TAMPERE UNIVERSITY OF TECHNOLOGY

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IDENTIFICATION AND CONTROL OF ELECTRO SERVO TEST BENCH

Master of Science Thesis

Thesis Topic approved by the Automation, Mechanical and Materials Engineering Faculty Council on 08 June 2011.

ABSTRACT

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When a control system is designed to follow a frequently changing reference signal, it can be referred to as a servo control design problem, in fact, systems designed to track reference signal are called tracking or servo systems. Admittedly, the main object of this thesis is to identify and control the electro servo motor.

The thesis is divided into three parts, firstly, detailed information about the ranges, properties, mechanical and electrical connections of components are provided. Then, in system identification discussion, model of the motor is identified and validate based on experimental tests on the system. Furthermore, proportional and phase lead controller are designed and tuned due to the proposed model and response of real system to meet the desired performance specifications.

Since to run the motor, the matlab simulink model should be developed and it should be compiled on d-Space control board through the control desk as an interface, the needed matlab Simulink model and control desk layout will be created.

The study indicates that, the proper method in our project which contributes to identify model of the motor. Based on numerous experimental tests, the linear model of the motor is proposed and it is validated throughout studying transient response of model and real system to typical identification inputs.

Furthermore, our closed loop analysis will look at some key properties of designed proportional and phase lead controller, they are stability, reference tracking performance and disturbance rejection performance.

The compensated phase provided by phase lead controller is studied, where the phase may be lost in process model or measurement device. Notice that, when the phase lead parameters apply in closed loop system include velocity feedback, it may make face the system to unknown oscillatory behavior, probably because of presence of noise and delay in velocity measurement, or it theoretically can help the system in sudden changing of reference input. Moreover, the response of both controllers in real system will be analyzed to figure out which of them provide the better transient response.

PREFACE

This Master of Science thesis was carried out at Hydraulic and Automation Department, the main goal is to identify and control the electro motor servo system. The material in text is mainly provided based on numerous tests on system and modern control concepts. The thesis has been done from mid of December 2010 to first of August 2011.

I would like to thanks so kindly my supervisor Dr. Reza Ghabchelo who highly contributes me during the project to approach the final results, i have learnt so many key knowledge in mechatronic and control engineering during this project. Also i wish to acknowledge to Professor Kalevi Huhtala who provides me the great opportunity to work on a subject that i like it.

Finally , i would like to thank my world Mona when without all her patience and understanding of my feelings, i never would have finalized my study, also i have special thanks to my father and mother whose love and support have never failed me.

AMIR SOLEIMANI ABYANEH
August 01, 2011

CONTENTS

ABBREVIATIONS

PMSM	Permanent Magnet Synchronize Motor
DS1103	d-Space Controller Board
CLP1103	Control Panel
Inc1	Incremental 1 (Port)
SBR2	Sensor Board Resolver 2 (Port)
DACH1	Digital to Analog Chanel 1
P.O	Percent of Overshoot
RGS	Random Gaussian Signal

Symbols

K_p	Proportional Controller Gain
T^{-1}	Corner Frequency
α	Attenuation Factor
M	Magnitude
φ	Phase
A	Amplitude in Sine
ω	Frequency
ω_n	Natural Frequency
ζ	Damping Factor
T_r	Rise Time
T_p	Peak Time
T_s	Settling Time
y_d	Response to Disturbance change
r	Reference
d	Disturbance

1. INTRODUCTION

The objective of this thesis is to study electro servo test bench. It includes identification of the motor (PMSM) by low order model, and feedback controller design. Both identification and controller design are based on experimental tests and basic control theory. The system consists of s electro servo motor, master derive motion control unit, d-Space control board, matlab simulink, control desk, and the load used in some implementations. The general block diagram of system is shown in figure 1.1.

The thesis is organized in the following manner; in section 2, the system components, electrical connections and mechanical connections are described in more details. In section 3, the principles of dynamic modeling through the frequency response method are studied, and thus the identified model is proposed. Furthermore, to validate the obtained model, transient response of the model is evaluated and compared with the transient response of real system throughout the step input. Moreover, the responses of the model and real system are to typical identification signal such as sum of Sine signal and Random Gaussian signal are studied. In section 4, the proportional controller and phase lead controller are designed, applied in simulation and tuned on the real system to meet the desired specification performance. Finally, the conclusion and overview of the thesis is described in section 5.

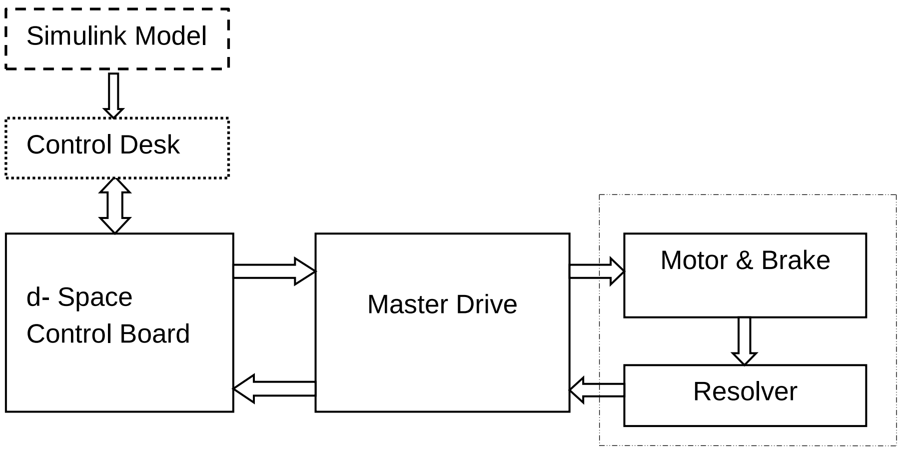

Figure 1.1*: Motor speed or position control system overall block diagram*

2. SYSTEM COMPONENTS

The purpose of this section is to provide some detailed information about the system components. Thus, their ranges, properties, mechanical connections and electrical connections are going to be discussed while, the principle of their operation is neglected to be presented.

2.1. Motor

Permanent magnet synchronies AC motor called PMSM is a three-phase servomotor with the technical data illustrated in table 2.1.

Table 2.1: *Technical data of Permanent magnet synchronies AC motor*

	Rated Speed(rpm)	Rated Power(Kw)	Rated Torque(Nm)	Moment of Inertia(kgm^2)	Rated Current(A)	Holding Torque(N)	U in (V)
PMSM	3000	0.85	2.6	0.35e-4	2.4	3.9	228

In our project, when we call the motor is connected to the “load”, it means according to the figure 2.1, the motor is mechanically connected to the gear box (pulley & belt) as well as the load, where the load is moved on the slop.

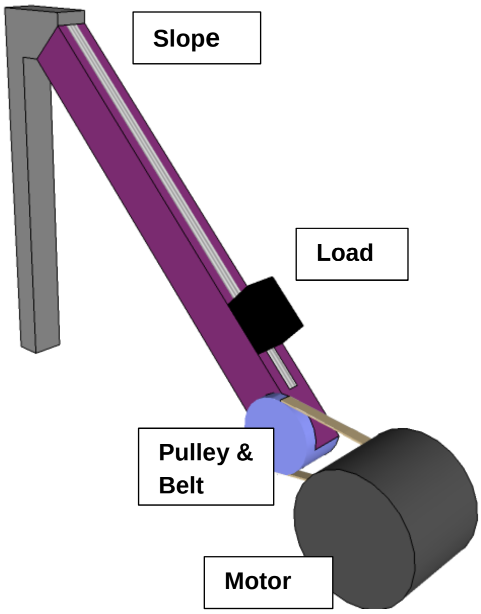

Figure 2.1: *Mechanical connections of the system connected to the load*

Also, when we call the motor with “no load”, it means the load is removed from the slop and the motor is working without any load and just connected to the gearbox. These terms “load” and “no load” later on are going to be used during the control design procedure at section 4. Notice that, during the identification procedure at section 3, the motor is entirely disconnected from the system shown in figure 2.1 and it is moved with no mechanical connections to any extra devices.

The motor is fed by Master drive which is discussed later on section 2.2. Moreover, the motor includes resolver, to create the position feedback, and send it to the master drive. Also, it consists the brake to stop the motor in emergency situation where the needed release signal is coming from the master derive. All electrical connections are shown in figure 2.2.

2.1.1. Resolver

Resolver provides position feedback for the controller, the method of functioning is inductive sampling sine/cosine evaluation for rotor, and it provides 1024 pulses for every 2π rad rotor revolution, the operating voltages is 5 V and can provide information up to 15000 rpm.

2.1.2. Brake

The function of brake is to stop the motor in emergency situation due to safety , the permanent magnet single-face brake works according to the closed-circuit current principle, the magnetic fields of the permanent magnet exerts the pulling force on the brakes armature plate, it means, in a zero current condition , the brake is closed thus preventing the motor shaft from turning, It is actuated by 24V DC, the current carrying coil generates an opposing field which cancel out the force exerted by the permanent magnet and releases the brake, for emergency stops or power failure , approximately 2000 braking operation can be carried out without casing excess wear on .

2.2. Master Drive and d-Space Board

The Master drive which is included the frequency convertor is the power electronic components for feeding highly dynamic three phase drives in output ranges illustrated in table 2.2:

Table 2.2*: Technical data of master drive*

	Voltage Input (V)	Frequency Input (Hz)	Frequency Convertor Output(kW)	Inverter Variable Output Frequency(Hz)
Master Drive Data	380-480	50-60	0.55-15	0-400

Furthermore, DS1103 (d-Space) as controller board is highly suitable for positioning systems and servo motors, also the CLP1103 connector panel serves as an interface between the DS1103 and all external hardware. Considering our project, all links which are connected to CLP1103 connector panel and DS1103 board are discussed below, and they are shown in figure 2.2.

Notice that, signals' voltage level conversion box is added to the system, because DS1103 control board's digital I/O signal level is 5V and motion control unit needs 24V at the terminal strip.

The following links are connected directly between master drive and motor:

- The master drive feeds the Permanent magnet synchronies AC motor according to the control signals coming from d-Space control board.
- Master drive receives the angular position provided by resolver from the motor.
- Master derive provide the needed voltage to release the brake located on the rotor.

Moreover, following links are connected between master drive and DS1103 board through CLP1103 connector panel as well as conversion box:

- The digital position output (SBR2-port) from Master Drive is taken and it is connected into the CLP1103 connector panel's Inc1-port.
- The digital I/O port is connected between the CLP1103 Connector Panel and conversion box.
- Slot 1 of master drive terminal strip gives 24V when the motion control unit is on, and it is used as a power supply for signal conversion box.
- Slot 2 is a ground.
- Slot 6 is inverter releaser and slot 8 is on/off, while both Slot 6 and 8 needs 24V signals which are coming from the conversion box.
- Slot 9 and 10 are for analog speed input command which are coming from the DS1103 board (DACH1-port on CLP 1103) while this analog command signal needs to be between +-10 V.

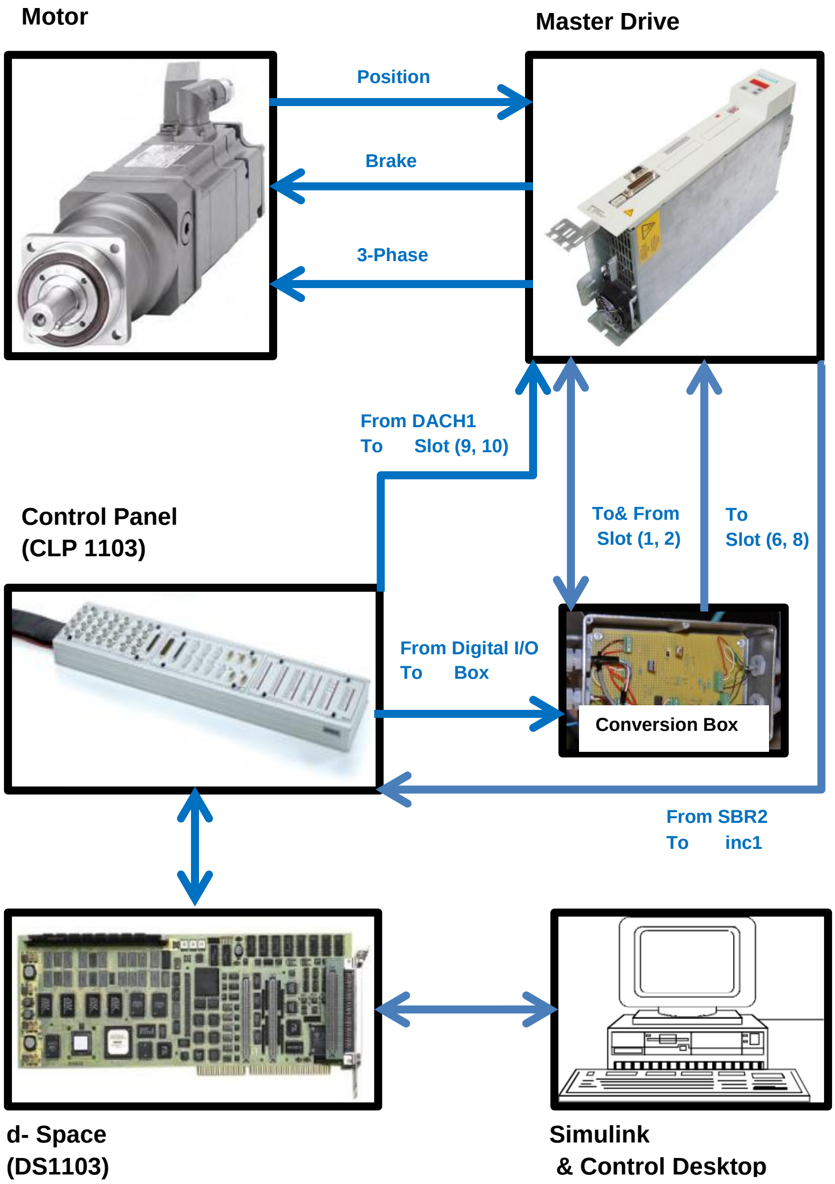

Figure 2.2*: Motor electrical connections block diagram in more details*

2.3. Matlab Simulink

Since to run the motor, the matlab simulink model should be developed and it should be compiled on d-Space control board through the control desk as an interface, the needed matlab Simulink model is created in this section.
Figure 2.3 shows the open loop model of the system described in section 2.1, where speed in rad/s is to input and speed measurement in rad/s is the output.

Moreover, the speed input into the d-Space control board should be normalized between [-1 to 1] that is corresponded to –314 rad/s to +314 rad/s. To do so, the input in rad/s should be multiplied by (1/314) to meet the needed unit.

Also the speed measurement of the motor is provided by the delta encoder block in specific unit and it should be converted to meet the rad/s. To do so, considering 1024 pulses which are corresponded to 2π rad revolution of the rotor with the sampling time equal to 0.001s, therefore, the output of delta encoder should be multiplied by $(\frac{2\pi}{1024\times 0.001})$ to meet the speed measurement of the system in rad/s.

Additionally to these sets of blocks, the required bit commends should be considered in matlab simulink model. Therefore , bit 0 to release the inverter, bit 2 to release the brake, and bit 4 to make on/off the driver of system are considered.

Notice that, the open loop model is only used for identification procedure at section 3, and the implementation that is used for controller design at section 4 will be throughout the closed loop model.

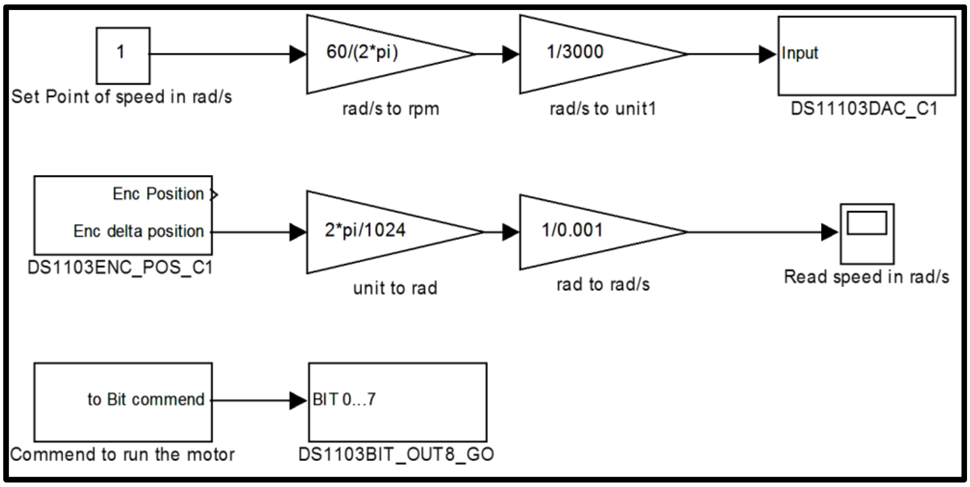

Figure 2.3*: Open loop control model at Simulink Matlab to run the motor in real time*

2.4. Control Desk

In our real time workshop, monitoring, controlling and recording values are performed throughout the control desk. For the first time of the system implementation, the DS1103 board should be detected by control desk software, from the menu bar in control desk software "Register" is selected in following the way, Platform ► Initialization ► Register. Then, from the "Register Board" dialog box that appears, "DS1103 PPC Controller Board" from the "Type" dropdown menu is selected and the "Port address" is set to 300. Then the "Register" button is selected. The "Register Board" dialog box is shown in figure 2.4.

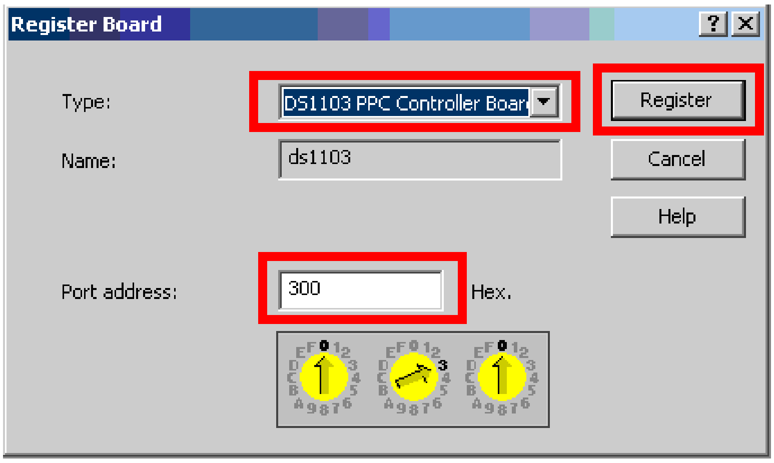

Figure 2.4: *Correct selections in register board dialog box.*

The DS1103 Board and DSP should now be reported as detected in the Control Desk "Platform" tab which is shown in figure 2.5.

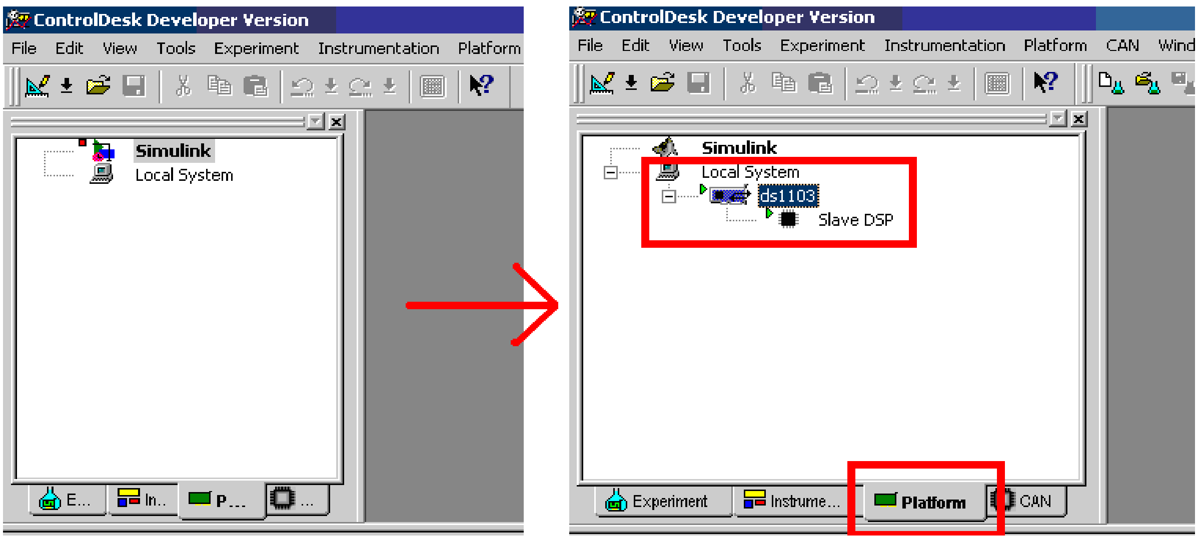

Figure 2.5: *DS1103 and slave DSP shown in the platform tab*

Later on, for any change in model no change is needed through the register and detecting of DS1103.

Furthermore, the model which was shown in figure 2.3 is built in following the way, Tools ▶ Real-Time Workshop ▶ Build Model. Then, "NameProject.sdf" file is uploaded into control desk software by selecting the file from upload tab in control desk menu.

Then, the layout should be created at control desk. The layout allows the values of constants, input blocks, and gain blocks from the Simulink model to be changed while the system is running, also we can record and visualize the input-output of the system in real time workshop. To create the layout, the "Edit Mode" is selected from the Instrumentation tab in menu bar. Some needed features are added by click, drag and release from the “Virtual Instrument” library into the control desk, figure 2.6 shows the required steps.

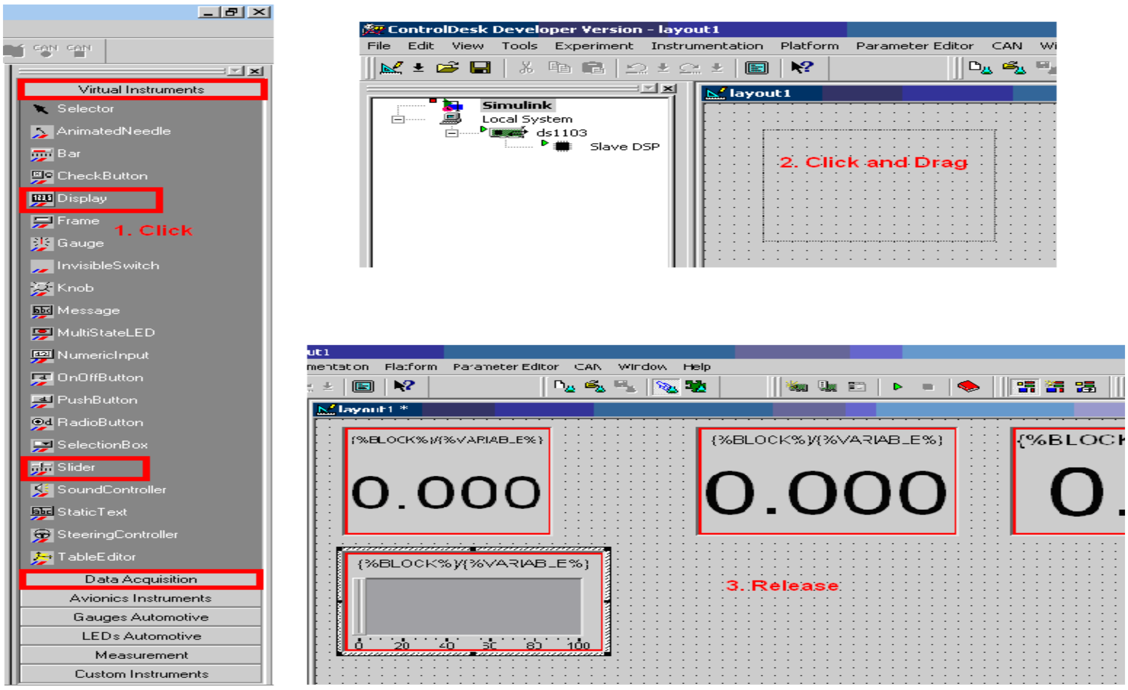

Figure 2.6: *Adding a virtual instrument*

To connect/assign variables from the Simulink model downloaded to the DS1103, the variable in the “.sdf” menu(s) are located and then they should be clicked as well as dragged into the corresponded virtual instrument tab at the layout. Generally, the values of constants and gain blocks in the model can be monitored and modified/controlled in real-time. The steps are shown in figure 2.7.

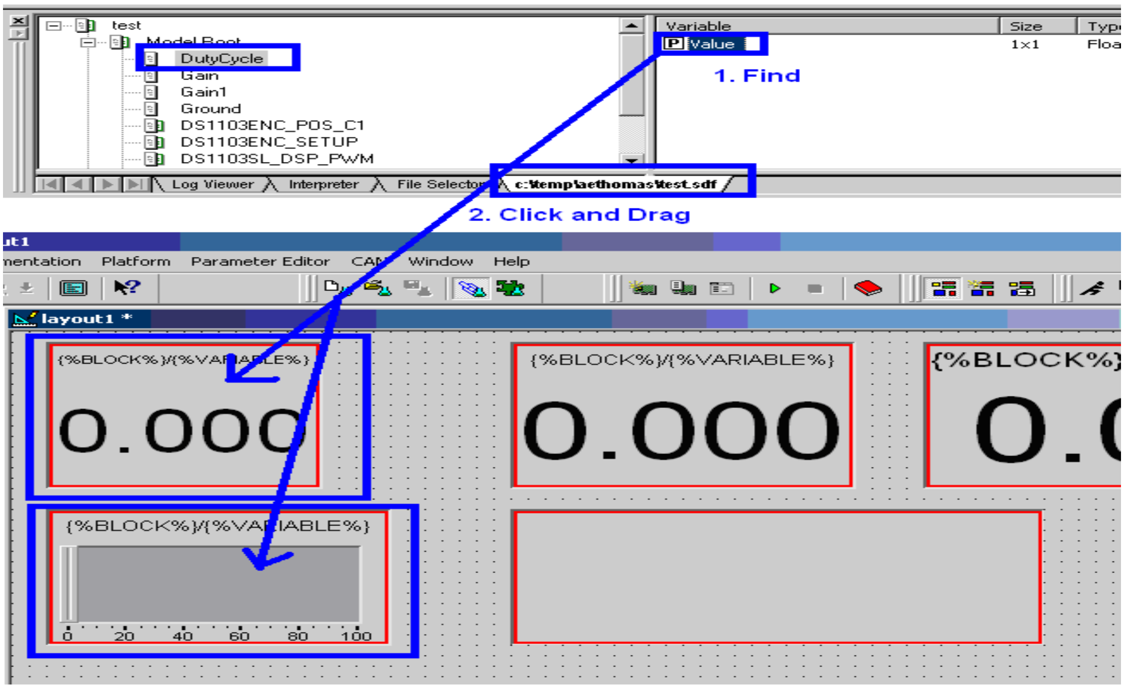

Figure 2.7: *Connecting variables to virtual instruments.*

Hence, The layout is created in control desk where it includes all the required parameters needed to be changed, visualizing interface and recoding tabs. Considering figure 2.8, the needed virtual instruments and accusation tabs are added:

- A: The tab to record the measured input or output of the system during real time workshop, assign the required time to save and locate the data which should be saved. Also, additional settings are allowed in properties of block.
- B: The blocks which its quantity can be changed during the real time workshop. any needed limitation can be assigned in properties.
- C: The required commends to run the motor. They include Bit commends, additionally, the required commend to reset the incremental encoder and commend to change between velocity and position control can be added.
- D: More visual instruments can be added into the layout.
- F: Record and visualize the needed measured input of output of the system.

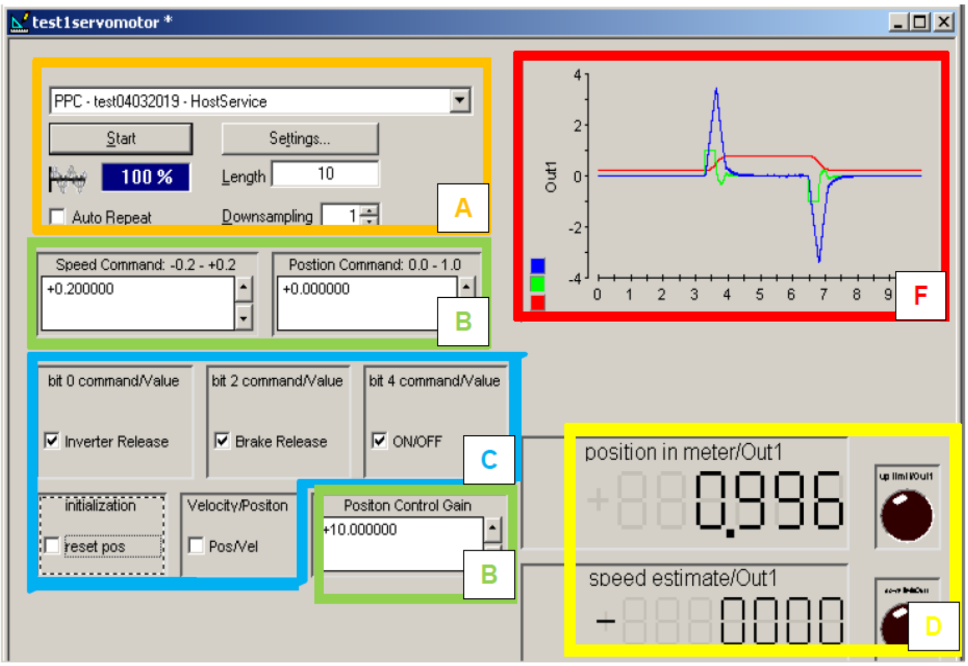

Figure 2.8*: The created control desktop layout to monitor, control and record the data of motor during real time operation*

Finally, to start the real-time workshop, the “Animation Mode” is selected from the same menu of “Edit Mode”.

Hence, the system which includes servo electro motor, master derive motion control unit, d-Space control board, matlab simulink, control desk, and possible connections of the load, can run in real time workshop. Notice that, by using the control desk layout interface, different elements can be monitored, recorded and controlled.

3. LINEAR MODEL IDENTIFICATION AND MODEL VALIDATION

“The main goal of feedback control is to make the output variable of dynamic process to follow a desired reference variable precisely. The first step to provide the controller is to develop a mathematical description called dynamic models of the process to be controlled, the term model means a set of differential equations that describe the dynamic behavior of the process, and a model can be derived using principles of physics or can be derived by testing of device “[1]

Following this section, to identify the model, frequency response method is chosen to be the main approach to the modeling. Then step response analysis is studied to validate the model by comparison of the transient specifications from model and step response of system. Finally the model is also validated by using random Gaussian noise signal and the sum of sinusoid signal as inputs.

3.1. Linear Model Identification

In our project, the identification is based on testing the device. After numerous experimental tests on the system, and using the variety of signal types, frequency response method is chosen to be the main approach to the modeling. Acquisition of the controller parameters during motor operations is a challenging task due to the inherent nonlinearity of motor dynamics [6]. Figure 3.1 shows the open loop of the system , notice that , (u0,y0) defines operating point of the system, where a linear model is fit to the system.

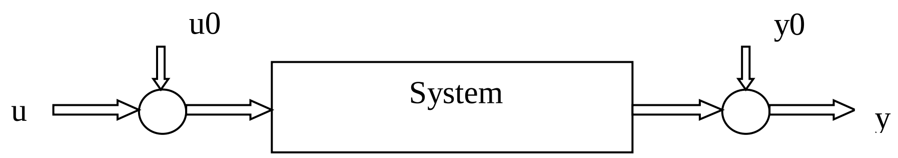

Figure 3.1*: General model of the system in which should be modeled*

3.1.1. Frequency Response Principle

In the cases for which a good model of the system does not exist and wish to determine the frequency-response magnitude and phase experimentally, the system can be excited with sinusoid varying in frequency. For a linear system, a sinusoidal input of a specific frequency, amplitude and phase results in an output that is also a sinusoid with the same frequency, but with different amplitude and phase.

To prove the main concept of frequency response by exciting the system to sinusoid input, we can approach the idea in the following way, the transfer function in Laplace domain between outputs in respect to an input is:

$$\frac{Y(s)}{U(s)} = G(s) \tag{1}$$

Considering the input $u(t)$ is a sine wave with amplitude A :

$$u(t) = A\sin(\omega_0 t)1(t) \tag{2}$$

Therefore, the Laplace transform of the output is obtained as:

$$Y(s) = G(s)\frac{A\omega_0}{s^2+{\omega_0}^2} \tag{3}$$

Also, a partial fraction expansion of the equation (3) is:

$$Y(s) = \frac{\alpha_1}{s - p_1} + \frac{\alpha_2}{s - p_2} + \ldots + \frac{\alpha_n}{s - p_n} + \frac{\alpha_o}{s + j\omega_o} + \frac{{\alpha^*}_o}{s - j\omega_o} \tag{4}$$

Where $p_1, p_2, \ldots, p_n$the poles of are$G(s)$,α_o is found by partial fraction expansion, and ${\alpha^*}_o$ is the complex conjugate of α_o . Then the time response corresponds to $Y(s)$ is presented as:

$$y(t) = \alpha_1 e^{p_1 t} + \alpha_2 e^{p_2 t} \ldots + \alpha_n e^{p_n t} + 2|\alpha_o|\sin(\omega_o t + \varphi), \quad t \geq 0 \tag{5}$$

Where:

$$\varphi = tan^{-1}\left[\frac{Im(\alpha_0)}{Re(\alpha_0)}\right] \tag{6}$$

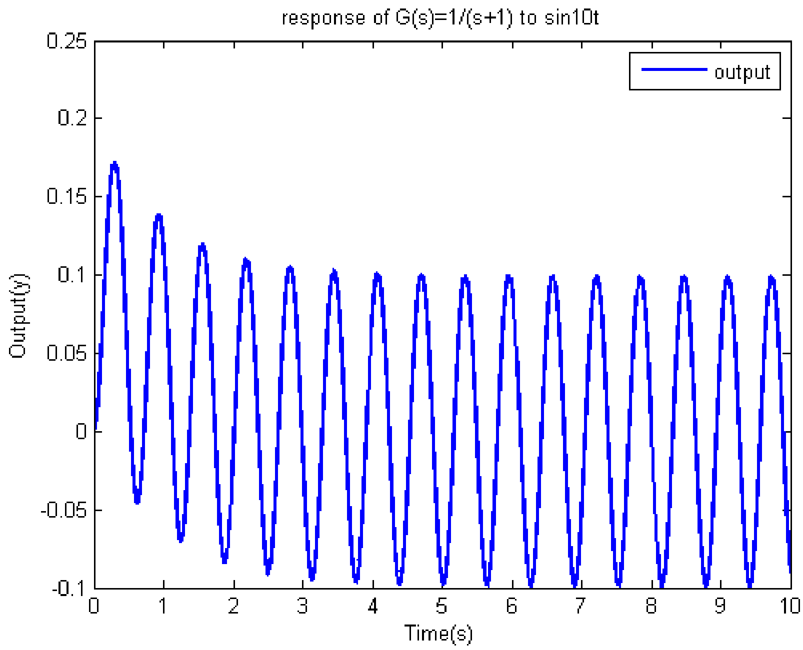

***Figure 3.2**: Response of the G(s) =1/(s+1) to sin10t*

If all the poles of the system represent stable behavior (the real parts $p_1, p_2, \dots, p_n < 0$,) the natural unforced response will die out eventually, this behavior is shown in figure 3.2, and therefore the steady state response of the system will be due to solely the sinusoid term in equation (5), which is caused by the sinusoid excitation. Thus, the remaining sinusoidal term in equation (5) can be expressed as:

$$y(t) = AM\sin(\omega_0 t + \varphi) \quad (7)$$

Where,

$$M = |G(j\omega_o)| = \sqrt{\{Re[G(j\omega_o)]\}^2 + \{Im[G(j\omega_o)]\}^2} \quad (8)$$

$$\varphi = tan^{-1}\frac{Im(G(j\omega_o))}{Re(G(j\omega_o))} \quad (9)$$

So, equations (8) and (9) present that a stable system with transfer function $G(s)$ excited by a sinusoid signal with unit amplitude and frequency ω_o will, after the response has reached steady-state, exhibit a sinusoidal output with a magnitude $M(\omega_o)$ and a phase $\varphi(\omega_o)$ at frequency ω_o . Note that, for linear systems, the output y is a sinusoid with same frequency as the input u and the magnitude ratio M and phase φ of the output are independent of amplitude A of the input. If the system being excited were a nonlinear or time-varying system, the output might contain frequencies other than the input frequency, and the output input ratio might be dependent on the input magnitude [1], hence, in our project, signals with small amplitude are going to be chosen to excite the system, therefore they will not excite nonlinear behaviors. Figure 3.3 shows the differences both at magnitude and phase among input and output of system with the same frequency.

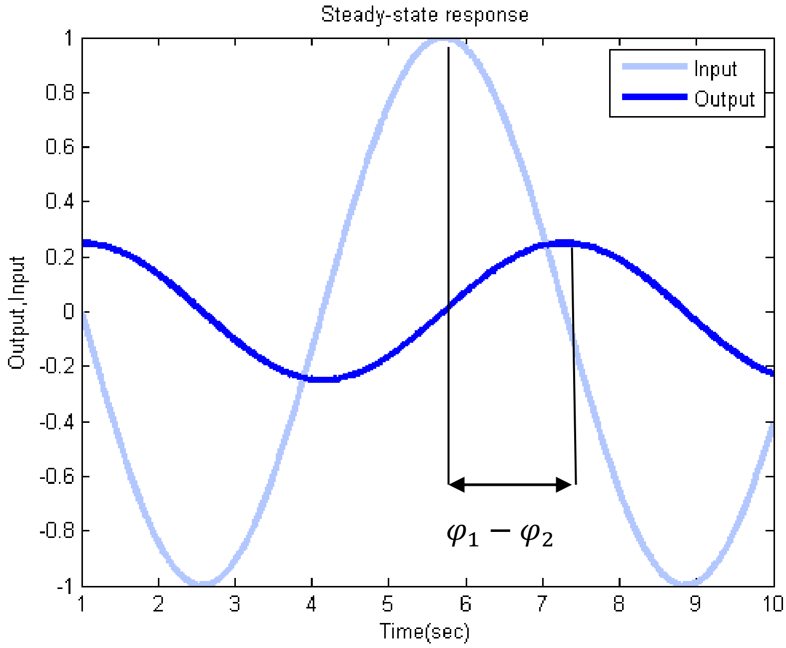

Figure 3.3*: Compare the two sinusoid signal with same frequency but different amplitude and phase*

3.1.2. Magnitude and Phase Shift

Since the principle of frequency response method to identify the linear and time-invariant system was discussed in 3.1.1, the following steps are applied to obtain the model of motor base on that:

Step 1: The motor is decoupled from all external loads and mechanical connections; In fact, it is identified when the motor s shaft is not connected to pulley and belt shown in figure 2.1.

The motor is excited by sinusoid input signals in variety of frequency ranges {10,20,30,40,50,60,90,100,150,200,230 rad/s}. Notice that output is not significant in the frequencies above 230 rad/s, and differentiating signal out of noise is not possible, so we limited the experiment to these frequencies. Admittedly, the input magnitude of the sinusoid input signals are maintained constant and equal to 15.7 rad/s. It means that, we use small signal perturbation to avoid exciting nonlinearities. Also we consider the phase of the sinusoid input signals equal to zero.

All links between different components of the system are connected according to the figure 2.2. Then, the open loop model with speed in rad/s as an input is created and built in matlab Simulink according to the figure 2.3. Also, the corresponded layout is created in control desk, while the real-time workshop codes should be uploaded from matlab Simulink model. Consequently by substituting the different speed signals at reference point and measure the speed response of the motor, the model can be identified. Furthermore, we study the system in two operating points: at 0rad/s and 157rad/s. Thus, the motor is excited in both operating point through the variety of frequency ranges, mostly because the possible different frequency response of the system should be studied in different operating point.

Step 2: The measured output has to be filtered to minimize the noise with high frequencies, it lets focus on data in required frequency band. Filtering the input and output data through the same filter does not change the input-output relationships for a linear system. Therefore, here, the input-output sinusoid data is filtered by fifth order pass band Butterworth filter in matlab.

To do so, the iddata object is created by input (u), output (y) data where they should be in same size, and Ts here is the sampling time which has the value equal to 0.001s:

```
data = iddata(y, u,Ts)
```

Then, the created iddata is filtered by:

```
fdata = idfilt (data,[wl wh])
```

Where the variable wl and wh is the lower and upper limit of the pass band, for instance considering the sinusoid input signal with w=10 rad/s, the wl and wh is substituted by 9 and 11 rad/s. [2]

Step 3: To obtain the magnitude M stated in equation (7), consider $u(t)$ and $y(t)$ which are the input and measured output sinusoid signal in rad/s:

$$u(t) = A_1 \sin(\omega_0 t + \varphi_1) \tag{10}$$

$$y(t) = A_2 \sin(\omega_0 t + \varphi_2) \tag{11}$$

When the system meets the steady state mode, the A_2 can be evaluated as maximum amplitude of measured sinusoid output signal in rad/s. Thus, considering the A_1=15.7 rad/s, therefore, the ratio of the magnitude change is obtained as:

$$ratio = \frac{A_2}{A_1} \tag{12}$$

Notice that, desired magnitude should be in dB, hence the magnitude is:

$$M = 20\log_{10}(ratio) \tag{13}$$

Step 4: To obtain the phase shift stated in equation (9), consider equation (14):

$$u(t)\, y(t) = A_1\, A_2\, \sin(\omega_0 t + \varphi_1) \sin(\omega_0 t + \varphi_2) \tag{14}$$

Using the rule of sum of sine and applying some modifications, equation (15) and (16) are obtained:

$$u(t)y(t) = -\frac{A_1\, A_2}{2} (\cos(2\omega_0 t{+}\varphi_1 + \varphi_2) - \cos(\varphi_1 - \varphi_2)) \tag{15}$$

$$\cos(2\omega_0 t{+}\varphi_1 + \varphi_2) + \cos(\varphi_1 - \varphi_2) = 2\frac{u(t)y(t)}{A_1 A_2} \tag{16}$$

Moreover, by taking the average from both side of the equation (16) the term $\cos(2\omega_0 t{+}\varphi_1 + \varphi_2)$ is equal to zero in one period and phase shift can be obtained by equation (17):

$$|\varphi_1 - \varphi_2| = \arccos(\, average\left(2\frac{u(t)y(t)}{A_1 A_2}\right)) \tag{17}$$

Where, A_1 is equal to 15.7 rad/s as the amplitude of input sinusoid signal, and A_2 is the amplitude of the measured output sinusoid signal through the same frequency as the input . Also, $u(t)$ and $y(t)$ in equation (17) are only one period of data when the system reach the steady state mode.

Hence the table 3.1 is obtained by implementing the step1 to 4, and figure 3.4 shows bode diagram which is plotted based on these derived data.

Table 3.1: *The corresponded data to bode diagram*

Input Frequency(rad/s)	Motor Rotational Speed operating point at 0 (rad/s)		Motor Rotational Speed operating point at 157 (rad/s)	
	Mag(dB)	Phase(Degree)	Mag(dB)	Phase(Degree)
10	0,05	-3,90	0,13	-2,37
20	0,09	-5,01	0,39	-2,60
30	0,14	-10,43	0,59	-8,00
40	0,12	-13,28	0,66	-13,17
50	-0,32	-23,28	0,22	-23,17
60	-1,38	-36,84	-0,84	-37,34
90	-4,95	-74,94	-4,27	-75,30
100	-5,83	-82,43	-5,21	-83,47
120	-7,50	-94,33	-6,91	-99,15
150	-9,59	-111,00	-9,02	-114,25
180	-11,27	-126,38	-10,86	-128,28
200	-12,46	-137,91	-11,92	-139,10
230	-14,05	-150,34	-13,52	-152,48

3.1.3. Derive Model

The bode is shown in figure 3.4 based on the data illustrated in table 3.1.

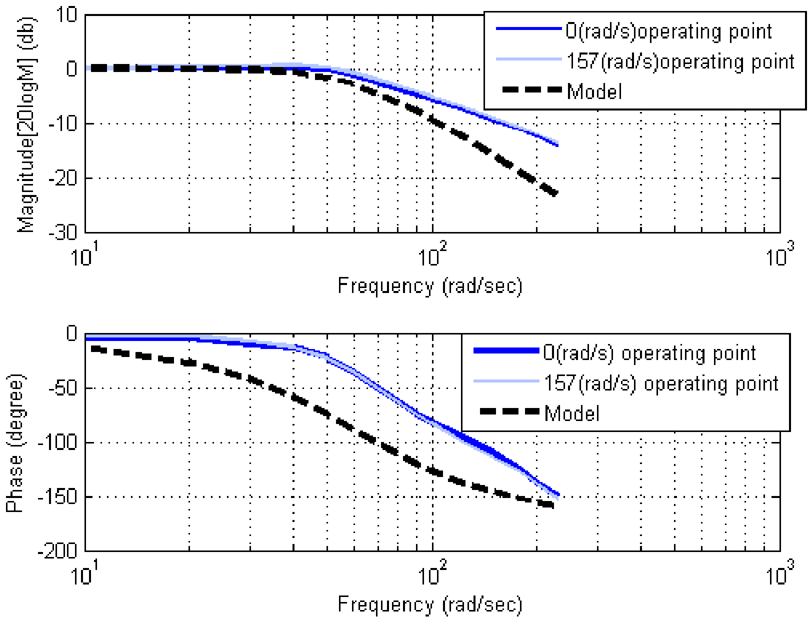

***Figure 3.4**: Compare the obtained bode diagram corresponded to table 3.1*

All the transfer functions for the kinds of not complicated systems are composed from these classes:

$$K_o(j\omega)^n \tag{18}$$

$$(j\omega + 1)^{\pm 1} \tag{19}$$

$$[(\frac{j\omega}{\omega_n})^2 + 2\zeta\left(\frac{j\omega}{\omega_n}\right) + 1]^{\pm 1} \tag{20}$$

Here in this approach, the equations (18), (19) and (20) are examined to fit bode diagram shown in figure 3.4, while individually or combination of them may be considered. Meanwhile, we select the pure second order model equation (20) which can present the bode plot in specific frequency band, shown in figure 3.4. The break point is at $\omega = \omega_n$, the magnitude should change 40 dB/decade and the phase changes should be $\pm$180 degree, where the transition through the break point region is changed by different damping ratio ζ . As figure 3.4 shows system behavior experimentally, the break point at ω= 57 to 61 rad/s in magnitude curve can be observed. Also the magnitude changes the slope 20dB in less than a decade. [1]

Moreover, the ζ can be approximated by considering a rough sketch of the transition at break point:

$$|G(j\omega)| = \frac{1}{2\zeta} \quad at \quad \omega = \omega_n \tag{21}$$

So the ζ is approximated 1.0 to 0.5. Consequently, according to the analysis of experimental bode plot, the system can be estimated as pure second order model:

$$\frac{1}{(\frac{s}{\omega_n})^2 + 2\zeta\left(\frac{s}{\omega_n}\right) + 1} \tag{22}$$

Where, $\omega_n = 60$ rad/s , $\zeta = 0.7$.
Figure 3.4 shows the comparison between the bode diagram of a model and the bode diagram from experimental testing. As it is clear, the model identifies the frequency below 60 rad/s quite well in amplitude and roughly in phase.

3.2. Model Validation

The purpose of this section is to validate the proposed model which was obtained by the frequency response method throughout the transient response method. In fact, after estimating the model, we can validate whether the model reproduces system behavior within acceptable bounds. We iterate between estimation and validation until we find the simplest model that best captures the system dynamics. [2]

To do so, step response of the real system and the response of the real system to specific inputs such as sum of sinusoid signal and Random Gaussian signals should be evaluated. Thus, the model is validated in time domain.

Notice that, considering the proposed model, based on iteration between estimation from frequency response method and validation from transient response method, the optimized coefficients of the model were presented in equation (22). And here the transient responses of the real system are going to be presented and show the validation of the proposed model.

3.2.1. Step Response

As a remarkable property which can be achieved by step response method to validate the model, it can be mentioned that, by applying the step change in our system as speed input in rad/s, the transient response specification of real system can be derived and compared by the transient response specification from the proposed model.

Hence, the step input through the bump test is applied into the system. Notice that, the amplitude of the step change in bump test should be large enough to provide the good signal to noise ratio and should not be so large to maintain the system in linear mode, so the output would not be saturated. Therefore the bump test is created with amplitude equal to 15.7 rad/s and 31.4 rad/s. Also, the enough time should be considered between each two step change to assure that the system is reach the steady state mode after each step change. Moreover, the bump test is applied into the system at two operating point equal to 0 rad/s and 157 rad/s. The bump test which is applied as speed input in rad/s is shown in figure 3.5.

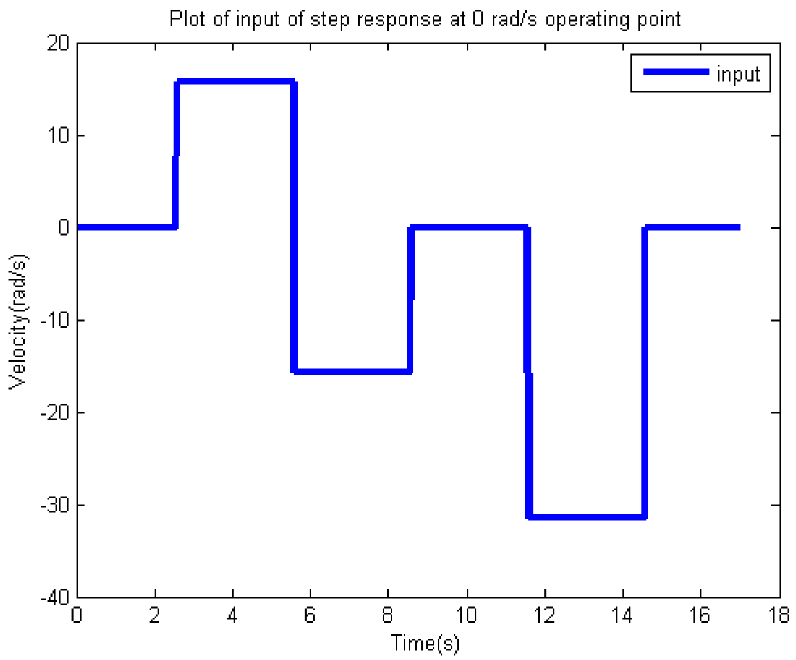

Figure 3.5*: Step change through the bump test as speed input in rad/s*

The bump test shown in figure 3.5 is applied as speed input in rad/s into the open loop implementation which was presented in figure 2.3, and the results for both operating point , at 0 rad/s are shown in figure 3.6, 3.7, 3.8, and for 157 rad/s are shown in figure 3.9, 3.10, 3.11.

Notice that, because the output of the system includes high frequency noise, to analyze the step response output more accurately, the average of the output is considered.

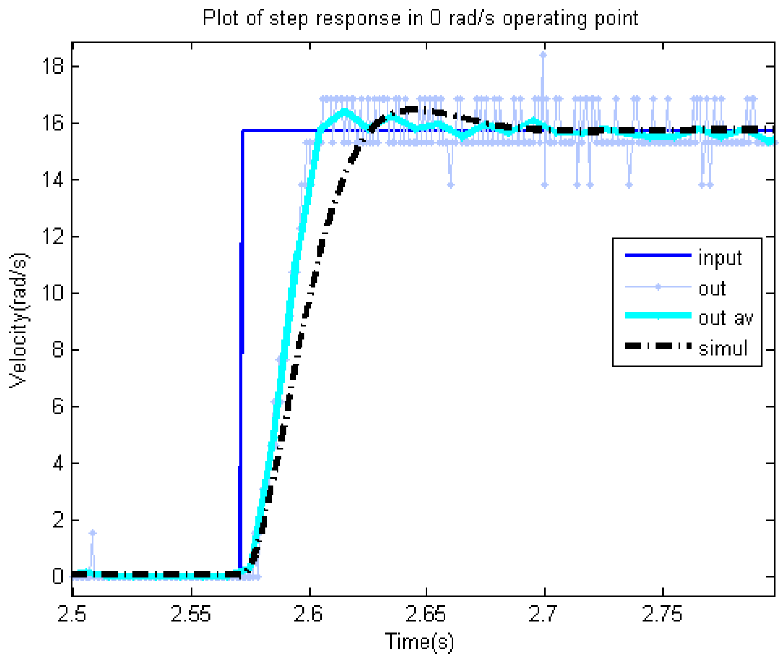

Figure 3.6*: Comparison of step response of the system in rad/s through the bump test at 0 rad/s operation point. For step amplitude equal to 15.7 rad/s in upward direction.*

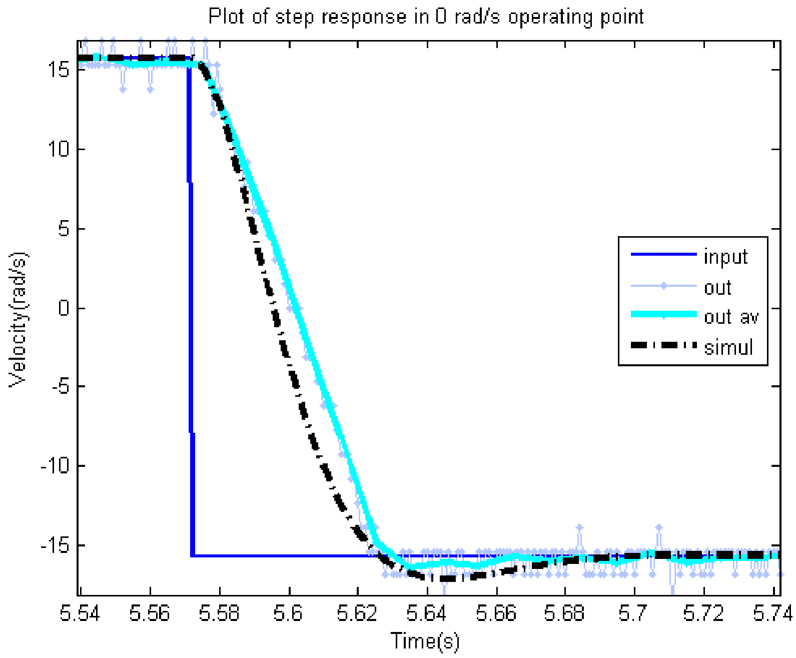

Figure 3.7*: Comparison of step response of the system in rad/s through the bump test at 0 rad/s operation point. For step amplitude equal to 31.4 rad/s in downward direction.*

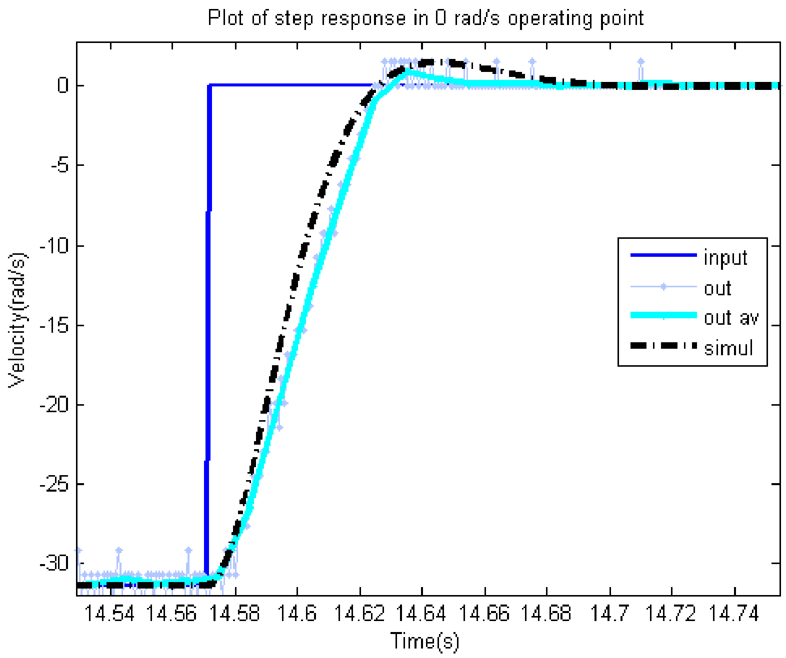

***Figure 3.8**: Comparison of step response of the system in rad/s through the bump test at 0 rad/s operation point for step amplitude equal to 31.4 rad/s in upward direction*

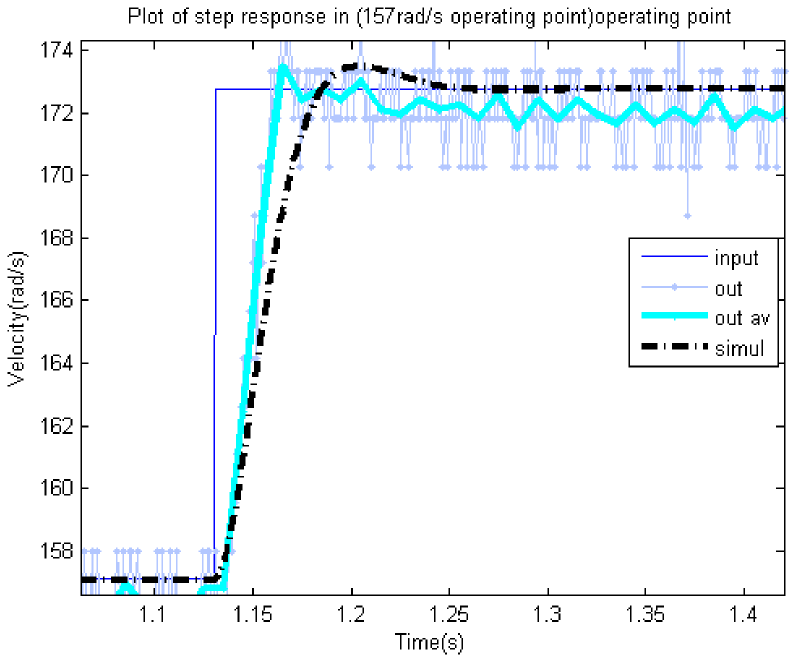

***Figure 3.9**: Comparison of step response of the system in rad/s through the bump test at 157 rad/s operation point. For step amplitude equal to 15.7rad/s in upward direction.*

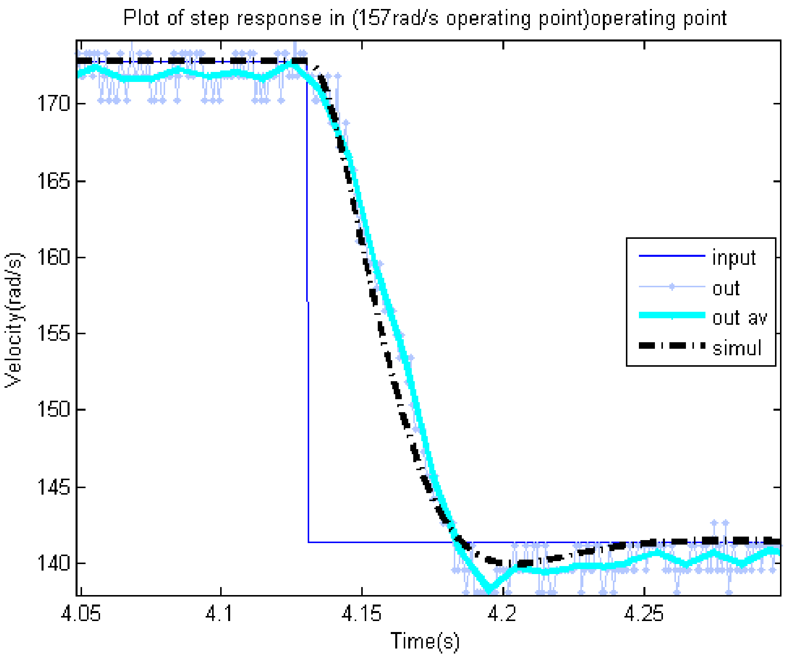

***Figure 3.10**: Comparison of step response of the system in rad/s through the bump test at 157 rad/s operation point. For step amplitude equal to 31.4rad/s in downward .*

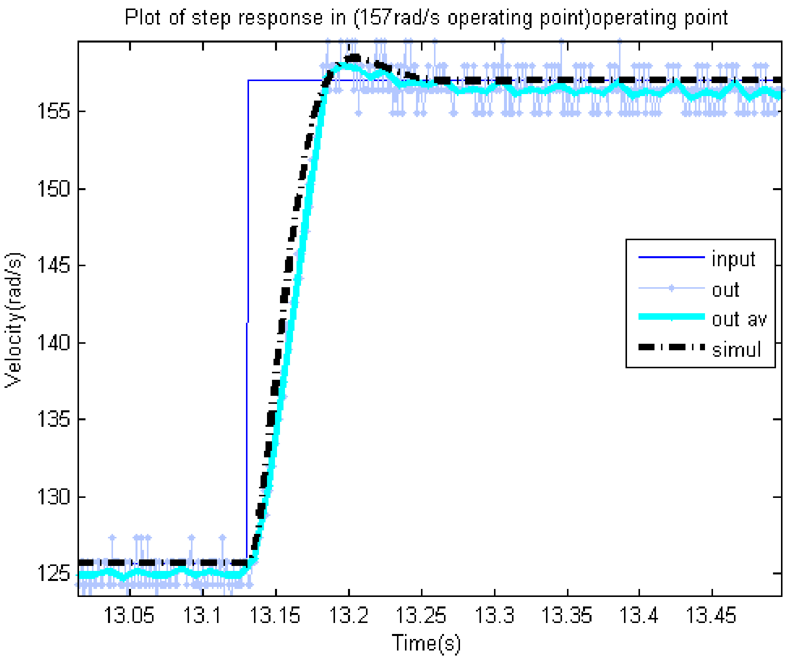

***Figure 3.11**: Comparison of step response of the system in rad/s through the bump test at 157 rad/s operation point for step amplitude equal to 31.4rad/s in upward .*

Therefore, to validate the proposed model from frequency response method, the transient specifications which were derived should be compared with those derived by the model. To do so, the following parameters are derived by pure second order lows:

Rise time; T_r is the time that takes the system to reach the vicinity of its new set point. In second order system, the rise time from $y = 0.1$ to $y = 0.9$ of output is approximately $\omega_n T_r = 1.8$. So an accurate approximation for pure second order system with no zeros can be considered as:

$$T_r \simeq \frac{1.8}{\omega_n} \tag{23}$$

Substituting the$\omega_n = 60\ rad/s$, the value of T_r is obtained equal to 0.03s.
Although, this equation is accurate for pure second order system, it can be considered for other system as rough approximation of relation among rise time and natural frequency.

***Over shoot* and *peak time*;** peak time T_p is the time required to reach the maximum overshoot point in the output, and overshoot M_p is the maximum amount of output of the system.
Considering , $\omega_d = \omega_n\sqrt{1-\zeta^2}$, it leads analytically to:

$$T_p = \frac{\pi}{\omega_d} \tag{24}$$

Substituting $\omega_n = 60\ rad/s$ and $\zeta = 0.7$ in ω_d, then the value of T_p is obtained equal to 0.07s.
Also, considering overshoot which is defined as:

$$M_p = e^{\frac{-\zeta\pi}{\sqrt{1-\zeta^2}}} \quad , \quad 0 \le \zeta \le 1 \tag{25}$$

The value of M_p is obtained equal to 0.05.

***Settling time*;** T_s is the time needed for system to decay to small value near steady state value, , for instance, when the response reaches the 0.01% of final value, the T_s can be approximated:

$$T_s = \frac{4.6}{\zeta\ \omega_n} \tag{26}$$

The value of T_s is obtained equal to 0.1 s.
Typically, it is difficult to determine more than three parameters from a step response unless the experimental conditions are exceptional. [3]

Hence, the comparison between derived transient response characteristics from the proposed model obtained by frequency response method and applying the step changes through the bump test into the system is shown in table 3.2.

***Table 3.2**: Compare derived transient response characteristics between the propose model from frequency response method and applying the step change into the system*

Motor Rotational speed Operation point(rad/s)	Input Amplitude(rad/s)	T_r	T_p	M_p	T_s
0	**15.7**	0.025s	0.05 s	3%	0.09 s
	31.4	0.04 s	0.07 s	5%	0.11 s
157	**15.7**	0.024 s	0.05 s	8%	0.10 s
	31.4	0.04 s	0.07 s	5%	0.14 s
Proposed Model based on equation (22)		0.03 s	0.07 s	5%	0.10 s

In conclusion, the system is stable by the step changes in both operating point, also table 3.2 illustrate that the transient specification of the model is near the transient specification of real step response in both operation point. However due to variable saturation of the motor causing nonlinearity properties there exists some differences. Furthermore the transient responses will be studied through the specific inputs to study more on the validation of the model.

3.2.2. Typical Identification Signal

After the model estimation based on frequency response method and validation based on step response of the system, the model is also analyzed and validated through the sum of sine signal and random Gaussian signal to find whether it captures the dynamic behavior of system in given frequency band or not [4].

The sum of sinusoid signal is applied as speed input in rad/s which is generating by matlab idinput command, where it has some remarkable properties such as:

- Sum of the sinusoid signal is typically used to identify the system.
- The frequencies are selected to be equally spread over the chosen grid, and each sinusoid is given a random phase.
- A number of trials are made, and the phases that give the smallest signal amplitude are selected, the amplitude is then scaled so as to satisfy the specifications of levels .[2]

Therefore to generate sum of the sinusoid signal at matlab, the following command is considered:

```
u = idinput (N,’sine’,band,levels)
u = idinput (10000,’sine’, 0.02, [-0.05 0.05])
```

Where:

- ”N” is the number of generated input data and the value of N is equal to 10000 means 10 second with sampling time equal to 0.001s.

- ”band” determine the lower and upper bound of the pass band, the frequencies are expressed in fractions of the Nyquist frequency it is normalized between 0 to 1, here by trial and error on the system the maximum band which can provide the considerable moving of the rotor is chose equal to 0.02. Figure 3.12 and 3.13 show the excited frequency band due to this selection at both operating points , thus [0 60] rad/s is provided ,where the (u1) is input and (y1) is output.
- Levels describe the amplitude required that normalized between [-1 to 1] corresponded -314 rad/s to +314 rad/s. Also similar to pervious amplitude limitation, it is chosen to be large enough to provide the good signal to noise ratio and small enough to not meet the saturation and cause the nonlinearity properties, thus it is considered to be 5% of maximum speed equal to 15 rad/s .

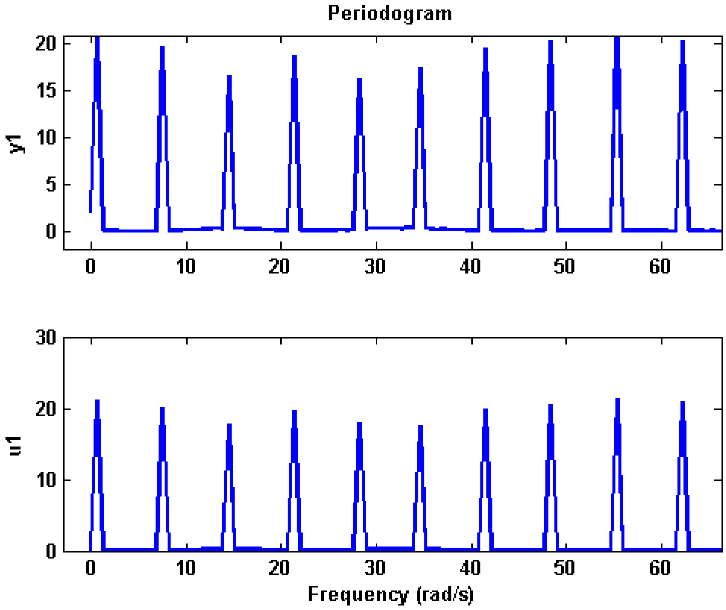

Figure 3.12*: Excitation frequency (sum of sine)y, at 0rad/s operating point*

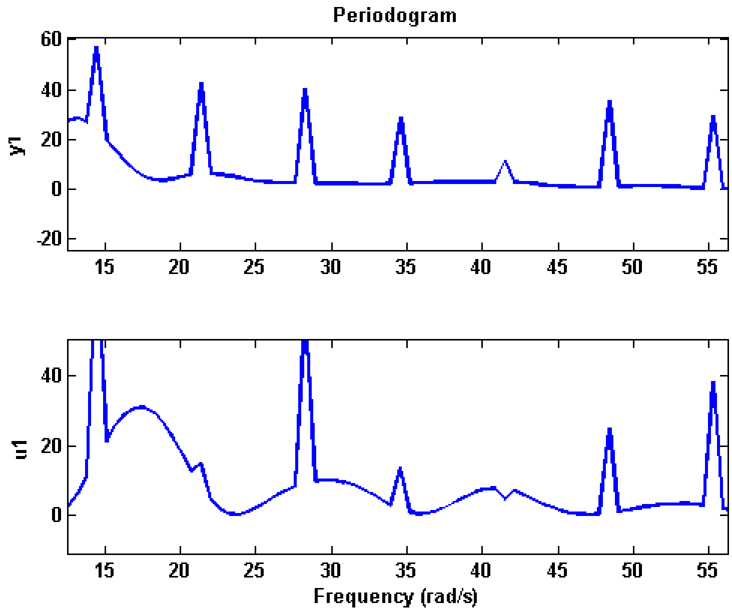

Figure 3.13*: Excitation frequency (sum of sine), at 157rad/s operating point*

Hence, considering the open loop model shown in figure 2.3, where the sum of sine signal is the speed input in rad/s and output of the system will be speed measurement in rad/s, Figure 3.14 shows the comparisons between the response of the real system at 0 rad/s operating point and the proposed model response.

Also, figure 3.15 shows the comparisons between the response of the real system at 157 rad/s operating point and the proposed model response.

Notice that, due to presence of high frequency noise in the output, the output data is averaged and plotted. Additionally, to find that how the model fit the output, we can compute the best fit percentage according to the equation (27):

$$Best\ Fit = (1 - (\frac{|y - \hat{y}|}{|y - \bar{y}|})) \quad (27)$$

In this equation, (y) is the measured output, ($\hat{y}$) is the simulated or predicted model output, and ($\bar{y}$) is the mean of (y). 100% corresponds to a perfect fit, and 0% indicates that the fit is no better than guessing the output to be a constant. Therefore according to the equation (27) and imply the matlab command, the model fit the output data approximately 30% for both operating point.

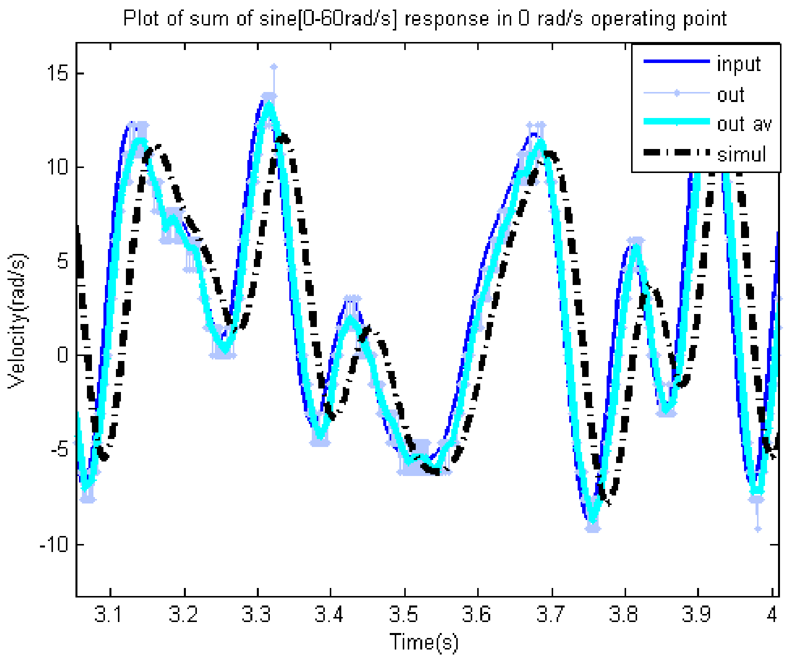

Figure 3.14: *Comparison of some part of output and the model response in rad/s at 0rad/s point through the sum of sine signal as speed input in rad/s*

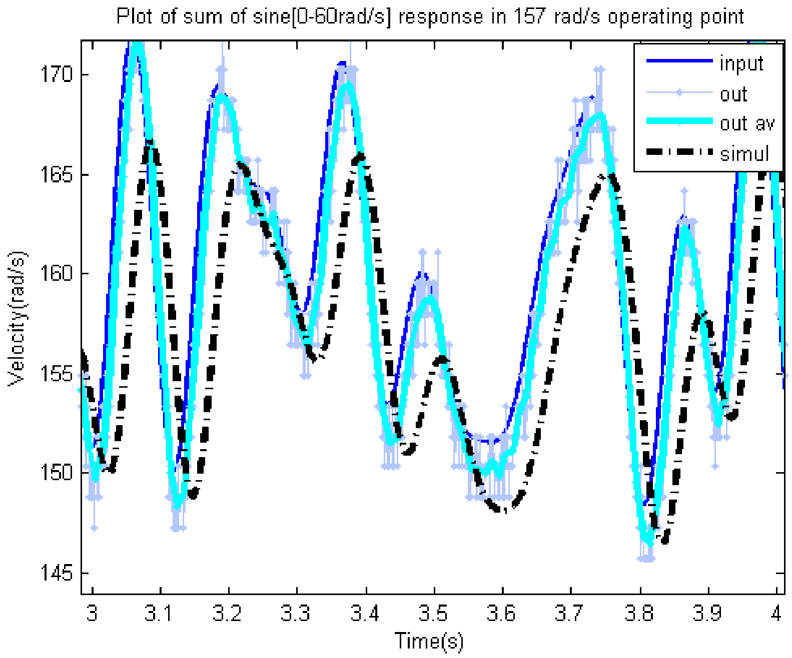

Figure 3.15*: Comparison of some part of output and the model response in rad/s at 157rad/s point through the sum of sine signal as speed input in rad/s*

Also, the random Gaussian signal is applied to the system as speed input in rad/s to validate the model, therefore the fowling command is considered:

```
u = idinput(N,'rgs',band,levels)
u = idinput (10000,' 'rgs'', 0.02, [-0.05 0.05])
```

Where, "N" and "band" are selected similar to sum of sine signal discussed before. Admittedly figure 3.16 and 3.17 show the excited frequency band at both operating point, where the (u1) is input and (y1) is output.

Notice that, the signal level is such that minimum level is the mean value of the signal minus one standard deviation, while maximum level is the mean value plus one standard deviation. Gaussian white noise with zero mean and variance one is thus obtained for levels = [-1, 1], here the level is considered also 0.05 which is corresponded to more than 30 rad/s. [2]

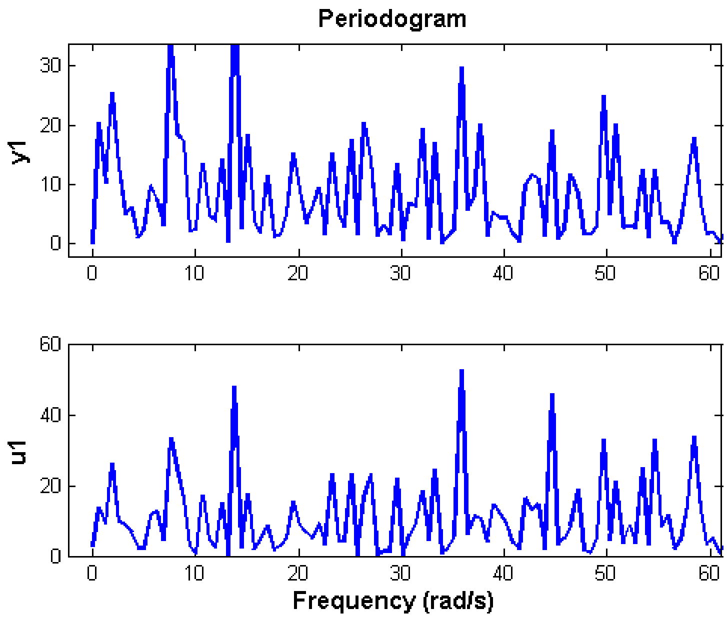

Figure 3.16*: Excitation frequency band of random Gaussian signal as speed input in rad/s, for 0rad/s operating point*

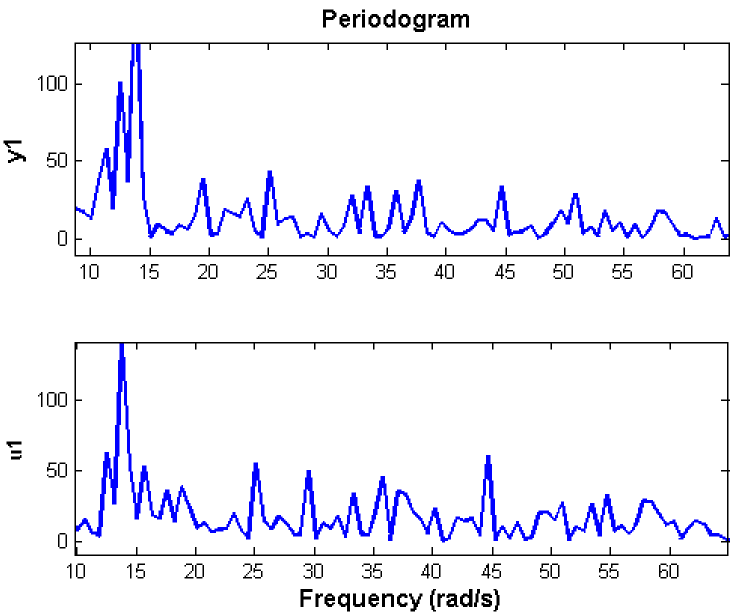

Figure 3.17*: Excitation frequency band of random Gaussian signal as speed input in rad/s, for 157rad/s operating point*

Hence, considering the open loop model shown in figure 2.3, where the random Gaussian signal is the speed input in rad/s and output of the system will be speed measurement in rad/s, Figure 3.18 shows the comparisons between the response of the real system at 0 rad/s operating point and the proposed model response. Also, figure 3.19 shows the comparisons between the response of the real system at 157 rad/s operating point and the proposed model response. Additionally, to find that how the model fit the output, we can compute the best fit according to the equation (27) and it equals to 50% for both operating point.

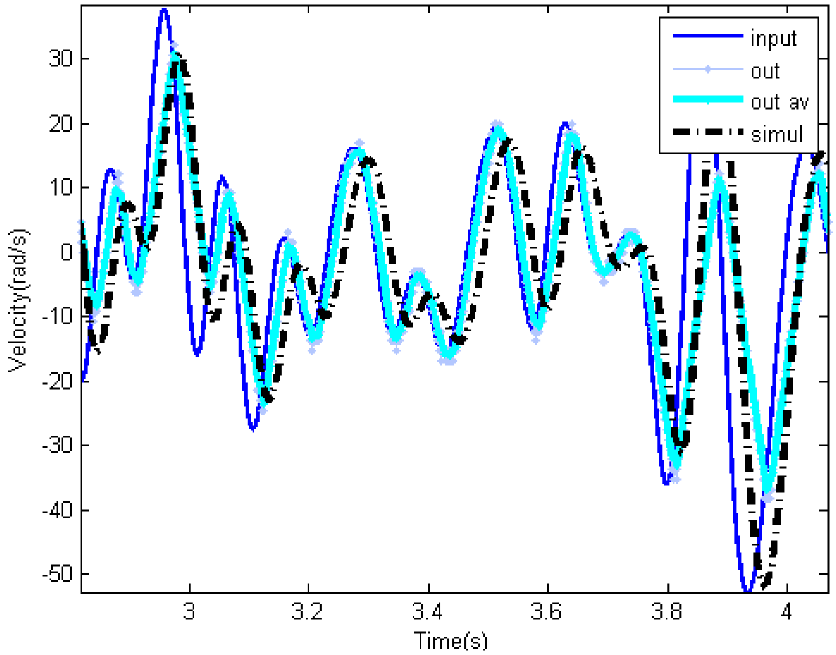

Figure 3.18: *Comparison of some part of output and the model response in rad/s at 0rad/s point through the random Gaussian signal as speed input in rad/s*

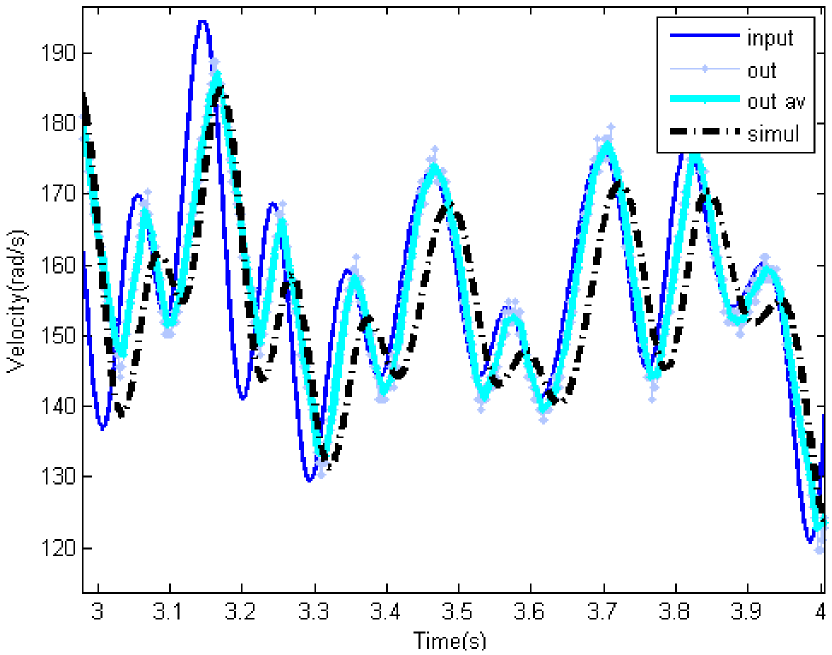

Figure 3.19: *Comparison of some part of output and the model response in rad/s at 157rad/s point through the random Gaussian signal as speed input in rad/s*

In a word, the system is estimated by pure second order model stated in equation (22) where the input and output of the system are in speed rad/s. The model is derived based on the frequency response method and it follows the frequency response of the real system in frequencies below the natural frequency at both operating point equal to 0 and 157 rad/s. Also the model is validated by the step, sum of sine and random Gaussian signals responses of the real system. In fact the model that we have estimated is optimized between frequency domain and time domain. For instance, adding a zero to the pure second order model can compensate the phase difference shown in figure 3.14, 3.15 and 3.18, 3.19, but it affects the transient response of real system, which makes it worse. Moreover, the system is stable in speed feedback mode, because it is not approach the natural oscillation by increasing the gain in loop gain means the closed loop model with only speed as feedback. Thus, there is no pole of system in origin or right hand side of s plane. So the pure second order model stated at equation (22) can represent the system.

3.3. Model of Position Include Saturation

Considering our case study, later on in section 4, we want to implement position control of the system through the proportional controller and phase-lead controller. Thus , the position model of the motor is needed to apply in closed loop implementation , furthermore the saturation effects of motor must be considered in simulation to obtain the response of the simulation which is more near the output of the real system.

3.3.1. Model of Position

The model of the motor presented in equation (22) is between speed as the input and output in rad/s, thus to obtain the model which is between position as input and output in rad, integrator should multiplied to equation (22). Thus, Equation (28) shows the model:

$$\frac{1}{(\frac{s}{\omega_n})^2 + 2\zeta(\frac{s}{\omega_n}) + 1} \times \frac{1}{s} \tag{28}$$

However, as the model presented in equation (28) is between position in rad as the input and output, to present model between position in meter as input and output equation (28) should be multiplied by conversion term, therefore the new model is:

$$\frac{9043}{s^3 + 84s^2 + 3600s} \tag{29}$$

Additionally, the block diagram of matlab simulation between control error and output in meter is shown in figure 3.20.

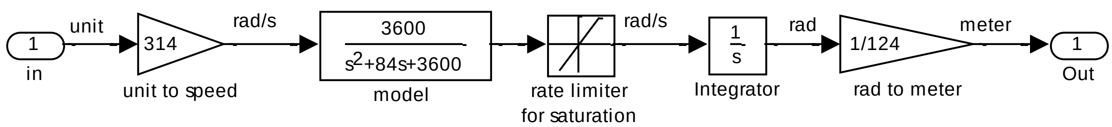

Figure 3.20: *Present the model of position in simulation model*

3.3.2. Saturation Effect

Speed of the motor is limited, and this fact is modeled in figure 3.21 by adding a rate limiter between integrator and the transfer function of the motor. The limited of derivative equal to 630 is derived by trial and error; In fact by comparison of step response of the proposed model and response of real system, this coefficient is obtained. [8]

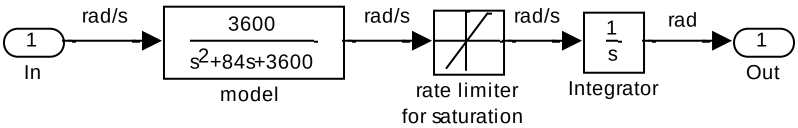

Figure 3.21: *Present the saturation effect of the motor in simulation model*

Also, to eliminate the speed of input signal at reference point, the rate limiter block is used in real implementation and simulation. Moreover, rate limiter makes the response of system more smother with lower overshot and provides the input signal which system can cope with [3]. The speed of the signal at reference input is reduced by rate limiter; the coefficient is considered by trial and error and for our system is assumed equal to 1.

4. CONTROLLER DESIGN

The purpose of this section is to design proportional and phase-lead controller throughout the closed loop position control of the electro motor servo system. Furthermore the obtained parameters from design stage, simulation and tuned on real system are going to be compered and analyzed.

4.1. Proportional Controller

For controller with proportional control action the relationship between the output of the controller $u(t)$ and the tracking error $e(t)$ is:

$$\frac{u(t)}{e(t)} = K_p \tag{30}$$

In fact, the proportional controller is essentially an amplifier with an adjustable gain.
Considering the position control block diagram of our system with proportional controller K_P, which is shown in figure 4.1, input is desired position in meter at reference point indicated by term (r), output is response of the system in meter that is indicated by (y) and disturbance in voltage which is indicated by (d).

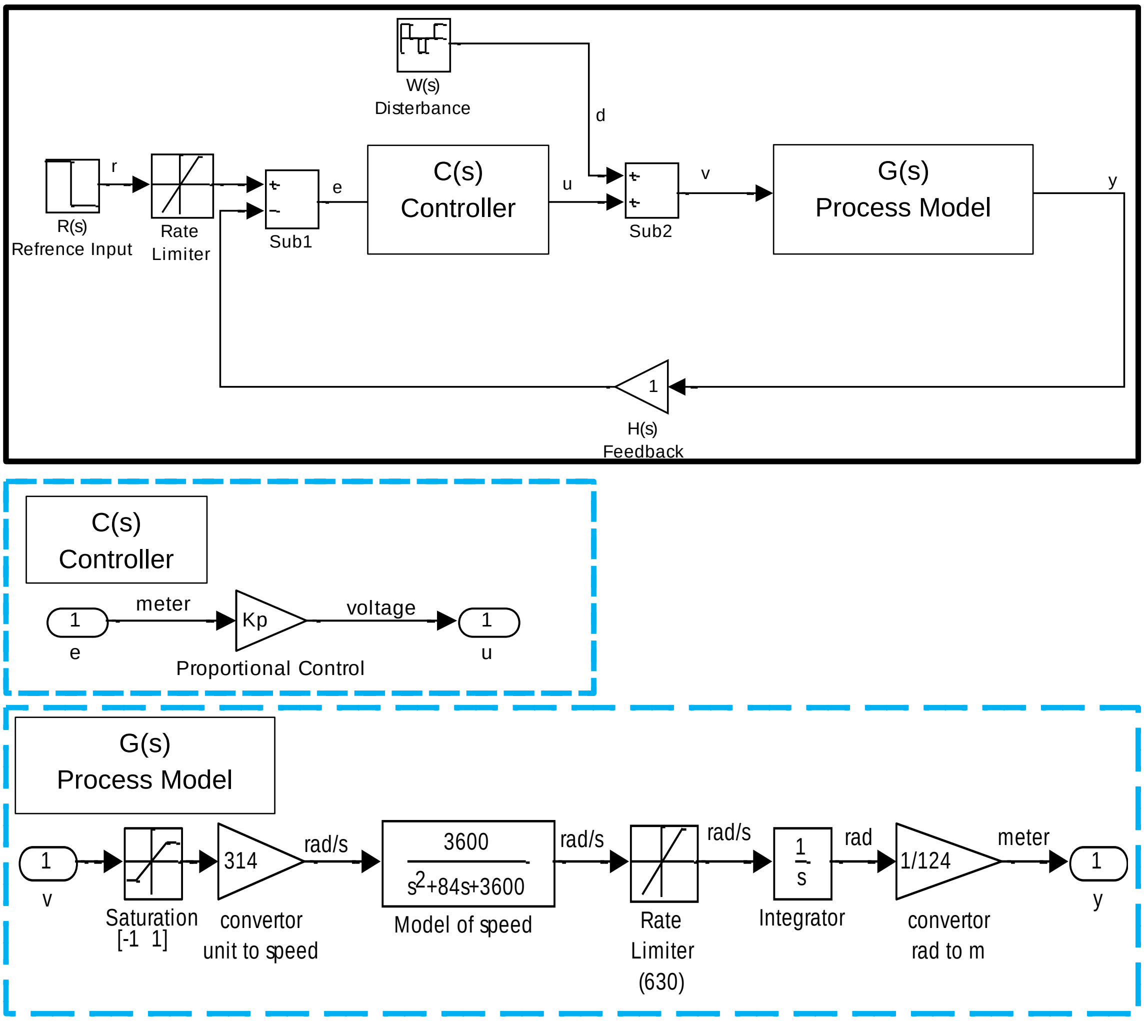

***Figure 4.1**: Proportional controller*

The main target to design controller is move the load to the desired position. Furthermore, table 4.1 shows the desired specifications should be met by designed controllers. [5]

***Table 4.1**: Assumed design specification*

Performance Measure	**Proportional Controller**
Settling time for step inputs with different amplitude at reference point.	Ts < 2 s
Percent Overshoot for step inputs with different amplitude at reference point.	Overshoot less than 25% No overshoot
Maximum Response for unite step inputs at disturbance point.	Y_d < 0.5 m

Hence, the proportional controller is going to be designed according to the table 4.1 needed performance specification and control laws. Then to study more on the designed K_P through the system response, the system is evaluated in the matlab simulation to figure out the proper K_P, and at the end, designed K_P are applied in real system to investigate their responses to select the better ones.

4.1.1. Design Proportional Controller

It is typically the case that implementing closed loop system provides faster response as the proportional gain is increased, and if there were no other factors, this is generally desirable.

However the response of higher order systems typically become less damped and eventually will become unstable as the gain is steadily increased. Therefore, there is a definite limit exists on how large the gain should be adjusted to eliminate the disturbance and sensitivity to parameters change.

In a word, feedback with proportional gain as controller changes the dynamic responses and with higher gain makes the system faster and less stable. Therefore, to choose the proper proportional gain, the stability condition and also the design specification are strongly considered. [1]

Thus, the following design steps are implemented to derive the proper proportional gains according to the required design specifications mentioned in table 4.1 and figure 4.1.

Step1: A stable system can be classified as a system type, defined to be the degree of the polynomial for which the steady state system error is a nonzero finite constant [1].

For instance, when the error to a ramp or first degree polynomial is a finite nonzero constant, such system is called type one that is the case here regarded to reference input. The type one system has a zero errors to step input at reference. Since C $(s) = K_P$ and $G(s) = \frac{9043}{s^3+84s^2+3600s}$, where they are shown in figure 4.1, the system error with the unity feedback case can be defined as:

$$Error_{closed\ loop} = \frac{1}{1 + C(s)G(s)}\, r - \frac{G(s)}{1 + C(s)G(s)}\, d \tag{31}$$

To derive the steady state error during the step input at reference point, neglecting the disturbance, the error is derived as:

$$E_r = \frac{1}{1 + K_P \frac{9043}{s^3 + 84s^2 + 3600s}}\, r \tag{32}$$

Substituting step input (called position) r $= \frac{1}{s}$ at reference point, and implementing the final value theorem to derive the steady state error, e_{ss} is:

$$\lim_{t\to\infty} e(t) = e_{ss} = \lim_{s\to 0} s \frac{s^3 + 84s^2 + 3600s}{s^3 + 84s^2 + 3600s + Kp9043} \frac{1}{s} \tag{33}$$

$$e_{ss} = 0$$

Also considering the system type regards to disturbance input with no input at reference point, this error is derived as:

$$E_d = \frac{\frac{9043}{s^3 + 84s^2 + 3600s}}{1 + Kp\frac{9043}{s^3 + 84s^2 + 3600s}} d \tag{34}$$

Substituting the disturbance input d $= \frac{1}{s}$, and implementing final theorem, e_{ss} is:

$$\lim_{t\to\infty} e(t) = e_{ss} = \lim_{s\to 0} s \frac{9043}{s^3 + 84s^2 + 3600s + K_P 9043} \frac{1}{s} \tag{35}$$

$$e_{ss} = \frac{1}{K_P}$$

Hence, studying the system type regards to step input at disturbance point, can lead us how large K_P contribute to reduce the disturbance error. So to meet the required disturbance response less than 0.5 m according to table 4.1, the K_P should be more than 2.

Step2: Since increasing the K_P make the system faster however it makes the system unstable with large K_P, therefore the stability of system must be studied to find out how large K_P can provide the system with good dynamic response also keep stability.
Considering the transfer function from output to input of closed loop control system:

$$\frac{y}{r} = \frac{C(s)G(s)}{1 + C(s)G(s)} \tag{36}$$

A necessary condition for stability of the system is that all of the roots of transfer function of system have a negative real part which in turn requires all the coefficients of characteristic polynomial be positive. Considering the characteristic equation:

$$1 + C(s)G(s) = 0 \quad (37)$$

$$s^3 + 84s^2 + 3600s + K_P 9043 = 0$$

Therefore, according to the Routh test, a system is stable if and only if all the elements in the first column of the Routh array are positive. Therefore implementing the Routh test: [5]

s^3	1	3600
s^2	84	$K_P 9043$
s^1	b	0
s^0	K9043	

Where b is derived as:

$$b = \frac{84 \times 3600 - Kp \ \times 9043}{84} \quad (38)$$

Noticeably, The case b=0 make the system marginally stable, means when the K_P is equal to 33.44, so in order to maintain the system stable, K_P must be less than 33.44, it corresponds the roots on jw- axis at $s = \pm j60$ .Also this critical K_P can be derived by study the root locus by Matlab, when the root locus of model G(s) is considered at figure 4.2, It confirms the maximum $K_P = 33.44$ that is corresponded to the roots located on Jw axis.

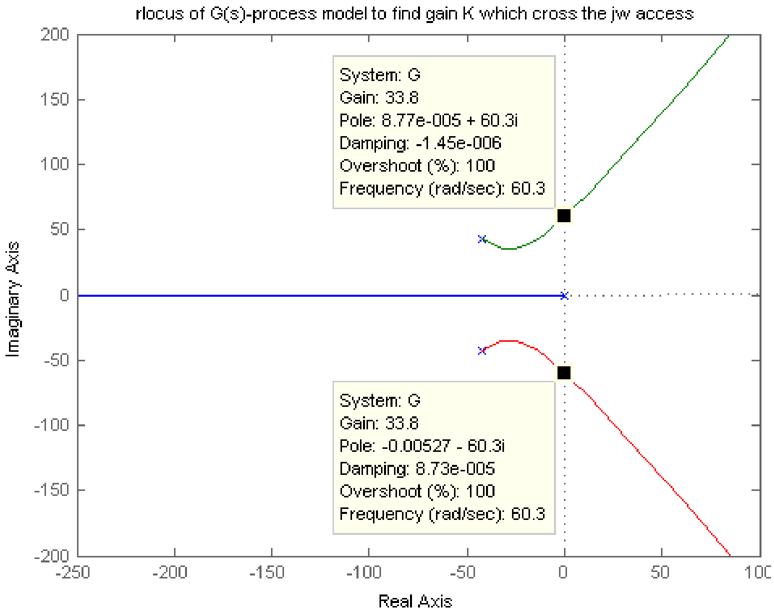

Figure 4.2: *Rlocus of G(s) in open loop system*

Consequently, the K_P must be less than 33 to maintain the system stable during the position feedback control.

Step 3: Based on required transient characteristics assigned in design specifications table 4.1, the computational program 4.1 is considered. The requirements are to provide over shoot less than 25%, and the settling time less than 2s.

```
t=0:0.01:3;
k=0;
for Kp=2:1:33;
    G=tf(9043,[1 84 3600 0]);         % process model
    tf_closed=feedback(Kp*G,1);
    y = step(tf_closed,t);            %step of closed loop
    s = 301; while y(s)>0.95&& y(s)<1.05;s=s-1;end;
    ts=(s-1)*0.01;                    %ts = settling time
    m=max(y);
       if m<1.25 && m> 0.9            %cond for over shoot
            if ts < 2                 %cond for settling time
            k=k+1;
            solution_step(k,1:3)=[Kp m ts];
            end;
       end;
end;
```

Program 4.1*: Computational program to derive the required Kp for P.O < 25% and Ts < 2s [7]*

Therefore, the proportional gains K_P are derived base on step response of closed loop system, the results shows in table 4.2.

Table 4.2*: The results of computation program based on required transient response*

Kp	over shoot	ts
2	1.00	0.70 s
3	1.00	0.44 s
4	1.00	0.31 s
5	1.00	0.22 s
6	1.00	0.15 s
7	1.01	0.10 s
8	1.06	0.15 s
9	1.10	0.14 s
10	1.14	0.13 s
11	1.19	0.20 s

Hence, according to the limitations on selection of K_P , from needed disturbance rejection, stability studies, and transient response requirements, they lead us to choose K_P more than 2 and less than 11. Additionally following the design steps, the relative stability studies based on derived K_P will be implemented.

Step 4: The relative stability of the compensated control system with proportional controller can be studied by open loop bode diagram to derive the phase and gain margin. The phase margin is that amount of additional phase lag at the gain crossover frequency required bringing system to the verge of instability; also the gain margin is the reciprocal of magnitude $|G(jw)|$ at the frequency which the phase angle is -180 degree. For robust performance, the phase margin should be more than 30 degree and the gain margin more than 6 dB. [7]

Notice that, when K_P is less than 9, phase and gain margins of the system is more than 59 degree. Hence, all K_P more than 2 and less than 9 meet the relative stability criteria. Moreover, the response of the system through the matlab simulation, are going to be studied in more details to select K_P more precisely.

4.1.2. Proportional Controller in Simulation

The proposed K_P gains from 4.1.1, are implemented in Simulink model to study the response of system in more details, the simulation model is based on figure 4.1.

Initially K_P equal to 2, 3 and 4 are applied on Simulink model to compare the system responses to step change in meter as input at reference point with different amplitudes, to evaluate which of them provide better response with no overshoot. Therefore, according to the figure 4.3, K_P equal to 3 provides the better transient response and would be proposed as initial guess for tuning procedure on real system.

In fact, when K_P is equal to 3, we meet no overshoot for different step changes correspond to 0.1, 0.05, 0.4, 0.7 meter at reference point. Notice that for K_P is equal to 4 we meet approximately 5% overshoot for step changes correspond to 0.4 and 0.7 meter, which is not desirable.

Also settling time is less than 1 s for K_P equal to 3, which is desirable in our case. Although K_P equal to 2, provides both overshoot and steeling time criterion but it makes system slowly. Hence K_P equal to 3 is the optimized K_P which is proposed in this controller design stage due to the simulation results.

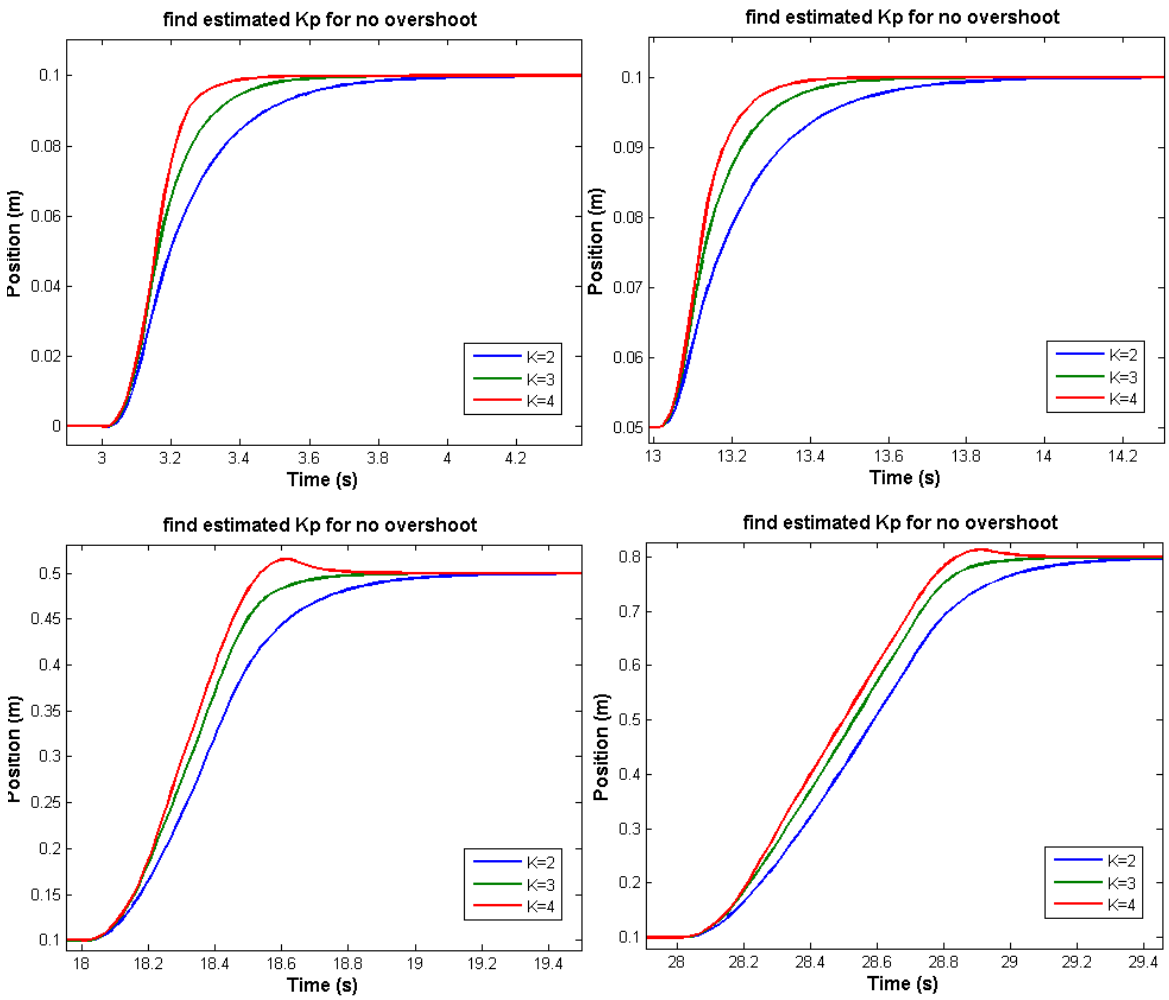

Figure 4.3: *Step responses by simulation for* K_P *=2, 3 and 4 to meet no over shoot. The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

Also K_P equal to 6, 8 and 10 are applied on Simulink model to compare the system responses to step change in meter as input at reference point with different amplitudes, and evaluate which of them provide better response with overshoot less than 25%. According to the figure 4.4, K_P equal to 6 provide the better transient response and would be proposed as initial guess for tuning procedure on real system.

In fact, when K_P is equal to 6, we meet overshoot less than 25% for different step changes correspond to 0.1, 0.05, 0.4, 0.7 meter at reference point. Notice that for K_P equal to 8 and 10, we meet approximately 45% overshoot for step changes correspond to 0.1, 0.05, and 0.4 meter, which are not desirable.

Also settling time is less than 1 s for K_P equal to 6, which is desirable in our case. Although K_P equal to 8, provides steeling time criterion but it makes system behave more oscillatory. Hence K_P equal to 6 is the optimized K_P which is proposed in this controller design stage due to the simulation results.

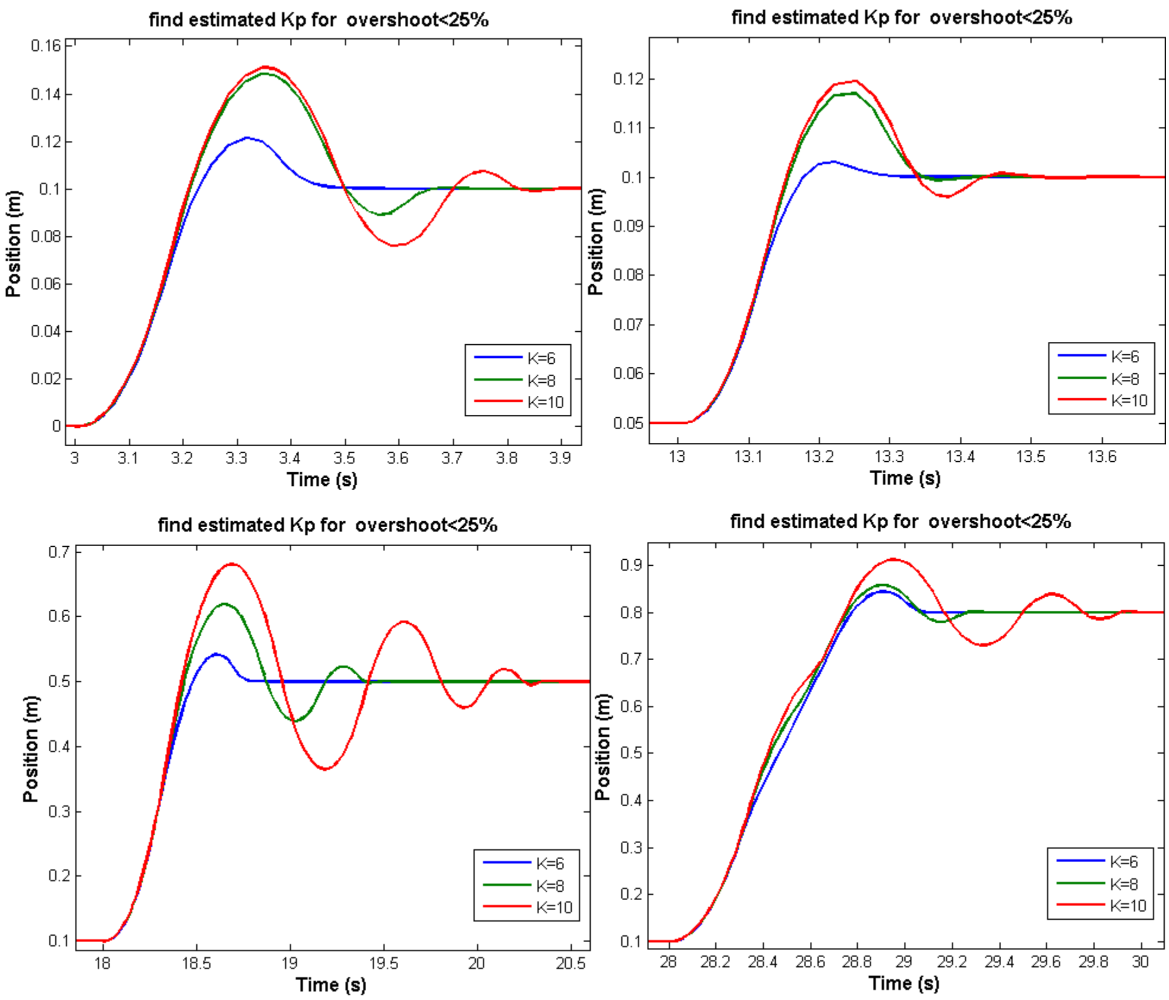

Figure 4.4: *Step responses by simulation for* K_P *=6, 8 and 10 to meet over shoot less than 25%. The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

4.1.3. Proportional Controller in Real System

Finally, the K_P optimized by simulation, is applied in real system, figure 4.5 shows the Simulink model implemented to provide closed loop position control throughout the proportional controller, the step input with different amplitude in meter is applied at reference point and output measured in meter is recorded by control desk presented in section 2.4.

Notice that, measured output is in unit which should be converted to meter. Since each 20640 increment is corresponded to 1.1 meter, thus it multiplied by 5e-5.

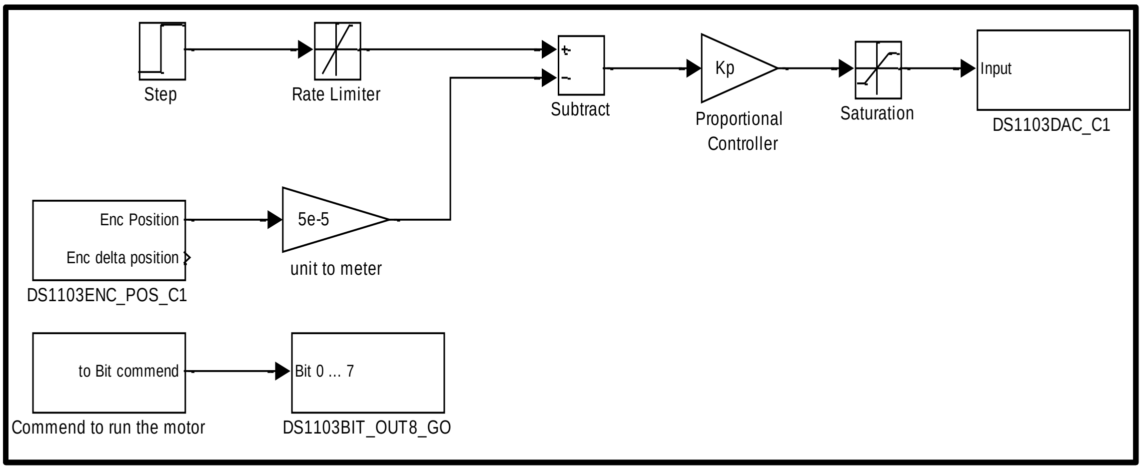

Figure 4.5*: The closed loop using proportional controller implemented in real system.*

Hence, step responses for K_P equal to 6, is shown in figure 4.6 and for K_P equal to 3 is shown in figure 4.7.

Although due to nonlinearity system shows the different transient behavior to different amplitudes, but the responses are acceptable according to the design specification table 4.1. Therefore the desired over shoot and settling time are provided for system with and without load mentioned in section 2.1.

Moreover, considering figure 4.6, when K_P is equal to 6, overshoot of the system with and without load is less than 25%. It reaches maximum amount equal to 17% for 0.1 meter step change in compare to other step changes. Also settling time of the system with and without load for K_P equal to 6 is less than 1 s. The system need maximum amount of 1 s to reach the 98% of final value for maximum amount of step change equal to 0.7 meter.

Hence K_P equal to 6 is the optimized K_P which is finalized in this controller design stage due to the real system results.

Additionally, considering figure 4.7, when K_P is equal to 3, the system with and without load meet no overshoot. Also settling time of the system with and without load for K_P equal to 3 is less than 1 s. The system need maximum amount of 1 s to reach the 98% of final value for maximum amount of step change equal to 0.7 meter. Notice that it can be observed that due to implementing smaller K_P in compare to before , the system acts slowly but the results still can acceptable according to the table 4.1.

Hence K_P equal to 3 is the optimized K_P which is finalized in this controller design stage due to the real system results.

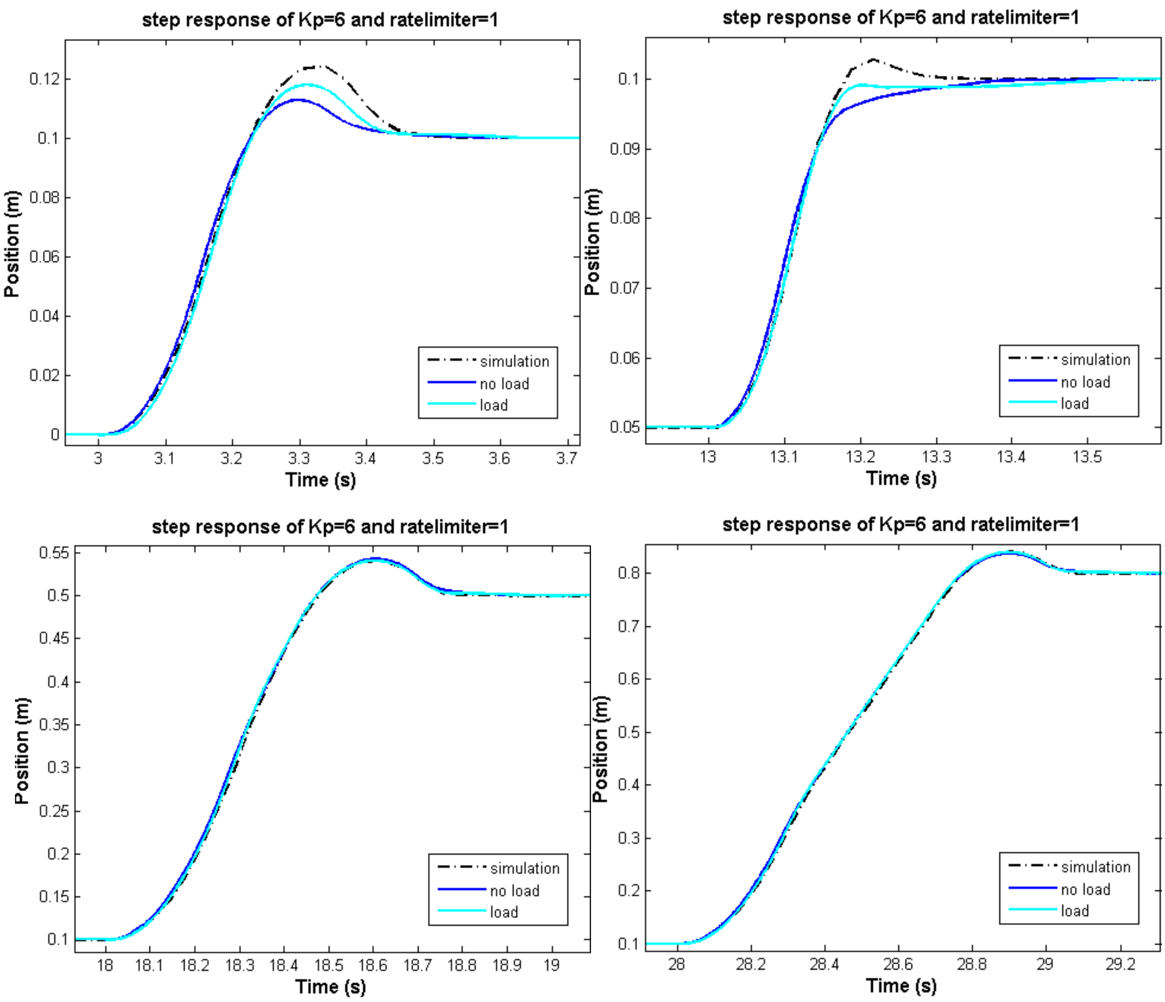

Figure 4.6*: Step responses of real system for* K_P *=6 to meet over shoot less than 25%. The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

In conclusion, table 4.3 shows the comparison between transient response characteristics among design, simulation and real system.

Since we consider linear model in design stage, the results are quite different from simulation, but it can be a valuable start point in design proportional controller. Also as there are some other nonlinearity in system which are not considered in simulation the results are different between simulation and real system but the differences are not considerable in this case and simulation can be quite useful to predict the system behavior to choose the desired K_P.

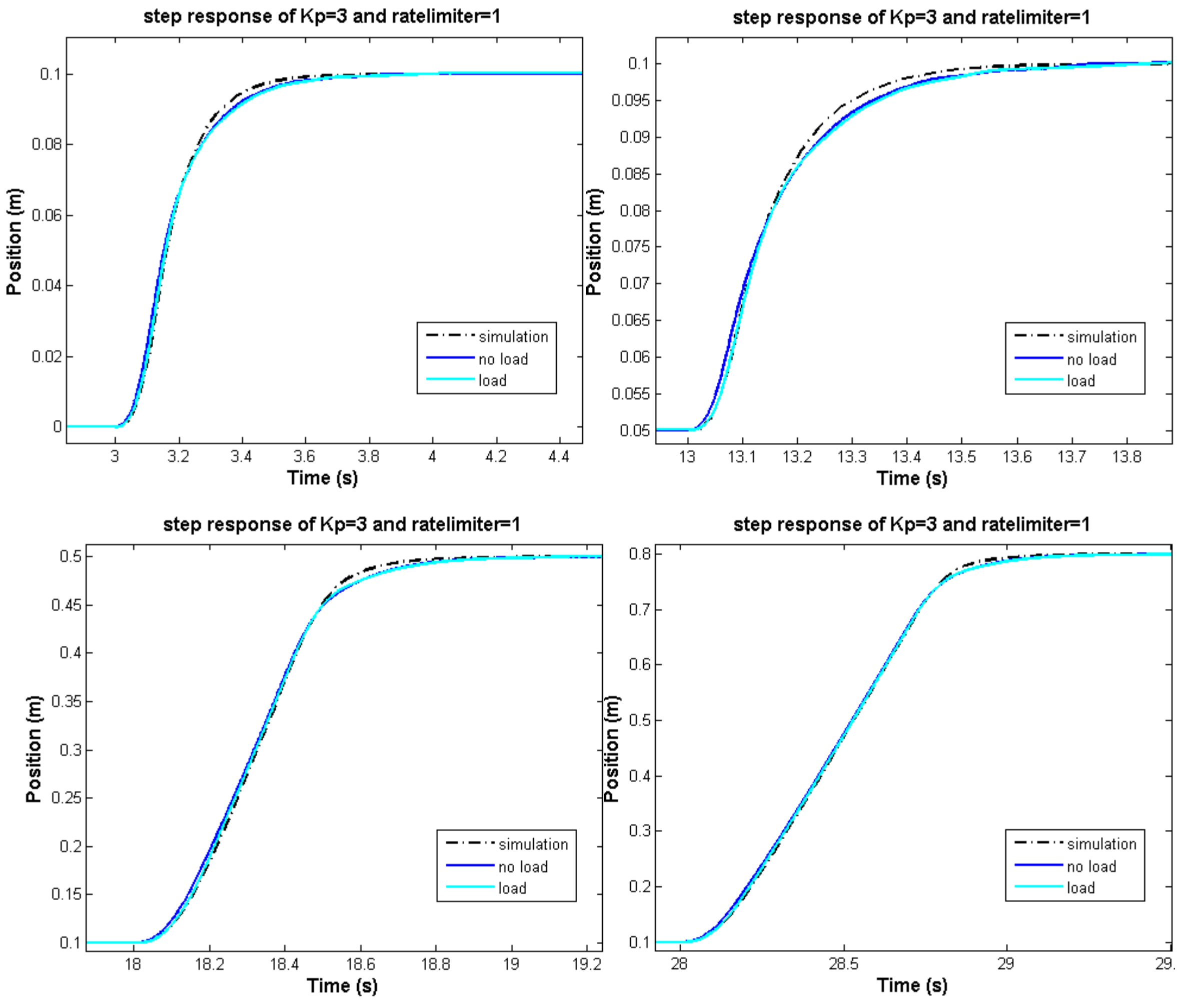

***Figure 4.7**: Step responses of real system for K_P =3 to meet no over shoot. The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

***Table 4.3**: Comparison of transient response of system with proportional controller K_P*

Step Change (meter)	Proportional Controller	Computed [Section 4.1.1]		Simulation [Section 4.1.2]		Real-Empty [Section 4.1.3]		Real-Load [Section 4.1.3]	
		t_s	P.O	t_s	P.O	t_s	P.O	t_s	P.O
0.1	$K_P = 3$	0.44 s	0	0.60 s	0	0.60 s	0	0.60 s	0
	$K_P = 6$	0.15 s	0	0.40 s	23%	0.40 s	12%	0.40 s	17%
0.05	$K_P = 3$	0.44 s	0	0.44 s	0	0.44 s	0	0.44 s	0
	$K_P = 6$	0.15 s	0	0.40 s	3%	0.40 s	0	0.40 s	1%
0.5	$K_P = 3$	0.44 s	0	0.70 s	0	0.70 s	0	0.70 s	0
	$K_P = 6$	0.15 s	0	0.75 s	8%	0.70 s	8%	0.75 s	8%
0.7	$K_P = 3$	0.44 s	0	1.00 s	0	1.00 s	0	1.00 s	0
	$K_P = 6$	0.15 s	0	1.00 s	5%	1.00 s	5%	1.00 s	5%

4.2. Phase Lead Controller

Lead compensation approximates the function of PD controls and acts mainly to speed up a response by lowering rise time. Considering the PD transfer function as:

$$C(s) = (T_D s + 1) \tag{39}$$

We use this compensation by locating $\frac{1}{T_D}$ so that the increased phase occurs in the vicinity of crossover frequency, thus increasing the phase margin. Note that, the magnitude of the compensation continues to grow with increasing frequency. This feature is undesirable because it amplifies the high frequency noise, therefore, in order to eliminate the high frequency amplification of the PD compensation a first order pole is added in the denominator at the frequencies substantially higher than the break point of the PD compensator. Thus the phase increase still occurs but the amplification at high frequency is eliminated, and the transfer function of new compensation is: [3]

$$C(s) = K\frac{Ts + 1}{\alpha Ts + 1} \quad \alpha < 1 \tag{40}$$

In a word, the phase lead controller is implemented to provide better transient response and compensate the phase, which may be lost in process model or measurement device. For instance, it happens when measurement device needs several sampling times to compute the output (like vision), or when the device itself has significant dynamics.

In our case, phase lead controller is applied within different implementations where the main object for both is to control the position of system according to the designed specification mentioned in table 3.1.

The implementation I of phase lead controller is shown in figure 4.8, where the desired position in meter is applied at reference point as an input indicated by the term (r) and the response of system in meter as an output indicated by the term (y), and the term (d) indicates disturbance in voltage. Notice that, the derivative action is applied on the control error which is indicated by (e). In fact, with the derivative in forward pass, a step change in the reference input, in theory, makes an intense initial pulse in the control signal which is undesirable.

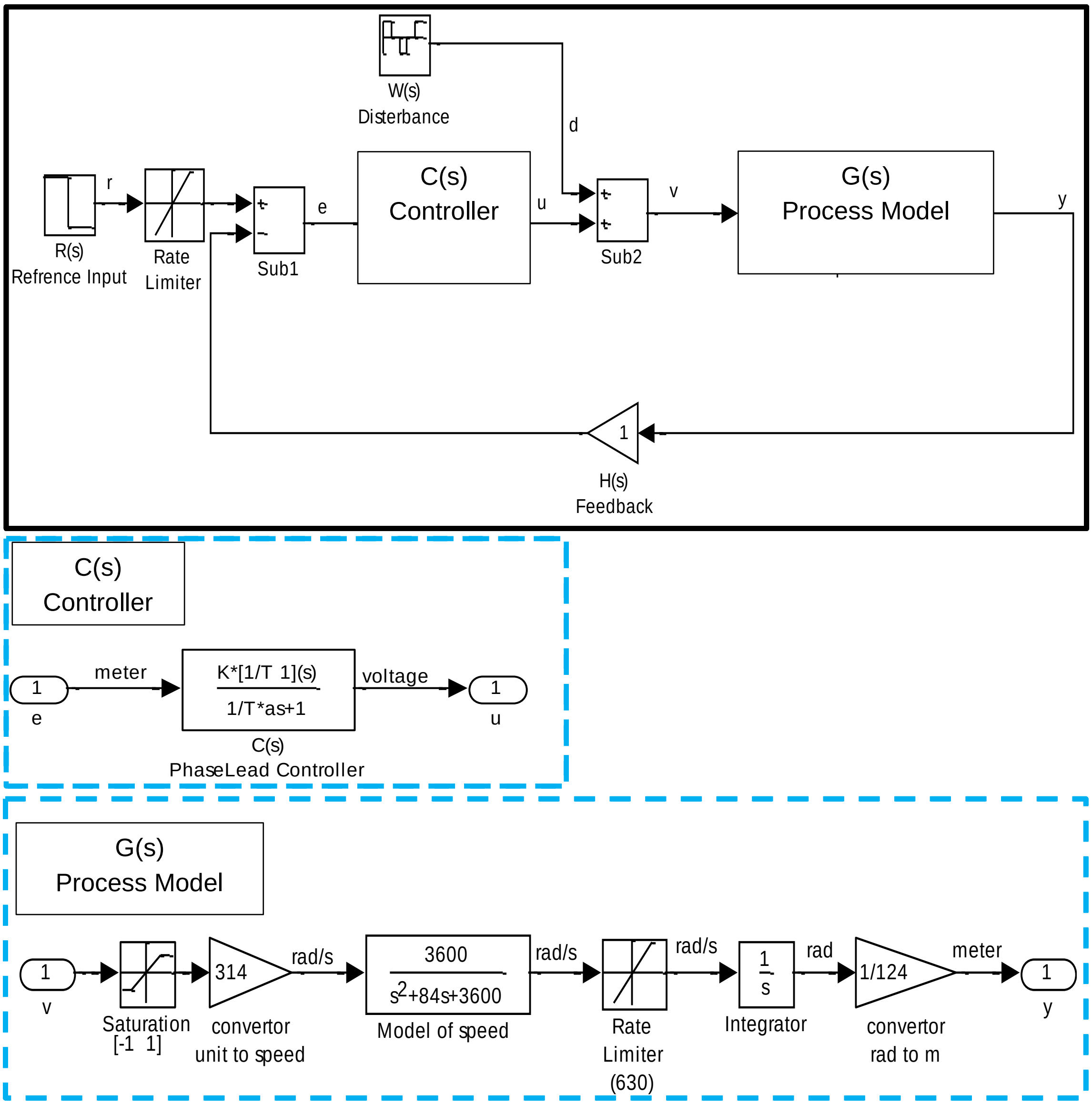

Figure 4.8: *Phase-lead controller. Implementation I*

Velocity feedback is very commonly used in positional servo systems, mostly because of sharp effect of derivative control on suddenly changing signals can be avoided by implementation II shown in figure 4.9. Thus term (D) is introduced into the feedback path and reference is not differentiated which is desirable result if the reference is subject to sudden changes.

Notice that, implementation I and II have the same characteristic equations. Considering the model from equation (29) and phase lead controller from equation (40), table 4.3 shows the transfer functions from (r) and (d) to (y) for both implementations.
Furthermore, if the system subjected to noise signals, velocity feedback may generate some difficulty as we will see in our case later, the result will be accentuation of the noise effects. [7]

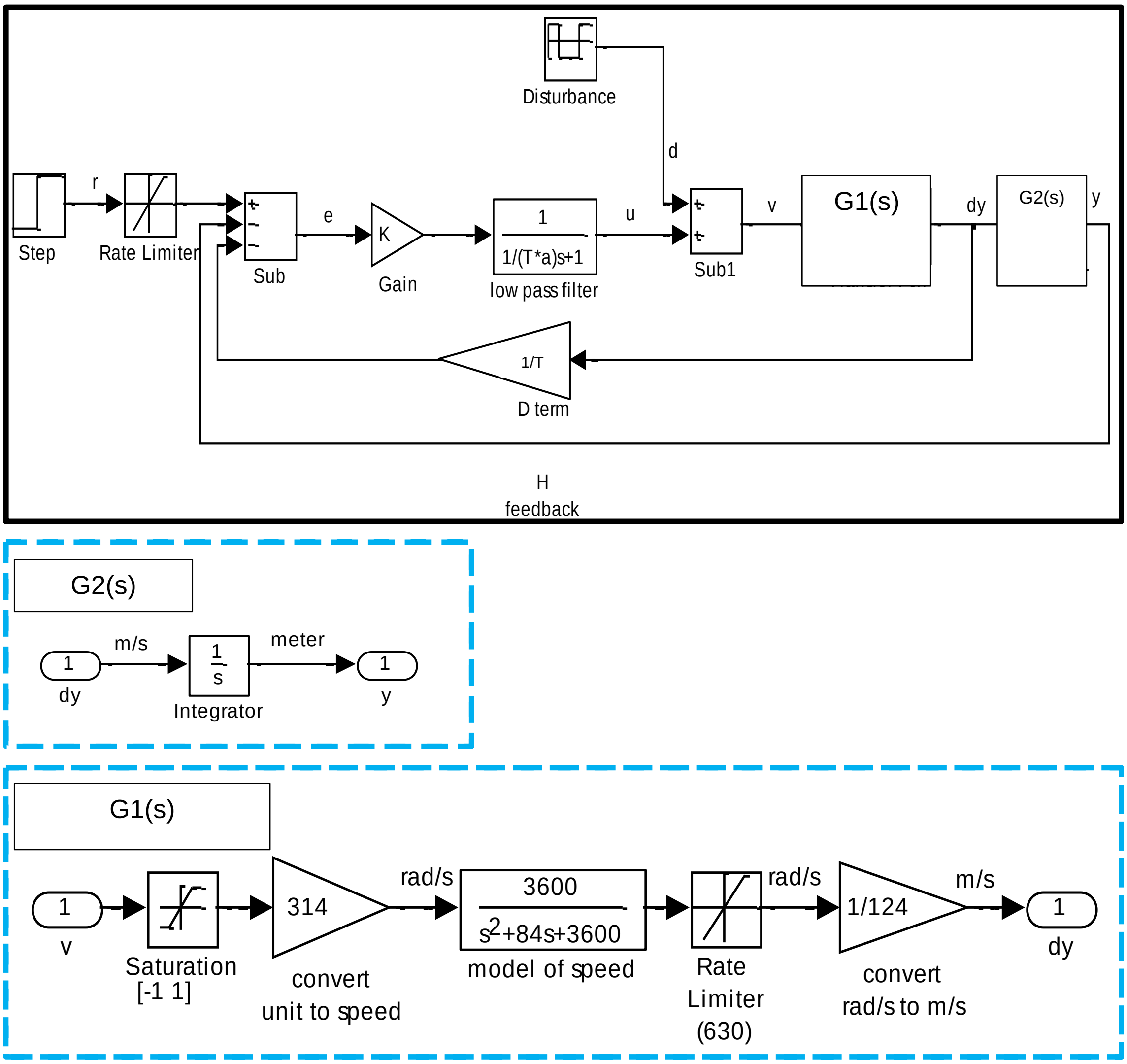

Figure 4.9: *Phase-lead controller. Implementation II*

Table 4.4*: The comparisons of transfer functions throughout implementations I and II.*

Relations	Implementation	Transfer functions
From r to y	**I**	$y = \frac{K(Ts+1)\times 9043}{(\alpha Ts+1)\times(s^3+84s^2+3600s)+K(Ts+1)\times 9043} r$
	II	$y = \frac{9043}{(\alpha Ts+1)\times(s^3+84s^2+3600s)+K(Ts+1)\times 9043} r$
From d to y	**I**	$y = \frac{(\alpha Ts+1)\times 9043}{(\alpha Ts+1)\times(s^3+84s^2+3600s)+K(Ts+1)\times 9043} d$
	II	$y = \frac{(\alpha Ts+1)\times 9043}{(\alpha Ts+1)\times(s^3+84s^2+3600s)+K(Ts+1)\times 9043} d$

Moreover , based on implementation I, the parameter of controller is going to be designed , then the designed controllers are tested throughout both implementation in matlab Simulink model and Finlay the result of real system are evaluated to figure out the better control parameters.

4.2.1. Design Phase Lead Controller

Initial phase lead parameters are going to be designed by following steps. Notice that, they are some initial guess, and the tuning must be done on the simulation model and finally on real system.

Step 1: The K from equation (40) as an open loop gain to satisfy required error and bandwidth requirements is determined in this stage. According to equation (35) the K should be more than 2 to lead the system to meet e_{ss}<0.5 meter, that is the error specification from disturbance point to output point.

Hence, to meet upper bandwidth compare to pure proportional controller, while the natural frequency of the model is equal to 60 rad/s based on equation (22), the value of K equal to 8 is chosen, therefore the cross over frequency will be 20 rad/s. Also to provide the response with no over shoot the K is considered as 4 which corresponded to cross over frequency equal to 10 rad/s.

Step 2: The phase margin (PM) of the uncompensated system should be evaluated using the value of K obtained from step1.

To do so, considering the open loop bode diagram of the system with K equal to 8 and the model from equation (29), the phase margin is 62 degree in cross over frequency equal to 20rad/s. Thus, to increase cross over frequency to 40 rad/s with 60 degree as gain margin, 30 degree margin is needed plus 5 degree extra margin.

Also, considering the open loop bode diagram of the system with K equal to 4 and the model from equation (29) the phase margin is 76 degree at cross over frequency equal to 10 rad/s. Thus, to increase cross over frequency to 25 rad/s with 60 degree as gain margin, 35 degree margin is needed plus 5 degree extra margin.

Step 3: The attenuation factor α is determined by using:

$$\sin(\varphi) = \frac{1-\alpha}{1+\alpha} \tag{41}$$

When $\varphi_{max} = 35$ degree, α is 0.27, and for $\varphi_{max} = 40$ degree, α is 0.21.

Step 4: The corner frequencies $\omega = \frac{1}{T}$ and $\omega = \frac{1}{\alpha T}$ should be determined. To do so, as the maximum phase lead angle φ_{max} occurs at the geometric mean of the two corner frequencies, or $\omega = \frac{1}{\sqrt{\alpha}T}$, the amount of modification in the magnitude curve at $\omega = \frac{1}{\sqrt{\alpha}T}$ due to inclusion of term$\frac{Ts+1}{\alpha Ts+1}$ is :

$$\left|\frac{1 + j\omega T}{1 + j\omega\alpha T}\right|_{\omega=\frac{1}{\sqrt{\alpha}T}} = \frac{1}{\sqrt{\alpha}} \tag{42}$$

Note that the magnitude in decibel is:

$$20log\frac{1}{\sqrt{\alpha}} \tag{43}$$

When α is 0.27, magnitude is -5.67dB and when α is 0.21 the magnitude is -6.70dB. Following the design procedure, according to the obtained magnitude, for K=8 it corresponds to ω =37 rad/s as new gain cross over frequency , also for K=4 it corresponds to $\omega = 22$ rad/s. noting that these frequencies corresponds to $\omega_c = \frac{1}{\sqrt{\alpha}T}$, thus the zero and pole are obtained by equations:

$$\frac{1}{T} = \sqrt{\alpha}\omega_c \tag{44}$$

$$\frac{1}{\alpha T} = \frac{\omega_c}{\sqrt{\alpha}}$$

Therefore, the compensators for both K are determined as:

$$C(s)_{k=8} = 8\frac{\frac{1}{19}s + 1}{\frac{1}{71}s + 1} \tag{45}$$

$$C(s)_{k=4} = 4\frac{\frac{1}{10}s + 1}{\frac{1}{50}s + 1} \tag{46}$$

Step 5: Open loop Boad diagram of the compensated system for K=4 and K=8 are drawn to check the cross over frequency and phase margin of compensated system, they are shown in figure 4.10 and 4.11.

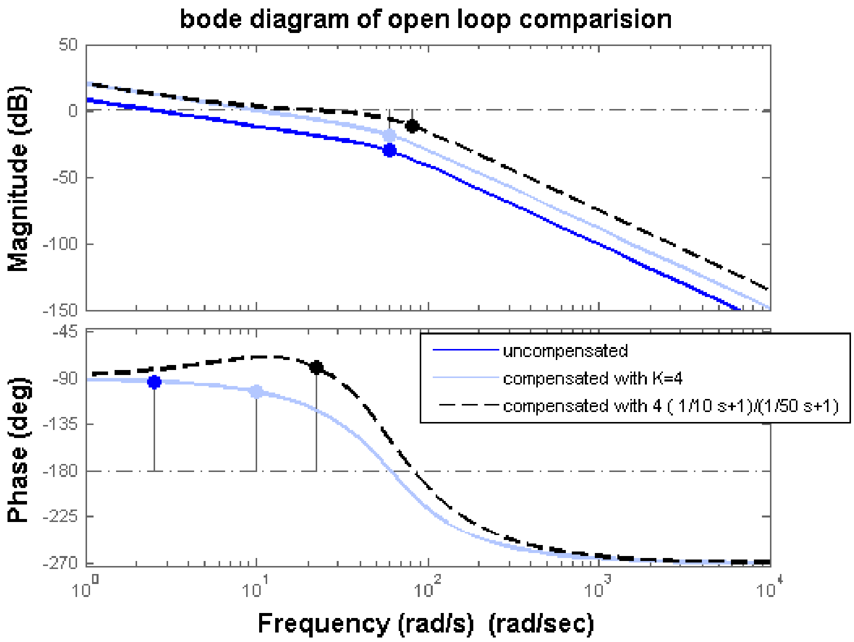

***Figure 4.10**: Comparisons of open loop Boad diagrams of phase lead controller stated in equation (46)*

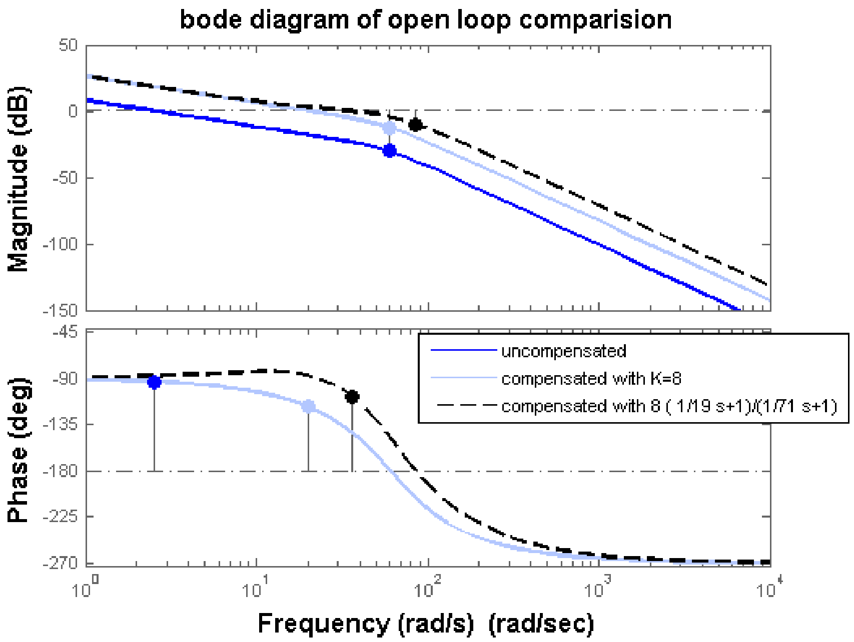

***Figure 4.11**: Comparisons of open loop Boad diagrams of phase lead controller stated in equation (45)*

Hence, figure 4.10 shows that the compensated cross over frequency is 22 rad/s with phase margin equal to 101 degree, and figure 4.11 shows that the compensated cross over frequency is 37 rad/s with phase margin equal to 70 degree . Both designed phase lead controller parameters provide the needed relative stability conditions, later on these parameters are applied in simulation throughout implementation I and II to study in more details.

4.2.2. Phase Lead Controller in Simulation

The proposed phase lead controller parameters obtained in 4.2.1, are implemented in Simulink model to study the response of system more precisely, the models are based on figure 4.8 and 4.9.

According to the simulation results shown in figure 4.12, the designed controller meets the desired specification performances. However the disturbance, friction, noise, delay and other unknown elements are not applied in this simulation results and by tuning the designed controller on real machine the final designed parameters are obtained.

Moreover, for phase lead controller in equation (45), overshoot is less than 25% for different step changes at two implementations I and II, while the implementation I shows more overshoot but still acceptable due to the desired performance specification from table 4.1. Also, settling time is less than 1 s for different step changes at both implementations.

Also, for phase lead controller in equation (46), system meets no overshoot for different step changes at two implementations, while implementation II makes the system behave more smoothly. Although settling time for implementation II is increased but still it confirms the design specifications.

Hence, in simulation results as it can be seen, the implementation II of phase lead controller with velocity feedback, states the better transient responses in compare to implementation I, but later on, it is shown that, due to unknown elements the behavior could be even worth in implementation II.

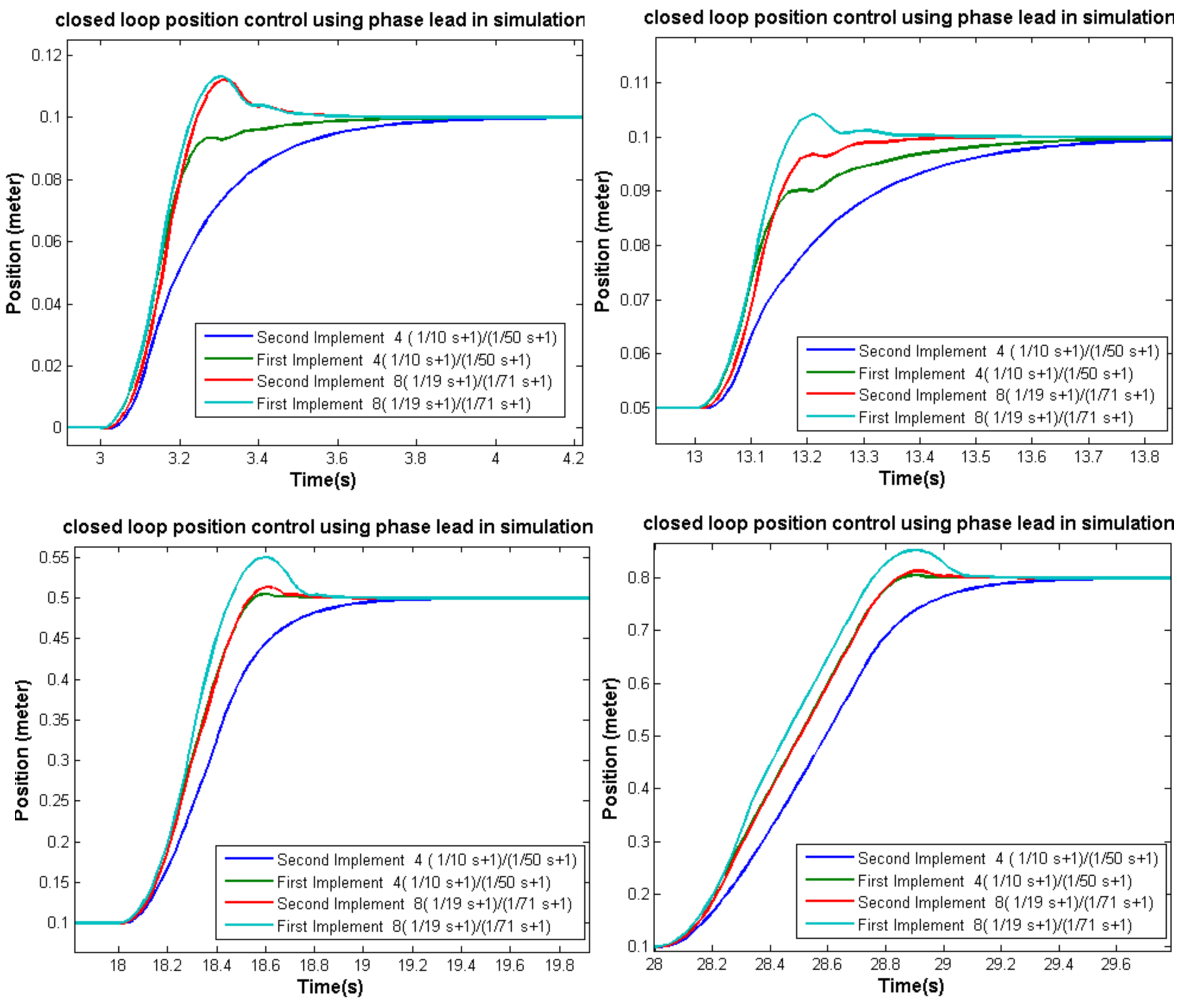

Figure 4.12: *Step responses by simulation with phase lead. The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

4.2.3. Phase Lead Controller in Real System

The position closed loop control system includes phase lead controller throughout implementation I is applied on matlab Simulink according to the figure 4.13, also implementation II is shown in figure 4.14.

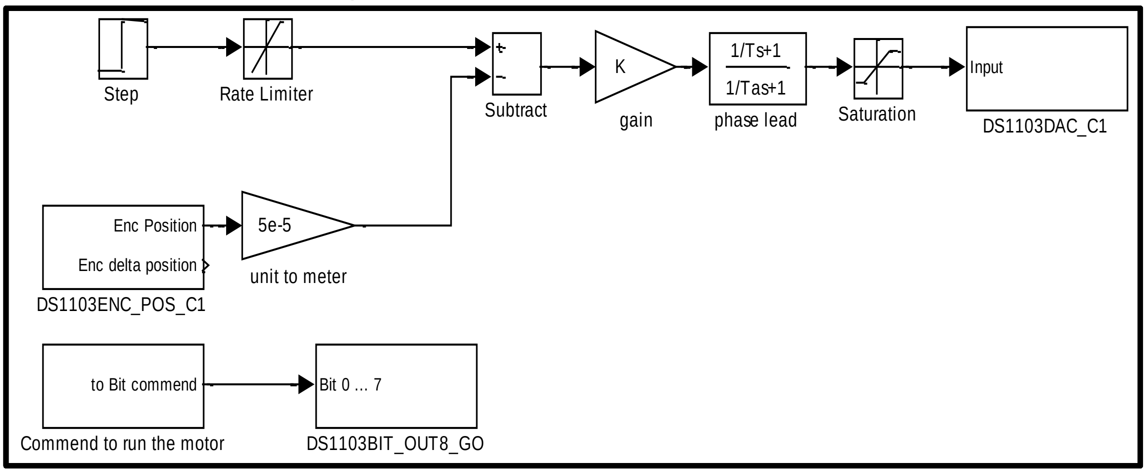

Figure 4.13: *Phase lead controller in real system. Implementation I*

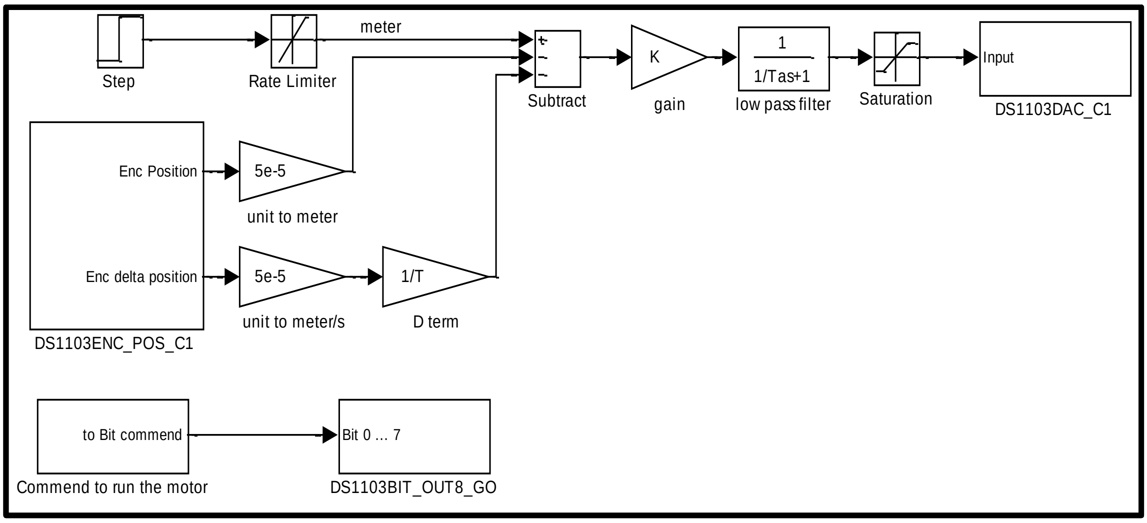

Figure 4.14: *Phase lead controller in real system. Implementation II*

Since the disturbances, noise, friction and other unknown elements are exist in real system, the tuning is needed to meet the desirable performance specifications. In our case, after using the initial guess obtained by designed procedure , as the unknown elements make the response of system oscillatory for larger gains through implementation II, it eliminates us to applied the faster controller , thus the gains are reduced to meet the better transient response in real time workshop.

$$3.4\frac{\frac{1}{10}s+1}{\frac{1}{58}s+1} \quad (47)$$

$$5\frac{\frac{1}{10}s+1}{\frac{1}{30}s+1} \quad (48)$$

Furthermore, the tuned parameters from implementation II are applied in implementation I and the comparison are shown in figure 4.15 and 4.16 to provide response with overshoot less that 25% and in figure 4.17 and 4.18 are to provide response with no over shoot.

Considering figure 4.15, the real system with and without load meet the required overshoot less than 25% for different step changes throughout Implementation I, while the maximum amount of overshoot is for step change correspond to 0.4 meter and it is 5%. Also, the settling time for different step changes is below 2 s, while the maximum amount of settling time is 1.1 s for maximum step changes correspond to 0.7 meter.

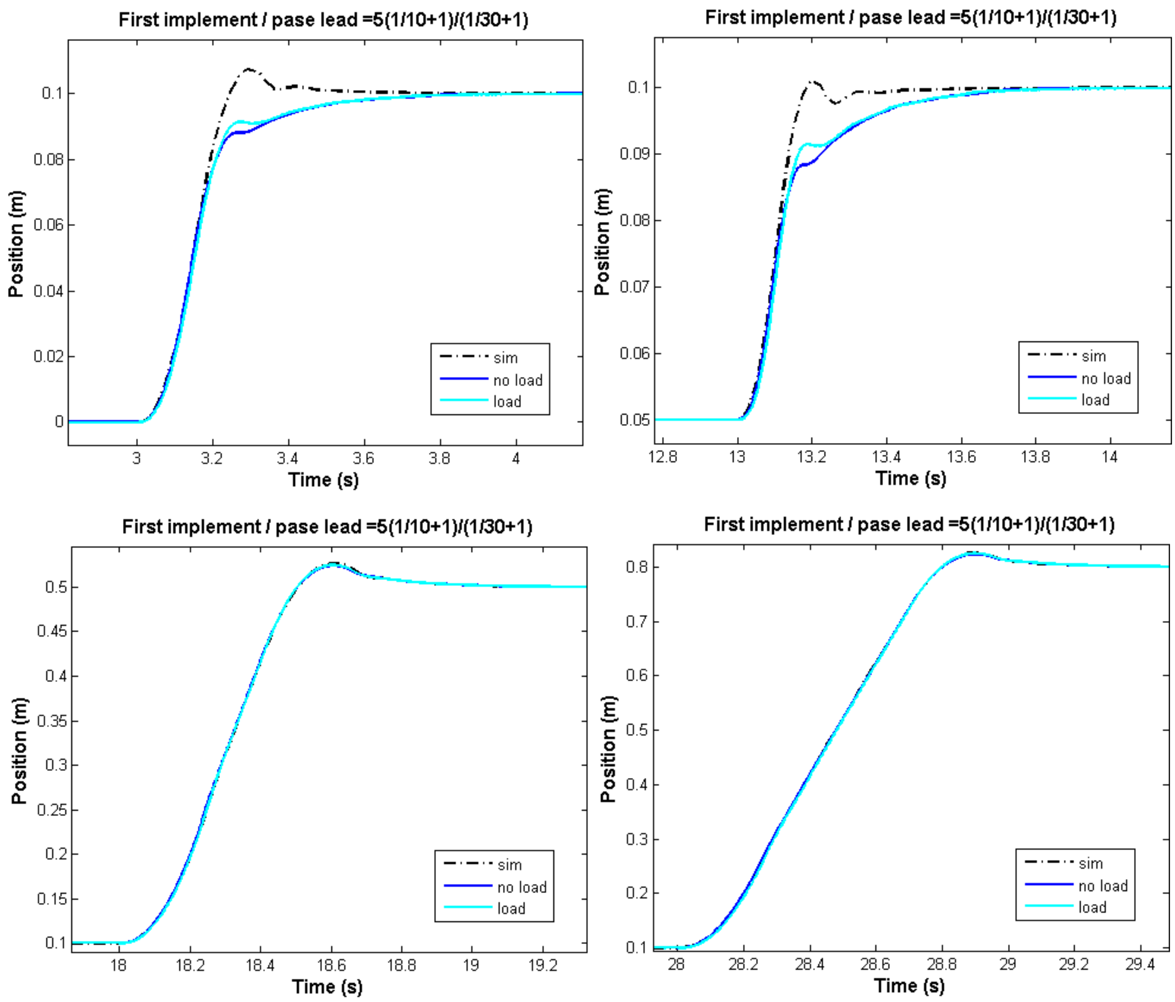

Figure 4.15*: Step responses by real system with phase lead controllers (Equation (48)).Implementation I. The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

Considering figure 4.16, the real system with and without load meet the required overshoot less than 25% for different step changes throughout implementation II, while the maximum amount of overshoot is for step change correspond to 0.1 meter and it is 23%.

Also, the settling time for different step changes is below 2 s, while the maximum amount of settling time is 1 s for maximum step changes correspond to 0.7 meter. Notice that, the real system in implementation I follow the simulation results, but in implementation II the result of the real system indicate considerable difference in transient response. In fact, Implementation II makes the system behave fast.

Thus, the real system result for tuned controller presented in equation (48) among implementation I and II, can meet the desired specification required in table 4.1.

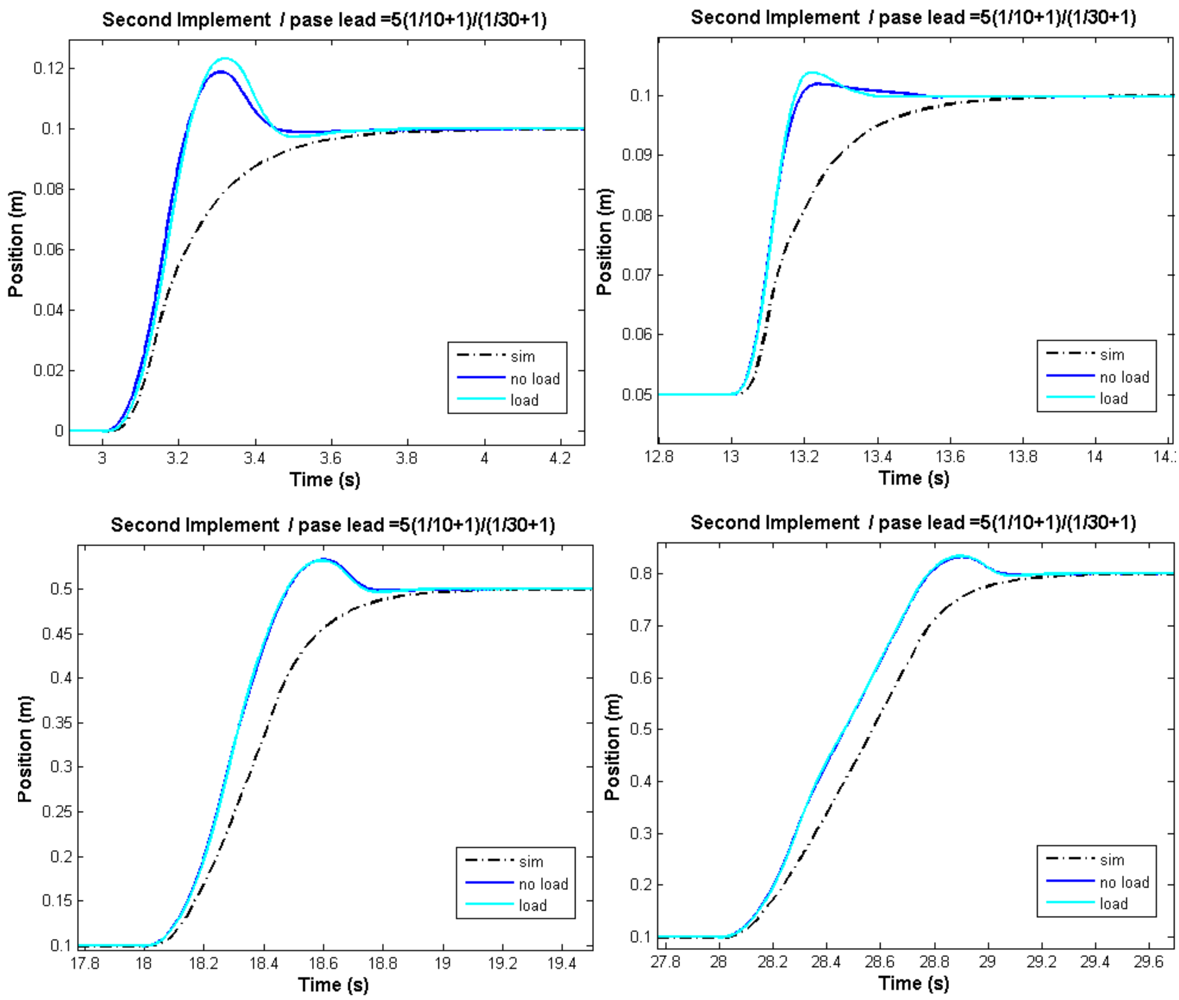

Figure 4.16: *Step responses of real system with phase lead controllers equation (48). Implementation II . The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

Considering figure 4.17, the real system with and without load meet no overshoot for different step changes throughout implementation I.

Also, the settling time for different step changes is below 2 s, while the maximum amount of settling time is 1.2 s for maximum step changes correspond to 0.7 meter.

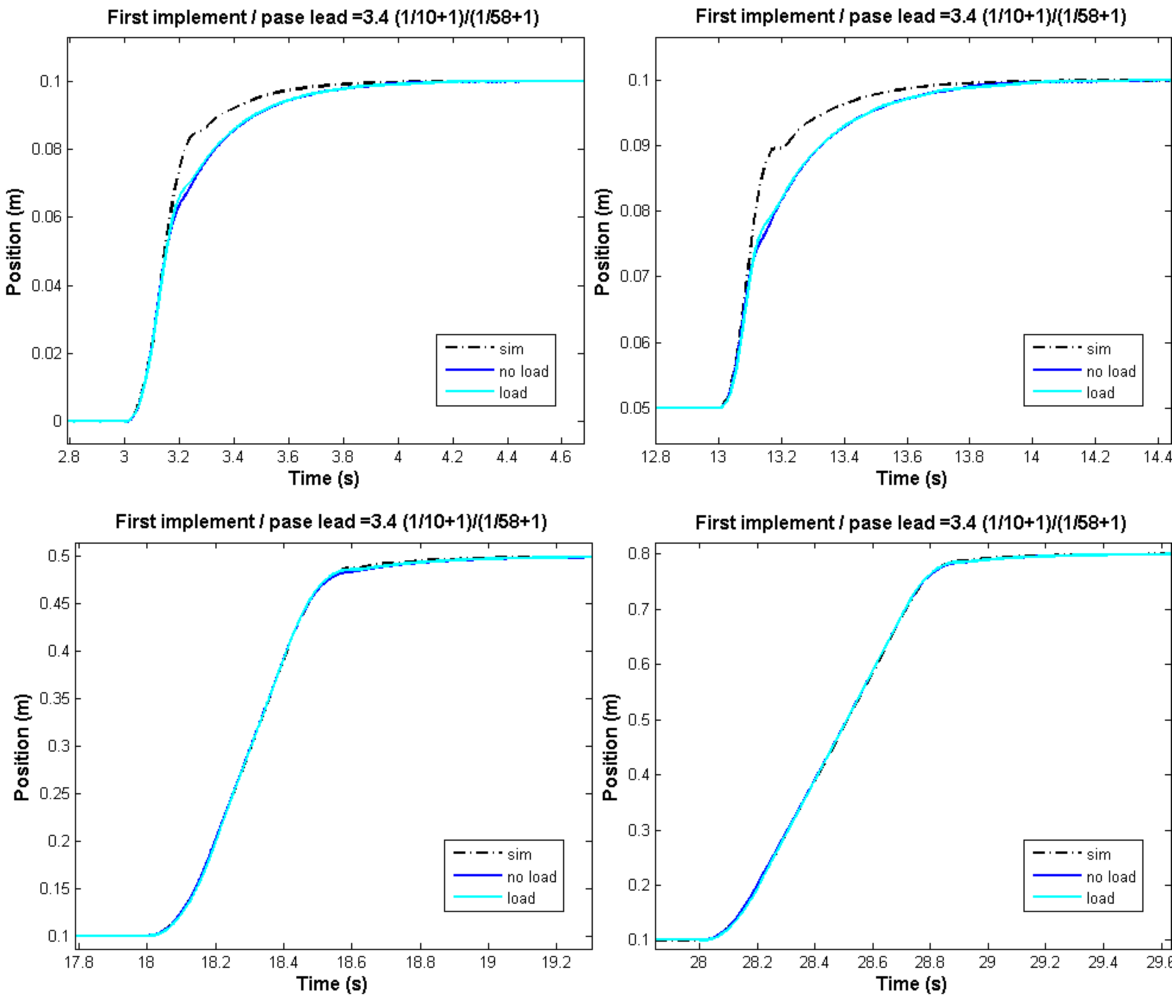

Figure 4.17: *Step responses by real system with phase lead controllers equation (47). Implementation I. The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

Considering figure 4.18, the real system, with and without load meet no overshoot for different step changes throughout implementation II.
Also, the settling time for different step changes is below 2 s, while the maximum amount of settling time is 1 s for maximum step changes correspond to 0.7 meter.
Notice that, the real system in implementation I follow the simulation results, but in implementation II the result of the real system indicate considerable difference in transient response. In fact, Implementation II makes the system behave fast.

Thus, the real system result for tuned controller presented in equation (47) among implementation I and II, can meet the desired specification required in table 4.1.

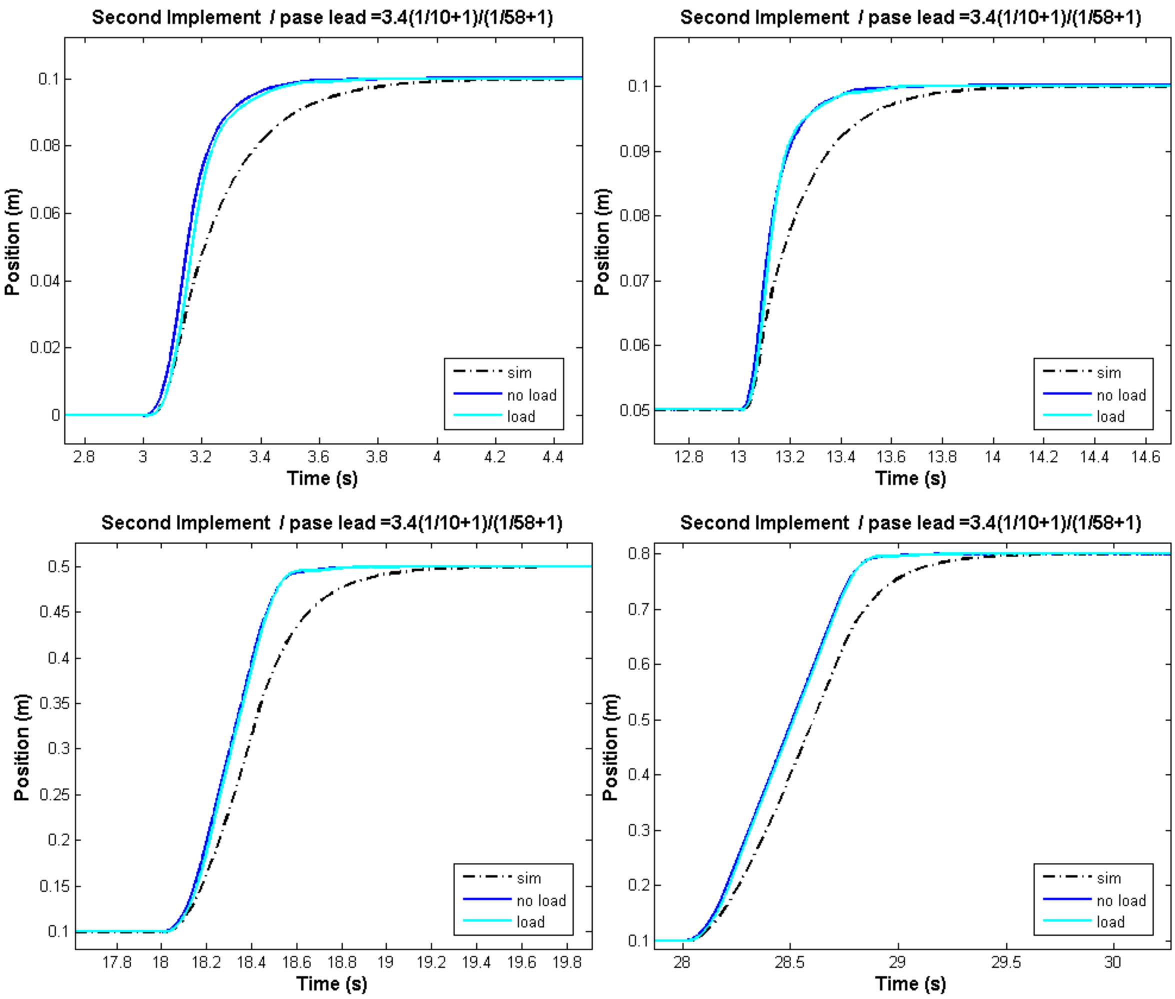

Figure 4.18*: Step responses by real system with phase lead controllers equation (47). Implementation II. The figures are the step change with 0.1, 0.05, 0.4, 0.7 meter.*

In conclusion, according to the table 4.5, although the second implementation face the system to unknown oscillatory behavior, probably because of presence of noise and delay in velocity measurement , it contributes the closed system to act more faster in comparison to first implementation with same parameters .

***Table 4.5**: Transient response comparisons between the proportional control and phase lead controller include both implementations*

Step Change (meter)	Phase-Lead Controllers	Implementation	Real-no Load Phase Lead		Real-Load Phase Lead		Real-no Load Proportional Controller		Real-Load Proportional Controller	
			t_s(s)	P.O	t_s(s)	P.O	t_s(s)	P.O	t_s(s)	P.O
0.1	Equation(47)	I	0.82	0.0	0.82	0.0	0.60	0	0.60	0
		II	0.48	0.0	0.48	0.0				
	Equation(48)	I	0.59	0.0	0.59	0.0	0.40	12%	0.40	17%
		II	0.56	18%	0.56	23%				
0.05	Equation(47)	I	0.68	0.0	0.68	0.0	0.44	0	0.44	0
		II	0.38	0.0	0.38	0.0				
	Equation(48)	I	0.50	0.0	0.50	0.0	0.40	0	0.40	1%
		II	0.4	2%	0.40	4%				
0.5	Equation(47)	I	0.70	0.0	0.70	0.0	0.70	0	0.75	0
		II	0.55	0.0	0.55	0.0				
	Equation(48)	I	0.74	5%	0.74	5%	0.70	8%	0.70	8%
		II	0.65	8%	0.65	8%				
0.8	Equation(47)	I	1.2	0.0	1.2	0.0	1.00	0	1.00	0
		II	1	0.0	1.0	0.0				
	Equation(48)	I	1.1	2%	1.1	2%	1.00	5%	1.00	5%
		II	1	5%	1 s	5%				

5. CONCLUSION

The main object of this thesis is to identify and control the servo electro motor, the permanent magnet synchronies AC motor is driven through master derive, d-Space control board, matlab simulink and control desk. In fact, the matlab simulink model is developed and it is compiled on d-Space control board throughout the control desk as an interface, then master drive feeds the motor due to the received control signals from the d-Space control board.

Since primarily step to provide controller is to develop a mathematical description called dynamic models of the process to be controlled, the model should be identified. To do so, after numerous experimental tests on the motor, frequency response method is chosen to be the main approach to the modeling. The motor is excited by sinusoid input signals in variety of frequency ranges from 0.1 to 230 rad/s at two different operating points 0 and 157rad/s. Notice that, it is implemented in an open loop control system when the input and output is speed in rad/s. Hence, the model is estimated based on the obtained Bode diagram. Moreover, the transient response of the model is evaluated and compared with the transient response of real system throughout the step input presented by the bump test. Also the responses of the model and real system are to typical identification signals such as sum of sinusoid signal and Random Gaussian Noise signal are studied.

Additionally, as we need the position model, the term integer should be considered into the model. Notice that, due the speed of the motor is limited by saturation effect, a rate limiter between integrator and the transfer function of the motor is also added to the simulation model.

The main target to design controller is move the load to the desired position while the system is unstable and cannot be controlled using open loop. Thus proportional controller and phase lead controller are designed, applied, and tuned based on assumed desired performance specifications to provide needed settling time, overshoot and disturbance rejection throughout closed loop control system. Furthermore, proportional controller and phase lead controller is implemented to provide better transient response. When phase lead controller contributes the system to compensate the phase, which may be lost in process model or measurement device.

In our project, the proportional controller gains which are derived by design and simulation can make the real system meet the desired performance specification. Also, phase lead parameters implemented in closed loop system with only position feedback can be approximated throughout the design and simulation, but when velocity feedback is added to closed loop system, the simulation cannot estimate the controller parameters precisely. Although the implementation II face the system to unknown oscillatory behavior, probably because of presence of noise and delay in velocity measurement , it contributes the closed system to act more faster in comparison to first implementation with same parameters .

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