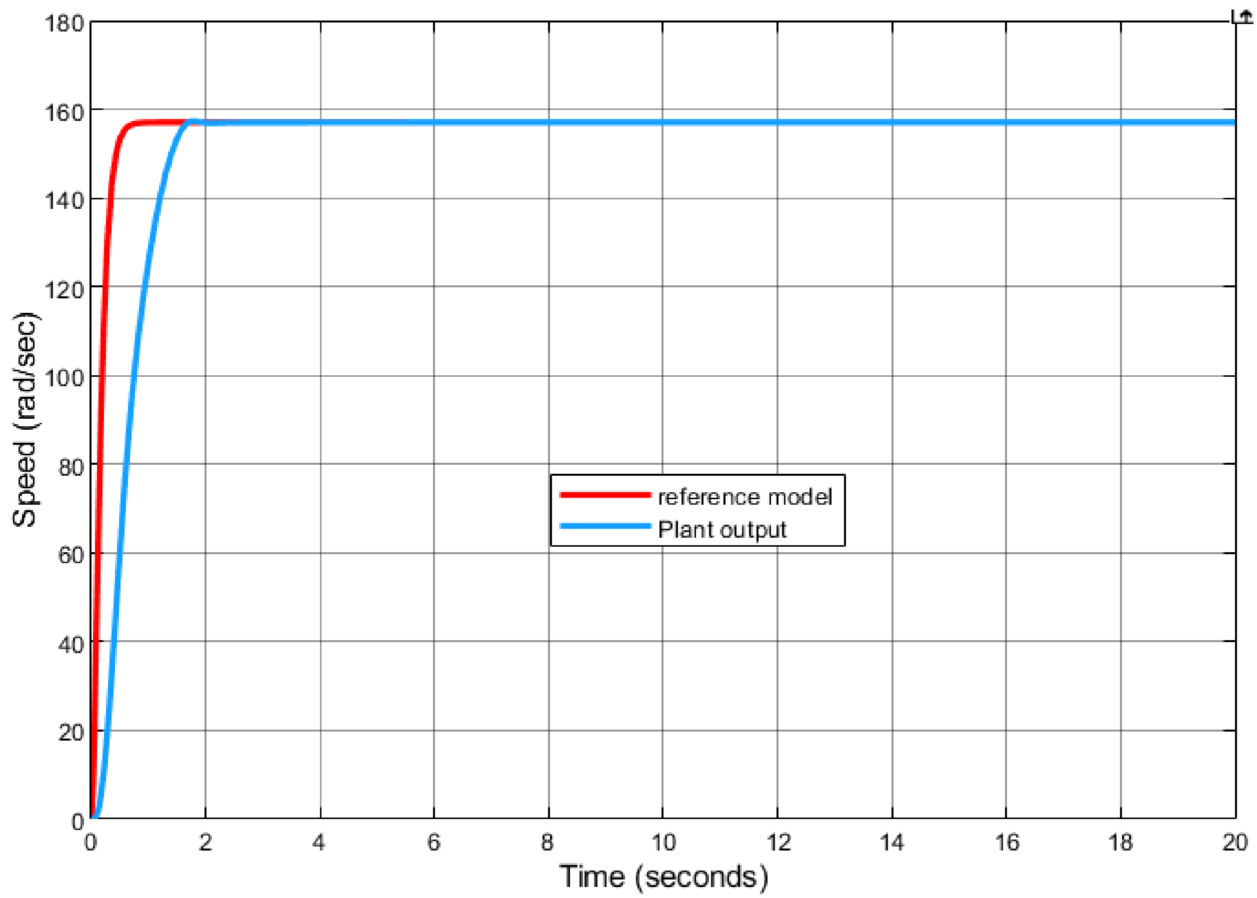

Figure 5.13: Rated speed response of the system with no load under MRAC based PID controller using modified MIT rule

This system has a rise time of 0.956 second, settling time of 2.828 second, 0.487% maximum overshoot and zero steady-state error.

5.6. Response of the system with load under MRAC based PID controller using MIT rule

As we have discussed in chapter three when load torque applied to the system it affects the speed response of the system. Here we have applied the load torque of 0.064 Nm to our system which is commanded to a unit step reference speed.

The Matlab/Simulink block diagram of subsystem contains a system under load torque condition for both MIT rule and modified MIT rule is shown in Figure 5.14.

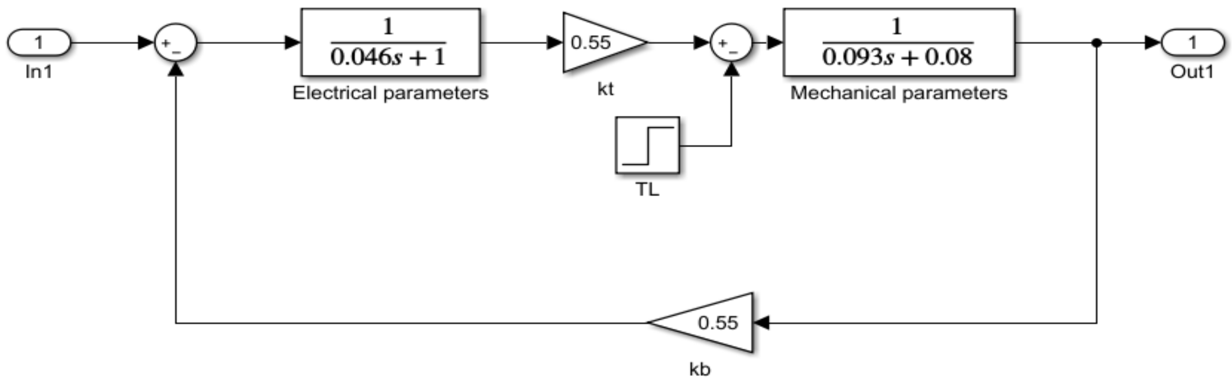

Figure 5.14: Matlab/Simulink model of the system with load

The block diagram of the system under load torque in combination with the controller in Matlab/Simulink is shown in Figure 5.15.

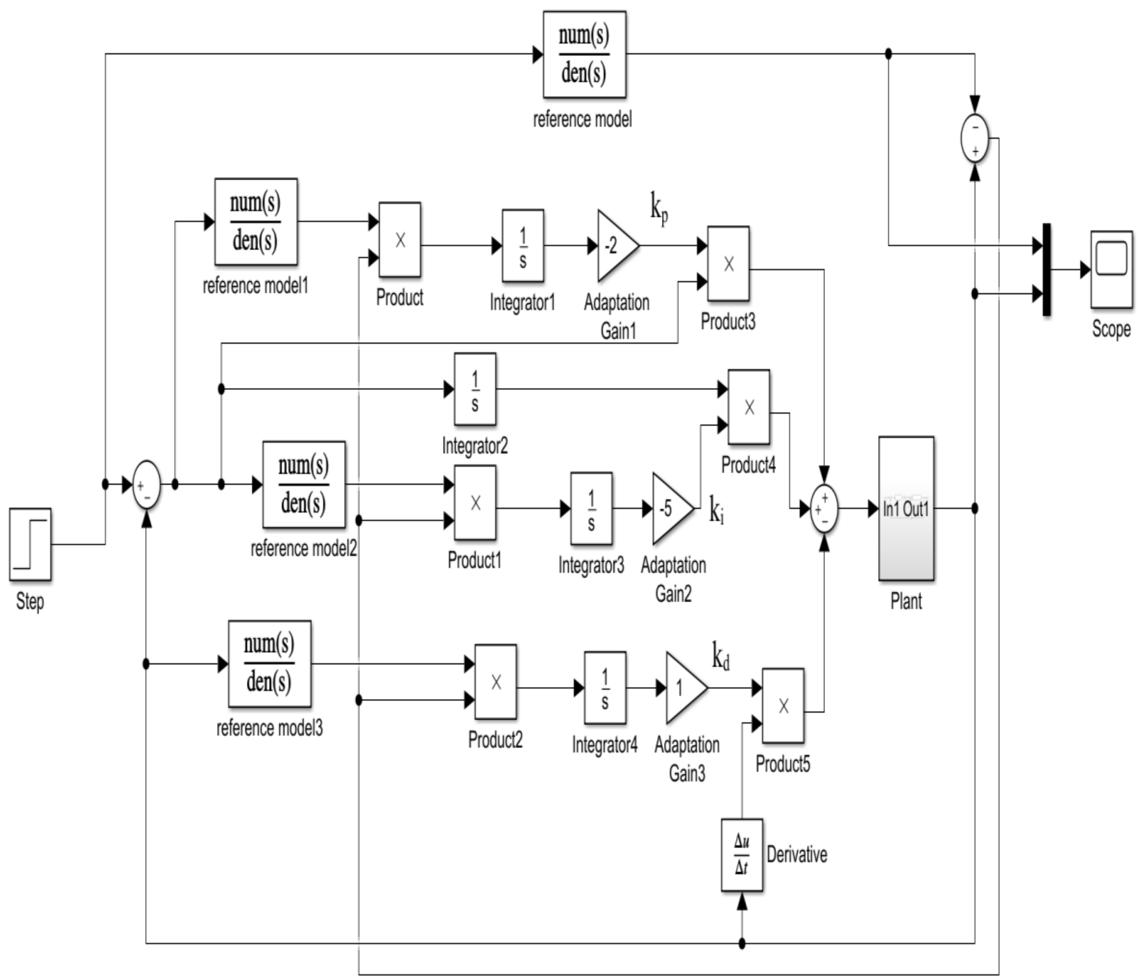

Figure 5.15: Matlab/ Simulink model of the system with load under MRAC based PID controller using MIT rule

Figure 5.16: shows the speed response of the system under load torque having a value of 0.064 Nm when the system is commanded at unit step reference speed.

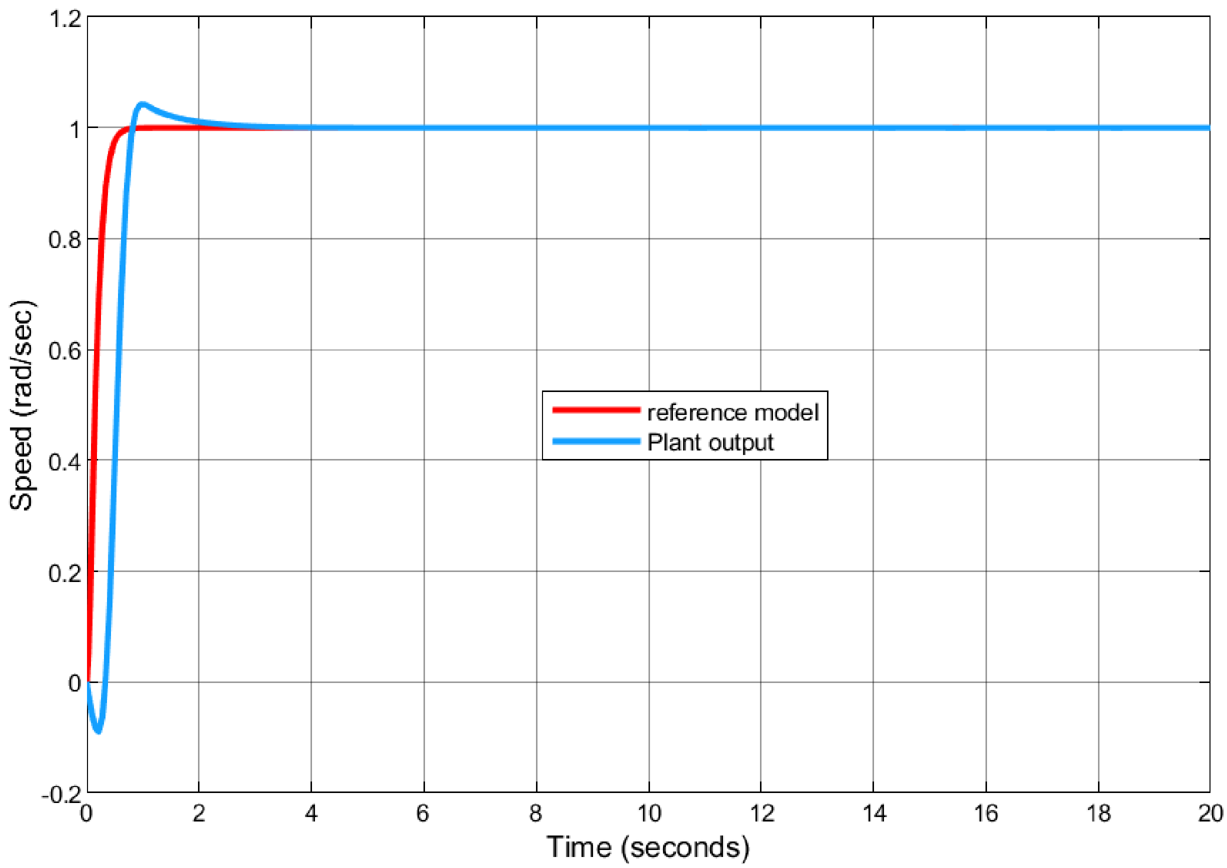

Figure 5.16: Step response of the system with load under MRAC based PID controller using MIT rule

As we see from Figure 5.16, when a load torque of 0.064 Nm is added to the system, the speed response of the system starts to decrease and reaches the minimum amplitude 0.08985 rad/sec at 0.211 second. Then after it have been increased to the desired speed within a rising time of 0.337 second, settling time of 4.103 second, 3.933% maximum overshoot and zero steady-state error. Here we have seen that the speed decline in high magnitude and have an overshoot when a load torque is applied.

5.7. Response of the system with load torque under MRAC based PID controller using modified MIT rule

We have also applied the same load torque of 0.064 Nm to our system which is commanded to a unit step reference speed.

The block diagram of the system under load torque in combination with the controller in Matlab/Simulink is shown in Figure 5.17.

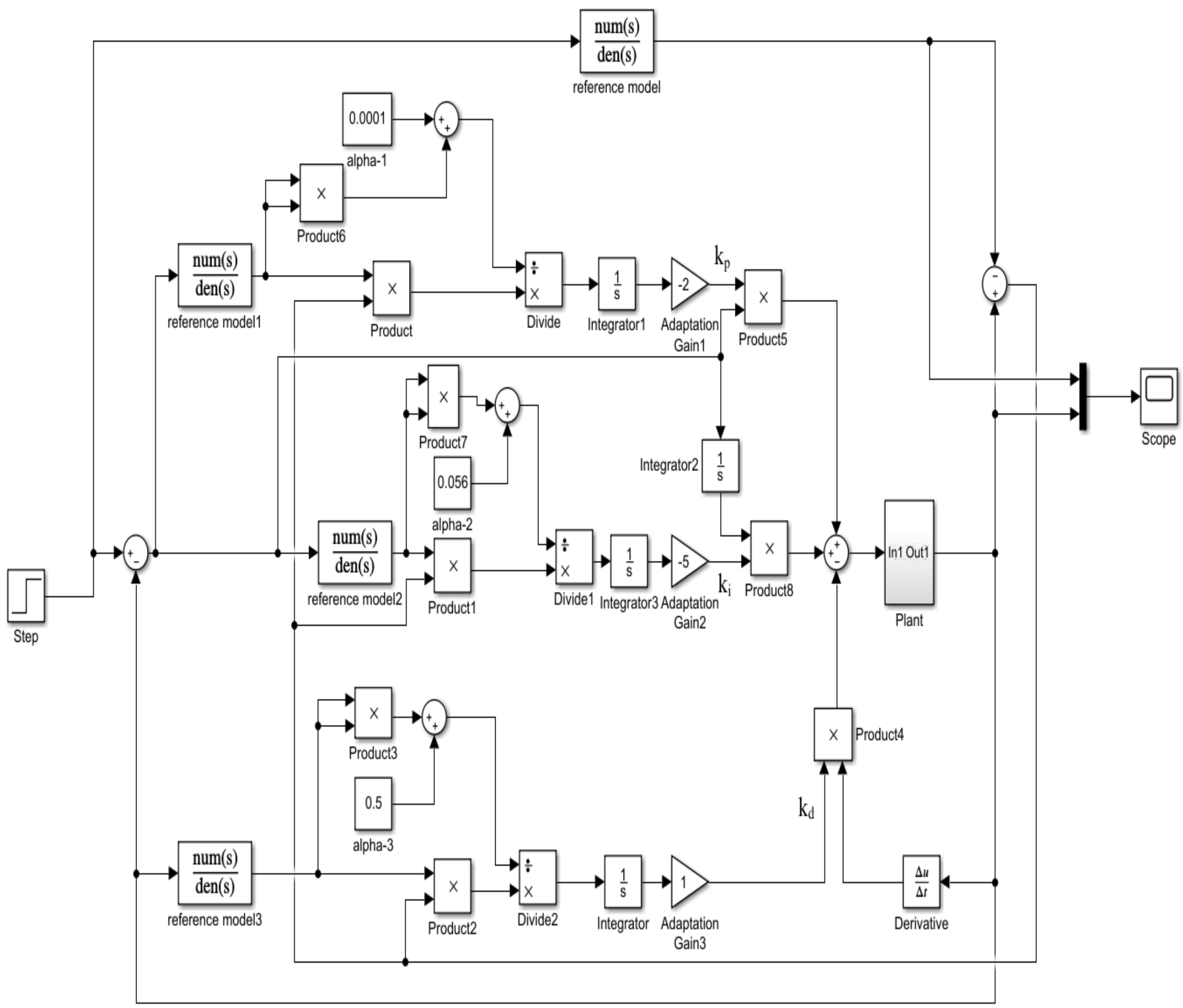

Figure 5.17: Matlab/ Simulink model of the system with load under MRAC based PID controller using modified MIT rule

The response is as shown in Figure 5.18.

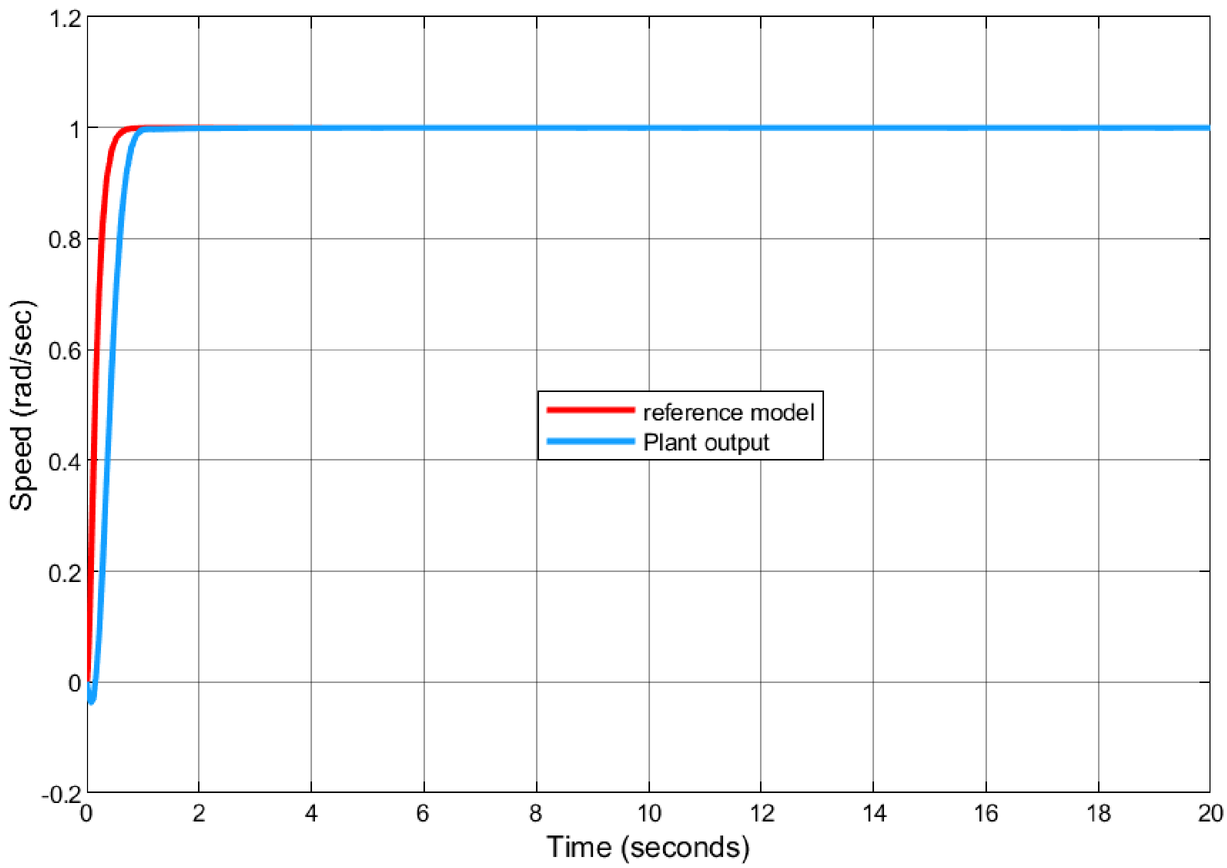

Figure 5.18: Step response of the system with load under MRAC based PID controller using modified MIT rule

Also, when a load torque of 0.064 Nm is applied to the system, the speed response of the system starts to decrease and reaches the minimum amplitude 0.03338 rad/sec at 0.096 second. Then after it have been increased to the desired speed within a rising time of 0.454 second, settling time of 3.001 second, 0.487% maximum overshoot and zero steady-state error. This indicates that when we compare MIT rule and modified MIT rule under load torque applied to the system, the modified MIT rule is preferred because the speed response reduced with less amplitude. That means at the starting time, the MIT rule based control needs high starting current than the modified MIT rule based control system. Moreover, after the speed of the system starts to increase to the reference speed the modified MIT rule based controlling system track the desired speed with the same overshoot as at no-load condition.

5.8. Response of the system with load and uncertain disturbance signal under MRAC based PID controller using MIT rule

In chapter three, we have mentioned that when load torque and uncertain output disturbance signal are applied to the system it affects the speed response of the system. Here we have applied the load torque of 0.064 Nm and uncertain output disturbance signal of 0.032 Nm to our system which is commanded to a unit step reference speed.

The Matlab/Simulink block diagram of subsystem contains a system under load torque and uncertain output disturbance signal condition for both MIT rule and modified MIT rule is shown in Figure 5.19.

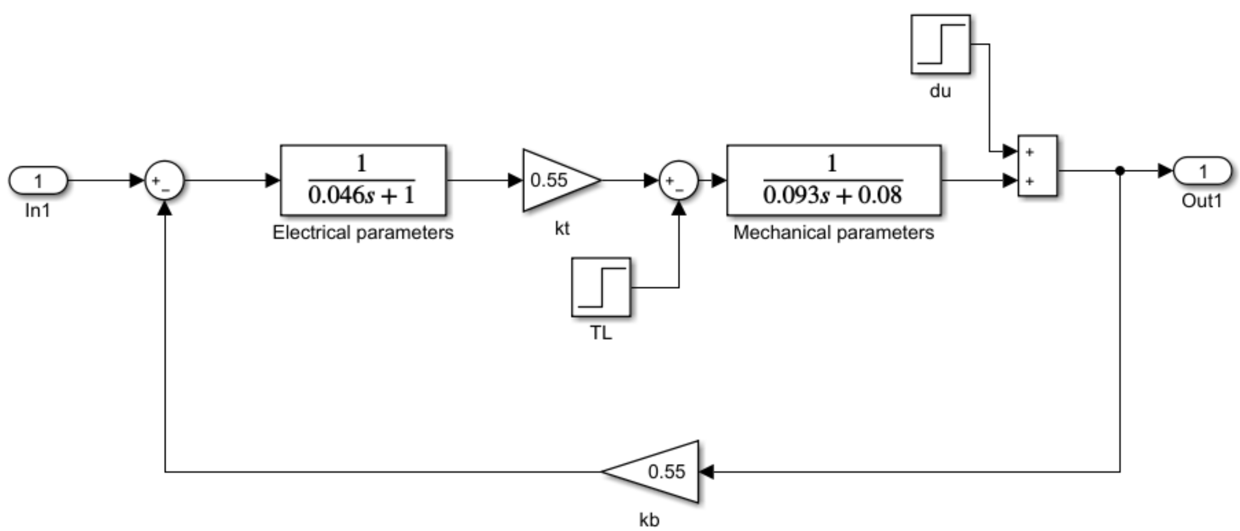

Figure 5.19: Matlab/Simulink model of the system with load torque and uncertain output disturbance signal

The block diagram of the system under load torque and uncertain output disturbance signal in combination with the controller in Matlab/Simulink is shown in Figure 5.20.

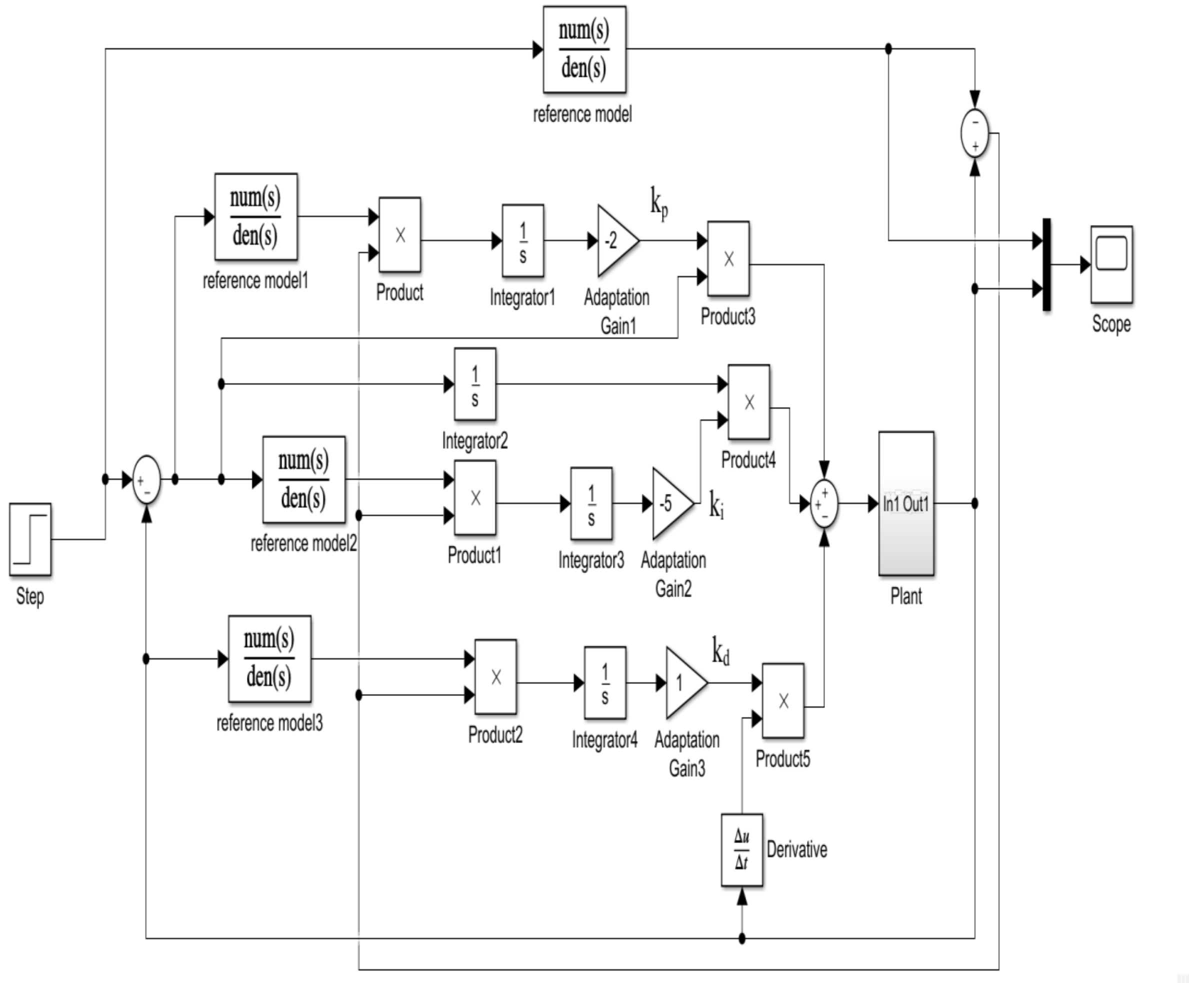

Figure 5.20: Matlab/ Simulink model of the system with load torque and uncertain output disturbance signal under MRAC based PID controller using MIT rule

Figure 5.21: shows the speed response of the system under load torque and uncertain signal when the system is commanded at unit step reference speed.

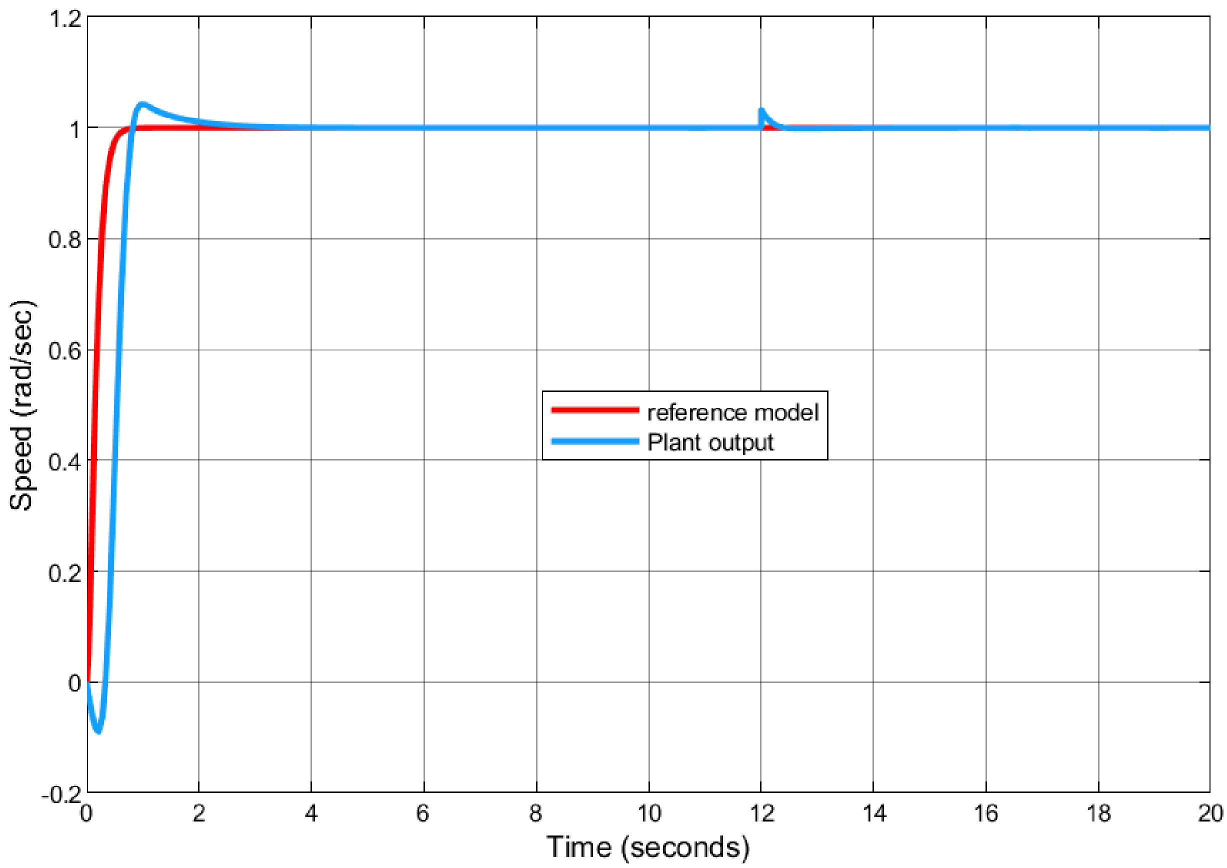

Figure 5.21: Step response of the system with load torque and uncertain output disturbance signal under MRAC based PID controller using MIT rule

As we see from Figure 5.21, when a load torque of 0.064 Nm and uncertain output disturbance signal of 0.032 Nm is added to the system respectively, the speed response of the system starts to decrease and reaches the minimum amplitude 0.08985 rad/sec at 0.211 second. Then after it have been increased to the desired speed within a rising time of 0.337 second, settling time of 4.103 second, 3.933% maximum overshoot and it come back to the reference speed for load torque applied. Also it becomes above the reference speed and reaches the maximum amplitude of 1.032 rad/sec at 12.004 second. Then after the speed becomes decline under the reference speed and reaches the steady-state response at 15.050 second.

5.9. Response of the system with load and uncertain disturbance signal under MRAC based PID controller using modified MIT rule

We have also applied the load torque of 0.064 Nm and uncertain output disturbance signal of 0.032 Nm to our system which is commanded to a unit step reference speed.

The block diagram of the system under load torque and uncertain output disturbance signal in combination with the controller in Matlab/Simulink is shown in Figure 5.22.

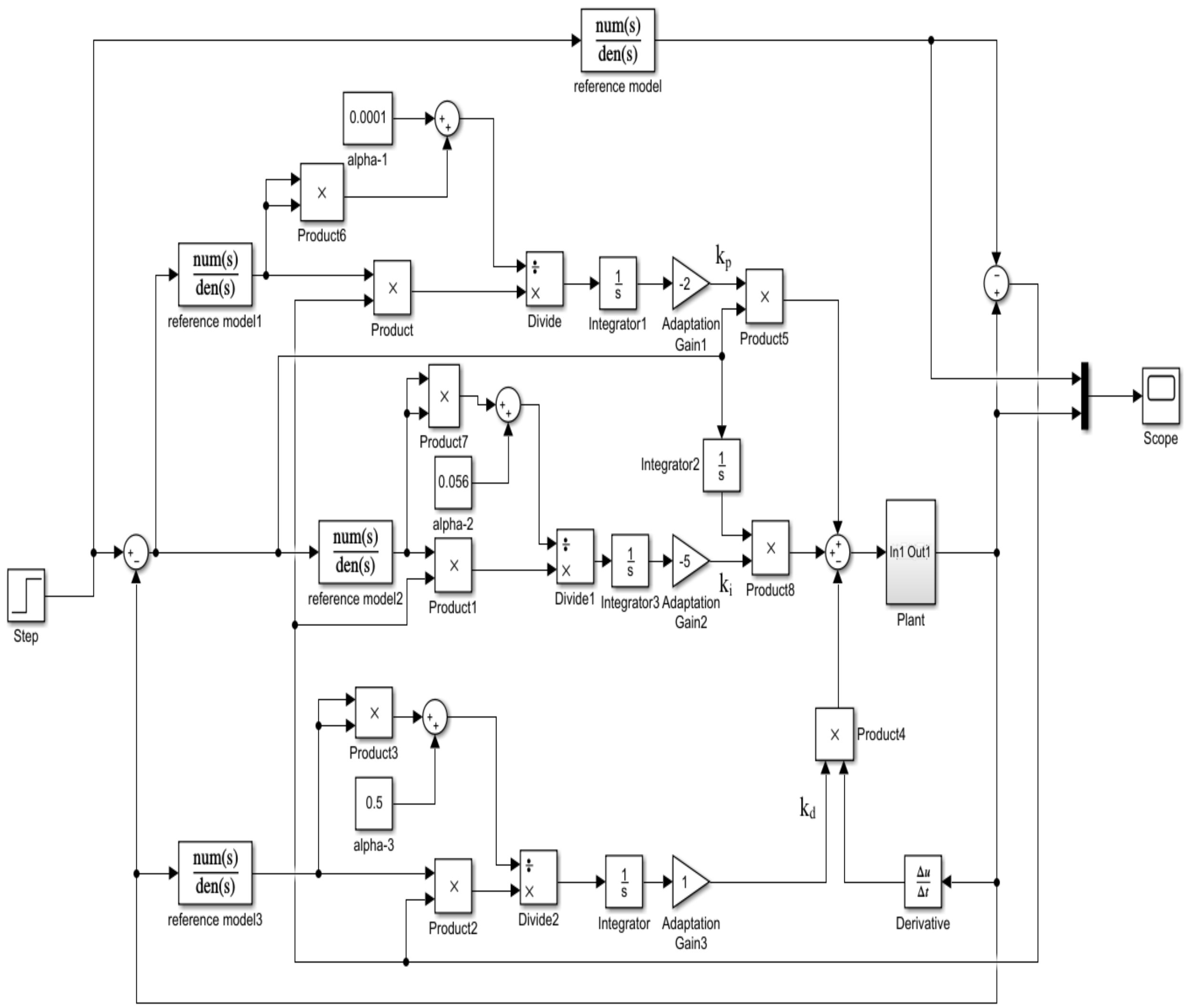

Figure 5.22: Matlab/ Simulink model of the system with load torque and uncertain output disturbance signal under MRAC based PID controller using modified MIT rule

The response is as shown in Figure 5.23.

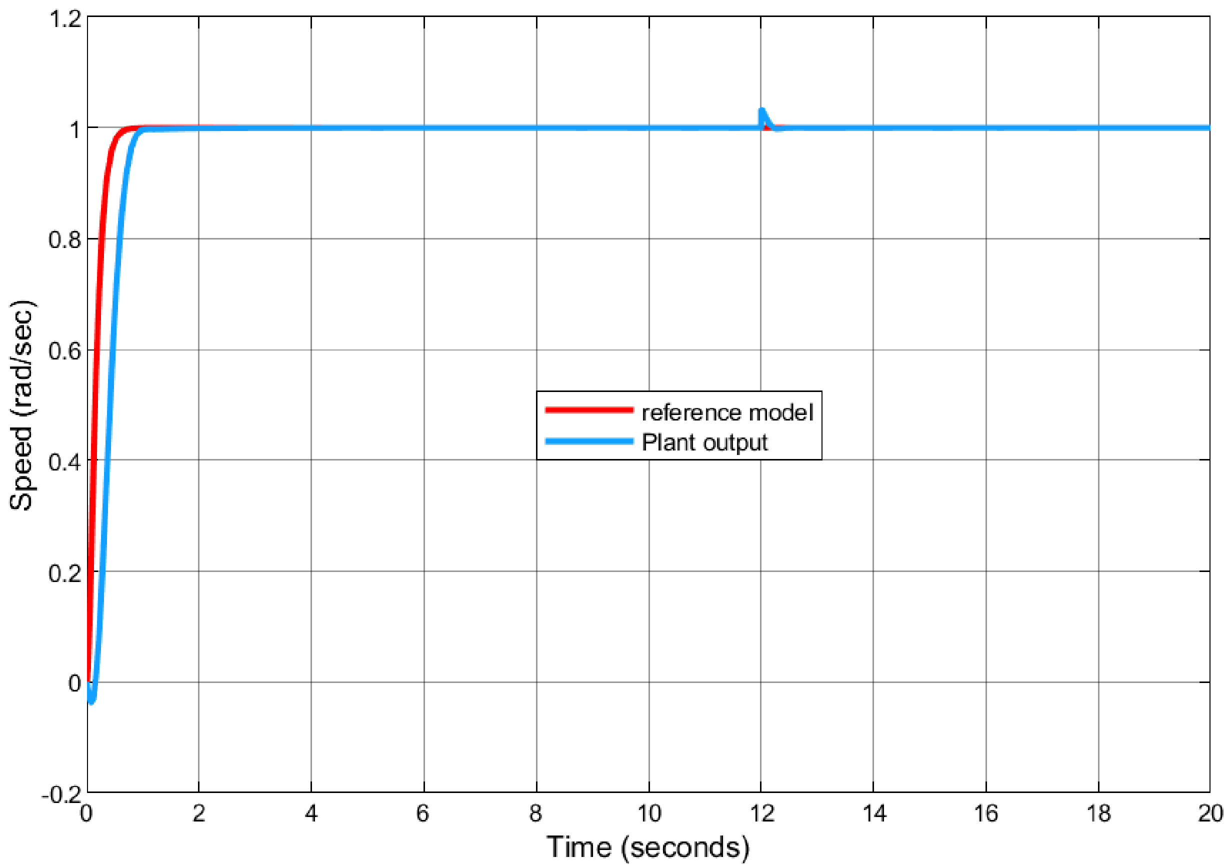

Figure 5.23: Step response of the system with load torque and uncertain output disturbance signal under MRAC based PID controller using modified MIT rule

Here we have seen that, when the same sudden load having the magnitude of 0.064 Nm and uncertain output disturbance signal of 0.032 Nm is added to the system respectively, the speed response of the system starts to decrease and reaches the minimum amplitude 0.03338 rad/sec at 0.096 second. Then after it have been increased to the desired speed within a rising time of 0.454 second, settling time of 3.001 second, 0.487% maximum overshoot and it come back to the reference speed in case of load torque is applied. Also it becomes above the reference speed and reaches the maximum amplitude of 1.032 rad/sec at 12.010 second. Then after the speed become decline under the reference speed and reaches the steady-state response at 12.560 second. When we have seen from the response of the system under load torque and uncertain output disturbance signal the modified MIT rule is preferred than the MIT rule because of the speed response is decreased with less amplitude, has fast settling time and has less overshoot.

Here the system response under the same axis when load torque and uncertain output disturbance signal applied in MIT and modified MIT rule is shown below.

Under the same axis the response are shown in Figure 5.24.

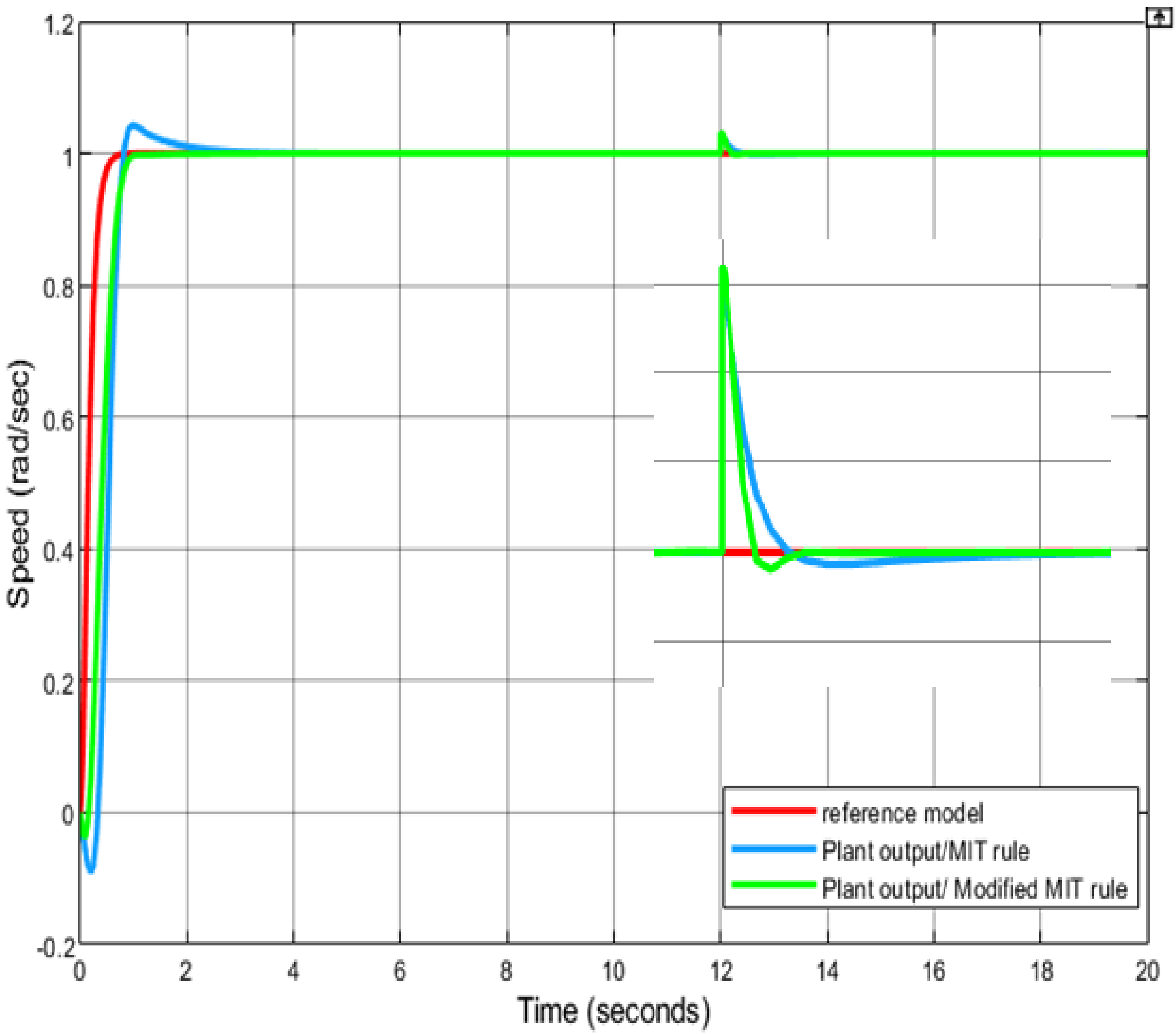

Figure 5.24: System response under same axis for load torque and uncertain output disturbance signal

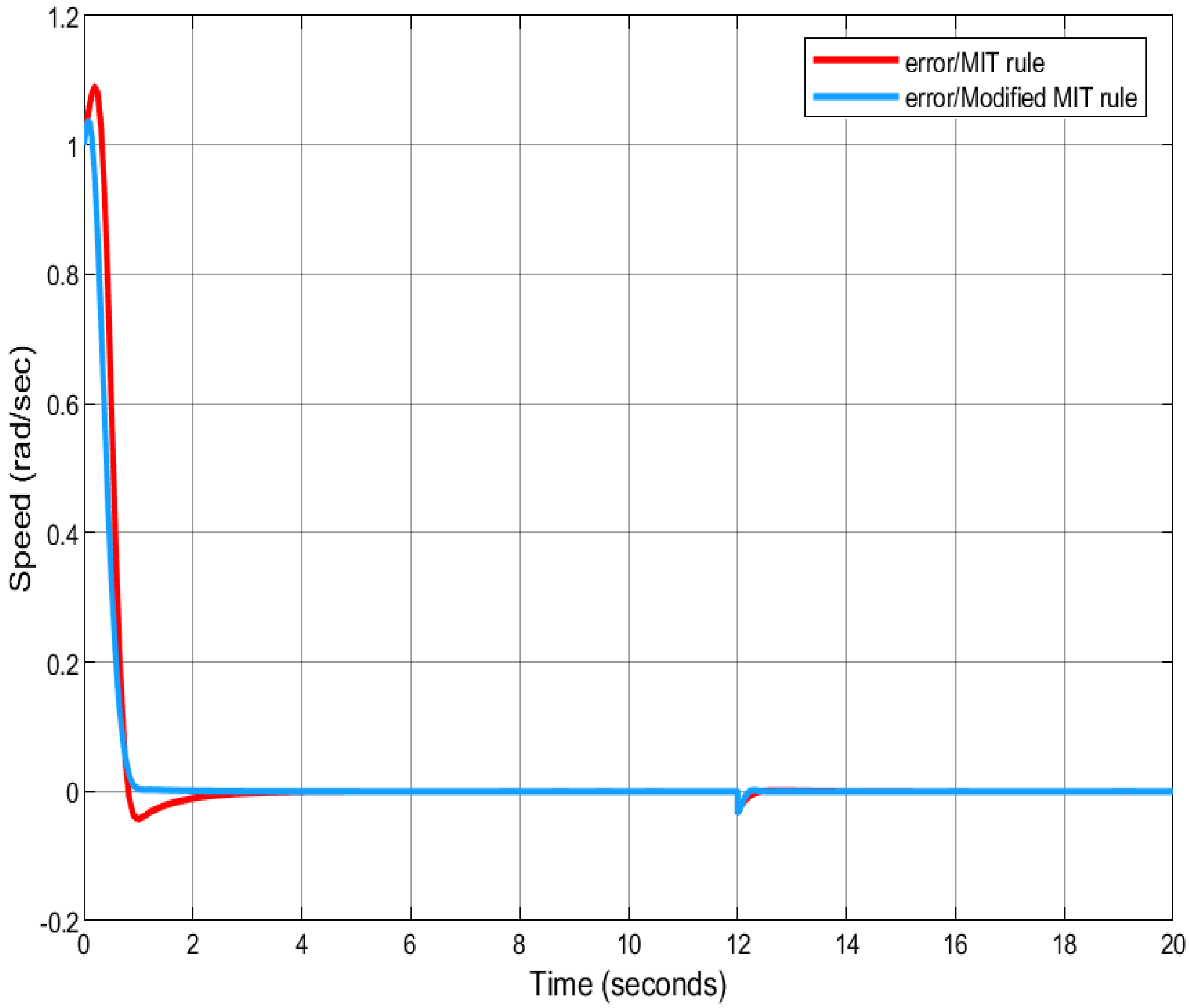

Figure 5.25: Step response error under the same axis for load torque and uncertain output disturbance signal

As we have seen from Figure 5.25 of the error response, the modified MIT rule converges the error to zero in fast. While the MIT rule is delayed to converges the error to zero with more disturbance. Although, when Plant parameter variations exist (load torque and uncertain output disturbance signal) MIT rule can produce a maximum magnitude of error that can take a time to converge to zero. This can be shown that MIT rule is more sensitive than the Modified MIT rule.

5.10. Summary on the simulation results

As we seen from the above response figures a variation of the motor speed with time, the summary of the speed response is given as shown in Table 5.2.

Table 5.2: Step response of the system in different parameters variations summary.

Unit step reference input under no-load				
S. No		MIT rule	Modified MIT rule	PID
1	Rise time	0.407 sec	0.473 sec	2.259 sec
2	Settling time	4.380 sec	3.110 sec	16.736 sec
3	Max. percent overshoot	0.508	0.487	8.152
When reference input increased to 2 rad/sec under no load				
1	Rise time	0.217 sec	0.438 sec	
2	Settling time	3.791 sec	3.100 sec	
3	Max. percent overshoot	14.368	0.505	
When sudden load torque is applied				
1	Time taken for Max. speed fall	0.211 sec	0.096 sec	
2	Max. Speed fall in magnitude	0.08985 rad/sec	0.03338 rad/sec	
3	Rise time	0.337 sec	0.454 sec	
4	Settling time	4.103 sec	3.001 sec	
5	Max. percent overshoot	3.933	0.487	
When output uncertain disturbance signal is happened				
1	Time taken to Max. speed rise	0.004 sec	0.010 sec	
2	Max. speed rise in magnitude	0.032 rad/sec	0.032 rad/sec	
3	Settling time	15.050 sec	12.250 sec	

Generally, From Table 5.2 we have seen that conventional PID controller for speed control of armature controlled separately excited DC motor has slow rising time, high overshoot and it takes more time to settle when we compare with model reference adaptive control based PID controller even under no loaded condition. This shows that the MRAC based PID controller is more preferred than the conventional PID controller for the system speed to track the reference speed in case of reducing the error.

Also, MRAC based PID controller for speed control of separately excited DC motor which discussed under armature controlled techniques of using MIT rule and modified MIT rule the steady-state speed are the same as that of the commanded reference speed for both methods. The system can follow the reference signal with less maximum percent overshoot under a modified MIT rule than the MIT rule for the no-load condition. Also when the load torque and uncertain output disturbance signal applied we have seen that the speed of the system decreased in high magnitude, has high rising time, has high overshoot and it take more time to settle under MIT rule than the modified MIT rule. These all indicate that in the MRAC based PID controller for speed of the system, the modified MIT rule is better than the MIT rule to get a stable system response under parameter variations.

5.11. Response of the system with no load under MRAC based PID controller using Lyapunov stability rule

MRAC with MIT rule and with a modified MIT rule adaptation mechanism does not guarantee the stability of controller parameters of the system. To ensure stability, it is possible to use MRAC with the Lyapunov stability theory adaptation mechanism. The difference between MRAC using the MIT rule and MRAC using the Lyapunov theory, as analyzed in chapter four, is the filter function is removed in the parameter adaptation rule when using the Lyapunov theory.

Then the parameter adaptation rule as follows:

$$k_p = -\frac{\gamma_p}{s} e\mathcal{E}$$

$$k_i = -\frac{\gamma_i}{s} e\mathcal{E}$$

$$k_d = \frac{\gamma_d}{s} ey$$

The block diagram is given in Figure 5.26.

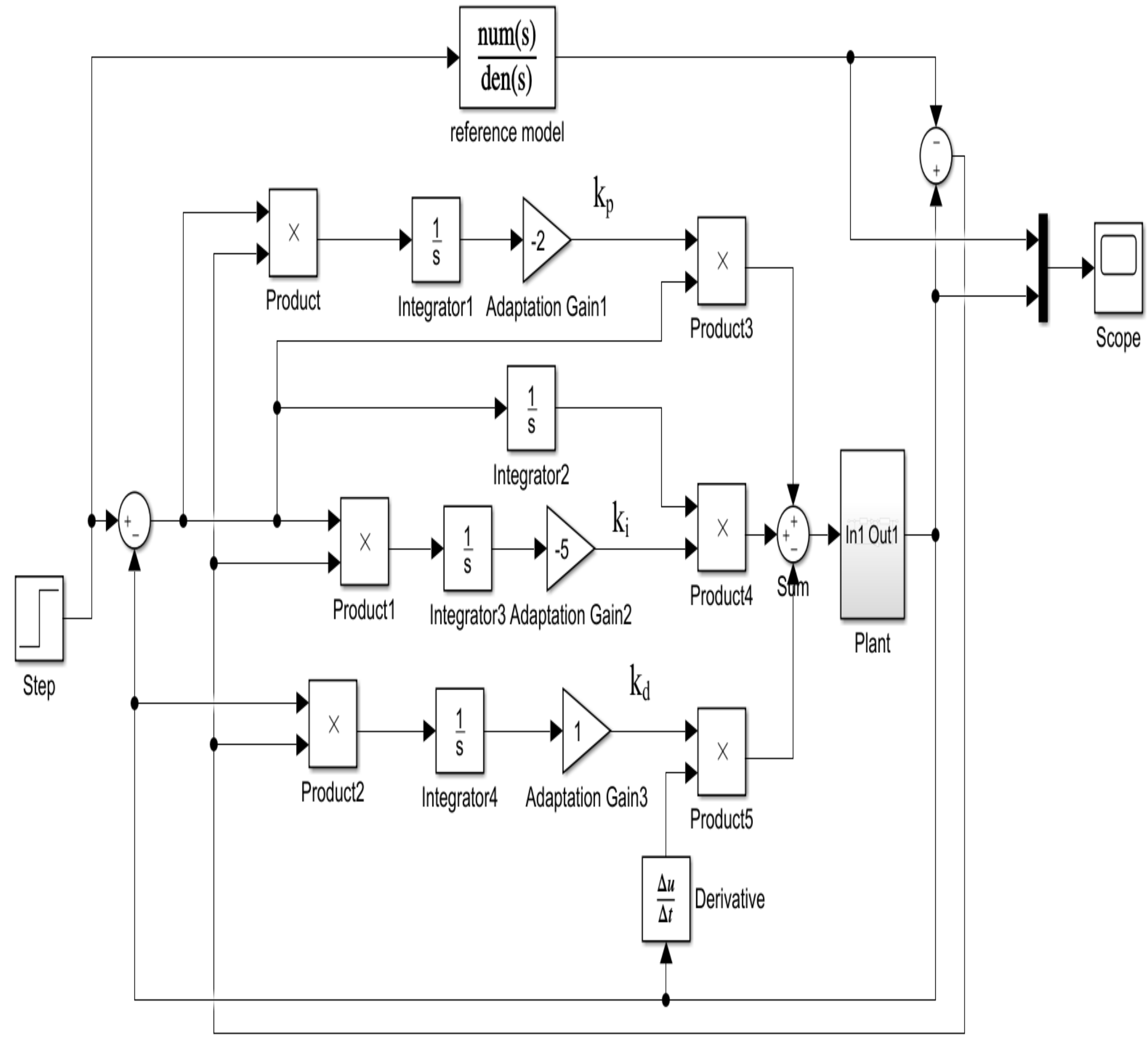

Figure 5.26: Matlab/Simulink model of system with no load under MRAC based PID controller using Lyapunov rule

The response is as shown in Figure 5. 27.

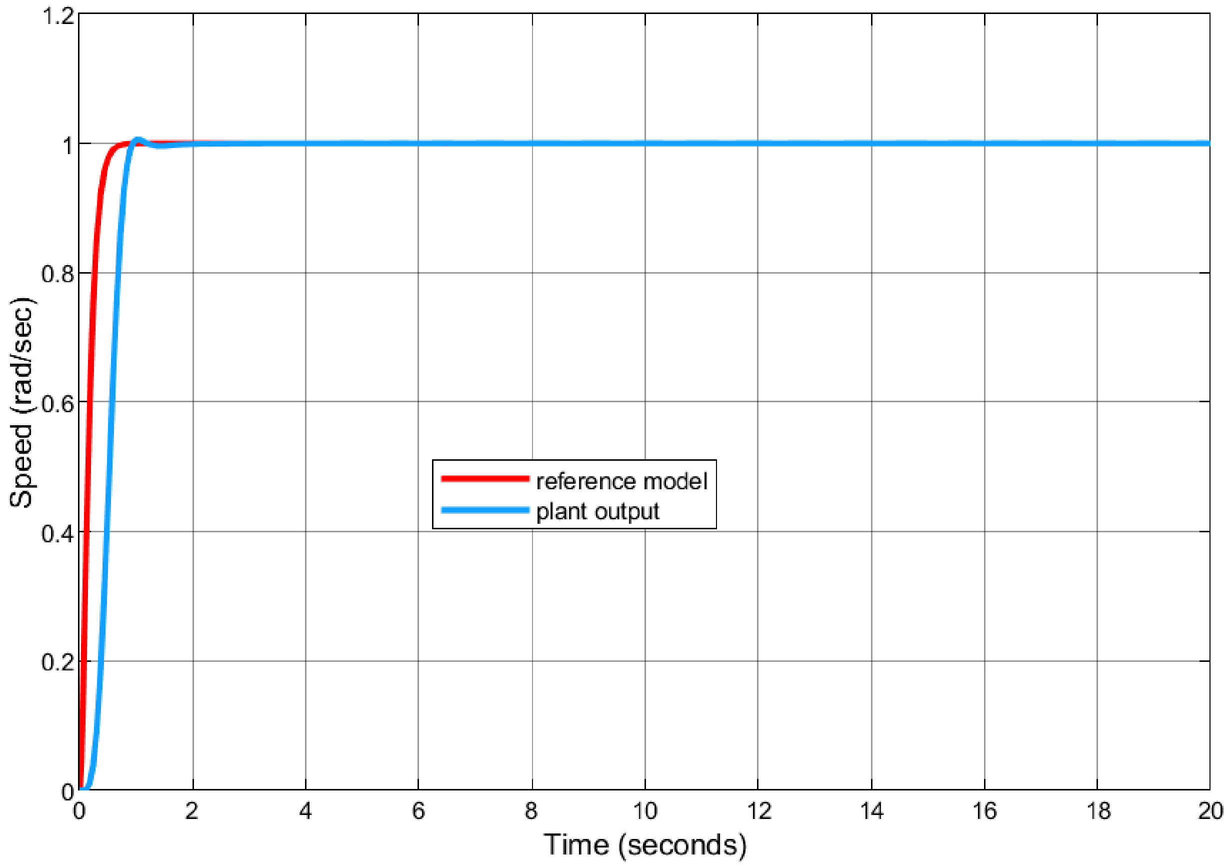

Figure 5.27: Step response of the system with no load under MRAC based PID controller using Lyapunov rule

The system has a rise time of 0.450 second, settling time of 3.600 second, 0.508% maximum overshoot and zero steady-state error.

The response is almost the same with MIT rule but this guarantees stability of the controller parameters.

Chapter Six

Conclusion and Future Works

6.1. Conclusion

This thesis basically explains the advantage of model reference adaptive control based PID controller over conventional control and especially modified MIT rule model reference adaptive control based PID controller. Using the PID controller as speed control of armature controlled separately excited DC motor is not satisfactory to the higher degree of accuracy condition.

Thus a model reference adaptive control especially the modified MIT rule is designed, to reduce the overshoot and decrease the tracking error so that it gives a better response and good result in all conditions than MIT rule.

Generally, the simulation results show that the conventional PID controller for speed control of armature controlled separately excited DC motor in Figure 5.3 has slow rising time, high overshoot and it takes more time to settle when we compare with model reference adaptive control based PID controller of Figure 5.7 and Figure 5.10. This shows that commonly MRAC based PID controller is more preferred than the conventional PID controller for the system output track the reference speed to reduce the error.

As well modified MIT rule of MRAC based PID controller for speed control of armature controlled separately excited DC motor has good performance under no-load condition since from figure 5.10 the system has 0.487% maximum overshoot with settling time of 3.110 second and when reference input increased as shown in figure 5.11 the response of the system has maximum overshot of 0.505% with settling time of 3.100 second which are better than the response shown in figure 5.7 and figure 5.8 of the MIT rule. Although under load torque and uncertain output disturbance signal the system response decreased in less magnitude, has fast settling time and has less overshoot under modified MIT rule. Thus the system response in all conditions: no-load, reference input increased, load torque and uncertain output disturbance signal exists, modified MIT rule of adaptive control shows less sensitive to the parameter variations and works better than MIT rule of adaptive control in order to reduce error to zero.

6.2. Future works

Recommendation for future works are recorded as follows:

- ❖ Considering the dynamic behavior of the system under flexible shaft.
- ❖ Performing experimental setup and analyzing the result on the lab room.

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